RADIO AND TELEVISION ENGINEERS' REFERENCE BOOK

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47 SPECIALIST CONTRIBUTORS

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EDITORS' FOREWORD

THE warm welcome given to the earlier editions of RADIO AND TELEVISION ENGINEERS' REFERENCE BOOK has shown clearly the value of a comprehensive and authoritative reference book, written at a practical engineering level. The very rapid progress being made in all branches of radio and television makes it essential, however, that such a book should be frequently and extensively revised.

In this third edition many new developments have been covered for the first time, and a number of sections completely re-written or re-cast to take the latest techniques fully into account. In particular, the rapid progress and practical application of semi-conductors have received careful attention. Section 26 (TRANSISTORS) has again been expanded to provide basic information on the latest types, and, covering modern applications, will be found new articles on GERMANIUM AND SILICON POWER RECTIFIERS (Section 26), TRANSISTOR D.C. CONVERTERS (Section 36), TRANSISTOR BROADCASTING RECEIVERS (Section 14), SERVICING TRANSISTORIZED EQUIPMENT (Section 39), etc. The practical use of transistors in modern transmitters is dealt with in the enlarged Section 7 (COMMUNICATION TRANSMITTERS), which also includes fresh information on low-power single-sideband transmitters, the choice of quartz crystals, V.H.F. scatter and meteor burst communication systems, etc.

Recent important advances in low-noise microwave devices, such as travelling wave tubes, parametric amplifiers and masers, are included in the revised Section 23 (VALVES). These developments are also taken into account in Section 19 (BROAD-BAND RADIO SYSTEMS), which includes the latest information on microwave communications.

The rapidly growing interest in RADIO ASTRONOMY AND SPACE-PROBE COMMUNICATION is reflected in the inclusion for the first time, of a section (Section 20) devoted entirely to this subject, and providing a concise summary of the present state of the art.

As in previous editions, particular importance has been attached to the provision of up-to-date information of practical value to those who are concerned with the production and servicing of domestic equipment. Section 14 (BROADCASTING RECEIVERS) has again been fully revised and expanded, and—apart from the transistor article already mentioned—now includes information on modern Car Radio Receivers. The increasing interest in good-quality reproduction is reflected in the new Section 37 (HIGH FIDELITY SOUND REPRODUCTION AND DISTRIBUTION) and in extensions to Sections 31 (MICROPHONES AND PICK-UPS) and 32 (LOUDSPEAKERS). Stereophonic reproduction is dealt with in Section 37, and its effect on recording techniques is described in the revised Section 34 (MAGNETIC AND DISC RECORDING).

Section 39 (RADIO RECEIVER INSTALLATION AND SERVICING) has been further expanded to include notes on dealing with printed-wiring panels, transistorized equipment and Band II alignment techniques, as well as a new sub-section SERVICING DOMESTIC TAPE RECORDERS. Section 40 (TELEVISION RECEIVER INSTALLATION AND SERVICING) has also been entirely re-cast to bring it into line with the latest practices.

Details of many new air navigation aids, including Doppler, have been added to Section 19 (AERONAUTICAL RADIO AND RADAR EQUIPMENT). New display devices which find increasing application in modern radar systems form part of the additions to Section 24 (CATHODE-RAY TUBES). The use of propagation forecasts for H.F. communications is described in a new article added to Section 44.

Section 46 (PROGRESS AND DEVELOPMENTS) has been entirely re-written to provide concise information on a number of interesting new developments and products resulting from the work of specific firms and organizations.

It is the aim of this Reference Book to bring together, within one convenient volume, the information required by engineers, technicians, radio amateurs and all who are engaged in the design, maintenance, technical sales, operation and servicing of modern radio and television transmitting and receiving equipment.

The Reference Book is arranged in forty-seven main sections, mostly dealing with a specific branch of the subject and written by authorities in the particular field concerned. A list of the main contributors appears on pages xiii–xx.

Throughout the book, emphasis has been placed on essentially practical information and theory, so as to ensure that the work will be of the utmost value to field engineers and technicians. Nevertheless, the instructor and student will also find much material—not readily available elsewhere—that will be of the greatest use in teaching or studying for careers in this rapidly expanding industry.

The Editors are indebted to the many contributors who have co-operated so whole-heartedly in the production of this new edition. Grateful acknowledgement is also made to the many leading firms and organizations who have kindly allowed their engineers to contribute to this work or who have assisted by supplying illustrations, information and data.

The Editors also wish to thank those readers who have kindly submitted comments or suggestions—many of these have been incorporated in this new edition.

W. E. P.
J. A. R.
J. P. H.

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SPECIALIST CONTRIBUTORS

L. S. A. **L. S. ALLARD,** B.Sc., A.Inst.P., was educated at Gravesend County Grammar School and graduated at Chelsea Polytechnic. He joined the Research Laboratories of the General Electric Company in 1941, and has worked on various aspects of cathode-ray-tube design. He is at present in charge of the Cathode-ray Tube Group.

S. W. A. **S. W. AMOS,** B.Sc.(Hons.), A.M.I.E.E., was born in Smethwick, Staffs., in 1915 and was trained as a teacher, obtaining a degree in Physics from the University of Birmingham in 1936 and the Ministry of Education Diploma in the following year. After four years as an Assistant Master in H.M. Dockyard School, Devonport, he joined the B.B.C. as a Maintenance Engineer, later transferring to the newly formed Engineering Training Department as an Instructor. In 1947 he joined the B.B.C. Technical Instructions Section, being appointed Assistant Editor in 1957 and Editor in 1959.

E. S. B. **E. S. BACON,** M.Sc.(Lond.), A.R.I.C., is Chief Supervisor of the Electrical Laboratories of the Ever Ready Co. (G.B.) Ltd.

C. H. B. **C. H. BANTHORPE,** technical director of Central Equipment Ltd., of which Derwent Television is a part. Throughout the war he was at the Telecommunications Research Establishment, after which he returned to Central Equipment to take charge of radar production and test, transferring to television shortly before the B.B.C. transmissions recommenced. For several years he has been works manager of the Derwent factory at Perivale, Middlesex, and has been responsible for the designing of their television receivers since 1951.

H. M. B. **H. M. BARLOW,** Ph.D., B.Sc.(Eng.), M.I.E.E., is Pender Professor of Electrical Engineering at University College, London, and specializes in high-frequency electrical engineering. His interest in this field of study started when in the First World War he joined as a Sub-Lieut. R.N.V.R. the Staff of the Signal School, Portsmouth, and worked on the reception of wireless signals under water. Subsequently, as a student of University College, London, he came under the influence of Sir Ambrose Fleming, and after practical experience in the engineering industry he was invited to join the staff of that College. The Second World War drew him into the Air Ministry as a Scientific Officer with the Telecommunications Research Establishment, and when the War ended he had become Superintendent of the Radio Department of the Royal Aircraft Establishment. By that time microwaves had found an important place in ultra-high-frequency technique, and Professor Barlow made a special study of their applications. He has written books on the subject entitled *Microwaves and Waveguides*, and, with Dr. A. L. Cullon, *Microwave Measurements*.

T. W. B. **T. W. BENNINGTON,** is a Research Engineer of the B.B.C. Engaged in short-wave radio since 1932 he joined the B.B.C. in 1934 and was for 15 years in the Department concerned with short-wave planning. Both there and in Research Department he has been concerned with the application of ionospheric data to long-distance transmission. He is the author of *Short-Wave Radio and the Ionosphere*.

E. W. B.-J. **E. W. BERTH-JONES,** B.Sc.(Eng.), was born in 1911, in North Wales, and was educated at Oundle and Manchester, receiving his B.Sc. degree in electrical engineering in 1933. His introduction to sound recording was in the laboratories of Messrs. British Acoustic Films Ltd., and he later acquired operational experience at Denham Studios with London Film Productions. He joined the staff of the Gramophone Company Ltd., Recording Department, in 1937. After specialized work for H.M. Government during the war, he returned to this company, under its new title of E.M.I. Studios Ltd., as Development Engineer, concerned principally with the introduction of magnetic tape recording and in the circuit design of new high-quality recording equipment for professional studio use, and later transferred to the reorganized Records Division of E.M.I. Ltd.

B. R. A. B. **B. R. A. BETTRIDGE,** A.M.Brit.I.R.E., Commercial Manager of the G.E.C. Semi-conductor Division. After many years of experience with thermionic valves, he became interested in germanium diodes and, since the introduction of transistors, has been engaged on the application of transistors and other semi-conductor devices to radio, television and electronics.

W. T. B. **W. T. BLACKBAND,** M.Sc., A.M.I.E.E., engaged on aerials and propagation research at R.A.E., Farnborough. Chairman of R.C.R.D.C. subcommittee on Radiofrequency Cables and N.A.T.O. committee on transmission lines. Member of British National Committee for I.G.Y. artificial satellite studies.

F. D. B. **F. D. BOLT,** B.Sc.(Eng.), M.I.E.E., graduated at Queen Mary College, London University, in 1929. He joined Standard Telephones & Cables Ltd. and worked on short-wave transmitter development between 1929 and 1934. Since 1934 he has worked for the B.B.C. on the planning and installation of transmitting equipment, particularly aerial and feeder systems. He was awarded the Fahie Premium of the Institution of Electrical Engineers jointly with Mr. F. C. McLean, in 1946. In 1951 he visited South Africa to advise the South African Broadcasting Corporation on the planning of a short-wave centre for that country. As Head of Aerial Unit, Planning and Installation Department of the B.B.C., he is responsible for the engineering of the aerials and feeder systems for all broadcast transmitting stations. Elected Member of the Institution of Electrical Engineers in 1953.

F. H. B. **F. H. BRITTAIN** joined the staff of the Research Laboratories of the General Electric Company early in 1926, and worked on absorption-type refrigerators. In 1928 he worked on photo-electric cells, transferring to the Acoustics Laboratory in 1930. During the war he worked on silicon for use in Radar sets, and returned to the Acoustics Laboratory in 1947.

J. E. B. **J. E. BRYDEN,** A.M.I.E.E., was born at Strood, Kent, in 1917 and educated at Sir Joseph Williamson's Mathematical School, Rochester. He enlisted in the Royal Artillery in 1939, and in 1941 was posted to the Military College of Science as an Instructor for Radar and Wireless, with the rank of Captain (R.A.O.C., later R.E.M.E.). After the war he joined the Staff of the Research Laboratories of the General Electric Company, and in 1951 was appointed Group Leader of the Radio Communications Group. In 1955 he joined the Engineering Department of Canadian Marconi Company, in Montreal, Canada; he is now Manager of the Radio Relay Department, and is responsible for Development, Systems and Sales Engineering. He has been intimately involved in the development of Broad-band Radio Systems since 1948.

R. H. B. **R. H. BURDICK**, A.C.G.I., A.M.I.E.E., was educated at Woolwich Polytechnic, gaining senior science scholarship to the Imperial College of Science, and, in 1930, A.C.G.I. He trained initially as a mechanical engineer, transferring to the electrical branch, and, after considerable experience on the design of telephone and radio equipment, he joined the telephone laboratory of Messrs. Siemens Brothers & Co. Ltd. He is at present with Marconi's Wireless Telegraph Co. Ltd., working on the development and design of radio and radar equipment.

V. J. C. **V. J. COOPER**, B.Sc.(Eng.), A.C.G.I., M.I.E.E., M.Brit.I.R.E., Assoc.I.R.E., is a member of Marconi's Wireless Telegraph Co. Ltd. and is at present Chief Engineer of the Advanced Development Group. For some years he has been responsible for the development of V.H.F. and U.H.F. transmission equipment and of television equipment in general.

J. H. C. **J. H. COZENS**, B.Sc., A.M.I.E.E., was educated at George Green's School and London University. After a short period in the teaching profession, spent two years on the staff of Telegraph Contruction & Maintenance Co. Ltd. Joined the Telegraph Condenser Co. Ltd. in 1932, taking charge of the Development Department in 1938. Since 1945 he has been Chief Technical Officer of this Company.

L. D. **L. DRISCOLL**, B.Sc., A.M.I.E.E., A.M.Brit.I.R.E., has been with Murphy Radio Ltd. in various capacities since 1935. At present he is in charge of their Radio Receiver Design Laboratory.

B. Z. de F. **B. Z. de FERRANTI**, B.Sc., B.E.(Hons.), Graduate I.E.E., Stud. I.E. Aust., was born in Sydney, Australia, in 1928, and educated at North Sydney Boys' High School and The University of Sydney. Worked on digital-computer techniques with C.S.I.R.O. before joining G.E.C., Coventry, in 1951. In 1953, joined the G.E.C. Research Laboratories, Wembley, and is at present working on electronic telephone exchanges.

E. A. F. **E. A. FIELDING**, B.Sc.Tech.(Hons.), A.M.C.T., M.I.E.E., A.I.R.E., was born in 1913, and educated at Oldham High School and Manchester University. He studied Electrical Engineering, specializing in Communications Engineering, and graduated with first-class honours in 1934. After graduation he secured employment with Salford Electrical Instruments Ltd., as a development engineer, and in 1938 became manager of the Quartz Crystal Department, a position he still holds. Since the end of 1951 he has also been manager of the Company's Chestergate works.

D. H. F. **D. H. FISHER**, A.M.I.E.E., was born in Halstead, Essex, in 1926, and received early education at Halstead Grammar School and Gosfield School, Essex. He commenced studies in Telecommunications Engineering at The Polytechnic, Regent Street, London, W., in 1942, and qualified with Polytechnic and Higher National Diplomas. He then joined E.M.I. Engineering Development Ltd., to work on defence projects and later television development, and subsequently with Pye Ltd., Engineering Department, specializing particularly in Television Receiver design and production, and since February 1952 on domestic receivers. Now Technical Director of Regentone Radio and Television Ltd. Contributor to the I.E.E. Television Convention, May 1952 and to the *Television Engineer's Pocket Book* (Newnes).

W. I. F. **W. I. FLACK,** Assoc.I.E.E., Senior Member I.R.E., Fellow of the Television Society, has been connected with the radio industry for twenty-five years. Received electrical training at the Regent Street Polytechnic. Worked on radar during the early part of the war, and was with the Telegraph Condenser Co. Ltd. for fourteen years. Joined the production engineering department of Radio and Allied Industries, Ltd., in 1957. Apart from developing new types of capacitors has also designed television receivers and a magnetic tape recorder for amateur constructors.

L. S. F. **L. S. FOSKETT** formerly served in the Royal Corps of Signals and was a member of the B.B.C. engineering staff. Joined the H.M.V. (The Gramophone Co.) Film Recording Unit and is now a member of the Sound Equipment Planning and Estimating Division of E.M.I. Sales and Service Ltd.

R. H. G. **R. H. GARNER,** B.Sc.(Eng.), A.M.I.E.E., A.M.Brit.I.R.E. Principal of The Technical College, Coatbridge, and present Chairman of the Scottish Section of the British Institution of Radio Engineers. Has wide interests in radio and electronic matters, especially in their applications to the Automatic Factory. Regular contributor to the Scottish radio press.

R. C. G. **R. C. GLASS,** M.A., B.Sc., A.M.I.E.E., is Lecturer in Applied Physics at Northampton College of Advanced Technology, London, and a Scholar of Trinity College, Dublin. After graduating with first class honours in physics, he was engaged in research on magnetrons at the Research Laboratories of the General Electric Company and on travelling-wave tube applications at the Research Laboratories of Decca Radar, being awarded an I.E.E. Premium for his magnetron work in 1951. He is also editor of the English edition of the Russian technical journal *Radiotekhnika* (Radio engineering).

F. J. G. **F. J. GRIMM,** A.M.Brit.I.R.E. of Pye Telecommunications Ltd.

J. P. H. **J. P. HAWKER** first became interested in radio communication as an amateur, obtaining a licence (2BUH, later G3VA) in 1936. Served in the Special Communications Units of the Royal Signals, 1941–46. Appointed as Assistant to the General Secretary of the Radio Society of Great Britain in 1947, and later became Assistant Editor of the *R.S.G.B. Bulletin*. Joined the Technical Books Department of George Newnes Ltd. in 1951.

P. J. **P. JONES,** was born in Cheshire in 1914, and educated at Chester City Grammar School and Liverpool Technical College. He began his career as an organic chemist with Messrs. British Insulated Callender's Cables Ltd. in research on natural and synthetic resin cable dielectrics, later developing an interest in electronic techniques for measurement and control in this field. Subsequently in 1950 he joined Aerialite Ltd., now being responsible for the design of radio and television receiver aerials, telecommunications cables, etc. He is the originator of numerous patents in the fields of high-frequency electronics, television aerials and accessories. He is actively interested in Amateur Radio communications, particularly on ultra-high frequency in the 70-, 24- and 3-cm. bands. Station call-sign G2JT.

P. R. K. **P. R. KELLER,** B.Sc., joined Marconi's Wireless Telegraph Co. Ltd. in 1944 after graduating at King's College, London University. Following a course of practical training at the Marconi College, he joined the Design and Development Division, and has specialized in V.H.F. transmitter-receiver equipment from

1946–1956. He is at present in charge of a section developing error-correcting telegraph equipment. He is a patentee and a part-time lecturer in the Engineering Department of the Mid-Essex Technical College. Author of *V.H.F. Radio Manual.*

J. M. K. **J. M. KIRK,** M.B.E., B.Sc.(Hons.), D.I.C., A.C.G.I., M.I.E.E., was born in Canton, China, and educated at Mill Hill School and Imperial College, London. Royal Corps of Signals, 1939–46. Member, since 1946, of the Radio Project Planning Section of Standard Telephones and Cables, Ltd. Amateur Transmitting Licence since 1935.

R. T. La. **R. T. LAKIN,** M.B.E., A.M.I.E.E., A.M.Br.I.R.E., is Chief Research Engineer, Whiteley Electrical Radio Co. Ltd. He has specialized in the design of loudspeakers and associated equipment, and is the author of several papers on loudspeakers, one of which was the subject for a series of Lectures for the Institution of Radio Engineers.

L. P. L. **L. P. LEARNEY** studied mathematics and physics as a full-time student at the Polytechnic, Regent Street, London, and joined the staff of Redifon Ltd. just after the outbreak of war. He assisted at first in the development of war-time military equipment, including beacon transmitters and radar training devices, and later as seuior engineer has developed a wide range of equipment for airborne, marine, mobile and fixed use, including receivers, and a valve and transistor transmitter and amplifiers for commercial and service application. He has also been engaged recently in standardization work.

R. T. Lo. **R. T. LOVELOCK,** A.M.I.E.E., first entered the drawing office of a telephone manufacturer, and became a designer-draughtsman of subscriber's and exchange apparatus. From there he transferred to a radio development laboratory to design and develop electronic instruments, and with the initiation of the British television service became a television development engineer. During the last war he was responsible for special measurement and design techniques within a laboratory, and a member of many Service–Industry liaison committees. More recently he was in charge of a laboratory responsible for material and component approval, for instrument calibration and maintenance, and for the clearance of special design and production problems. In this latter position he took the opportunity to investigate and utilize the statistical design of experiments and control of production processes, and to contribute to the work of several B.S.I. committees.

H. A. McG. **H. A. McGHEE,** Grad.I.E.E., was educated at the Technical College, Coatbridge, and the Royal Technical College, Glasgow. He is the author and joint author of a number of papers on television technique, and is a member of the engineering staff of Pye Ltd., Cambridge. Author of *Industrial Television.*

R. D. A. M. **R. D. A. MAURICE,** Ing.-Dr., Ing.E.S.E., A.M.I.E.E., received his general education in England and California. Engineering degree, France (1933) and doctorate of engineering from Sorbonne (1952). Graduate trainee (1933) and subsequently development engineer with E.M.I. Ltd. Joined B.B.C. Research Department (1939). Became head of Receiver and Measurements Section, now deputy head, Television Group.

E. M. **E. MOLLOY,** formerly General Editor, Technical Books Department, George Newnes Ltd. Trained Royal Technical College, Salford. Obtained King's Prize in Applied Mechanics, 1911. After practical experience with Lea Recorder Co. Ltd., and United Brassfounders & Engineers Ltd., joined staff of *Electrical Engineering*. Assistant Examiner, Inventions Dept., Ministry of Munitions, 1917–19. In 1919 joined staff of Sir Isaac Pitman & Sons Ltd., and in 1924 was appointed Technical Editor. Joined staff of George Newnes Ltd. in 1929, retired in 1958.

J. M. **J. MOIR,** M.I.E.E., M.I.R.E., Technical Director of Goodmans Industries and formerly in the Electronic Engineering Department of the B.T.H. Co. as Head of the Section dealing with the development and engineering of audio-frequency and communication equipment, telemetering units, and with architectural acoustical problems, particularly in the sound film field. Privately a high-fidelity sound reproduction enthusiast. Member of the Institution of Electrical Engineers and the American Acoustical Society–Institute of Radio Engineers.

E. W. M. **E. W. MORTIMER,** joined the Garrard Engineering & Manufacturing Co. Ltd. in 1919, and served a five-year apprenticeship to precision engineering. He was responsible for the design of the first Garrard Record Changer, and has since been actively engaged on the development of record reproducing equipment. He has travelled extensively, studying the export requirements for gramophone equipment.

L. A. M. **L. A. MOXON,** B.Sc.(Eng.), A.M.I.E.E., was educated at Clifton College and the City and Guilds Engineering College, obtaining the London University B.Sc. degree in 1929. After two years research under the auspices of the D.S.I.R. he joined the staff of Murphy Radio Ltd., where he was responsible for development and research in connection with broadcast reception. In 1941 he joined H.M. Signal School, Portsmouth, where he was concerned with the development of radar receivers. He is now a member of the Royal Naval Scientific Service.

W. E. P. **W. E. PANNETT,** A.M.I.E.E., was with Cable and Wireless Ltd. and Marconi's Wireless Telegraph Co. Ltd. for many years. From 1921 to 1928 he was engaged in pioneer research and development work on microphones and reproducers at 2LO, the original London broadcasting station, and communication recording apparatus. From 1928 until 1956 he was successively occupied in charge of the construction and administration of some of the early Beam transmitting stations, the design of transmitters, station planning and installation design. He was latterly in charge of the Power Equipment and Installation Division of Marconi's Wireless Telegraph Co. Ltd. He is author of *Radio Installations: Their Design and Maintenance* and *Radio Engineering Formulæ and Calculations*.

R. D. P. **R. D. PETRIE,** M.B.E., A.M.I.M.E., A.M.I.E.E., was born in 1904 and educated at Hull Technical College. He trained firstly as a mechanical engineer, and later as an electrical engineer. Joined Gaumont British Picture Corporation in 1929, and was appointed Area Engineer for North of England in 1930 and later, the same year, Divisional Engineer for the West End of London and the South Coast Division. He joined the B.B.C. in 1935 and served in the Operational Department a year before transferring to the Equipment Department. Later joined the then Station Design and Installation Department, supervising installation of war-time control room and studios, and later designing reproducing equipment. In 1947 he joined the

Studio and Equipment Section of the Designs Department. Elected to Associate Membership of the Institution of Mechanical Engineering in 1934, and to Associate Membership of the Institution of Electrical Engineers in 1940. Member of the Acoustics Group of the Physical Society.

W. C. R. **W. C. RIDDIFORD** is holder of City and Guilds certificates in Radio I, II, III and IV. Employed from 1927 to 1945 as service and civilian instructor in Radio Engineering to the R.A.F. Recently Senior Lecturer in Radio Engineering to Airways Training College and author of *Radio Reference*.

F. W. J. S. **F. W. J. SAINSBURY**, A.C.G.I., D.I.C., is a Whitworth Scholar and is at present with the Receiver Development Division of Marconi's Wireless Telegraph Co. Ltd.

D. H. C. S. **D. H. C. SCHOLES** joined the Marconi Company in 1933. In 1940 he left the Research Department of that Company to join the London Engineering Staff of the B.B.C. Later in 1940 he volunteered for and was commissioned in the Signals Branch of the R.A.F., but gave up this appointment to work for M.A.P. at R.A.E., Farnborough. In 1941 he joined the Royal Navy as an Air Engineer Officer and served at sea, abroad and at home on various engineering duties connected with airborne radio and radar. Promoted to Lt.-Commander 1943, and served the year following the war in the Admiralty. He joined Plessey Ltd. as a Senior Engineer 1946, and was appointed Head of Radio Laboratory in 1948, Chief Radio Engineer 1950, and Chief Engineer in Telecommunications Division 1952. In 1953 he assumed the duties of Sales Manager in that Division also.

J. R. S. **J. R. SHAKESHAFT**, read for the Natural Sciences Tripos at St. John's College, Cambridge, from 1949 to 1952. Since that date he has been doing research in radio-astronomy at the Cavendish Laboratory, Cambridge. He took a Ph.D. and was elected to a fellowship at St. John's College in 1957.

R. J. S. **R. J. SLAUGHTER**, B.Sc.(Hons.), is employed by the Telegraph Construction & Maintenance Co. Ltd. After initial experience on the chemical and physical aspects of cable testing, particularly submarine cable, he spent a short period with the power cable research group. In 1949, after graduating in physics, he specialized in high-frequency cables, and is now in charge of the high-frequency cable research section.

D. F. U. **D. F. URQUHART**, has for the past five years been in charge of Engineering Laboratory and Electronic Development at Erie Resistor Ltd. Previously employed on production. His technical education was at Acton Technical College. In 1916 he was invalided from the Forces, and employed in the Proof and Experimenting Department of Armstrong Whitworth Ltd., then engaged on heavy electrical engineering with various firms in the north-eastern area. In 1930 he spent eighteen months in automatic telephones, followed by engagement as a brewer's electrical engineer until war broke out. During the early part of war he was a lecturer in the R.A.F. Technical School. Then followed two years in valve manufacture, after which he took his present employment. Radio enthusiast since 1920.

V. V. **V. VALCHERA** is Technical Director of Valradio Ltd. and Associated Companies Dar-val Engineering Ltd. and Valmade Ltd. He has been mainly responsible for the development of heavy-duty vibrators and vibrator converters in this country, and has played a prominent part in the development of large-screen projection television receivers for the home and export markets.

A. H. B. W. **A. H. B. WALKER,** B.Sc.(Eng.), D.I.C., A.C.G.I., M.I.E.E., graduated with first-class honours at City and Guilds (Eng.) College in 1934. He joined the Rectifier Engineering Department of the Westinghouse Brake & Signal Co. Ltd., designing all types of rectifier equipment, and originated various novel constant-voltage and constant-current systems based on transductors and ferro-resonance. In 1938 he joined the Westinghouse Research Laboratory, and has since been concerned with the continuous development of copper oxide, selenium, and germanium rectifiers of all types. While specializing on the umerous applications of metal rectifiers in electronic circuitry, voltage multipliers, etc., he has also developed electronic and magnetic servo controllers for large rectifiers. Other original work includes the "Stabilistor" (A.C. voltage stabilizer). the "Phase Converter" (static single-phase to three-phase power converter) which was widely used during the war for driving three-phase motorized machine tools in single-phase districts, and more recently the "Transbooster" constant-voltage system. In 1952 he was appointed Joint-Chief of the Westinghouse Research Laboratory.

T. W. **T. WORSWICK,** M.Sc., A.M.I.E.E., was trained at Regent Street Polytechnic and City and Guilds. He joined the B.B.C. as a student apprentice in 1934, working at Alexandra Palace 1936–38. He then transferred to the Lines Department Television Section, and is at present in charge of Engineering Designs Department Television Apparatus Section.

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ELECTRICAL MEASURING INSTRUMENT

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A special version of the Universal AvoMeter having a **sensitivity of 20,000 ohms per volt on all D.C. ranges, and 1,000 ohms per volt on A.C. ranges from 100V upwards.** Accuracy on D.C. current ranges and A.C. ranges is to B.S. 89/1954 for 5-inch scale-length industrial portable instruments, and on D.C. voltage ranges is 2% of indication from full-scale to half-scale deflection, and 1% of full-scale value below half-scale.

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Accessories are available for extending the above ranges.

Size : 8½" × 7¼" × 4½".

The Model 7 Universal AvoMeter

Size : 8" × 7¼" × 4½".

A multi-range A.C./D.C. instrument of B.S. 89/1954 accuracy, providing 50 ranges of readings on a 5" hand calibrated scale. Range selection is by means of two rotary switches, for A.C. and D.C. respectively. A press button provides an additional range for each value of current and voltage shown on the switch knobs. Current consumption at full-scale deflection is 1mA. or 2mA. according to whether the press button is used or not. The total resistance of the meter is 500,000 ohms. An automatic cut-out provides protection against damage through severe overload.

CURRENT : A.C. and D.C. 0 to 10 amps.
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RESISTANCE : Up to 40 megohms.
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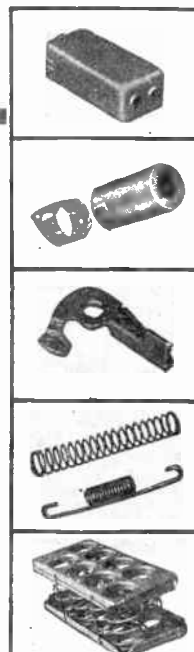

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1. FORMULÆ AND CALCULATIONS

1. FORMULÆ AND EXAMPLES

RESISTORS AND VOLTAGE DIVIDERS

Voltage, Current, Power and Resistance Relationships

The relationships between the current flow in a resistor or a circuit containing resistance, the applied voltage, the power dissipated in it and the resistance are determined from simple transformations of Ohm's Law.

Current flow: $I = E/R = P/E = \sqrt{P/R}$

Volt drop: $E = IR = P/I = \sqrt{PR}$

Power dissipated: $P = EI = I^2R = E^2/R$

Resistance: $R = E/I = P/I^2 = E^2/P$

where
I = current, amperes;
E = voltage;
P = power, watts;
R = resistance, ohms.

Resistance of a Conductor

Given the dimensions and specific resistance (resistance per unit cube) of a conductor, its resistance

$$R = \rho l/A$$
$$= 1{\cdot}273\ \rho l/d^2 \text{ for circular conductors}$$

CHANGE OF RESISTANCE WITH TEMPERATURE

$$R_T = R_0\ (1 + \alpha T)$$

where
R = resistance, ohms;
R_T = resistance at temperature $T°$, ohms;
R_0 = resistance at 0°, ohms;
ρ = specific resistance of conductor, ohm-cm.;
l = length of conductor, cm.;
A = cross-sectional area of conductor, sq. cm.;
d = diameter of conductor, cm.;
α = temperature coefficient of resistance.

Resistances in Series and Parallel

Resistances are added when in series; conductances are added when in parallel.

Equivalent resistance in series:

$$R_e = R_1 + R_2 + R_3 + \ldots \text{ etc.}$$

Equivalent conductance in series:

$$1/G_e = 1/G_1 + 1/G_2 + 1/G_3 + \ldots \text{ etc.}$$

TABLE 1.—RESISTIVITIES AND TEMPERATURE COEFFICIENTS OF METALS AND ALLOYS

	Resistivity ($\mu\Omega$-*cm.*)	*Temp. Coefficient of Resistivity at* 20° *C.*
Aluminium	2·83	0·004
Brass	6·73	0·002
Copper, annealed	1·72	0·0039
Copper, hard drawn	1·78	0·0038
Eureka (Constantan)	49·1	±0·0002
Gas carbon	5,000	−0·0005
German silver	29·2	0·00027
Gold	2·44	0·0034
Iron, annealed	9·66	0·0057
Lead	22	0·0042
Manganin	44·8	±0·00002
Mercury	96	0·00089
Molybdenum	5·69	0·0045
Nichrome	112	0·00017
Nickel	8·7	0·0047
Nickel silver	27·6	0·00026
Phosphor bronze	9·4	0·003
Platinum	10·6	0·0038
Silver	1·64	0·004
Tin	11·5	0·0042
Tungsten	5·61	0·0045
Zinc	5·86	0·0037

Equivalent resistance in parallel:

$$1/R_e = 1/R_1 + 1/R_2 + 1/R_3 + \ldots \text{ etc.}$$

Two resistances in parallel:

$$R_e = R_1R_2/(R_1 + R_2)$$

Equivalent conductance in parallel:

$$G_e = G_1 + G_2 + G_3 + \ldots \text{ etc.}$$

where R_e = equivalent resistance, ohms;
G_e = equivalent conductance, mhos;
$R_1R_2R_3$ = resistances in series or parallel;
$G_1G_2G_3$ = conductances in series or parallel.

Voltage-reducing Resistors

Resistance required in series with a resistive load to reduce voltage from E_0 to E_L, assuming negligible internal resistance of source.

$$R_s = (E_0 - E_L)/I$$

where R_s = resistance in series, ohms;
E_0 = voltage of supply;
E_L = voltage across load;
I = load current, amperes.

EXAMPLE 1.—*A stepped resistor is required to control the current from a 230-volt rectifier to a load of 150 ohms resistance in steps of 1, ½ and ¼ amperes. Calculate each step of resistance and the dissipation in watts for which each section must be graded.*

Total resistance in circuit, including load resistance :

At 1 ampere

$$R_L + R_1 = E_0/I_1$$
$$= \frac{230}{1} = 230 \text{ ohms}$$

At ½ ampere

$$R_L + R_1 + R_2 = E/I_2$$
$$= \frac{230}{0{\cdot}5} = 460 \text{ ohms}$$

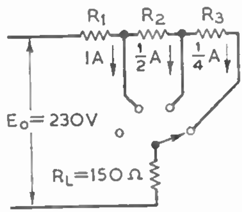


FIG. 1.—EXAMPLE 1.

At ¼ ampere

$$R_L + R_1 + R_2 + R_3 = E/I_3$$
$$= \frac{230}{0{\cdot}25} = 920 \text{ ohms}$$

Hence section resistances :

$$R_1 = 230 - 150 = \underline{80 \text{ ohms}}$$
$$R_2 = 460 - 230 = \underline{230 \text{ ohms}}$$
$$R_3 = 920 - 460 = \underline{460 \text{ ohms}}$$

Watts dissipated in each section :

$$\text{In } R_1,\ P_1 = 1^2 \times 80 = \underline{80 \text{ watts}}$$
$$\text{In } R_2,\ P_2 = 0{\cdot}5^2 \times 230 = \underline{58 \text{ watts}}$$
$$\text{In } R_3,\ P_3 = 0{\cdot}25^2 \times 460 = \underline{29 \text{ watts}}$$

Voltage Dividers

Values of sectional resistances for anode supply or positive tappings:

$$R_1 = E_1/I_0$$
$$R_2 = (E_2 - E_1)/(I_0 + I_1)$$
$$R_3 = (E_3 - E_2)/(I_0 + I_1 + I_2)$$

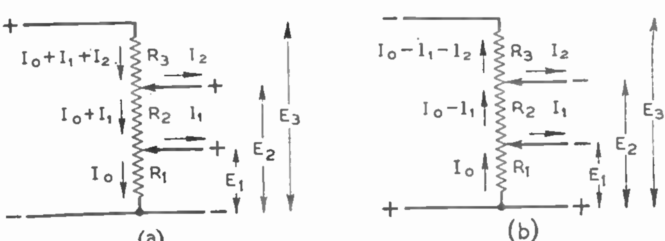


FIG. 2.—VOLTAGE DIVIDERS.

Values of sectional resistances for negative tappings (Fig. 2 (*b*)):

$$R_1 = E_1/I_0$$
$$R_2 = (E_2 - E_1)/(I_0 - I_1)$$
$$R_3 = (E_3 - E_2)/(I_0 - I_1 - I_2)$$

EXAMPLE 2.—*A tapped voltage divider is to be connected across a 200-volt supply to provide outputs at:* (a) 20 *mA at* 100 *volts, and* (b) 30 *mA at*

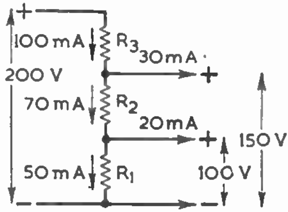


FIG. 3.—EXAMPLE 2.

150 *volts. Find the resistance of each section of the divider, the total resistance and the dissipation in watts of each section, if the bleeder current passed by the divider is to be 50 mA on load.*

SECTION R_1	SECTION R_2	SECTION R_3
Current:		
$I_1 = 50$ mA	$I_2 = 50 + 20 = 70$ mA	$I_3 = 50 + 30 + 20 = 100$ mA
Volt drop:		
$E_1 = 100$ volts	$E_2 = 150 - 100 = 50$ volts	$E_3 = 200 - 150 = 50$ volts
Resistance:		
$R_1 = \frac{100}{0.05} = \underline{2{,}000 \text{ ohms}}$	$R_2 = \frac{50}{0.07} = \underline{715 \text{ ohms}}$	$R_3 = \frac{50}{0.1} = \underline{500 \text{ ohms}}$

Watts dissipated:

$I_1^2R_1 = 0.05^2 \times 2{,}000 = \underline{5 \text{ watts}}$ $\quad I_2^2R_2 = 0.07^2 \times 715 = \underline{3.5 \text{ watts}}$ $\quad I_3^2R_3 = 0.1^2 \times 500 = \underline{5 \text{ watts}}$

Total resistance of divider:

$$R = R_1 + R_2 + R_3$$
$$= 2{,}000 + 715 + 500 = \underline{3{,}215 \text{ ohms}}$$

INDUCTORS

Inductance

Inductance is defined as the property of a circuit or coil which tends to oppose any change in the value of current, and is expressed by the fundamental equation

$$L = -e\Big/\frac{di}{dt}$$

$$e = -L\frac{di}{dt}$$

The voltage induced between two circuits

$$e = -M\frac{di}{dt}$$

Energy stored in inductance

$$W = LI^2/2$$

where L = coefficient of self-inductance, henrys;
M = coefficient of mutual inductance, henrys;
e = back e.m.f. produced by change in magnetic flux linkages, volts;
$\frac{di}{dt}$ = rate of change of current, ampere/second;
W = energy stored, joules;
I = current, amperes.

Inductance in a D.C. Circuit

In a circuit having inductance and resistance the instantaneous current i after a time t from application of the voltage E

$$i = (E/R)(1 - e^{-Rt/L})$$
$$= I(1 - e^{-Rt/L})$$

After a time t from removal of the voltage, E, without breaking the circuit

$$i = (E/R)e^{-Rt/L}$$
$$= Ie^{-Rt/L}$$

EXAMPLE 1.—*The coil of a signalling relay has an inductance of 250 mH and a resistance of 100 ohms. If a steady signal of 30 volts is applied, find the time taken for the current to reach 50 per cent of its maximum value.*

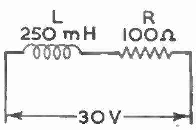


FIG. 4.—EXAMPLE 1.

$$i/I = 1 - e^{-Rt/L}$$

$$0{\cdot}5 = 1 - e^{-100t/0{\cdot}25}$$

$$e^{-400t} = 0{\cdot}5$$

Taking logs of both sides

$$-400t \log_{10}e = \log_{10} 0{\cdot}5$$
$$-400t \times 0{\cdot}4343 = -0{\cdot}301$$
$$t = 1{\cdot}73 \times 10^{-3} \text{ seconds or } \underline{1{\cdot}73 \text{ milliseconds}}$$

Time Constant

This is the time taken for the current to reach $1 - 1/e$ or 63·2 per cent of its full value on application of the voltage, or to fall to $1/e$ or 36·8 per cent of its full value on removal of the voltage.

$$T = L/R$$

where I = full value of current, amperes;
i = instantaneous value of current, amperes;
E = applied voltage;
R = resistance in series, ohms;
t = time, seconds;
T = time constant, seconds;
L = inductance, henrys;
e = base of Naperian logarithms (= 2·7183).

Inductance in an A.C. Circuit

Voltage self-induced in a circuit

$$E = \omega LI$$

Voltage induced between two circuits

$$E = \omega MI$$

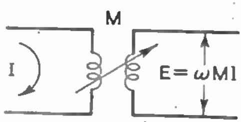


FIG. 5.—INDUCTANCE IN AN A.C. CIRCUIT

where E = R.M.S. value of induced voltage;
I = R.M.S. value of primary current, amperes;
L = inductance, henrys;
ω = $2\pi \times$ frequency (c/s).

Inductance of Coils

General formula for coils

$$L = \phi T/10^8 I$$
$$= 4\pi\mu AT^2/10^9$$

where L = inductance, henrys;
ϕ = number of flux lines linking turns of coil;
T = number of turns of coil;
I = current, amperes;
μ = permeability of medium through which flux passes;
A = cross-sectional area of path, sq. cm.;

Inductance of Single Layer Radio-frequency Coils

Nagaoka's formula

$$L = k(\pi dT)^2/10^3 l$$

where L = inductance, μH;
d = diameter of coil, cm.,
l = length of coil, cm.;
T = total number of turns;
k = factor depending on the ratio d/l (see Table 2).

TABLE 2.—FACTOR k

d/l	0·1	0·2	0·4	0·6	0·8	1·0	2·0	3·0	4·0	5·0
k	0·959	0·920	0·850	0·789	0·735	0·688	0·526	0·429	0·365	0·320

EXAMPLE 2.—*Calculate the inductance of a single-layer air-cored cylindrical coil 4 cm. diameter, wound with fifty turns of wire 0·05 cm. diameter and a spacing between turns of 0·05 cm. Take the shape factor as 0·735. What will be the inductance if a finer-gauge wire is used to increase the number of turns by 50 per cent within the same winding length?*

$$d = 4 \text{ cm.}$$
$$T = 50$$
$$l = 50 \times (0{\cdot}05 + 0{\cdot}05) = 5 \text{ cm.}$$
$$L = k(\pi d T)^2/1{,}000l$$
$$= \frac{0{\cdot}735 \times (\pi \times 4 \times 50)^2}{5 \times 10^3} = \underline{58\ \mu\text{H}}$$

If the number of turns is increased 50 per cent in the same winding space:

$$L \propto T^2,\ l \text{ remaining constant}$$
$$L_2/L_1 = (T_2/T_1)^2$$
$$= \left(\frac{1{\cdot}5}{1}\right)^2 = 2{\cdot}25$$

i.e., the inductance is increased $\underline{2{\cdot}25 \text{ times}}$ to 131 μH.

Iron-cored Coils

Iron circuit almost completely closed, with small air gap.

$$L = T^2 \mu A/l$$

where
L = inductance, henrys;
T = total number of turns;
μ = permeability of air gap (= 1);
A = cross-section of iron at gap, sq. cm.;
l = length of gap, cm.

CAPACITORS

Capacitance

The property of a circuit or capacitor to build up a charge on application of a voltage is defined quantitatively as the charge in coulombs acquired per volt applied.

$$C = Q/E$$
$$E = Q/C$$

Energy stored

$$W = CE^2/2$$

where
Q = charge, coulombs;
C = capacitance, farads;
E = applied e.m.f., volts;
W = energy, joules (watt-seconds).

Capacitor in a D.C. Circuit

In a circuit having capacitance and resistance, the instantaneous charging current after a time t from application of the voltage

$$i = (E/R)e^{-t/CR}$$
$$= Ie^{-t/CR}$$

At $t = 0$ $\qquad i = E/R$

Instantaneous current when discharging through resistance

$$i = -(E/R)e^{-t/CR}$$
$$= -Ie^{-t/CR}$$

Time Constant

This is the time taken for the voltage to reach $1 - 1/e$ or 63·2 per cent of its full value on applying the voltage, or to fall to $1/e$ or 36·8 per cent of its initial value on removing the voltage.

$$T = CR$$

where
i = instantaneous current, amperes;
I = maximum current, amperes;
E = applied voltage;
R = resistance in series, ohms;
t = time, seconds;
T = time constant, seconds;
C = capacitance, farads;
e = base of Naperian logarithms (= 2·7183).

Capacitor in an A.C. Circuit

$$E = I/\omega C$$

where
E = r.m.s. voltage;
I = r.m.s. current, amperes;
C = capacitance, farads;
ω = 2π × frequency (c/s).

EXAMPLE 1.—*What is the peak working voltage to which a capacitor of 0·003 μF is subjected when a current of 10 mA at 1,000 c/s flows?*

R.m.s. voltage:

$$E = I/\omega C$$
$$= \frac{0{\cdot}01}{2\pi \times 1{,}000 \times 3 \times 10^{-9}} = 531 \text{ volts r.m.s.}$$

Peak voltage:

$$E = \sqrt{2}E \text{ r.m.s.}$$
$$= \sqrt{2} \times 531 = \underline{751 \text{ volts peak}}$$

Capacitors in Series and Parallel

Series :

$$C_T = \frac{1}{1/C_1 + 1/C_2 + 1/C_3 + \ldots \text{ etc.}}$$

$$= C_1C_2/(C_1 + C_2) \text{ for two condensers in series.}$$

Parallel :

$$C_T = C_1 + C_2 + C_3 + \ldots \text{ etc.}$$

where C_T = effective total capacitance;
$C_1C_2C_3$ = capacitances of individual condensers.

Capacitance of Multi-plate Capacitor

$$C = 0{\cdot}0884\ n\kappa A/d$$

where C = capacitance, $\mu\mu$F;
n = number of sheets of dielectric between plates;
κ = dielectric constant of dielectric;
A = surface area of one plate, sq. cm.;
d = distance between plates (thickness of dielectric, cm.).

This formula does not take into account fringing effect. If the dielectric is made up partly of a solid and partly of air, treat as two capacitors in series, having dielectric constants κ_1 and κ_2 and dielectric thicknesses d_1 and d_2. Then

$$C_1 = 0{\cdot}0084\ \kappa_1 A/d_1$$
$$C_2 = 0{\cdot}0884\ \kappa_2 A/d_2$$
$$C_T = C_1C_2/(C_1 + C_2)$$

Dielectric Loss

The effect of dielectric loss on the circuit is expressed by treating the capacitor as having equivalent resistance either in series or in shunt with it.

Equivalent series resistance :

$$R_s = \tan\psi/\omega C = (\text{P.F.})/\omega C$$

using the approximation for Power Factor given on page 1–11.

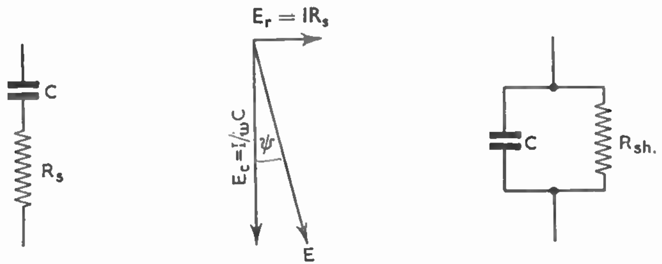


FIG. 6.—DIELECTRIC LOSS.

Equivalent shunt resistance:

$$R_{sh} = 1/\omega C \tan\psi = 1/(\omega C \times \text{P.F.})$$

Power loss:

$$\begin{aligned} P_L &= E^2/R_{sh} = \omega C E^2 \tan\psi \\ &= \omega C E^2 \times \text{P.F.} \\ &= I^2 R_s = (I^2 \times \text{P.F.})/\omega C. \end{aligned}$$

To convert a series loss resistance into its equivalent shunt value and vice versa

$$\begin{aligned} R_s &= 1/R_{sh}(\omega C)^2 \quad \text{when} \quad R_{sh} \gg 1/\omega C \\ R_{sh} &= 1/R_s(\omega C)^2 \quad \text{when} \quad R_s \ll 1/\omega C \end{aligned}$$

Power Factor of a Capacitor

The power factor is a measure of the dielectric losses in a capacitor. The losses are expressed in terms of the angle by which the current falls short of being 90° out of phase with the applied voltage.

$$\begin{aligned} \text{P.F.} &= \sin\psi \\ &\simeq \tan\psi = \omega CR \text{ for small values of } \psi \text{ in practice.} \end{aligned}$$

where
ψ = loss angle;
$\omega = 2\pi \times$ frequency (c/s);
C = capacitance, farads;
R = equivalent series loss resistance, ohms.

TABLE 3.—DIELECTRIC CONSTANTS AND POWER FACTORS OF INSULATING MATERIALS

Material	*Dielectric Constant at 1 Mc/s and 25° C.* *	*Power Factor at 1 Mc/s*
Bakelite	4·3–5·4	0·03–0·06
Ebonite	2·8	0·006
Frequentite	6	0·0006
Mica, ruby	5·4	0·0003
Micanite	7	Poor
Mycalex	7·5	0·001
Oil, cable	2·3	0·0008
Polystyrene	2·6	0·00007
Porcelain	5–6	0·0075–0·009
Pyrex	4·5	0·0002
Quartz	3·8	0·0002
Shellac	3·5	0·002
Water, distilled	78	0·04

* The dielectric constant falls slightly with increase of frequency. The power factor may either rise or fall with different materials as frequency increases.

B

EXAMPLE 2.—*A capacitor of 200 pF is subjected to an effective voltage of 1,000 at a frequency of 3 Mc/s. The dielectric loss and resistance of the plates and connectors are together equivalent to a series resistance of 10 ohms. Calculate:* (a) *the power factor, and* (b) *the power loss.*

(*a*)
$$\omega C = 2\pi \times 3 \times 10^6 \times 200 \times 10^{-12}$$
$$= 0{\cdot}00377$$
$$\text{Power factor} = \omega CR$$
$$= 0{\cdot}00377 \times 10$$
$$= \underline{0{\cdot}0377}$$

(*b*) Power loss:
$$P_L = \omega CE^2 \times \text{P.F.}$$
$$= 0{\cdot}00377 \times 10^6 \times 0{\cdot}0377$$
$$= \underline{142 \text{ watts}}$$

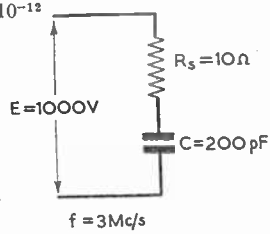


FIG. 7.—EXAMPLE 2.

A.C. APPLICATIONS OF OHM'S LAW

Impedance and Reactance

The opposition to the current flow in an A.C. circuit is called the impedance, and is measured in ohms. It is compounded of resistance and reactance, the latter arising from the presence of inductance or capacitance, or both. Resistance and reactance are also both measured in ohms.

Reactance due to inductance: $X_L = 2\pi fL$

Reactance due to capacitance: $X_C = 1/2\pi fC$

Impedance:
$$Z = R + j(X_L - X_C)$$
$$|Z| = \sqrt{R^2 + (X_L - X_C)^2}$$

where
L = inductance, henrys;
C = capacitance, farads;
R = resistance, ohms.

CONDUCTANCE, SUSCEPTANCE AND ADMITTANCE

Conductance is the reciprocal of resistance: $G = 1/R$
Susceptance is the reciprocal of reactance: $B = 1/X$
Admittance is the reciprocal of impedance: $Y = 1/Z$

Ohm's Law for A.C. Circuits

The various transformations for A.C. circuits are:

Voltage:	$E = IZ$	$= I/Y$
Current:	$I = E/Z$	$= EY$
Impedance:	$Z = E/I$	$= \sqrt{R^2 + X^2}$
Admittance:	$Y = 1/Z$	$= \sqrt{G^2 + B^2}$
Conductance:	$G = R/Z^2$	$= R/(R^2 + X^2)$
Susceptance:	$B = -X/Z^2$	$= -X/(R^2 + X^2)$
Power:	$P = EI \cos\phi$	$= I^2Z \cos\phi$
	$= I^2R$	$= E^2 \cos\phi/Z$

where
E = voltage;
I = current, amperes;
Z = impedance, ohms;
Y = admittance, mhos;
R = resistance, ohms;
X = reactance, ohms;
G = conductance, mhos;
B = susceptance, mhos;
P = power, watts;
ϕ = phase angle between current and voltage.

Impedances in Series and Parallel

Series: $Z = Z_1 + Z_2 + Z_3 + \ldots$ etc.

Parallel: $1/Z = 1/Z_1 + 1/Z_2 + 1/Z_3 + \ldots$ etc.

$Z = Z_1Z_2/(Z_1 + Z_2)$ for two impedances in parallel.

Series: $1/Y = 1/Y_1 + 1/Y_2 + 1/Y_3 + \ldots$ etc.

Parallel: $Y = Y_1 + Y_2 + Y_3 + \ldots$ etc.

where $Z_1Z_2Z_3$ = impedances in series or parallel;
$Y_1Y_2Y_3$ = admittances in series or parallel.

Z and Y are complex quantities, which must be added or subtracted vectorially by resolving into their real and imaginary parts.

$$Z = R + jX = \sqrt{R^2 + X^2}$$
$$Y = G + jB = \sqrt{G^2 + B^2}$$

Impedances (and reactances) are added vectorially when in series;
Admittances (and susceptances) are added vectorially when in parallel.

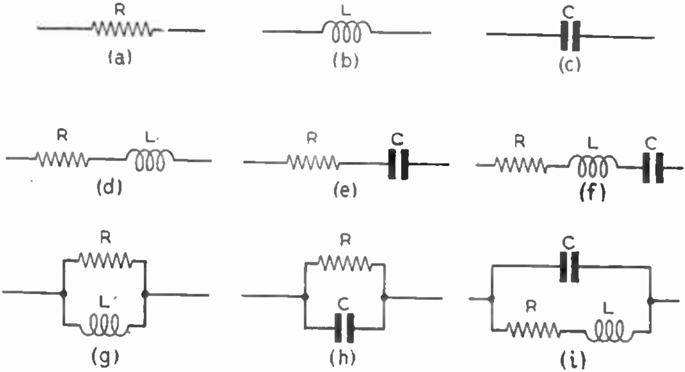


FIG. 8.—RESISTANCES AND REACTANCES IN SERIES AND PARALLEL.

Resistances and Reactances in Series and Parallel

Table 4.—Impedance and Phase Angle Formulæ

Fig. 8 Reference	*Magnitude of Impedance, $\lvert Z \rvert$*	*Phase Angle, ϕ*
(a)	R	0
(b)	X_L	$\pi/2$
(c)	X_c	$-\pi/2$
(d)	$\sqrt{R^2 + X_L^2}$	$\arctan (X_L/R)$
(e)	$\sqrt{R^2 + X_c^2}$	$-\arctan (X_c/R)$
(f)	$\sqrt{R^2 + (X_L - X_c)^2}$	$\arctan [(X_L - X_c)/R]$
(g)	$R/\sqrt{1 + (R/X_L)^2}$	$\arctan (R/X_L)$
(h)	$R/\sqrt{1 + (R/X_c)^2}$	$-\arctan (R/X_c)$
(i)	$\sqrt{(R^2 + X_L^2)/[(1 - X_L/X_c)^2 + (R/X_c)^2]}$	$\arctan [X_L(1 - X_L/X_c) - R^2/X_c]/R$

where X_L = reactance due to inductance ($= 2\pi fL$), ohms;
X_c = reactance due to capacitance ($= 1/2\pi fC$), ohms.

Example 1.—*A circuit consists of two branches in parallel, one containing a coil of inductance* 120 *mH in series with a resistance of* 150 *ohms, and the other containing a capacitor of* 0·2 *μF in series with a resistance of* 200 *ohms. If a voltage of* 100 *at a frequency of* 1,000 *c/s is applied, determine:* (a) *the branch currents;* (b) *their phase angles with respect to the applied voltage; and* (c) *the total current that will flow.*

Branch impedances:

$$\begin{aligned} Z_1 &= R_1 + j\omega L \\ &= 150 + j(2\pi \times 1{,}000 \times 120 \times 10^{-3}) \\ &= 150 + j753 \end{aligned}$$

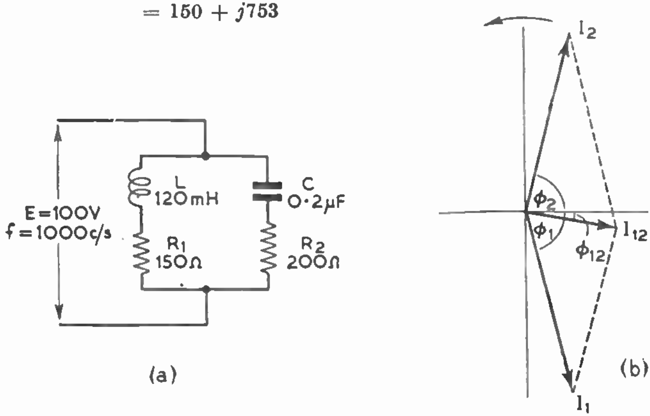


Fig. 9.—Example 1.

$$
\begin{aligned}
Z_2 &= R_2 - j/\omega C \\
&= 200 - j\left(\frac{1}{2\pi \times 1{,}000 \times 0{\cdot}2 \times 10^{-6}}\right) \\
&= 200 - j797
\end{aligned}
$$

Scalar values of impedances:

$$
\begin{aligned}
Z_1 &= \sqrt{R_1^2 + (\omega L)^2} = \sqrt{150^2 + 753^2} = 767 \text{ ohms} \\
Z_2 &= \sqrt{R_2^2 + (1/\omega C)^2} = \sqrt{200^2 + 797^2} = 822 \text{ ohms}
\end{aligned}
$$

Branch currents:

$$
I_1 = E/|Z_1| = \frac{100}{767} = \underline{0{\cdot}1305 \text{ amperes}}
$$

$$
I_2 = E/|Z_2| = \frac{100}{822} = \underline{0{\cdot}1217 \text{ amperes}}
$$

Phase angles of branch currents

$$
\phi_1 = \text{arc tan } (X_L/R_1) = \text{arc tan } \frac{753}{150} \quad \phi_1 = \underline{+\,79^\circ \text{ approx.}}
$$

$$
\phi_2 = \text{arc tan } (X_c/R_2) = \text{arc tan } \frac{797}{200} \quad \phi_2 = \underline{-\,76^\circ \text{ approx.}}
$$

Admittance of both branches in parallel:

$$
\begin{aligned}
Y_{12} &= Y_1 + Y_2 \\
&= \frac{1}{150 + j\,753} + \frac{1}{200 - j\,797} \\
&= \frac{150 - j\,753}{150^2 + 753^2} + \frac{200 + j\,797}{200^2 + 797^2} \\
&= (2{\cdot}54 \times 10^{-4}) - j(1{\cdot}28 \times 10^{-3}) + (2{\cdot}96 \times 10^{-4}) + j(1{\cdot}18 \times 10^{-3}) \\
&= (5{\cdot}50 \times 10^{-4}) - j(1{\cdot}00 \times 10^{-4})
\end{aligned}
$$

Total current:

$$
\begin{aligned}
I_{12} &= EY_{12} \\
&= 100\sqrt{(5{\cdot}50 \times 10^{-4})^2 + (1{\cdot}00 \times 10^{-4})^2} \\
&= \underline{0{\cdot}056 \text{ amperes}}
\end{aligned}
$$

EXAMPLE 2.—*Calculate the impedance of a combination of an inductance of 150 μH, a capacitance of 200 pF and a resistance of 10 kΩ in parallel at a frequency of 1 Mc/s.*

Branch A:

Inductive reactance

$$
\begin{aligned}
X_1 &= 2\pi f L \\
&= 2\pi \times 10^6 \times 150 \times 10^{-6} = 300\pi
\end{aligned}
$$

Susceptance

$$B_1 = -1/X_1$$

$$= -\frac{1}{300\pi} = -1{\cdot}061 \times 10^{-3}$$

Branch B:

Capacitive reactance

$$X_2 = -1/2\pi fC$$

$$= -\frac{1}{2 \times 10^6 \times 200 \times 10^{-12}} = -\frac{10^4}{4\pi}$$

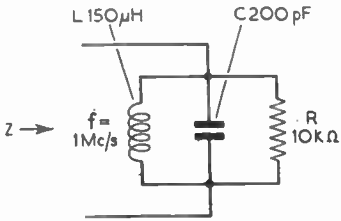


FIG. 10.—EXAMPLE 2.

Susceptance

$$B_2 = -1/X_2$$

$$= \frac{4\pi}{10^4} = 1{\cdot}257 \times 10^{-3}$$

Branch C:

Conductance

$$G_3 = 1/R$$

$$= \frac{1}{10^4} = 10^{-4}$$

Combined admittance

$$Y = jB_1 + jB_2 + G_3$$

$$= G_3 + j(B_1 + B_2)$$

$$= 10^{-4} + j(-1{\cdot}061 \times 10^{-3} + 1{\cdot}257 \times 10^{-3})$$

$$= 10^{-4} + j(1{\cdot}96 \times 10^{-4})$$

$$|Y| = \sqrt{10^{-8} + (3{\cdot}84 \times 10^{-8})}$$

$$= 2{\cdot}2 \times 10^{-4} \text{ mhos}$$

Combined impedance

$$|Z| = 1/|Y|$$

$$= \frac{1}{2{\cdot}2 \times 10^{-4}} = \underline{4{,}550 \text{ ohms}}$$

EXAMPLE 3.—*An e.m.f. of* 100 *volts at a frequency of* 2 *Mc/s is applied to a circuit consisting of:* (a) *an inductance* $L = 50\ \mu H$ *in series with a resistance* $R = 25$ *ohms in parallel with* (b) *a capacitance* $C = 100\ \mu\mu F$. *Find the combined impedance, the magnitudes of the currents through each branch, the total current and the phase relationships of the currents to the voltage.*

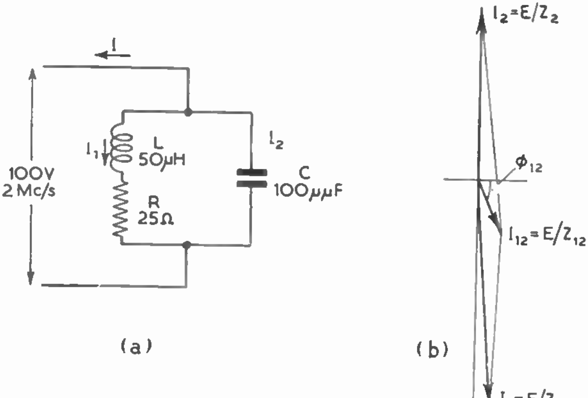


FIG. 11.—EXAMPLE 3.

Impedance of Branch A :

$$\begin{aligned} Z_1 &= R + jX_L = R + j\omega L \\ &= 25 + j(2\pi \times 2 \times 10^6 \times 50 \times 10^{-6}) \\ &= 25 + j628 \end{aligned}$$

Impedance of Branch B :

$$\begin{aligned} Z_2 &= -jX_c = -j/2\pi fC \\ &= -j\frac{1}{2 \times 2 \times 10^6 \times 10^{-10}} = -j\,796 \text{ ohms} \\ X_c &= 796 \text{ ohms} \end{aligned}$$

Then impedance of parallel combination :

$$\begin{aligned} Z_{12} &= Z_1 Z_2/(Z_1 + Z_2) \\ &= (R + jX_L)(-jX_c)/[R + j(X_L - X_c)] \\ &= X_c^2 R/[R^2 + (X_L - X_c)^2] \\ &\qquad + j[X_c(X_L X_c - R^2 - X_L^2)]/[R^2 + (X_L - X_c)^2] \\ &= \frac{796^2 \times 25}{25^2 + (628 - 796)^2} + j\frac{796(628 \times 796 - 25^2 - 628^2)}{25^2 + (628 - 796)^2} \\ &= 547 + j2880 \end{aligned}$$

Combined impedance :

$$|Z_{12}| = \sqrt{547^2 + 2{,}880^2} = \underline{2{,}940 \text{ ohms}}$$

Branch Currents :

Branch A $\quad I_1 = E/Z_1 = \dfrac{100}{628} = \underline{0{\cdot}159 \text{ amperes}}$

Branch B $\quad I_2 = E/Z_2 = \dfrac{100}{796} = \underline{0{\cdot}216 \text{ amperes}}$

Total $\quad I_{12} = E/Z_{12} = \dfrac{100}{2{,}940} = \underline{0{\cdot}034 \text{ amperes}}$

Phase Angles:

Between I_1 and E

$$\phi_1 = \text{arc tan}\,\frac{628}{25}$$

$$= \underline{87\tfrac{1}{2}^\circ \text{ approx.}} \quad I_1 \text{ lagging behind } E.$$

Between I_2 and E

$$\phi_2 = \text{arc tan}\,\frac{796}{0}$$

$$= \underline{90^\circ} \quad I_2 \text{ leading on } E.$$

Between I_{12} and E

$$\phi_{12} = \text{arc tan}\,\frac{2{,}880}{547}$$

$$= \underline{79^\circ \text{ approx.}} \quad I_2 \text{ lagging behind } E.$$

These results are illustrated vectorially in Fig. 11 (*b*).

CIRCUIT THEOREMS

Equivalence of Series and Parallel Complex Circuits

Resistance and reactance elements in parallel can be converted into series equivalents and vice versa by the use of the following relationships. These relationships are valid only for the particular frequency for which they are calculated.

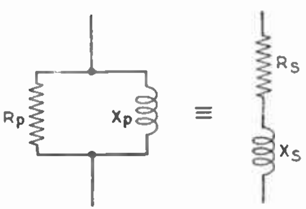


Fig. 12.—Equivalence of Series and Parallel Complex Circuits.

$$R_p = R_s + X_s^2/R_s$$
$$X_p = X_s + R_s^2/X_s$$
$$R_s = R_p/\{1 + (R_p/X_p)^2\}$$
$$X_s = X_p/\{1 + (X_p/R_p)^2\}$$

where

R_p = equivalent resistance of parallel circuit, ohms;
R_s = equivalent resistance of series circuit, ohms;
X_p = equivalent reactance of parallel circuit, ohms;
X_s = equivalent reactance of series circuit, ohms.

Example 1.—*Translate the circuit shown in Fig.* 13 *into its simplest series equivalent for a frequency of* 1 *Mc/s.*

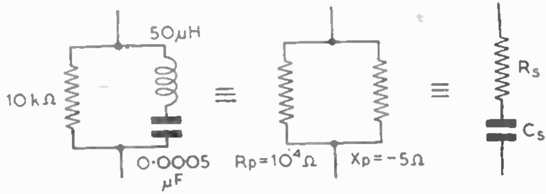


FIG. 13.—EXAMPLE 1.

Considering the right-hand branch:

Inductive reactance

$$X_L = 2\pi fL = 6{\cdot}28 \times 10^6 \times 50 \times 10^{-6} = 314 \text{ ohms}$$

Capacitive reactance

$$X_c = 1/2\pi fC = \frac{1}{6{\cdot}28 \times 10^6 \times 5 \times 10^{-10}} = 319 \text{ ohms}$$

Effective reactance

$$X_p = X_L - X_c = 314 - 319 = -5 \text{ ohms}$$

Equivalent series resistance

$$R_s = R_p/\{1 + (R_p/X_p)^2\}$$

$$= \frac{10^4}{1 + (10^4/-5)^2} = \underline{0{\cdot}0025 \text{ ohms}}$$

Equivalent series reactance

$$X_s = X_p/\{1 + (X_p/R_p)^2\}$$

$$= \frac{-5}{1 + (-5/10^4)^2} = -5 \text{ ohms}$$

This is equivalent to a capacitance C_s in series:

$$C_s = -1/2\pi fX_s \times 10^6 \ \mu\text{F}$$

$$= \frac{10^6}{2\pi \times 10^6 \times 5} = \underline{0{\cdot}032 \ \mu\text{F}}$$

Star and Mesh Networks

A network of star-connected impedances Z_a, Z_b, Z_c is equivalent to a mesh-connected network Z_1, Z_2, Z_3 between terminals AB, BC, CA when

$$Z_a = Z_1Z_3/(Z_1 + Z_2 + Z_3)$$
$$Z_b = Z_1Z_2/(Z_1 + Z_2 + Z_3)$$
$$Z_c = Z_2Z_3/(Z_1 + Z_2 + Z_3)$$

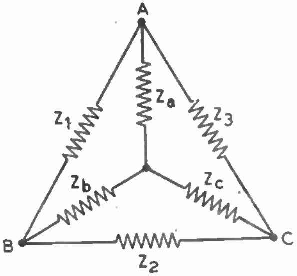


FIG. 14.—STAR AND MESH NETWORK.

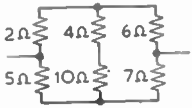


FIG. 15.—EXAMPLE 2.

For the converse,

$$Z_1 = Z_a + Z_b + Z_aZ_b/Z_c$$
$$Z_2 = Z_b + Z_c + Z_bZ_c/Z_a$$
$$Z_3 = Z_c + Z_a + Z_aZ_c/Z_b$$

EXAMPLE 2.—*Find the equivalent resistance of the resistance network shown in Fig.* 15.

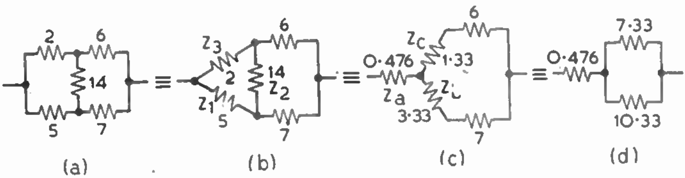


FIG. 16.—SOLUTION OF EXAMPLE 2.

The circuit can be simplified successively as in Fig. 16 (*a*) and (*b*) and the mesh-connected portion converted into the equivalent star as in (*c*). Then the impedance of the arms of the star are:

$$Z_a = \frac{5 \times 2}{5 + 14 + 2} = 0{\cdot}476 \text{ ohms}$$

$$Z_b = \frac{5 \times 14}{5 + 14 + 2} = 3{\cdot}33 \text{ ohms}$$

$$Z_c = \frac{14 \times 2}{5 + 14 + 2} \doteq 1{\cdot}33 \text{ ohms}$$

Re-arranging (*c*) as in (*d*)

$$Z_e = 0{\cdot}476 + \frac{7{\cdot}33 \times 10{\cdot}33}{7{\cdot}33 + 10{\cdot}33} = \underline{4{\cdot}77 \text{ ohms}}$$

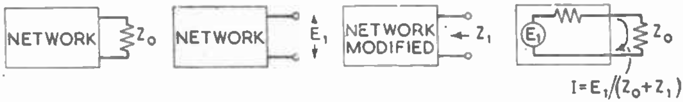


FIG. 17.—THEVENIN'S THEOREM.

Thevenin's Theorem

The current in any branch Z_0 of a network is the same as if it were connected to a generator of e.m.f. E_1 and internal impedance Z_1.

$$I = E_1/(Z_0 + Z_1)$$

where E_1 = voltage appearing across the branch when open-circuited;
Z_1 = impedance of network between branch terminals, all sources of e.m.f. in the network being replaced by their internal impedances.

EXAMPLE 3.—*Find, by the use of Thevenin's Theorem, the current indicated by the milliammeter in the unbalanced bridge network of Fig.* 18.

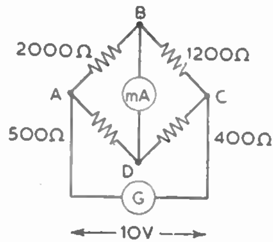


FIG. 18.—EXAMPLE 3.

Remove branch BD, Fig. 19 (*a*), in which it is required to know the current. Let E_2 = p.d. across branch AB and E_3 = p.d. across branch AD. Then the e.m.f. appearing across branch BD

$$\begin{aligned} E_1 &= E_2 - E_3 \\ &= \left(\frac{2{,}000}{2{,}000 + 1{,}200} \times 10\right) - \left(\frac{500}{500 + 400} \times 10\right) \\ &= 0{\cdot}694 \text{ volts} \end{aligned}$$

Impedance of network between B and D (Fig. 19 (*b*))

$$\begin{aligned} Z &= \frac{2{,}000 \times 1{,}200}{2{,}000 + 1{,}200} + \frac{500 \times 400}{500 + 400} \\ &= 972 \text{ ohms} \end{aligned}$$

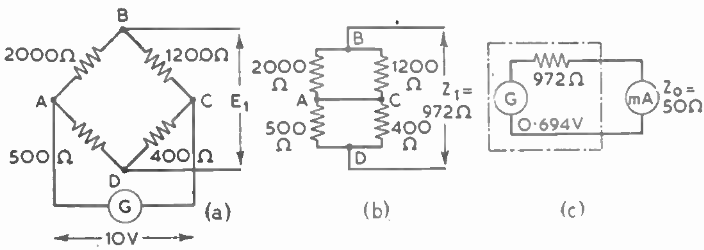


FIG. 19.—SOLUTION OF EXAMPLE 3.

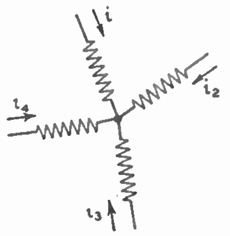


FIG. 20.—KIRCHHOFF'S LAW.

Sum of Currents.

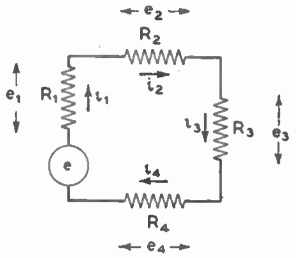


FIG. 21.—KIRCHHOFF'S LAW.

Sum of Products of Current and Resistance.

The equivalent network, Fig. 19 (c), is then a generator of e.m.f. = 0·694 volt and internal resistance Z_1 = 972 ohms. Hence the current in branch BD

$$\begin{aligned} I &= E_1/(Z_1 + Z_0) \\ &= \frac{0{\cdot}694}{972 + 50} \\ &= 0{\cdot}00068 \text{ ampere or } \underline{0{\cdot}68 \text{ mA}} \end{aligned}$$

Kirchhoff's Laws

(1) The algebraic sum of the currents which meet at a junction point in a circuit at any instant is zero (Fig. 20).

$$i_1 + i_2 + i_3 + \ldots = \Sigma i = 0$$

(2) The algebraic sum of the products of current and resistance of each part of a closed circuit is equal to the e.m.f. acting in the circuit (Fig. 21).

$$i_1R_1 + i_2R_2 + i_3R_3 + \ldots = \Sigma iR = e$$

N.B.—These laws apply to instantaneous and vector quantities, but not to r.m.s. values.

EXAMPLE 4.—*Calculate the currents flowing in each branch of the network shown in Fig.* 22 (a).

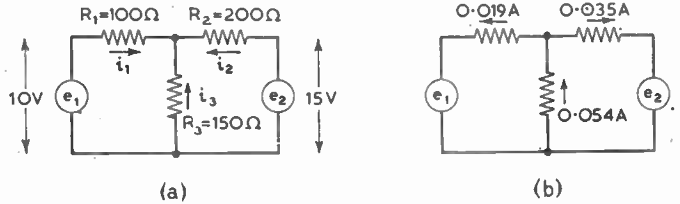


FIG. 22.—EXAMPLE 4.

Applying Kirchhoff's Laws and assuming directions of currents shown in Fig. 22 (*a*), we obtain three simultaneous equations, from which i_1, i_2 and i_3 can be evaluated.

$$i_1 + i_2 + i_3 = 0 \quad . \quad . \quad . \quad . \quad . \quad (1)$$

$$e_1 + i_1R_1 - i_3R_3 = 0 \quad . \quad . \quad . \quad . \quad . \quad (2)$$

$$e_2 + i_2R_2 - i_3R_3 = 0 \quad . \quad . \quad . \quad . \quad . \quad (3)$$

Subtracting (3) from (2) to eliminate i_3

$$e_1 - e_2 + i_1R_1 - i_2R_2 = 0$$

$$10 - 15 + 100i_1 - 200i_2 = 0$$

$$i_2 = 0{\cdot}5i_1 - 0{\cdot}025 \quad . \quad . \quad (4)$$

Multiplying (1) through by R_2 and subtracting (3) to eliminate i_2

$$i_1R_2 - e_2 + i_3(R_2 + R_3) = 0$$

$$200i_1 - 15 + 350i_3 = 0$$

$$i_3 = 0{\cdot}0429 - 0{\cdot}571i_1 \quad . \quad . \quad (5)$$

Substituting these values of i_2 and i_3 in (1) to evaluate i_1

$$i_1 + 0{\cdot}5i_1 - 0{\cdot}025 - 0{\cdot}571i_1 + 0{\cdot}0429 = 0$$

$$i_1 = \underline{-\ 0{\cdot}0194 \text{ ampere}}$$

Substituting this value of i_1 in (4) to evaluate i_2

$$i_2 = (0{\cdot}5 \times -\ 0{\cdot}0194) - 0{\cdot}025$$

$$= \underline{-\ 0{\cdot}0347 \text{ ampere}}$$

Substituting the value of i_1 in (5) to evaluate i_3

$$i_3 = 0{\cdot}0429 - (0{\cdot}571 \times -\ 0{\cdot}0194)$$

$$= \underline{+\ 0{\cdot}054 \text{ ampere}}$$

GAINS AND LOSSES

Power Levels

In communication systems the decibel ($\frac{1}{10}$ Bel) is the practical unit of power, voltage or current level. It is used in various ways to define :

(*a*) a difference in level (gain or loss) of power or voltage between two points, e.g., the output and input of an amplifier or transmission line;

(*b*) a change in level (increase or decrease), e.g., in the output of an amplifier;

(*c*) a definite level by reference to some arbitrary level, called zero level.

Zero level of power (European standard) = 1 mW
Zero level of power (U.S.A. standard) = 6 mW
Zero level of acoustic loudness = 0·0007 dynes/sq. cm. at 400 c/s

A difference or change in level is expressed as so many decibels up or down, or so many decibels gain or loss.

A particular level is expressed as so many decibels above or below zero level.

Power Gain or Loss

Power gain or loss, decibels :

$$N = 10 \log_{10} (P_2/P_1) \text{ is the gain}$$
$$-N = 10 \log_{10} (P_1/P_2) \text{ is the loss}$$

Conversely, a power ratio :

$$P_2/P_1 = \text{antilog } (N/10)$$

where P_1 = power input or reference level;
P_2 = power output.

The overall gain of a system is the sum of the individual gains of each link, in decibels, gains being reckoned + and losses as − gains.

When $P_2/P_1 < 1$ (a loss), calculation is simplified by inverting the ratio and prefixing by the negative sign.

$$N = -10 \log_{10} (P_1/P_2)$$

Voltage Gain or Loss

Voltage or current gain or loss, decibels :

$$N = 20 \log_{10} (E_2/E_1) = 20 \log_{10} (I_2/I_1)$$

Conversely, a voltage ratio :

$$E_2/E_1 = \text{antilog}_{10} (N/20)$$

Example 1.—*What is the percentage efficiency of power transfer in a transmission line which has a loss of 2 db ?*

Loss in decibels $N = -10 \log_{10} (P_1/P_2)$

Hence
$$\begin{aligned} P_1/P_2 &= \text{antilog } (-N/10) \\ &= \text{antilog } (2/10) \\ &= 1{\cdot}585 \end{aligned}$$

Efficiency, per cent
$$\begin{aligned} &= 100\, P_2/P_1 \\ &= 100 \times \frac{1}{1{\cdot}585} = \underline{63{\cdot}1 \text{ per cent}} \end{aligned}$$

Example 2.—*The power output from an amplifier is increased from 2 to 5 watts. What is the change in level in decibels ?*

$$\begin{aligned} N &= 10 \log_{10} (5/2) = 10 \log_{10} 2{\cdot}5 \\ &= 3{\cdot}979 \text{ or } \underline{4 \text{ db gain approx.}} \end{aligned}$$

Example 3.—*If the energy gain of an aerial array over a single element is 12·6, what is the gain in decibels ?*

$$N = 10 \log_{10} 12{\cdot}6 = 10 \times 1{\cdot}1004 = \underline{11 \text{ db approx.}}$$

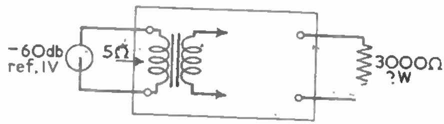
EXAMPLE 4.—*The field strength of a desired signal at a receiver is 20 μV and that of an interfering signal is 9 μV. What is the signal/interference ratio in decibels?*

Signal interference ratio = 20/9

$$N = 20 \log_{10} (20/9) = 20 \log_{10} 2{\cdot}222 = \underline{6{\cdot}94 \text{ db}}$$

EXAMPLE 5.—*The output level of a microphone is 60 db below 1 volt when working into a resistance of 5 ohms. Find the gain to be provided by an amplifier to produce an output of 2 watts into 3,000 ohms.*

FIG. 23.—EXAMPLE 5.



Input level = − 60 db reference 1 volt

Input voltage:

$$E_1 = \text{antilog } (N/20) = \text{antilog } (-60/20) = 0{\cdot}001 \text{ volt}$$

Power input:

$$P_1 = E_1^2/R_1 = 0{\cdot}001^2/5 = 2 \times 10^{-7} \text{ watt}$$

Gain to be provided:

$$N = 10 \log_{10} (P_2/P_1) = 10 \log_{10} [2/(2 \times 10^{-7})]$$
$$= 10 \log_{10} 10^7 = \underline{70 \text{ db}}$$

Impedance Changes

If the input and output impedances are not equal, voltage gain or loss must be specified with reference to the impedances. Then

$$N = 20 \log_{10} (E_2/E_1) + 10 \log_{10} (Z_1/Z_2)$$

where $Z_1 Z_2$ = input and output impedances, respectively, provided Z_1 and Z_2 have the same phase angle.

Power gain or loss can be specified without reference to impedance.

EXAMPLE 6.—*An amplifier provides a gain of 40 db with an input of 0·5 volt across 2 MΩ and an output load resistance of 4,000 ohms. Determine the voltage across the load and the power output.*

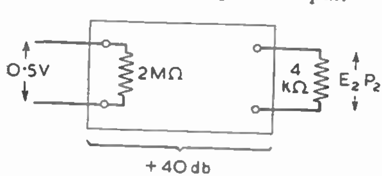


FIG. 24.—EXAMPLE 6.

$$N = 20 \log_{10} (E_2/E_1) + 10 \log_{10} (R_1/R_2)$$

Output voltage :

$$E_2 = E_1 \text{ antilog } \left(\frac{N - 10 \log_{10} (R_1/R_2)}{20}\right)$$

$$= 0{\cdot}5 \text{ antilog } \left(\frac{40 - 10 \log_{10} (2 \times 10^6/4{,}000)}{20}\right)$$

$$= 0{\cdot}5 \text{ antilog } 0{\cdot}65 = \underline{2{\cdot}23 \text{ volts}}$$

Power output :

$$P_2 = E_2^2/R_2 \text{ watts} = \frac{2{\cdot}23^2 \times 1{,}000}{4{,}000} \text{ mW}$$

$$= \underline{1{\cdot}24 \text{ mW}}$$

EXAMPLE 7.—*The voltage gain of an amplifier stage is equal to* $g_m R_L R_a/(R_L + R_a)$. *Determine the stage power gain in decibels, given that*

Mutual conductance, g_m = 1·8 *mA/volt*;
Anode A.C. resistance. R_a = 5,000 *ohms*;
Load resistance, R_L = 20,000 *ohms*;
Input resistance = 1 *M*Ω.

$$g_m = 1{\cdot}8 \times 10^{-3} \text{ ampere/volt}$$

VOLTAGE GAIN

$$m = g_m R_L R_a/(R_L + R_a)$$

$$= \frac{1{\cdot}8 \times 10^{-3} \times 20{,}000 \times 5{,}000}{20{,}000 + 5{,}000}$$

$$= \frac{18}{2{\cdot}5} = 7{\cdot}2$$

POWER GAIN IN DB

$$N = 20 \log_{10} (E_2/E_1) + 10 \log_{10} (R_1/R_2)$$
$$= 20 \log_{10} 7{\cdot}2 + 10 \log [10^6/(2 \times 10^4)]$$
$$= 20 \log_{10} 7{\cdot}2 + 10 \log 50$$
$$= \underline{34{\cdot}14 \text{ db}}$$

Volume Units

In audio-frequency work the volume unit (VU) is the volume level above zero reference level, and is taken to be 1 mW into 600 ohms.

ELECTRICAL TOLERANCES

Electrical tolerance is the maximum deviation allowable from the nominal value, usually expressed as a percentage of the nominal value.

$$p = 100\ r/R$$

where R = nominal value;
r = deviation from nominal value.

Resistances or Inductances in Series (or Capacitors in Parallel).

Overall Tolerance (Per Cent)

$$p = (p_1R_1 + p_2R_2 + \ldots)/(R_1 + R_2 + \ldots)$$

Two resistances or inductances in parallel (or two capacitors in series)

$$p = \frac{p_1R_2(1 + p_2/100) + p_2R_1(1 + p_1/100)}{R_1(1 + p_1/100) + R_2(1 + p_2/100)}$$

$$\simeq (p_1R_2 + p_2R_1)/(R_1 + R_2) \text{ for small tolerances}$$

where p_1p_2, etc. = percentage tolerances of individual resistances;
R_1R_2, etc. = nominal values of individual resistances.

Example 1.—*Two resistors of* 500 *and* 1,000 *ohms have respectively tolerances of* $\pm 2\frac{1}{2}$ *and* ± 5 *per cent. What is the maximum percentage variation in resistance of the combination when connected:* (a) *in series;* (b) *in parallel?*

(*a*)
$$p = (p_1R_1 + p_2R_2)/(R_1 + R_2)$$
$$= \frac{(2{\cdot}5 \times 500) + (5 \times 1{,}000)}{500 + 1{,}000} = \underline{4{\cdot}2 \text{ per cent}}$$

(*b*)
$$p = (p_1R_2 + p_2R_1)/(R_1 + R_2)$$
$$= \frac{(2{\cdot}5 \times 1{,}000) + (5 \times 500)}{500 + 1{,}000} = \underline{3{\cdot}3 \text{ per cent}}$$

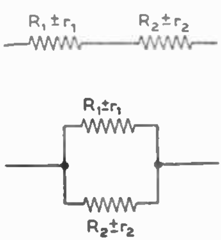


Fig. 25.—Overall Tolerance.

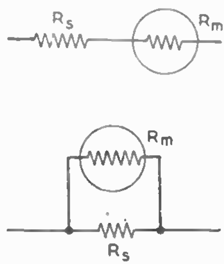


Fig. 26.—Meter Shunts.

Meter Resistances

Voltmeter and Series Resistance

Overall accuracy per cent :

$$p = (p_m R_m + p_s R_s)/(R_m + R_s)$$

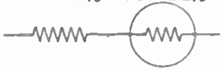
Ammeter and Shunt Resistance

$$p = (p_{sh} - p_m)R_m/(R_m + R_{sh})$$

where p_m = percentage overall accuracy of meter alone;
p_s = percentage tolerance of series resistance;
p_{sh} = percentage tolerance of shunt resistance;
R_m = nominal resistance of meter alone;
R_s = nominal value of series resistance;
R_{sh} = nominal value of shunt resistance.

Example 2.—*The accuracy of a moving coil meter is ±1 per cent and the resistance of the coil is 20 ohms. What is the maximum possible overall error when the meter is used: (a) as a voltmeter in series with a resistance of 198 ohms ±2 per cent; (b) as a milliammeter shunted by a resistance of 2·22 ohms ±2 per cent?*

(a)
$$p = (p_m R_m + p_s R_s)/(R_m + R_s)$$
$$= \frac{(1 \times 20) + (2 \times 198)}{20 + 198}$$
$$= \underline{1{\cdot}9 \text{ per cent}}$$



(b)
$$p = (p_{sh} - p_m)R_m/(R_m + R_{sh})$$
$$= \frac{(2 - 1) \times 10}{10 + 2{\cdot}22} = \underline{0{\cdot}82 \text{ per cent}}$$

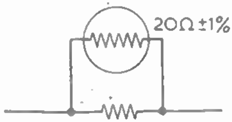


Fig. 27.—Example 2.

Frequency Tolerance

Frequency tolerance is the maximum deviation allowable above or below the nominal frequency, expressed as a percentage of the nominal frequency.

$$p = 100\Delta f/f$$

where f = nominal frequency;
Δf = deviation from nominal frequency.

When sum and difference frequencies are produced by modulation, de-modulation or heterodyning, the maximum percentage divergence of the resulting frequencies

$$p = (p_0 f_0 + p_m f_m)/(f_0 + f_m) \text{ for the upper (sum) frequency}$$
$$p = (p_0 f_0 + p_m f_m)/(f_0 - f_m) \text{ for the lower (difference) frequency}$$

where p_0 = percentage variation of signal frequency;
p_m = percentage variation of modulating frequency;
f_0 = nominal signal frequency;
f_m = nominal modulating frequency.

EXAMPLE 3.—*A received signal of frequency 16 Mc/s having a maximum variation of ±0·01 per cent is applied to the mixer stage of a receiver, where it is heterodyned by a local oscillator at a frequency of 15·4 Mc/s ±0·005 per cent. What is the maximum percentage variation of the difference frequency ?*

$$p = (p_0 f_0 + p_m f_m)/(f_0 - f_m)$$

$$= \frac{(0{\cdot}01 \times 16) + (0{\cdot}005 \times 15{\cdot}4)}{16 - 15{\cdot}4}$$

$$= \underline{0{\cdot}395 \text{ per cent}}$$

FREQUENCY AND WAVELENGTH

The velocity of electromagnetic wave propagation in a given medium is constant, and in space is equal to that of light, $c = 2{\cdot}998 \times 10^8$ metres/second approximately. For all practical applications c is taken as 3×10^8 metres/second. The frequency, in terms of velocity (c) and wavelength (λ) in metres,

$$= c/\lambda$$

More useful forms are :

$$f \text{ (in c/s)} = 3 \times 10^8/\lambda$$
$$f \text{ (in kc/s)} = 3 \times 10^5/\lambda$$
$$f \text{ (in Mc/s)} = 300/\lambda$$

Useful datum points for mentally converting frequency and wavelength are given in Table 5A.

TABLE 5A.—FREQUENCY–WAVELENGTH CONVERSION

f in Mc/s .	1	3	10	30	100	300	1,000	3,000
λ in metres	300	100	30	10	3	1	0·3	0·1

Table 5B on the following pages gives conversion points throughout the range 10·0–99·9 in 0·1 steps, and may be used for both wavelength-to-frequency and frequency-to-wavelength conversion provided that the correct units and position of the decimal points are ascertained from Table 5A.

Table 5B.

Wavelength/Frequency Conversion Table

	0	1	2	3	4	5	6	7	8	9
10	30,000	29,703	29,412	29,126	28,846	28,571	28,302	28,037	27,778	27,523
11	27,273	27,027	26,786	26,549	26,316	26,087	25,862	25,641	25,424	25,210
12	25,000	24,793	24,590	24,390	24,193	24,000	23,809	23,622	23,437	23,256
13	23,077	22,901	22,727	22,556	22,388	22,222	22,059	21,898	21,739	21,583
14	21,429	21,277	21,127	20,979	20,833	20,690	20,548	20,408	20,270	20,134
15	20,000	19,867	19,737	19,608	19,480	19,355	19,231	19,108	18,987	18,868
16	18,750	18,634	18,518	18,405	18,293	18,182	18,072	17,964	17,857	17,752
17	17,647	17,544	17,442	17,341	17,241	17,143	17,045	16,949	16,854	16,760
18	16,667	16,575	16,483	16,393	16,304	16,216	16,129	16,043	15,957	15,873
19	15,790	15,707	15,625	15,544	15,464	15,385	15,306	15,228	15,151	15,075
20	15,000	14,925	14,851	14,778	14,706	14,634	14,563	14,493	14,423	14,354
21	14,286	14,218	14,151	14,084	14,019	13,954	13,889	13,825	13,762	13,699
22	13,636	13,575	13,513	13,453	13,493	13,333	13,274	13,216	13,158	13,100
23	13,043	12,987	12,931	12,875	12,820	12,766	12,712	12,658	12,605	12,552
24	12,500	12,448	12,397	12,346	12,295	12,245	12,195	12,146	12,097	12,048
25	12,000	11,952	11,905	11,858	11,811	11,765	11,719	11,673	11,628	11,583
26	11,538	11,494	11,450	11,407	11,364	11,321	11,278	11,236	11,194	11,152
27	11,111	11,070	11,029	10,989	10,949	10,909	10,870	10,830	10,791	10,753
28	10,714	10,676	10,638	10,601	10,563	10,526	10,489	10,453	10,417	10,381
29	10,345	10,309	10,274	10,239	10,204	10,169	10,135	10,101	10,067	10,033
30	10,000	9,967	9,934	9,901	9,868	9,836	9,804	9,772	9,740	9,709
31	9,677	9,646	9,615	9,585	9,554	9,524	9,494	9,464	9,434	9,404
32	9,375	9,346	9,317	9,288	9,259	9,231	9,202	9,174	9,146	9,109
33	9,091	9,063	9,036	9,009	8,982	8,955	8,929	8,901	8,876	8,850
34	8,823	8,798	8,772	8,746	8,721	8,696	8,671	8,645	8,621	8,596
35	8,571	8,547	8,523	8,499	8,375	8,451	8,427	8,403	8,380	8,356
36	8,333	8,310	8,287	8,264	8,242	8,219	8,197	8,174	8,152	8,130
37	8,108	8,086	8,065	8,043	8,021	8,000	7,979	7,958	7,937	7,916
38	7,895	7,874	7,853	7,833	7,813	7,792	7,772	7,752	7,732	7,712
39	7,692	7,672	7,653	7,633	7,614	7,595	7,576	7,557	7,538	7,519
40	7,500	7,481	7,463	7,444	7,426	7,407	7,389	7,371	7,353	7,335
41	7,317	7,299	7,282	7,264	7,246	7,229	7,211	7,194	7,177	7,160
42	7,143	7,126	7,109	7,092	7,075	7,059	7,042	7,026	7,009	6,993
43	6,977	6,961	6,944	6,928	6,912	6,897	6,881	6,865	6,850	6,834
44	6,818	6,803	6,787	6,772	6,757	6,742	6,727	6,711	6,696	6,682
45	6,667	6,652	6,637	6,622	6,608	6,593	6,579	6,565	6,550	6,536
46	6,522	6.508	6,494	6,479	6,666	6,452	6,438	6,424	6,410	6,397
47	6,388	6,369	6,356	6,343	6,329	6,316	6,302	6,289	6,276	6,263
48	6,250	6,237	6,224	6,211	6,198	6,186	6,173	6,160	6,148	6,135
49	6,122	6,110	6,097	6,085	6,073	6,061	6,048	6,036	6,024	6,012
50	6,000	5,988	5,976	5,964	5,952	5,941	5,929	5,917	5,905	5,894
51	5,882	5,871	5,859	5,848	5,836	5,825	5,814	5,803	5,791	5,780
52	5,769	5,758	5,747	5,736	5,725	5,714	5,703	5,692	5,682	5,671
53	5,660	5,650	5,639	5,629	5,618	5,607	5,597	5,587	5,576	5,566
54	5,556	5,545	5,535	5,525	5,515	5,505	5,494	5,485	5,474	5,464

Examples:

27·2 metres = 11,029 kc/s.
2720 metres = 110·29 kc/s.
272 kc/s = 1102·9 metres.
2·72 Mc/s = 110·29 metres.
27·2 cm = 1102·9 Mc/s.

49·8 metres = 6,024 kc/s.
498 kc/s = 602·4 metres.
4·98 Mc/s = 60·24 metres.
498 metres = 602·4 kc/s.
49·8 cm = 602·4 Mc/s.

Table 5b (*contd.*)

Wavelength/Frequency Conversion Table

	0	1	2	3	4	5	6	7	8	9
55	5,455	5,445	5,435	5,425	5,415	5,405	5,396	5,386	5,376	5,367
56	5,357	5,347	5,338	5,329	5,319	5,310	5,300	5,291	5,282	5,272
57	5,263	5,254	5,245	5,236	5,227	5,217	5,208	5,199	5,190	5,181
58	5,172	5,164	5,155	5,146	5,137	5,128	5,119	5,111	5,102	5,093
59	5,085	5,076	5,068	5,059	5,050	5,042	5,034	5,025	5,017	5,008
60	5,000	4,992	4,983	4,975	4,967	4,959	4,951	4,942	4,934	4,926
61	4,918	4,910	4,902	4,894	4,886	4,878	4,870	4,862	4,854	4,846
62	4,839	4,831	4,823	4,815	4,808	4,800	4,792	4,785	4,777	4,769
63	4,762	4,754	4,747	4,739	4,732	4,724	4,717	4,710	4,702	4,695
64	4,687	4,680	4,673	4,666	4,658	4,651	4,644	4,637	4,630	4,622
65	4,615	4,608	4,601	4,594	4,587	4,580	4,573	4,566	4,559	4,552
66	4,546	4,539	4,532	4,525	4,518	4,511	4,504	4,498	4,491	4,484
67	4,477	4,471	4,464	4,458	4,451	4,444	4,438	4,431	4,425	4,418
68	4,412	4,405	4,399	4,392	4,386	4,380	4,373	4,367	4,360	4,354
69	4,348	4,342	4,335	4,329	4,323	4,316	4,310	4,304	4,298	4,292
70	4,286	4,279	4,273	4,267	4,261	4,255	4,249	4,243	4,237	4,231
71	4,225	4,219	4,213	4,207	4,202	4,196	4,190	4,184	4,178	4,172
72	4,167	4,161	4,155	4,149	4,144	4,138	4,132	4,126	4,121	4,115
73	4,110	4,104	4,098	4,093	4,087	4,081	4,076	4,071	4,065	4,060
74	4,054	4,048	4,043	4,038	4,032	4,027	4,021	4,016	4,011	4,005
75	4,000	3,995	3,989	3,984	3,979	3,973	3,968	3,963	3,958	3,952
76	3,947	3,942	3,937	3,932	3,927	3,922	3,916	3,911	3,906	3,901
77	3,896	3,891	3,886	3,881	3,876	3,871	3,866	3,861	3,856	3,851
78	3,846	3,841	3,836	3,831	3,826	3,822	3,817	3,812	3,807	3,802
79	3,797	3,793	3,788	3,783	3,778	3,774	3,769	3,764	3,759	3,754
80	3,750	3,745	3,741	3,736	3,731	3,727	3,722	3,718	3,713	3,708
81	3,704	3,699	3,694	3,690	3,685	3,681	3,676	3,672	3,667	3,663
82	3,658	3,654	3,649	3,645	3,641	3,636	3,632	3,628	3,623	3,619
83	3,614	3,610	3,606	3,601	3,597	3,593	3,589	3,584	3,580	3,576
84	3,571	3,567	3,563	3,559	3,554	3,550	3,546	3,542	3,538	3,534
85	3,529	3,525	3,521	3,517	3,513	3,509	3,505	3,501	3,496	3,492
86	3,488	3,484	3,480	3,476	3,472	3,468	3,464	3,460	3,456	3,452
87	3,448	3,444	3,440	3,436	3,433	3,429	3,425	3,421	3,417	3,413
88	3,409	3,405	3,401	3,397	3,394	3,390	3,386	3,382	3,378	3,375
89	3,371	3,367	3,363	3,359	3,356	3,352	3,348	3,344	3,341	3,337
90	3,333	3,330	3,326	3,322	3,319	3,315	3,311	3,307	3,304	3,300
91	3,297	3,293	3,289	3,286	3,282	3,279	3,275	3,271	3,268	3,264
92	3,261	3,257	3,254	3,250	3,247	3,244	3,240	3,236	3,233	3,229
93	3,226	3,222	3,219	3,215	3,212	3,208	3,205	3,202	3,198	3,195
94	3,191	3,188	3,185	3,181	3,178	3,175	3,171	3,168	3,165	3,161
95	3,158	3,154	3,151	3,148	3,145	3,141	3,138	3,135	3,131	3,128
96	3,125	3,122	3,118	3,115	3,112	3,109	3,106	3,102	3,099	3,096
97	3,093	3,090	3,086	3,083	3,080	3,077	3,074	3,070	3,067	3,064
98	3,061	3,058	3,055	3,052	3,049	3,046	3,043	3,040	3,036	3,033
99	3,030	3,027	3,024	3,021	3,018	3,015	3,012	3,009	3,006	3,003

Examples:

84·3 metres = 3,559 kc/s.
843 metres = 355·9 kc/s.
8·43 Mc/s = 35·59 metres.
8430 Mc/s = 3·559 cm.
8430 metres = 35·59 kc/s.

57·5 metres = 5·217 Mc/s.
575 kc/s = 521·7 metres.
5·75 Mc/s = 52·17 metres.
575 Mc/s = 52·17 cm.
57·5 kc/s = 5217 metres.

RESONANT CIRCUITS

Resonance

Resonance occurs when the inductive and capacitive reactances of a tuned circuit are equalized, i.e., when

$$X_L = X_c$$
$$\omega L = 1/\omega C$$

RESONANT FREQUENCY

$$\omega_r = 1/\sqrt{LC}$$
$$f_r = 1/2\pi\sqrt{LC}$$

where $\omega = 2\pi \times$ frequency (c/s);
$\omega_r = 2\pi \times$ resonant frequency (c/s);
$L =$ inductance, henrys;
$C =$ capacitance, farads;
$f_r =$ resonant frequency, c/s.

If f is in kc/s, L in μH and C in μF

$$f_r\text{kc/s} = 159/\sqrt{LC}$$
$$\lambda\,\text{m} = 1885\sqrt{LC}$$

Series Resonant Circuit

MAGNIFICATION FACTOR

$$Q = X_L/R = X_c/R \text{ at resonance}$$

Impedance is a minimum and current a maximum at resonance.

Then
$$Z = R$$
$$I = E/R$$

IMPEDANCE OFF RESONANCE

$$Z = \sqrt{R^2 + (X_L - X_c)^2}$$
$$= R\sqrt{1 + (QF)^2}$$

CURRENT OFF RESONANCE

$$I = I_r/\sqrt{1 + (QF)^2}$$

for excitation of constant amplitude

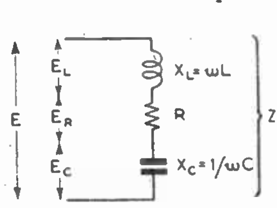


FIG. 28.—SERIES RESONANT CIRCUIT.

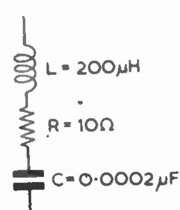


FIG. 29.—EXAMPLE 1.

PHASE ANGLE OF CURRENT

$$\phi = \text{arc tan}\ (X/R)$$
$$= \text{arc tan}\ (QF)$$

CIRCUIT VOLTAGES AT RESONANCE

$$E_L = E_c = EQ$$
$$= EX_L/R = EX_c/R$$

where Z = impedance of circuit, ohms;
R = resistance of coil and capacitor, ohms;
Q = magnification factor;
$F = 2\Delta\omega/\omega_r = 2\Delta f/f_r$, where Δf = frequency deviation;
E = voltage across circuit;
E_L = voltage across coil;
E_c = voltage across capacitor.
$X = X_L - X_c$

EXAMPLE 1.—*A tuned circuit consists of an inductance of* 200 μH *and resistance* 10 *ohms in series with a capacitor of* 0·0002 μF. *What is:* (a) *the* Q *value of the circuit;* *and* (b) *the relative amplitude of the current in decibels* 5 *kc/s above resonance compared with the current at resonance?*

Magnification factor:

$$Q = \omega_r L/R$$

Since

$$\omega_r = 1/\sqrt{LC}$$

$$Q = (1/\sqrt{LC})(L/R)$$
$$= (1/R)\sqrt{L/C}$$
$$= \frac{1}{10}\sqrt{\frac{200 \times 10^{-6}}{2 \times 10^{-10}}}$$
$$= \underline{100}$$

Resonant frequency

$$f_r\ \text{kc/s} = 159/\sqrt{LC} = \frac{159}{200 \times 2 \times 10^{-4}} = 795\ \text{kc/s}$$

$$\text{Factor } F = 2\Delta f/f_r = \frac{2 \times 5}{795} = 0{\cdot}0126$$

Relative current 5 kc/s above resonance:

$$I/I_r = 1/\sqrt{1 + (QF)^2}$$
$$= \frac{1}{\sqrt{[1 + (100 \times 0{\cdot}0126)^2]}} = 0{\cdot}623$$

Relative current in decibels

$$= 20 \log_{10} (I/I\ )$$
$$= -\ 20 \log_{10} (I_r/I)$$
$$= -\ 20 \log_{10} 1{\cdot}61$$
$$= -\ 4{\cdot}1\ \text{db or}\ \underline{4{\cdot}1\ \text{db down}}$$

EXAMPLE 2.—*A tuned circuit is made up of an inductance of 200 μH and resistance 10 ohms in series with a capacitor of 250 pF. Calculate the impedance at the resonant frequency and at 20 kc/s above and below the resonant frequency.*

IMPEDANCE AT RESONANCE

$$Z_r = R$$
$$= \underline{10 \text{ ohms}}$$

At 20 kc/s above and below resonant frequency

$$\omega = 2\pi(f_r \pm 2 \times 10^4)$$
$$= 2\pi\{(1/2\pi\sqrt{LC}) \pm 2 \times 10^4\}$$
$$= (1/\sqrt{LC}) \pm 4\pi \times 10^4$$
$$= \frac{1}{\sqrt{200 \times 10^{-6} \times 250 \times 10^{-12}}} \pm 4\pi \times 10^4$$
$$= (4{\cdot}47 \pm 0{\cdot}126) \times 10^6$$

$$X_L = \omega L$$
$$= 4{\cdot}596 \times 200 = 919 \text{ ohms for the upper frequency}$$
$$4{\cdot}344 \times 200 = 870 \text{ ohms for the lower frequency}$$

$$X_c = 1/\omega C$$
$$= 10^6/(4{\cdot}596 \times 250) = 870 \text{ ohms for the upper frequency}$$
$$10^6/(4{\cdot}344 \times 250) = 920 \text{ ohms for the lower frequency}$$

IMPEDANCE OFF RESONANCE

$$Z = R + j(X_L - X_c)$$

At 20 kc/s above : $Z = 10 + j(919 - 870) = \underline{10 + j50 \text{ ohms}}$

At 20 kc/s below : $Z = 10 + j(870 - 920) = \underline{10 - j50 \text{ ohms}}$

Parallel Resonant Circuit

RESONANT FREQUENCY

$$= (1/2\pi)\sqrt{(1/LC) - (R^2/L^2)} \text{ when } R_c = O \text{ (Fig. 31)}$$
$$= 1/2\pi\sqrt{LC} \text{ when } R^2 \text{ is small in comparison with } L/C.$$

MAGNIFICATION FACTOR

$$Q = R_d/X_L = R_d/X_c = X_L/R = X_c/R$$

at resonance where $R = R_L + R_c$

Impedance is a maximum and current a minimum at resonance.

Then

$$Z_d = R_d$$
$$= (\omega_r L)^2 R = L/CR$$
$$\doteq \omega_r LQ \quad = Q/\omega_r C$$
$$I = E/R_d$$

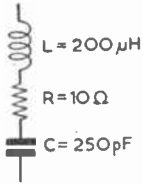


FIG. 30.—EXAMPLE 2.

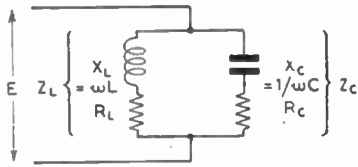


FIG. 31.—PARALLEL RESONANT CIRCUIT.

IMPEDANCE OFF RESONANCE

$$Z_d = Z_L Z_c/(Z_L + Z_c)$$
$$\simeq jX_L/(1 - X_L/X_c) \text{ if resistance is negligible}$$

CURRENT OFF RESONANCE

$$I = E/Z_d$$

CIRCUIT CURRENTS AT RESONANCE

$$I_L = I_c = IQ$$

where Z_d = dynamic impedance, ohms;
R_d = dynamic resistance, ohms;
Z_L = impedance of inductive branch, ohms;
Z_c = impedance of capacitive branch, ohms.

EXAMPLE 3.—*A parallel resonant circuit is to be used to tune the anode circuit of a radio-frequency amplifier. The coil has an inductance of 100 μH and a resistance of 5 ohms. What is:* (a) *the value of capacitance required to tune to a frequency of 1 Mc/s, and* (b) *the effective impedance of the combination?*

Capacitance required to resonate to 1 Mc/s:

$$C = (1/\omega^2 L) \times 10^6$$
$$= \frac{10^6}{(2\pi \times 10^6)^2 \times 100 \times 10^{-6}}$$
$$= 0{\cdot}000254\ \mu F \text{ or } \underline{254\ pF}$$

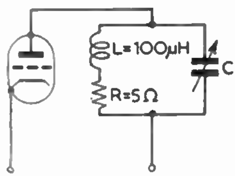


FIG. 32.—EXAMPLE 3.

Dynamic impedance of circuit:

$$Z_d = L/CR \ (L \text{ and } C \text{ can be in } \mu H \text{ and } \mu F)$$
$$= \frac{100}{254 \times 5 \times 10^{-6}}$$
$$= \underline{78{,}740 \text{ ohms}}$$

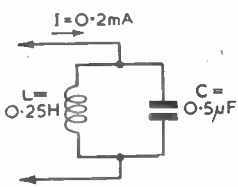


FIG. 33.—EXAMPLE 4.

EXAMPLE 4.—*A parallel resonant circuit has an inductance* L = 0·25 *henrys and a capacitance* C = 0·5 μF. *The resistance of the circuit is negligible by comparison. Determine the dynamic impedance of the circuit and the voltage across it at* 400, 450 *and* 500 *c/s when a current of* 0·2 *mA is flowing.*

At 400 c/s: $\omega = 2\pi \times 400 = 2{,}513$; $\omega^2 = 6{\cdot}31 \times 10^6$
At 450 c/s: $\omega = 2\pi \times 450 = 2{,}826$; $\omega^2 = 7{\cdot}99 \times 10^6$
At 500 c/s: $\omega = 2\pi \times 500 = 3{,}141$; $\omega^2 = 9{\cdot}86 \times 10^6$

LC product:

$$LC = 0{\cdot}25 \times 0{\cdot}5 \times 10^{-6}$$
$$= 0{\cdot}125 \times 10^{-6}$$

Impedance of circuit:

$$Z = j\omega L/(1 - \omega^2 LC)$$

At 400 c/s: $$Z = \frac{j(2{,}513 \times 0{\cdot}25)}{1 - (6{\cdot}31 \times 10^6 \times 0{\cdot}125 \times 10^{-6})} = \underline{j2{,}977 \text{ ohms}}$$

At 450 c/s: $$Z = \frac{j(2{,}826 \times 0{\cdot}25)}{1 - (7{\cdot}99 \times 0{\cdot}125)} = \underline{j353{,}250 \text{ ohms}}$$

At 500 c/s: $$Z = \frac{j(3{,}141 \times 0{\cdot}25)}{1 - (9{\cdot}86 \times 0{\cdot}125)} = \underline{-j3{,}380 \text{ ohms}}$$

Voltage across circuit: $E = IZ$

At 400 c/s: $E = 0{\cdot}0002 \times 2{,}977 = \underline{0{\cdot}6 \text{ volt}}$

At 450 c/s: $E = 0{\cdot}0002 \times 353{,}250 = \underline{70{\cdot}6 \text{ volts}}$

At 500 c/s: $E = 0{\cdot}0002 \times -3{,}380 = \underline{-0{\cdot}67 \text{ volt}}$

COUPLED CIRCUITS

MUTUAL INDUCTANCE $\quad M = k\sqrt{L_1 L_2}$

COUPLING COEFFICIENT $\quad k = M/\sqrt{L_1 L_2}$

EFFECTIVE INDUCTANCE OF TWO CIRCUITS COUPLED BY MUTUAL INDUCTANCE

The combined inductance is the total linkage per unit current flowing through the combination.

Coils in series (Fig. 34 (*a*)):

Aiding $\quad L = L_1 + L_2 + 2M$

Opposing $\quad L = L_1 + L_2 - 2M$

Coils in parallel (Fig. 34 (*b*)):

Aiding $\quad L = (L_1 L_2 - M^2)/(L_1 + L_2 - 2M)$

Opposing $\quad L = (L_1 L_2 - M^2)/(L_1 + L_2 + 2M)$

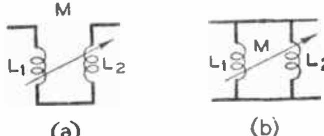


FIG. 34.—INDUCTANCE OF TWO COUPLED CIRCUITS.
(*a*) Coils in series. (*b*) Coils in parallel.

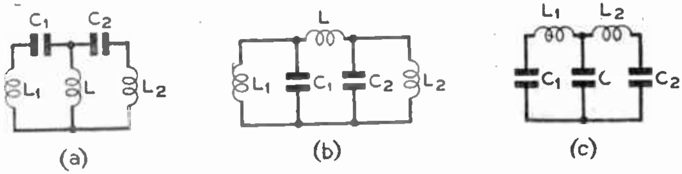
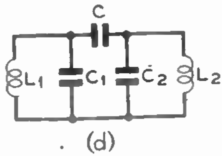


FIG. 35.—INDUCTANCE OF COUPLED CIRCUITS.

(*a*) Shunt inductive.
(*b*) Series inductive.
(*c*) Shunt capacitive.
(*d*) Series capacitive.



Shunt inductive coupling (Fig. 35 (*a*)):

$$k = L/\sqrt{(L + L_1)(L + L_2)}$$

Series inductive coupling (Fig. 35 (*b*)):

$$k = \sqrt{L_1L_2/(L + L_1)(L + L_2)}$$

Shunt capacitive coupling (Fig. 35 (*c*)):

$$k = \sqrt{C_1C_2/(C + C_1)(C + C_2)}$$

Series capacitive coupling (Fig. 35 (*d*)):

$$k = C/\sqrt{(C + C_1)(C + C_2)}$$

where M = mutual inductance, henrys;
k = coupling coefficient ($k = 1$ for 100 per cent coupling);
L_1 = inductance of primary circuit, henrys;
L_2 = inductance of secondary circuit, henrys;
L = inductance of coupling element, henrys;
C_1 = capacitance of primary circuit, farads;
C_2 = capacitance of secondary circuit, farads;
C = capacitance of coupling element, farads.

EXAMPLE 1.—*Two inductance coils of 250 and 175 μH are coupled together. What are the maximum and minimum values of inductance obtainable when the coils are connected in series aiding and series opposing, if the coupling coefficient is:* (a) 0·8; (b) 0·3?

Working in μH, we have

$$k = 0{\cdot}8$$

$$M_1 = k_1\sqrt{L_1L_2} = 0{\cdot}8\sqrt{250 \times 175} = 168\ \mu H$$

$$k = 0{\cdot}3$$

$$M_2 = (k_2/k_1)M_1 = \frac{0{\cdot}3}{0{\cdot}8} \times 168 = 63\ \mu H$$

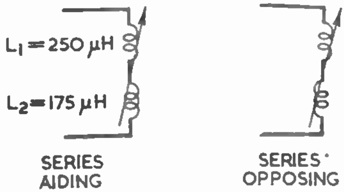


FIG. 36.—EXAMPLE 1.

Maximum and minimum values of inductance :

(a) $L = L_1 + L_2 \pm 2M = 250 + 175 \pm (2 \times 168) = \underline{761 \text{ and } 89\ \mu H}$

(b) $L = L_1 + L_2 \pm 2M = 250 + 175 \pm (2 \times 63) = \underline{551 \text{ and } 299\ \mu H}$

Tuned Circuits Coupled by Mutual Inductance

Maximum energy is transferred between coupled tuned circuits at a critical value of the coupling, when the resistance of the secondary circuit thrown back into the primary is equal to the resistance of the primary, i.e., when

$$(\omega M)^2/R_2 = R_1$$

Then
$$k = 1/\sqrt{Q_1Q_2}$$

When both circuits are tuned to the same frequency f_r and the coupling exceeds the critical value, two frequencies exist at which response is maximum.

$$f_1 = f_r/\sqrt{1 + k}$$
$$f_2 = f_r/\sqrt{1 - k}$$

EXAMPLE 2.—*Two coupled circuits tuned to a frequency of* 500 *kc/s each contain a coil of inductance* 275 *μH and resistance* 10 *ohms. Determine the percentage coupling necessary for maximum transfer of energy.*

Q value of coils

$$Q_1 = Q_2 = \omega L/R$$
$$= \frac{2\pi \times 500 \times 10^3 \times 275 \times 10^{-6}}{10}$$
$$= 86{\cdot}4$$

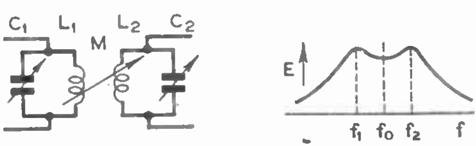


FIG. 37.—COUPLED TUNED CIRCUITS.

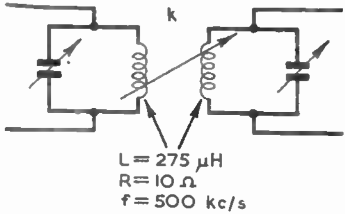


FIG. 38.—EXAMPLE 2.

Critical coupling for maximum energy transfer

$$k = 1/\sqrt{Q_1 Q_2}$$

$$= \sqrt{\frac{1}{86 \cdot 4^2}} = 0{\cdot}0116 \text{ or } \underline{1{\cdot}2 \text{ per cent}}$$

EXAMPLE 3.—*Two tuned circuits, loosely coupled, are adjusted to a frequency of 581 kc/s. The coupling coefficient is then increased to 15 per cent. What are the two frequencies to which the combination will have maximum response if K is greater than its critical value?*

Lower frequency:

$$f_1 = f_0/\sqrt{1+k} = \frac{581}{\sqrt{1+0{\cdot}15}} = \underline{542 \text{ kc/s}}$$

Upper frequency:

$$= f_0/\sqrt{1-k} = \frac{581}{\sqrt{1-0{\cdot}15}} = \underline{631 \text{ kc/s}}$$

ATTENUATORS AND FILTERS

Attenuators

Four terminal networks of resistances designed to produce a specified loss when inserted between two impedances Z_1 and Z_2 to which the input and output impedances of the attenuator are matched.

Attenuators are either unbalanced (elements arranged unsymmetrically) or balanced (elements arranged symmetrically). The basic unbalanced types are the T and π networks (Figs. 39 (*a*) and 40 (*a*)). Corresponding balanced types are the H and □ networks (Figs. 39 (*b*)

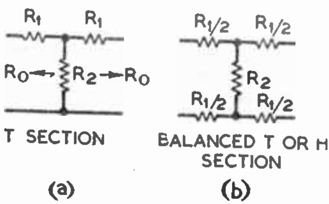


FIG. 39.—T AND BALANCED T ATTENUATORS.

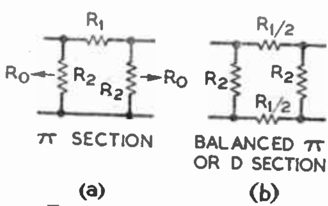


FIG. 40.—π AND BALANCED π ATTENUATORS.

and 40 (*b*)), the elements of which have the equivalent values shown in the diagrams.

T-type (Fig. 39);

$$(a)\ R_1 = R_0(n-1)/(n+1)$$
$$(b)\ R_2 = 2R_0n/(n^2-1)$$

π-type (Fig. 40):

$$(a)\ R_1 = R_0(n^2-1)/2n$$
$$(b)\ R_2 = R_0(n+1)/(n-1)$$

where R_1R_2 have the significance shown in the structural diagrams;
R_0 = terminal resistance of circuit, ohms;
n = ratio input voltage/output voltage, e_1/e_2
(attenuation in decibels $\alpha = 20 \log_{10} n$).

The series elements of the balanced T or balanced π section each have half the value of those of the unbalanced section.

Recurrent Networks

The total attenuation is the sum of the attenuations of the individual T or π sections; the elements having the values shown in the structural diagram.

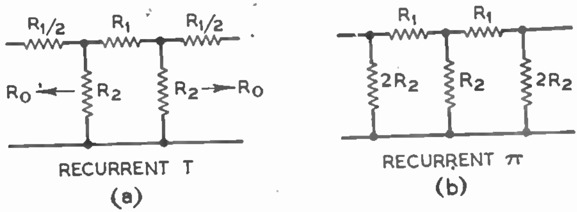


FIG. 41.—RECURRENT ATTENUATOR NETWORKS. (2 SECTIONS).

EXAMPLE 1.—*A recurrent T type attenuator is required to produce a loss of 30 db in three 10-db steps and to have a characteristic impedance of 100 ohms. Determine the values of the series and shunt elements.*

Each section must produce an attenuation of 10 db. Hence

$$n = \text{antilog } \alpha/20$$
$$= \text{antilog } 10/20 = 3{\cdot}162$$

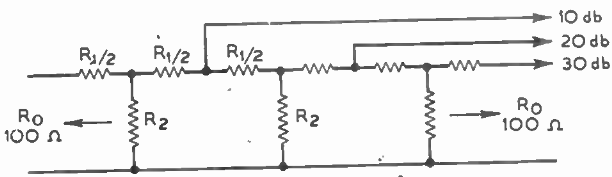


FIG. 42.—EXAMPLE 1.

ELEMENTS

$$R_1/2 = R_0(n-1)/(n+1) = \frac{100(3\cdot162 - 1)}{3\cdot162 + 1} = \underline{52 \text{ ohms}}$$

$$R_2 = 2R_0n/(n^2-1) = \frac{2 \times 100 \times 3\cdot162}{3\cdot162^2 - 1} = \underline{70\cdot3 \text{ ohms}}$$

Filters

Electrical wave filters are four-terminal networks composed of inductive and capacitive elements, which pass a required continuous band of frequencies and attenuate all others. Filters are classified broadly according to their pass-band characteristics as shown in Table 6.

TABLE 6.—CLASSIFICATION OF FILTERS

Type	*Pass Band*
Low pass	All frequencies below a critical frequency f_c
High pass	All frequencies above a critical frequency f_c
Band pass	All frequencies between two critical frequencies f_1 and f_2
Band elimination	All frequencies outside two critical frequencies f_1 and f_2

TYPE

Low pass constant-k:
- T-section (Fig. 43)
- π-section (Fig. 44)

Low pass, derived:
- Series-derived T (Fig. 45)
- Shunt-derived π (Fig. 46)

$$\left\{\begin{aligned} f_c &= 1/\pi\sqrt{LC} \\ L &= R_0/\pi f_c \\ C &= L/R_0^2 \\ L_1 &= aL \\ L_2 &= bL \\ C_1 &= L_2/R_0^2 \\ C_2 &= L_1/R_0^2 \end{aligned}\right.$$

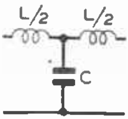


FIG. 43.—T SECTION.

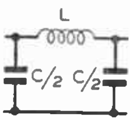


FIG. 44.—π SECTION.

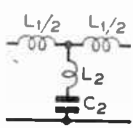


FIG. 45.—SERIES-DERIVED T SECTION.

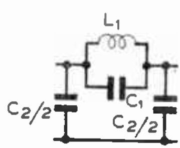


FIG. 46.—SHUNT-DERIVED π SECTION.

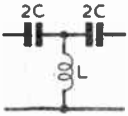


FIG. 47.—HIGH-PASS T SECTION.

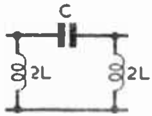


FIG. 48.—HIGH-PASS π SECTION.

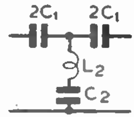


FIG. 49.—HIGH-PASS SERIES-DERIVED T SECTION.

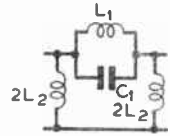


FIG. 50.—HIGH-PASS SHUNT-DERIVED π SECTION.

High pass, constant-k:
T-section (Fig. 47)
π-section (Fig. 48)

High pass, derived:
Series-derived T (Fig. 49)
Shunt-derived π (Fig. 50)

$$\begin{cases} f_c = 1/4\pi\sqrt{LC} \\ L = R_0/4\pi f_c \\ C = L/R_0^2 \\ L_1 = L/b \\ L_2 = L/a \\ C_1 = L_2/R_0^2 \\ C_2 = L_1/R_0^2 \end{cases}$$

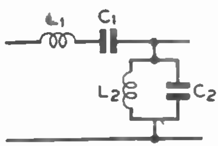


FIG. 51.—BAND-PASS CONSTANT k.

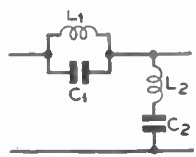


FIG. 52.—BAND ELIMINATION CONSTANT k.

Band pass, constant-k (Fig. 51)

$$\begin{cases} L_1 = R_0/\pi(f_1 - f_2) \\ L_2 = R_0(f_1 - f_2)/4\pi f_1 f_2 \\ C_1 = L_2/R_0^2 \\ C_2 = L_1/R_0^2 \end{cases}$$

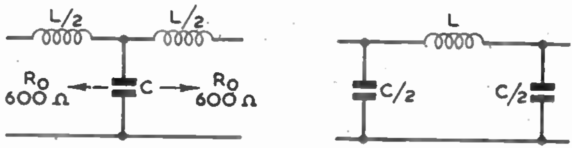
Band elimination, constant-k (Fig. 52)

$$\begin{cases} L_1 = R_0(f_1 - f_2)/\pi f_1 f_2 \\ L_2 = R_0/4\pi(f_1 - f_2) \\ C_1 = L_2/R_0^2 \\ C_2 = L_1/R_0^2 \end{cases}$$

where $\left.\begin{matrix} L\ L_1 L_2 \\ C\ C_1 C_2 \end{matrix}\right\}$ have the significance shown in the structural diagrams;

$a = \sqrt{1 - (f_c/f_m)^2}$ (low pass); $\sqrt{1 - (f_m/f_c)^2}$ (high-pass);
$b = (1 - a^2)/4a$;
f_c = cut-off frequency, c/s;
$f_1 f_2$ = upper and lower cut-off frequencies, c/s;
f_m = frequency of maximum attenuation, c/s;
R_0 = terminal resistance of circuit, ohms.

EXAMPLE 2.—*Design a low pass constant-*k *filter for a 600-ohm line, to pass speech frequencies up to 2,500 c/s, employing:* (a) *a T section,* (b) *a π section.*



(a) $f_c = 2500$ c/s (b)

FIG. 53.—EXAMPLE 2.

T SECTION

Series inductance elements

$$L = R_0/\pi f_c = \frac{600}{2{,}500\pi} = 0{\cdot}0765 \text{ henry or } \underline{76{\cdot}5 \text{ mH}}$$

$$\text{Series elements each} = L/2 = \underline{38{\cdot}25 \text{ mH}}$$

Shunt capacitance :

$$C = L/R_0{}^2 = \frac{0{\cdot}0765 \times 10^6}{600^2} = \underline{0{\cdot}212\ \mu\text{F}}$$

π SECTION

Series inductance element : $L = \underline{76{\cdot}5 \text{ mH}}$

Shunt capacitance elements $= C/2 = \underline{0{\cdot}106\ \mu\text{F}}$

EXAMPLE 3.—*A constant*-k *band pass filter is required for a multi-channel telegraph circuit having a characteristic impedance of* 600 *ohms, to pass a band of frequencies* 120 c/s *wide at a mid-frequency of* 660 c/s. *Calculate the values of the series and shunt elements.*

$$\text{Upper cut-off frequency} = 660 + \frac{120}{2} = 720 \text{ c/s}$$

$$\text{Lower cut-off frequency} = 660 - \frac{120}{2} = 600 \text{ c/s}$$

SERIES INDUCTANCE ELEMENT

$$L_1 = R_0/\pi(f_1 - f_2) = \frac{600}{120\pi} = \frac{5}{\pi} = \underline{1{\cdot}59 \text{ henrys}}$$

SHUNT INDUCTANCE ELEMENT

$$L_2 = R_0(f_1 - f_2)/4\pi f_1 f_2$$

$$= \frac{600 \times 120}{4\pi \times 720 \times 600}$$

$$= \frac{1}{24\pi} = 0{\cdot}01325 \text{ henry or } \underline{13{\cdot}25 \text{ mH}}$$

C

SERIES CAPACITANCE ELEMENT

$$C_1 = L_2/R_0^2 = \frac{0{\cdot}01325 \times 10^6}{600^2} = \underline{0{\cdot}0368\ \mu F}$$

SHUNT CAPACITANCE ELEMENT

$$C_2 = L_1/R_0^2 = \frac{1{\cdot}59 \times 10^6}{600^2} = \underline{4{\cdot}42\ \mu F}$$

EXAMPLE 4.—*Determine the elements of a constant-*k *band elimination filter to suppress heterodyne whistles between* 8,500 *and* 9,000 *c/s and to have a terminal impedance of* 2,000 *ohms.*

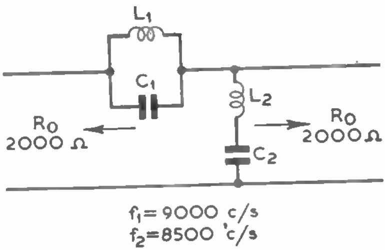


FIG. 54.—EXAMPLE 4.

SERIES INDUCTANCE

$$L_1 = R_0(f_1 - f_2)/\pi f_1 f_2 = \frac{2{,}000\ (9{,}000 - 8{,}500)}{\pi \times 9{,}000 \times 8{,}500}$$

$$= 0{\cdot}00416 \text{ henry or } \underline{4{\cdot}16\ \text{mH}}$$

SHUNT INDUCTANCE

$$L_2 = R_0/4\pi(f_1 - f_2) = \frac{2{,}000}{4\pi \times 500} = \frac{1}{\pi} = 0{\cdot}318 \text{ henry or } \underline{318\ \text{mH}}$$

SERIES CAPACITANCE

$$C_1 = L_2/R_0^2 = \frac{0{\cdot}318 \times 10^6}{2{,}000^2} = \underline{0{\cdot}0795\ \mu F}$$

SHUNT CAPACITANCE

$$C_2 = L_1/R_0^2 = \frac{0{\cdot}00417 \times 10^6}{2{,}000^2} = \underline{0{\cdot}00104\ \mu F}$$

THERMIONIC AMPLIFIERS

Valve Constants

The three important constants which determine the electrical performance of an amplifying valve are:

Amplification factor $\mu = -\frac{de_a}{de_g}$ (i_a constant)

Mutual conductance (transconductance) $g_m = \frac{di_a}{de_g}$ (e_a constant)

Dynamic resistance (A.C. resistance) $R_a = \frac{de_a}{di_a}$ (e_g constant)

$= \mu / g_m$

where $\frac{de_a}{de_g}$ = rate of change of anode voltage (e_a) with respect to grid voltage (e_g);

$\frac{di_a}{de_g}$ = rate of change of anode current (i_a) with respect to grid voltage (e_g);

$\frac{de_a}{di_a}$ = rate of change of anode voltage (e_a) with respect to anode current (i_a).

GRID BIAS (see Fig. 55)

For Class A amplifiers: $E_g = -E_a/2\mu$
For Class B amplifiers: $E_g = -E_a/\mu$
For Class C amplifiers: $E_g = -C_1E_a/\mu + C_2(E_a/\mu - e_{gp})$
Self-bias voltage: $E_g = -I_aR_g$
A.C. component of voltage across bias resistor: $e_g = I_aZ_g$

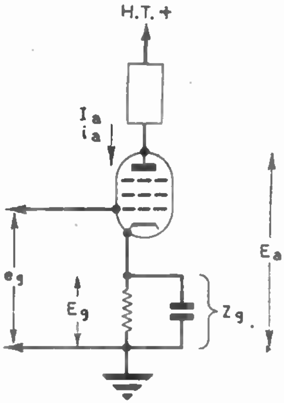


FIG. 55.—GRID BIAS VALVE CONSTANTS.

where E_g = grid bias voltage;
E_a = anode D.C. voltage;
I_a = D.C. component of anode current, amperes;
i_a = A.C. component of anode current, amperes;
e_g = signal voltage applied to grid;
e_{gp} = peak value of signal voltage.

Anode Current

A.C. component of anode current:

$$i_a = \mu e_g/(R_a + Z_L)$$

where i_a = A.C. component of anode current;
R_a = anode A.C. resistance of valve, ohms;
Z_L = anode load impedance, ohms.

Classification of Amplifiers

Amplifiers are classified according to the grid-bias setting and the amplitude of the signal voltage applied to the control grid. In the accompanying table the valve characteristic referred to is the anode-current/grid-bias characteristic.

Suffix 1 denotes that no grid current flows during any part of the cycle.

Suffix 2 denotes that grid current flows for at least part of the cycle. In Class C amplifiers grid current is always present, and a suffix is unnecessary.

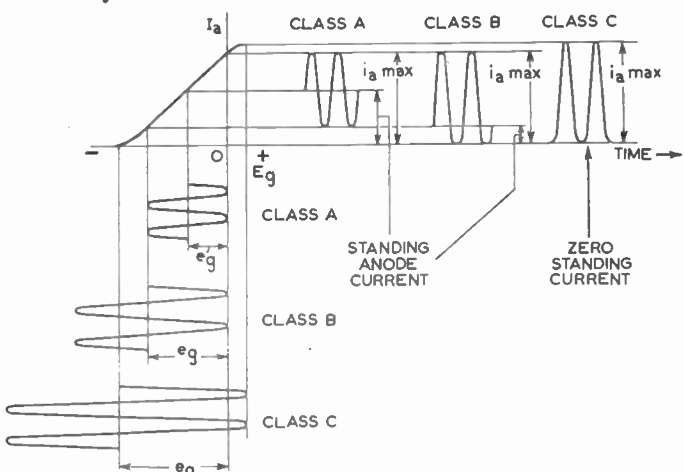


Fig. 56.—Class A, B, and C Amplification.

Class	*Grid-bias Setting*	*Signal Voltage Swing*	*Anode Current Flow*	*Performance*
A_1	Centre point of characteristic	Confined to linear portion of characteristic	Whole of cycle	Output undistorted. Gain high. Power conversion efficiency low (25% max.)
A_2	Above centre point of characteristic	Extends into upper bend of characteristic	Whole of cycle	Output almost undistorted. Gain less but efficiency higher than Class A_1
AB_1	Below centre point of characteristic	Extends into lower bend of characteristic	Cuts off for small portion of negative half-cycle	Output in push–pull practically undistorted. Gain less but efficiency higher than Class A_2
AB_2	Centre point of characteristic	Extends into lower and upper bends of characteristic	Cuts off for small portion of negative half-cycle	Slight harmonic distortion in push–pull. Gain less but efficiency higher than Class AB_1
B_1	Near lower bend of characteristic	Extends beyond lower bend of characteristic	Cuts off for greater part of negative half-cycle	Small harmonic distortion in push–pull. Gain less than Class AB_2. Efficiency 78·5% max.
B_2	Near lower bend of characteristic	Extends into lower and upper bends of characteristic	Cuts off for greater part of negative and small portion of positive half-cycles	Some harmonic distortion in push-pull. Gain less but efficiency higher than Class B_1
C	Beyond lower bend of characteristic	Extends well beyond lower and upper bends of characteristic	Cuts off for whole of negative and part of positive half-cycles	Considerable harmonic distortion. Gain low, but power conversion efficiency a maximum (up to 80%)

VOLTAGE AMPLIFICATION (see Fig. 57)

General case :

$$m = e_2/e_1 = \mu Z_L/(R_a + Z_L)$$

For maximum voltage gain : $Z_L \gg R_a$

For maximum power gain : $Z_L = R_a$

where m = voltage gain;
e_1 = input r.m.s. voltage;
e_2 = output r.m.s. voltage.

EXAMPLE 1.—*Two valves have respectively the following characteristics:* (a) $\mu = 30$; $R_a = 75$ $k\Omega$; (b) $\mu = 20$; $R_a = 20$ $k\Omega$. *Which will give the greater voltage gain in an amplifier with a load impedance of* 25 $k\Omega$, *and what will be the relative gain expressed in decibels ?*

Voltage gain of (*a*)

$$m_1 = \mu R_L/(R_a + R_L) = \frac{30 \times 2{\cdot}5 \times 10^4}{(2{\cdot}5 \times 10^4) + (7{\cdot}5 \times 10^4)} = 7{\cdot}5$$

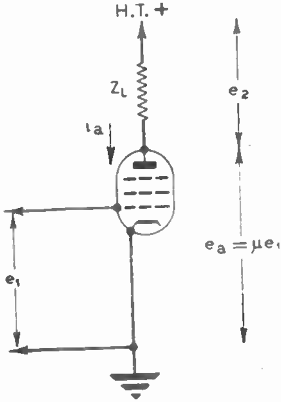


FIG. 57.—VOLTAGE AMPLIFICATION.

Voltage gain of (*b*) .

$$m_2 = \frac{20 \times 2{\cdot}5 \times 10^4}{(2 \times 10^4) + (2{\cdot}5 \times 10^4)} = 11{\cdot}11$$

Hence (*b*) has the greater gain. The gain of (*b*) relative to (*a*)

$$= 20.\log_{10}(m_2/m_1) = 20 \log_{10} \frac{11{\cdot}11}{7{\cdot}5} = \underline{3{\cdot}4 \text{ db}}$$

Tuned Anode Coupling

At resonance : $Z_L = L/CR$

Hence gain at resonance : $m = \mu/(1 + R_a RC/L)$

Gain off resonance $m = \dfrac{\mu L}{\sqrt{(L + R_a RC)^2 + R_a^2(\omega/\omega_0^2 - 1/\omega)^2}}$

where
L = inductance of tuning coil, henrys;
C = capacitance of tuning capacitor, farads;
R = R.F. resistance of tuning circuit, ohms;
$\omega_0 = 2\pi \times$ resonant frequency;
$\omega = 2\pi \times$ frequency (off resonance).

EXAMPLE 2.—*The tuned anode circuit of a radio-frequency amplifier contains an inductance of* 200 μH *and resistance* 1·5 *ohms. The valve has* *an amplification factor of 15 and an anode A.C. resistance of 50 kΩ. What is the capacitance required to tune the circuit to 500 kc/s and the voltage amplification at this frequency?*

(*a*) Capacitance required to tune to 500 kc/s:

$$f = 1/2\pi\sqrt{LC}$$

$$C = 1/(2\pi f)^2 L \text{ farads}$$

$$= \frac{10^6}{(2\pi \times 5 \times 10^5)^2 \times 200 \times 10^{-6}}\,\mu\text{F}$$

$$= \frac{10^{-2}}{2\pi^2} = \underline{0{\cdot}00051\ \mu\text{F}}$$

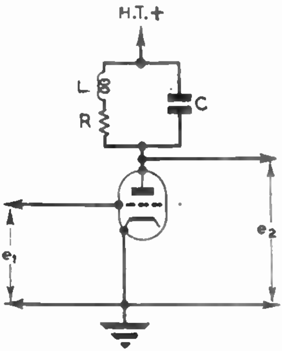


Fig. 58(*a*).—Tuned Anode Coupling.

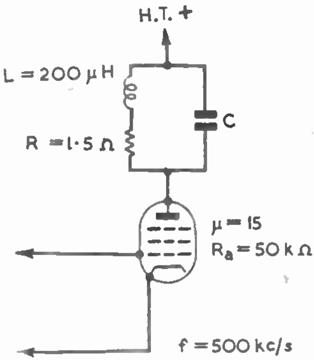


Fig. 58(*b*).—Example 2, Tuned Anode Coupling.

(*b*) Voltage amplification at 500 kc/s:

$$m = \mu/(1 + R_a RC/L)$$

$$= \frac{15}{1 + [5 \times 10^4 \times 1{\cdot}5 \times 5{\cdot}1 \times 10^{-10}/(2 \times 10^{-4})]} = \underline{12{\cdot}6}$$

Audio-frequency Amplifiers *

Resistance Coupling (Fig. 59)

Gain:

$$m = \mu R_L/(R_a + R_L)$$

* Formulæ for Figs. 59 and 61, assume that value of grid leak $\gg R_L$ and R in parallel.

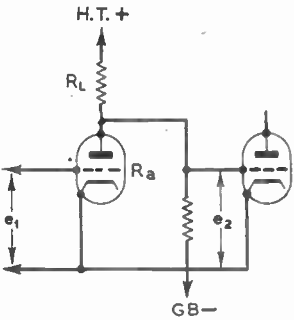


Fig. 59.—Resistance Coupling.

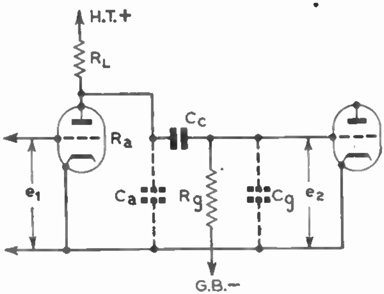


Fig. 60.—Resistance-Capacitance Coupling.

Resistance-Capacitance Coupling (Fig. 60)

Amplification is a maximum at medium frequencies of the audio frequency range, and the gain for triodes is then

$$m = \mu R_e/R_a$$

where m = voltage gain of stage;
R_e = equivalent resistance formed by the anode resistance, load coupling resistance and grid leak in parallel
$= 1/(1/R_a + 1/R_L + 1/R_g)$;
R_a = value of anode resistance.

For pentodes, where $R_a \gg R_L$ and R_g

$$R_e \simeq 1/(1/R_L + 1/R_g)$$

Low frequencies

The ratio of gain at low frequency to gain at medium frequency

$$m_1/m = 1/(1 - jf_1/f)$$

where f = frequency considered, c/s;
$f_1 = 1/2\pi C_c R_{e1}$ (the frequency at which the reactance of $C_c = R_{e1}$);
C_c = capacitance of coupling condenser, farads;
$R_{e1} = R_g + R_a R_L/(R_a + R_L)$ (the equivalent resistance formed by R_g in series with $R_a R_L$ in parallel).

High frequencies

The ratio of gain at high frequency to gain at medium frequency

$$m_2/m = 1/(1 + jf/f_2)$$

where f = frequency considered, c/s;
$f_2 = 1/2\pi C_c R_{e2}$ (the frequency at which the reactance of $C_c = R_{e2}$);
$R_{e2} = 1/(1/R_a + 1/R_L + 1/R_g)$ (the equivalent resistance formed by R_a, R_L and R_g in parallel).

Gain at higher frequencies, e.g., video-frequency amplification, is greatly improved by inserting an inductance L in series with the load-coupling resistance. Then the gain

$$m \simeq (\mu/R_a)\sqrt{R_L^2 + (\omega L)^2}/\sqrt{(1 - \omega^2 LC)^2 + (\omega CR_L)^2}$$

where L = inductance, henrys;

C = total anode and grid capacitance, $C_a + C_g$.

EXAMPLE 3.—*Determine the gain of a resistance–capacitance coupled a.f. pentode amplifier stage at frequencies of* 50, 1000 *and* 10,000 *c/s, given the following data:*

Voltage amplification factor	= 200
Valve anode resistance	= 100 kΩ
Anode load resistance	= 250 kΩ
Capacitance of coupling capacitor	= 0·01 μF
Grid-leak resistance	= 0·5 MΩ

Equivalent resistance at 50 c/s:

$$R_{e1} = R_g + R_aR_L/(R_a + R_L)$$

$$= 500 + \frac{100 \times 250}{100 + 250} = 571{\cdot}4 \text{ k}\Omega$$

Equivalent resistance at 1,000 c/s (mid-frequency) and 10,000 c/s:

$$R_{e2} = 1/(1/R_a + 1/R_L + 1/R_g)$$

$$= \frac{1}{1/100 + 1/250 + 1/500} = 62{\cdot}5 \text{ k}\Omega$$

Voltage gain at 1,000 c/s:

$$m = \mu R_{e2}/R_a$$

$$= \frac{200 \times 62{\cdot}5}{100} = 125$$

$$m \text{ (db)} = 20 \log_{10} 125 = \underline{42 \text{ db}}$$

Voltage gain at 50 c/s:

Reference frequency—

$$f_1 = 1/2\pi C_c R_{e1}$$

$$= \frac{1}{2\pi \times 0{\cdot}01 \times 10^{-6} \times 571{\cdot}4 \times 10^3} = 27{\cdot}9$$

$$m_1/m = 1/(1 - jf_1/f)$$

$$= \frac{1}{1 - j27{\cdot}9/50} = \frac{1}{1 - j0{\cdot}558}$$

$$= \frac{1 + j0{\cdot}558}{1 + 0{\cdot}312} = 0{\cdot}76 + j0{\cdot}425$$

$$= \sqrt{0{\cdot}76^2 + 0{\cdot}425^2} = 0{\cdot}87$$

$$m_1 = 0{\cdot}87m = 0{\cdot}87 \times 125 = 109$$

$$m_1\text{(db)} = 20 \log_{10} 109 = \underline{41 \text{ db}}$$

Voltage gain at 10,000 c/s:

From the expressions for f_1 and f_2—

$$f_2/f_1 = R_{e1}/R_{e2}$$

$$f_2 = \frac{27 \cdot 9 \times 571 \cdot 4}{62 \cdot 5} = 255$$

$$m_2/m = 1/(1 + jf/f_2)$$

$$= \frac{1}{1 + j10{,}000/255} = \frac{1}{1 + j39 \cdot 2}$$

$$= \frac{1 - j39 \cdot 2}{1 + 1535} = 0 \cdot 00065 - j0 \cdot 255$$

$$= \sqrt{0 \cdot 00065^2 + 0 \cdot 255^2} = 0 \cdot 26$$

$$m_2 = 0 \cdot 26 \times 125 = 32 \cdot 5$$

$$m_2(\text{db}) = 20 \log_{10} 32 \cdot 5 = \underline{30 \text{ db}}$$

CHOKE COUPLING (Fig. 61)

Gain :

$$m = \mu\sqrt{R_L^2 + X_L^2}/\sqrt{(R_L + R_a)^2 + X_L^2}$$

$$= \mu X_L/\sqrt{R_a^2 + X_L^2} \text{ if resistance of choke is very small}$$

where R_L = anode choke resistance, ohms;
X_c = capacitive reactance of coupling capacitor, ohms;
X_L = inductive reactance of coupling choke, ohms.

TRANSFORMER COUPLING (Fig. 62)

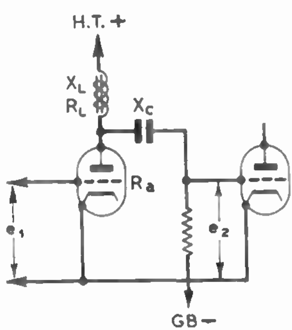


FIG. 61.—CHOKE COUPLING.

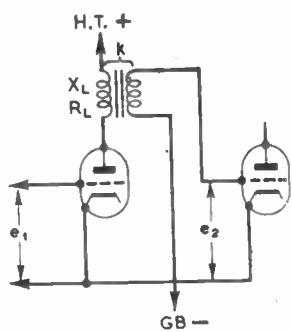


FIG. 62.—TRANSFORMER COUPLING.

Gain at medium audio frequencies:

$$m = \mu k R_L/(R_a + R_L)$$

Gain at low audio frequencies:

$$m = \mu k/\sqrt{\{(R_a + R_L)/R_L\}^2 + (R_a/X_{L_1})\}^2}$$

Gain at high audio frequencies:

$$m = \mu k R_L/\sqrt{(R_a + R_L)^2 + X_{L_0}^2}$$

Turns ratio for maximum power gain:

$$k = \sqrt{R_a/R_L}$$

where k = ratio of primary to secondary turns;
X_{L_0} = leakage reactance of transformer ($= \omega L_0$), ohms;
X_{L_1} = reactance of primary winding ($= \omega L_1$), ohms;
R_L = effective resistance of transformer, referred to primary, ohms.

Power Amplification

Maximum power output is obtained when
$R_L = R_a = R_a/k^2$ when the load is transformer coupled

$$k = \sqrt{R_a/R_L}$$

Turns ratio of speech output transformer:

$$k = \sqrt{R_0/Z_s}$$

Optimum load impedance for an output stage:

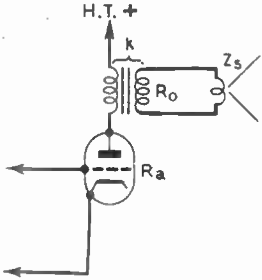


Fig. 63.—Output Transformer.

For a triode $= 2R_a$ to $3R_a$
For a pentode $= \frac{1}{4}R_a$ to $R_a/10$
For two valves in push–pull $= 2 \times R_0$ of a single valve (Class A)
For two valves in parallel $= \frac{1}{2} \times R_0$ of a single valve

where R_L = anode load resistance, ohms;
R_a = A.C. resistance of valve, ohms;
R_0 = optimum load resistance, ohms;
Z_s = nominal impedance of speech coil at 400 c/s, ohms.

Example 3.—*A 5 ohm loudspeaker is coupled by a transformer to the output stage of an amplifier. What turns ratio is required to obtain maximum undistorted power output with:* (a) *a triode having an A.C. resistance of 5,000 ohms,* (b) *a pentode of 50,000 ohms?*

(*a*) Optimum load resistance:

$$R_L = 2R_a = 2 \times 5{,}000 = 10{,}000 \text{ ohms}$$

Turns ratio $= \sqrt{R_L/R_a} = \sqrt{\dfrac{5}{10{,}000}} = \dfrac{1}{44{\cdot}7}$

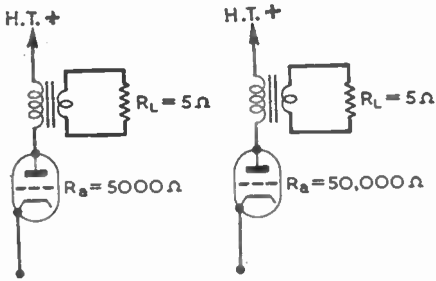


FIG. 64.—EXAMPLE 3.

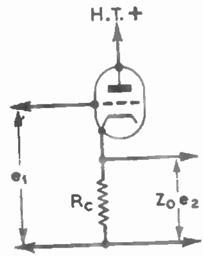


FIG. 65.—CATHODE FOLLOWER.

(b) Optimum load resistance :

$$R_L = R_a/4 = \frac{50,000}{4} = 12,500 \text{ ohms}$$

Turns ratio $$= \sqrt{\frac{5}{12,500}} = \frac{1}{50}$$

Cathode Follower

The amplification factor m is always <1 :

$$m = \mu/\{\mu + 1 + (R_a/R_c)\}$$

Output impedance :

$$Z_0 = m/g_m$$

A.C. resistance to H.T. fluctuations :

$$R_d = R_a + R_c(\mu + 1)$$

where Z_0 = output impedance, ohms;
R_c = cathode resistance, ohms;
R_d = dynamic resistance, ohms;
g_m = mutual conductance, amperes/volt.

EXAMPLE 4.—*From the following data for a cathode follower, determine* (a) *the voltage amplification,* (b) *the A.C. resistance to H.T. variations, and* (c) *the output impedance :* $\mu = 50$, $g_m = 5$ *mA/volt,* $R_c = 15$k. *What would be* (d) *the voltage amplification if the same valve were used in a conventional amplifier having an anode load resistance of* 15 *k*Ω?

(a) As a cathode follower :

$$R_a = \mu/g_m = \frac{50}{0{\cdot}005} = 10^4$$

Voltage amplification :

$$m = \mu/(\mu + 1 + R_a/R_c)$$

$$= \frac{50}{50 + 1 + 10^4/(1{\cdot}5 \times 10^4)} = \underline{0{\cdot}968}$$

(*b*) A.C. resistance :

$$R = R_a + R_c(\mu + 1)$$

$$= 10^4 + \{(1{\cdot}5 \times 10^4)(50 + 1)\} = \underline{7{\cdot}75 \times 10^5 \text{ ohms}}$$

(*c*) Output impedance :

$$Z_0 = m/g_m = \frac{0{\cdot}968}{0{\cdot}005} = \underline{194 \text{ ohms}}$$

(*d*) As an anode-loaded amplifier :

$$m = \mu R_L/(R_a + R_L)$$

$$m' = \frac{50 \times 1{\cdot}5 \times 10^4}{10^4 + (1{\cdot}5 \times 10^4)} = \underline{30}$$

Negative Voltage Feedback

Frequency distortion, amplitude distortion and phase distortion are reduced by the factor $1/(1 + \beta m)$.

Voltage gain without feedback : $m = e_2/e_1$

Voltage gain with feedback : $m' = m/(1 + \beta m)$
$= 1/(1/m + \beta)$

Loss in decibels with feedback $= 20 \log_{10} (1 + \beta m)$

The relative loss with feedback at any two frequencies, for which the gains without feedback are m_1 and m_2

$$= 20 \log_{10} \{m_1(1 + \beta m_2)/m_2(1 + \beta m_1)\}$$

where e_1 = input voltage without feedback ;
e_2 = output voltage without feedback ;
β = fraction of voltage fed back from output to input in reverse phase.

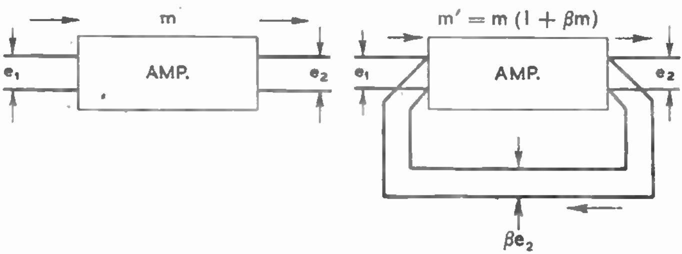


FIG. 66(*a*).—NEGATIVE VOLTAGE FEEDBACK.

EXAMPLE 5.—*An amplifier has a voltage gain of 50 at 1,000 c/s and 5 at 50 c/s. What is the relative loss in output volume at the lower frequency, expressed in decibels:* (a) *without feedback,* (b) *if 5 per cent of the output voltage is fed back to the input in reverse phase.*

(*a*) Relative gain at 50 c/s without feedback:

$$\alpha = 20 \log_{10} (m_2/m_1) = 20 \log_{10} \frac{5}{50}$$

$$= - 20 \log_{10} \frac{50}{5} = - 20 \text{ db or } \underline{20 \text{ db loss}}$$

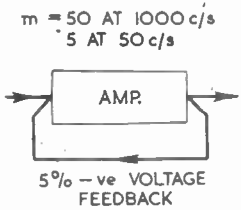


FIG. 66(*b*).—EXAMPLE 5.

(*b*) Gain with feedback:

At 1,000 c/s $\qquad m_1' = m/(1 + \beta m)$

$$= \frac{50}{1 + 50\,(0{\cdot}05)} = 14{\cdot}3$$

At 50 c/s $\qquad m_2' = \dfrac{5}{1 + 5\,(0{\cdot}05)} = 4$

Relative gain at 50 c/s with feedback

$$= 20 \log_{10} (m_2/m_1)$$

$$= - 20 \log_{10} (m_1/m_2)$$

$$= - 20 \log_{10} \frac{14{\cdot}3}{4} = - 11 \text{ db gain or } \underline{11 \text{ db loss}}$$

Grounded-Grid Amplifier

The earthed grid acts as a screen between the anode or output circuit and the cathode or input circuit, so that neutralization is unnecessary in radio-frequency amplifiers. Inherent negative feedback confers a high degree of linearity. The circuit is, therefore, particularly suitable for V.H.F., F.M. transmitters and receivers.

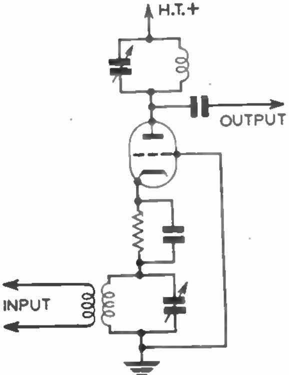


FIG. 67.—GROUNDED-GRID R.F. AMPLIFIER.

Voltage gain:

$$m = (\mu + 1)R_L/(R_a + R_L)$$

Input-circuit resistance:

$$R_i \simeq (1/\mu R_a)(1 + R_L/R_a)$$

Output-circuit resistance—

$$R_o \simeq R_a$$

Optimum load resistance when $\mu \gg 1$—

$$R_L = R_a\sqrt{1 + g_m R_i}$$

where μ = voltage amplification factor of valve;
R_L = load impedance between anode and H.T. +;
R_a = valve-anode resistance;
R_i = valve-input resistance at operating frequency;
g_m = mutual conductance of valve.

AMPLIFIER NOISE

The principal sources of noise generated in an amplifier are: thermal agitation noise (Johnson effect); valve noise (shot effect); induced grid noise in resistors, valves and leads, and A.C. mains hum.

Random noise produced by thermal agitation and shot effect is distributed evenly over the frequency spectrum, and is therefore proportional to the bandwidth. Induced grid noise and A.C. mains hum can be effectively eliminated by careful design and suitable smoothing filters.

Noise voltages from the various sources must be added in quadrature, so that total noise voltage:

$$e_t^2 = e_1^2 + e_2^2 + e_3^2 + \ldots \text{ etc.}$$

$$e_t = \sqrt{e_1^2 + e_2^2 + e_3^2 + \ldots} \text{ etc.}$$

Thermal Agitation Noise

This is expressed as a voltage in series with a given resistor, the equivalent resistance of a tuned circuit at resonance or the radiation resistance of an aerial, as the case may be.

$$e^2_{ta} = 4kT\Delta fR$$

where k = Boltzmann's constant = $1{\cdot}37 \times 10^{-23}$ joules/° K;
T = temperature, ° K;
Δf = frequency bandwidth, c/s;
R = resistance, ohms.

At normal temperatures it is sufficiently accurate to take $4kT = 1{\cdot}6 \times 10^{-20}$. Hence

$$e^2_{ta} = 1{\cdot}6 \times 10^{-20}\Delta fR$$

$$e_{ta} = 1{\cdot}26 \times 10^{-10}\sqrt{\Delta fR}$$

EQUIVALENT THERMAL AGITATION NOISE RESISTANCE

Resistor, R	R
Resistors R_1, R_2 in series	$R_1 + R_2$
Resistors R_1, R_2 in parallel	$R_1R_2/(R_1 + R_2)$
Tuned circuit at resonance	$\omega LQ = Q/\omega c$
Coupled aerial circuit	n^2R_r

where L = tuned circuit inductance, henrys;
C = tuned circuit capacitance, farads;
Q = circuit magnification factor;
n = voltage step-up ratio;
R_r = radiation resistance of aerial.

Valve Noise

Valve-noise voltage is expressed in terms of an equivalent resistance at normal temperature.

$$e_v = 1{\cdot}26 \times 10^{-10}\sqrt{\Delta f \,.\, R_{eq}}$$

EQUIVALENT VALVE NOISE RESISTANCE OF AMPLIFIERS AND MIXERS

Triode amplifier . .	$2{\cdot}5/g_m$
Triode mixer . .	$4/g_c$
Pentode amplifier . .	$\{I_a/(I_a + I_{g2})\}(2{\cdot}5g_m + 20I_{g2}/g_m^2)$
Pentode mixer . .	$\{I_a/(I_a + I_{g2})\}(4/g_c + 20I_{g2}/g_c^2)$
Multi-grid mixer . .	$20I_a\,(I_c - I_a)/I_c g_c^2$

where g_m = mutual conductance, grid-anode;
g_c = conversion transconductance;
I_a = average anode current;
I_{g2} = average screen grid current;
I_c = average cathode current.

EXAMPLE 1.—*Calculate the total noise voltage and the signal-to-noise ratio at the grid of the first R.F. stage (Fig. 67(a)) of a receiver, given the following data:*

Input impedance	= 75 ohms
Tuning capacitance	= 200 pF
Circuit Q value	= 75
Signal frequency	= 15 Mc/s
Band-width	= 10 kc/s
Signal voltage in aerial feeder	= 5 μV
Valve mutual conductance	= 2 mA/V.

Equivalent resistance of tuned circuit

$$R_c = Q/\omega C$$

$$= \frac{75}{2\pi \times 15 \times 10^6 \times 200 \times 10^{-12}} = 3{,}980\ \Omega$$

Equivalent valve noise resistance

$$R_v = 2{\cdot}5/g_m$$

$$= \frac{2{\cdot}5}{2 \times 10^{-3}} = 1{,}250\ \Omega$$

Total noise voltage at grid input

$$e_t = 1{\cdot}26 \times 10^{-10}\sqrt{\Delta f(R_c + R_s)}$$
$$= 1{\cdot}26 \times 10^{-10}\sqrt{10{,}000\,(3{,}980 + 1{,}250)}$$
$$= 9{\cdot}2 \times 10^{-7} \text{ volts} = \underline{0{\cdot}92\ \mu V}$$

Voltage step-up ratio at input

$$n = \sqrt{R_c/R_i}$$
$$= \sqrt{\frac{3{,}980}{75}} = 7{\cdot}3$$

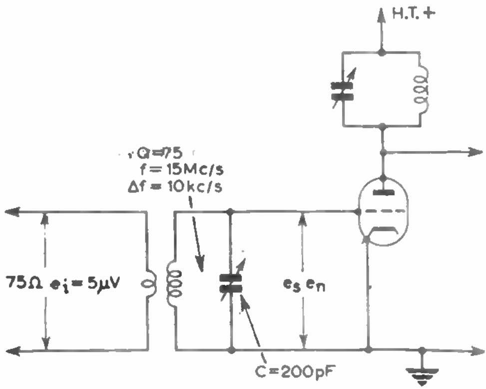


FIG. 67(*a*).—EXAMPLE 1.

When aerial and feeder are correctly matched to the receiver input circuit, the signal input voltage at the receiver input is half the aerial voltage.

Signal input voltage to grid

$$e_s = ne_{ae}/2$$
$$= \frac{7{\cdot}3 \times 5}{2} = 18{\cdot}25\ \mu V$$

Signal-to-noise ratio

$$N(\text{db}) = 20 \log_{10} (e_s/e_n)$$
$$= 20 \log_{10} \frac{18{\cdot}25}{0{\cdot}92} = \underline{60 \text{ db approx.}}$$

Noise Factor

The basis for the comparison of noisiness in receivers. It is defined by the ratio:

$$\frac{S_i/N_i}{S_o/N_o}$$

where S_i/N_i = signal-to-noise power ratio at input;

S_o/N_o = signal-to-noise power ratio at output.

The noise factor of an ideal amplifier would be unity (0 db). Actual values for multi-stage amplifiers and receivers may range from 2 to 10 (3 to 10 db).

RECEIVERS

Definitions of Performance

SENSITIVITY

The input in decibels above or below 1 μV required to produce a given power output. The usual standards of reference of output are:

Communications receivers

2 mW into 600 ohms

Broadcast receivers

50 mW (with input modulated to an average depth of 30 per cent)

$$s = 20 \log_{10} E_1$$

The modern tendency with communications and V.H.F. receivers is to define receiver sensitivity on an "absolute" basis in terms of noise factor (db). Noise factor has been defined as follows:

> The noise factor of a linear receiver, or any other linear four-terminal network having its input terminals connected to an impedance of stated value and temperature, is the number of times by which an addition to the noise power available from this impedance must exceed the thermal noise at signal frequency in order to double the noise power available from the output terminals of the network, provided that all sources of noise give the same frequency spectrum at the output terminals.

In practice, the noise factor is usually measured by means of a noise generator, which is adjusted to give a 2 : 1 change in the noise output from the receiver.

SELECTIVITY

Usually specified by the output voltage attenuation in decibels at several representative frequencies off resonance compared with the maximum voltage at resonance. Frequently plotted as a resonance curve for a band of frequencies.

$$s = 20 \log_{10} (E_1/E_r)$$

where E_r = output voltage at resonance;
E_1 = output voltage at a given frequency off resonance.

EXAMPLE 1.—*Tests on a receiver with a signal generator showed that an input of 5 μV was required to produce a standard output when the receiver was tuned to the test signal and an input of 125 μV when it was tuned 2 kc/s off resonance. What is the sensitivity expressed in decibels and the selectivity defined by the attenuation in decibels off resonance?*

Sensitivity with reference to zero level (1 μV)

$$s = 20 \log_{10} (5/1)$$
$$= 20 \times 0{\cdot}699 = \underline{14 \text{ db above } 1\ \mu\text{V}}$$

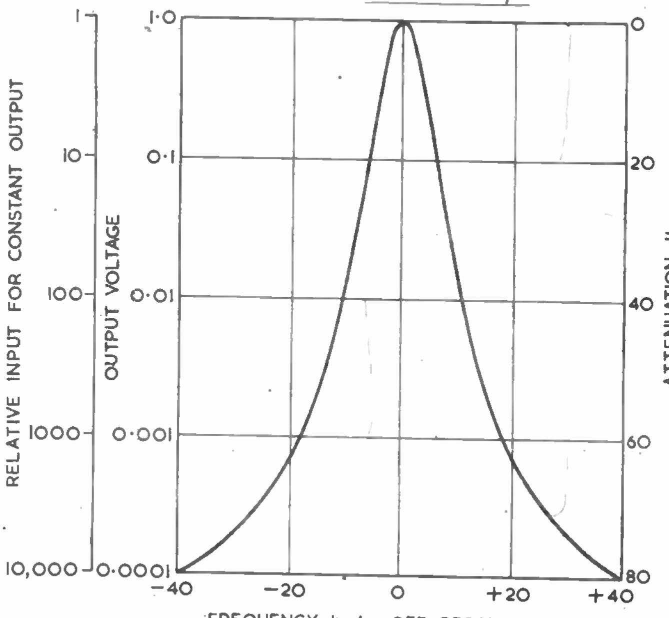


FIG. 68.—TYPICAL SELECTIVITY CURVE.

Attenuation at 2 kc/s off resonance

$$= 20 \log_{10} (125/5)$$
$$= 20 \times 1{\cdot}3979 = \underline{28 \text{ db}}$$

Gain

The ratio of the output voltage applied to a load designed to match the impedance of the output circuit to the input voltage from the aerial, expressed in decibels.

$$m = 20 \log_{10} (E_2/E_1) + 10 \log_{10} (R_1/R_2)$$

where E_1 = input voltage from aerial;
E_2 = output voltage across load;
R_1 = input resistance, ohms;
R_2 = output resistance, ohms.

EXAMPLE 2.—*A communication receiver requires an input of 10 μV across 100 $k\Omega$ to produce a power output of 2 mW into a 600-ohm line. Determine the sensitivity in decibels to an input reference level of 1 μV and the overall gain in decibels.*

FIG. 69.—EXAMPLE 2.

SENSITIVITY

$$s = 20 \log_{10} E = 20 \log_{10} 10 = \underline{20 \text{ db above } 1\ \mu V}$$

OUTPUT VOLTAGE

$$E = \sqrt{PR} = \sqrt{2 \times 10^{-3} \times 600} = \sqrt{1{\cdot}2} = 1{\cdot}098$$

GAIN

$$m = 20 \log_{10} (E_2/E_1) + 10 \log_{10} (R_1/R_0)$$
$$= 20 \log \frac{1{\cdot}098}{10^{-5}} + 10 \log \frac{10^5}{600}$$
$$= 20 \log (1{\cdot}098 \times 10^5) + 10 \log (1{\cdot}667 \times 10^2) = \underline{123 \text{ db}}$$

Signal-to-noise Ratio

Defined by the ratio of the fully modulated signal voltage at maximum volume to the average noise expressed in decibels.

$$= 20 \log_{10} (E_s/E_n)$$

where E_s = signal voltage;
E_n = noise voltage.

Frequency Range

Maximum and minimum frequencies of tuning range

$$f_{min.} = \sqrt{\alpha} f_{max.}$$

Overall range :

$$f_{max.} - f_{min.} = (1 - 1/\sqrt{\alpha}) f_{min.}$$

where $f_{max.}$ = maximum frequency of range;
$f_{min.}$ = minimum frequency of range;
σ = ratio of minimum to maximum tuning capacitance.

EXAMPLE 3.—*A receiver is tuned by a 300-pF capacitor having a minimum capacitance of 15 pF. What is the minimum frequency to which the receiver will tune if the maximum frequency is 1 Mc/s? If a fixed 300-pF capacitor is connected in parallel with the variable capacitor, what will be the new maximum and minimum frequencies?*

(*a*) On the higher band

$$\alpha = C_{min.}/C_{max.} = \frac{15}{300} = 0{\cdot}05$$

Minimum frequency :

$$f_{min.} = \sqrt{\alpha} f_{max.} = \sqrt{0{\cdot}05} \times 1 = 0{\cdot}224 \text{ Mc/s or } \underline{224 \text{ kc/s}}$$

(*b*) On the lower band :

$$C^1_{min.} = 15 + 300 = 315 \text{ pF}$$

$$C^1_{max.} = 300 + 300 = 600 \text{ pF}$$

$$\alpha^1 = C^1_{min.}/C^1_{max.} = \frac{315}{600} = 0{\cdot}525$$

Since $$f \propto 1/\sqrt{C},$$

$$f^1_{max.}/f_{max.} = \sqrt{C_{min.}/C^1_{min.}}$$

Hence the maximum frequency :

$$f^1_{max.} = f_{max.}\sqrt{C_{min.}/C^1_{min.}} = 1\sqrt{\frac{15}{315}} = 0{\cdot}218 \text{ Mc/s or } \underline{218 \text{ kc/s}}$$

Minimum frequency :

$$f^1_{min.} = \sqrt{\alpha^1} f^1_{max.} = 0{\cdot}7253 \times 218 = \underline{158 \text{ kc/s}}$$

Superheterodyne Receivers

Intermediate Frequency (I.F.)

$$f_i = f_r - f_0 \quad \text{or} \quad f_o - f_r$$

Interference signals. Interference may be produced by :

(*a*) A second-channel received signal, above or below the local oscillator frequency, beating with the local oscillator, when the frequency difference is equal to the intermediate frequency.

$$f_r = f_0 - f_i \text{ (2nd channel to an I.F.} = f_r - f_0)$$

or $$f_r = f_0 + f_i \text{ (2nd channel to an I.F.} = f_0 - f_r)$$

(*b*) Two received signals equally displaced either side of the tuned frequency, whose frequency difference is equal to the intermediate frequency, beating together independently of the local oscillator.

$$f_{r1} = f_r + f_i/2$$
$$f_{r2} = f_r - f_i/2$$

(*c*) A signal frequency differing from harmonics of the intermediate frequency by up to ± 10 kc/s, depending on the selectivity of the receiver.

$$f_r = n f_i \pm (0 \text{ to } 10 \text{ kc/s})$$

where f_r = frequency of received signal;
f_0 = frequency of local oscillator;
f_i = intermediate frequency;
n = harmonic multiple of intermediate frequency.

EXAMPLE 4.—*At what frequencies would unwanted signals be expected to produce spurious responses in a superheterodyne receiver tuned to receive a signal of* 15 *Mc/s with the local oscillator adjusted to* 15·45 *Mc/s?*

(*a*) A second-channel signal at a frequency f_r, such that

$$f_i = f_r - f_0 \text{ or } f_r = f_i + f_0 = 15{\cdot}45 + 0{\cdot}45 = \underline{15{\cdot}9 \text{ Mc/s}}$$

(*b*) Two received frequencies f_{r1} and f_{r2} equally displaced either side of the resonant frequency, producing an independent beat:

$$f_{r1} = f_r + f_i/2 = 15 + 0{\cdot}45/2 = \underline{15{\cdot}225 \text{ Mc/s}}$$
$$f_{r2} = f_r - f_i/2 = 15 - 0{\cdot}45/2 = \underline{14{\cdot}775 \text{ Mc/s}}$$

(*c*) Frequencies differing from the lower harmonics of the intermediate frequency by up to 10 kc/s.

Second intermediate-frequency harmonic:

$$= 2f_i \pm (0 \text{ to } 10) = (2 \times 450) \pm (0 \text{ to } 10)$$

or frequencies between <u>890 and 910 kc/s</u>

Third intermediate-frequency harmonic:

$$f_r = 3f_i \pm (0 \text{ to } 10) = (3 \times 450) \pm (0 \text{ to } 10)$$

or frequencies between <u>1·34 and 1·36 Mc/s</u>

OSCILLATORS

Valve Oscillators

TABLE 7.—CONDITIONS FOR SELF-OSCILLATION

	Tuned Grid Circuit (*Fig.* 70 (a))	*Tuned Anode Circuit* (*Fig.* 70 (b))
Minimum mutual inductance to sustain self-oscillation .	$M = CR/g_m$	$M = (CR + L/R_a)/g_m$
Oscillation frequency . .	$f = 1/2\pi\sqrt{LC}$	$f = 1/2\pi\sqrt{LC}$

where M = minimum mutual inductance, henrys;
C = tuning capacitance, farads;
R = R.F. resistance of tuned circuit, ohms;
R_a = A.C. resistance of valve, ohms;
L = tuning inductance, henrys;
g_m = mutual conductance of valve, amperes/volt.
f = oscillation frequency, c/s.

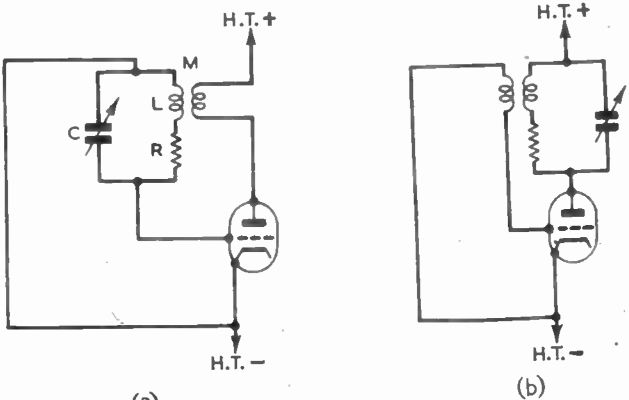


FIG. 70.—VALVE OSCILLATORS.

EXAMPLE 1.—*A parallel tuned circuit consisting of a coil of 150 mH inductance and 60 ohms resistance and a capacitor of 0·05 μF is to be used with a valve oscillator having a mutual conductance of 4 mA/volt and an A.C. resistance of 15 kΩ. What is the minimum value of mutual inductance required to sustain oscillation when used as:* (a) *a tuned anode circuit;* *and* (b) *a tuned grid circuit? What is the frequency of oscillation in each case?*

MINIMUM VALUE OF M

(a) As a tuned anode circuit:

$$M_1 = (CR + L/R_a)/g_m$$

$$= \frac{(0{\cdot}05 \times 10^{-6} \times 60) + [150 \times 10^{-3}/(1{\cdot}5 \times 10^4)]}{4 \times 10^{-3}}$$

$$= 3{\cdot}25 \times 10^{-3} \text{ henry or } \underline{3{\cdot}25 \text{ mH}}$$

(b) As a tuned grid circuit:

$$M_2 = CR/g_m$$

$$= \frac{0{\cdot}05 \times 10^{-6} \times 60}{4 \times 10^{-3}}$$

$$= 0{\cdot}75 \times 10^{-3} \text{ henry or } \underline{0{\cdot}75 \text{ mH}}$$

EFFECTIVE INDUCTANCE OF TUNED CIRCUIT

$$L_1{}^1 = L - M_1 = 150 - 3{\cdot}25 = 146{\cdot}75 \text{ mH}$$

$$L_2{}^1 = L - M_2 = 150 - 0{\cdot}75 = 149{\cdot}25 \text{ mH}$$

OSCILLATION FREQUENCY

$$f_1 = 1/2\pi\sqrt{L_1{}^1C}$$

$$= \frac{1}{2\pi\sqrt{146{\cdot}75 \times 10^{-3} \times 0{\cdot}05 \times 10^{-6}}}$$

$$= \underline{1{,}865 \text{ c/s}}$$

$$f_2 = f_1\sqrt{L_1{}^1/L_2{}^1} = 1{,}865\sqrt{\frac{146{\cdot}75}{149{\cdot}25}}$$

$$= \underline{1{,}850 \text{ c/s}}$$

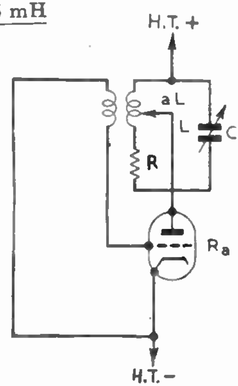


FIG. 71.—ANODE TAP.

Anode Tap

To obtain maximum power output, the anode load impedance must be made equal to the anode A.C. resistance. This condition is realized when the fraction of the inductance tapped from the earthy end

$$a = \sqrt{RR_aC/L}$$

Quartz Crystal Oscillators

	X-cut Plate	Y-cut Plate
Frequency of thickness vibration	$f_c = 2{\cdot}87/t$	$f_c = 2{\cdot}44/t$
Temperature coefficient per ° C.	$\alpha = -2{\cdot}1 \times 10^{-5}$	$\alpha = +9 \times 10^{-5}$

Frequency Variation with Temperature :

$$\Delta f_e = \alpha T f_e \times 10^6$$

where f_e = frequency, Mc/s;
t = thickness of plate, mm.;
α = temperature coefficient, c/s per c/s per ° C.;
T = temperature change, ° C.

EXAMPLE 2.—*What is the natural frequency of an X-cut quartz crystal ground to a thickness of 2·5 mm. and the maximum frequency variation, if the ambient temperature varies between the normal 15° C. and 30° C.?*

Natural frequency

$$= 2{\cdot}87/t = \frac{2{\cdot}87}{2{\cdot}5} = \underline{1{\cdot}15\ \mathrm{Mc/s}}$$

Maximum frequency variation :

$$\Delta f_e = \alpha T f_e = (-\,2{\cdot}1 \times 10^{-5})(30 - 15)(1{\cdot}15 \times 10^6) = \underline{-\,363\ \mathrm{c/s}}$$

Master Oscillators for Transmitters

Frequency tolerance as percentage :

$$T = \pm\, (\Delta f_e/f_e) \times 100$$

Frequency multiplication. The fundamental or transmitted frequency:

$$f_0 = n f_e$$

where f_0 = fundamental frequency of transmitter, Mc/s;
Δf_e = frequency deviation, Mc/s;
n = total number of successive frequency multiplications;
T = frequency tolerance as percentage.

EXAMPLE 3.—*A crystal oscillator cut for a frequency of 1,500 kc/s and having a temperature coefficient of −2 c/s/Mc/s/° C. drives a transmitter operating at 15 Mc/s. If the crystal is subjected to a temperature increase of 5° C., what is the altered transmitter frequency and the percentage frequency change?*

Change of oscillator frequency at 1·5 Mc/s :

$$\Delta f_e = -\,2 \times 5 \times 1{\cdot}5 = -\,15\ \mathrm{c/s}$$

Change of fundamental frequency :

$$\Delta f_0 = -\,15 \times \frac{15}{1{\cdot}5} = -\,150\ \mathrm{c/s}$$

Altered transmitter frequency :

$$f_0' = 15{\cdot}0 - 0{\cdot}00015 = \underline{14{\cdot}99985\ \mathrm{Mc/s}}$$

Percentage frequency change :

$$= \frac{0{\cdot}00015}{15} \times 100 = \underline{0{\cdot}001\ \text{per cent}}$$

TRANSMITTERS

Power Amplifiers

Driving power: $P_{g.av.} = E_{g.av.} I_{g.av.}$

Power input to anode: $P_i = E_a I_a$

Anode conversion efficiency: $\eta_a = P_0/P_i = P_0/(P_0 + P_a)$

Anode dissipation: $P_a = P_i - P_0 = P_0(1/\eta_a - 1)$
$= P_i(1 - \eta_a)$

where
$P_{g.av.}$ = average driving power;
$E_{g.av.}$ = average value of grid voltage;
$I_{g.av.}$ = average value of grid current, amperes;
P_i = anode power input;
P_0 = radio-frequency power output;
P_a = power dissipated at anode;
η_a = anode conversion efficiency of amplifier.

Average values of η_a:

Class A amplifiers	0·2 –0·3
Class B audio-frequency amplifiers	0·35–0·65
Class B radio-frequency amplifiers	0·6 –0·7
Class C radio-frequency amplifiers	0·65–0·85

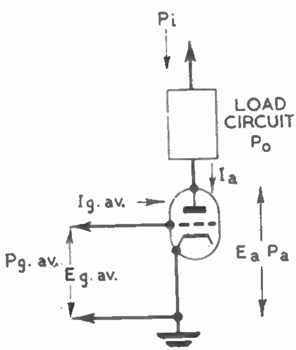


FIG. 72.—POWER AMPLIFIER.

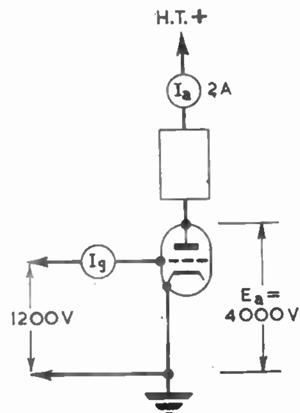


FIG. 73.—EXAMPLE 1.

EXAMPLE 1.—*The power amplifier of a transmitter takes a D.C. anode current of 2 amperes at 4,000 volts and delivers 6 kW of radio-frequency power to the load circuit with a driving voltage on the grid of 1,200 volts r.m.s. and a grid current of 300 mA. Calculate the anode conversion efficiency, anode dissipation and driving power.*

Power input to anode:

$$P_i = E_a I_a = 4{,}000 \times 2 = 8{,}000 \text{ watts or } 8 \text{ kW}$$

Anode conversion efficiency :

$$\eta_a = P_0/P_i = \frac{6}{8} = 0{\cdot}75 \text{ or } \underline{75 \text{ per cent}}$$

Anode dissipation :

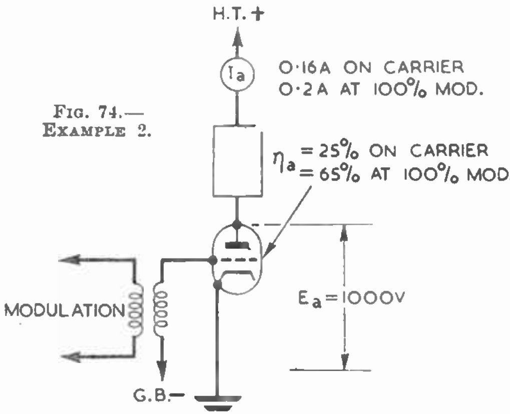
$$P_a = P_i - P_0 = 8 - 6 = \underline{2\ \text{kW}}$$

Driving power :

$$P_{g.av.} = I_{g.av.}E_{g.av.} = 0{\cdot}3 \times 1{,}200 = \underline{360 \text{ watts}}$$

EXAMPLE 2.—*A Class C grid-modulated radio-frequency amplifier has an anode efficiency of* 65 *per cent at* 100 *per cent modulation and takes an anode current of* 0·2 *ampere at* 1,000 *volts. When the carrier is quiescent,*

FIG. 74.—EXAMPLE 2.



the efficiency falls to 25 *per cent and the current to* 0·16 *ampere. Calculate the anode dissipation under both conditions and the average dissipation, if the average depth of modulation is* 30 *per cent and the dissipation decreases directly with depth of modulation.*

Anode dissipation :

$$P_a = P_i(1 - \eta_a)$$

When fully modulated:

$$P_a = (0{\cdot}2 \times 1{,}000)(1 - 0{\cdot}65) = \underline{70 \text{ watts}}$$

When unmodulated :

$$P_a = (0{\cdot}16 \times 1{,}000)(1 - 0{\cdot}25) = \underline{120 \text{ watts}}$$

Decrease in dissipation at 100 per cent modulation $= 120 - 70$
$= 50$ watts

Decrease in dissipation at 30 per cent modulation $= 50 \times \frac{30}{100}$
$= 15$ watts

So average dissipation at 30 per cent modulation $= 120 - 15$
$= \underline{105 \text{ watts}}$

Modulated Amplifiers (Modulated at Anode)

Power input to anode $P_i = P_0(1 + m^2/2)/\eta_a$

Voltage variation at anode : $E_a = \dfrac{\mu E'_g X_L}{R_m\sqrt{1 + \{X_L(R_a + R_m)/R_aR_m\}^2}}$

Depth of modulation : $m = E_m/E_c = I_m/I_c$

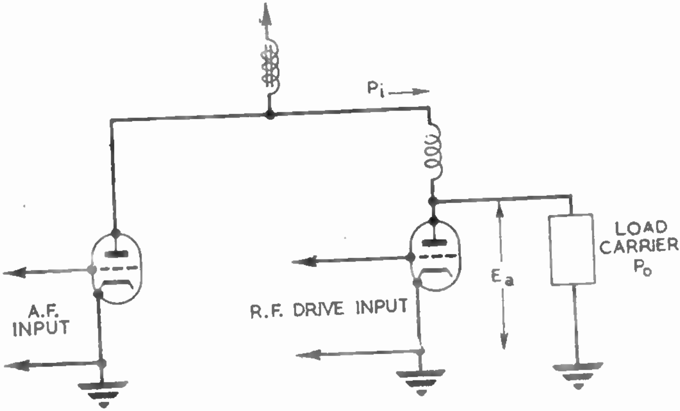


Fig. 75.—Modulated Amplifier.

Amplitude Modulators

Power input to anode : $P = P_0 m^2/2\eta_a\eta_m$

where m = depth of modulation;
E_c = r.m.s. value of carrier voltage;
E_m = r.m.s. value of modulating voltage;
μ = amplification factor of modulator valve;
E'_g = peak voltage applied to grid of modulator valve;
X_L = reactance of modulation choke ($= 2\pi fL$);
R_a = anode resistance of modulated amplifier valve;
R_m = anode resistance of modulator valve;
η_m = anode conversion efficiency of modulator.

Example 3.—*A transmitter is modulated by anode choke control. Given the following data, calculate the anode voltage variation at the* *modulated amplifier when an input of* 10 *volts peak at a frequency of* 1,000 *c/s is applied to the modulator grid:*

Modulator: $\mu = 8$, $R_m = 2{,}500$ *ohms*
Modulated amplifier: $R_a = 10{,}000$ *ohms*
Choke inductance = 20 *henrys*

If the peak value of the unmodulated radio-frequency voltage at the amplifier anode is 530 *volts, what is the percentage modulation?*

(*a*) Reactance of choke:

$$X_L = 2\pi fL = 2\pi \times 1{,}000 \times 20 = 1{\cdot}256 \times 10^5$$

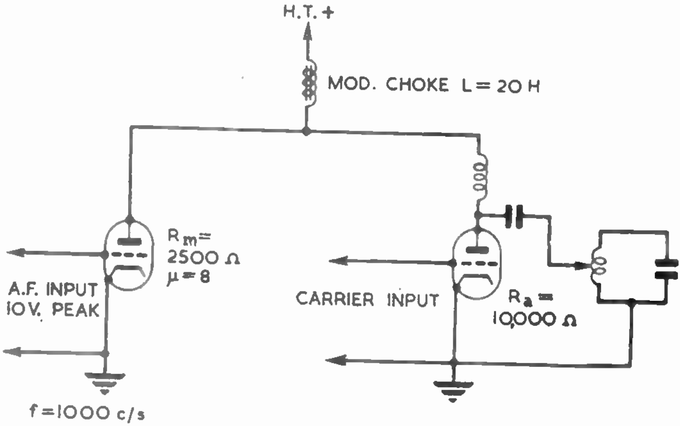


FIG. 76.—EXAMPLE 3.

Anode voltage variation:

$$E_a = \frac{\mu E'_g X_L}{R_m\sqrt{1 + \{X_L(R_a + R_m)/R_aR_m\}^2}}$$

$$= \frac{8 \times 10 \times 1{\cdot}256 \times 10^5}{2{,}500\sqrt{1 + \{(1{\cdot}256 \times 10^5)(10{,}000 + 2{,}500)/(10{,}000 \times 2{,}500)\}^2}}$$

$$= \underline{64 \text{ volts peak}}$$

(*b*) Depth of modulation:

$$m = E_m/E_c = \frac{64}{530} = 0{\cdot}12 \text{ or } \underline{12 \text{ per cent}}$$

Modulation Transformers

Turns ratio for maximum power output (for fixed input):

$$k = \sqrt{Z_m/Z_a}$$

For distortionless output the anode load impedance of the modulator must be twice the dynamic impedance of the modulated amplifier.

$$k = \sqrt{2Z_m/Z_a}$$

where Z_m = anode impedance of modulator;
Z_a = anode impedance of modulated amplifier.

Load and Aerial Power

Power developed in tuned load circuit: $P_0 = E_0 I_0$
Power dissipated in tuned load circuit: $P_L = I_0^2 R_L$
Aerial power: $P_{ae} = I_{ae}^2 R_e$
Aerial current: $I_{ae} = \sqrt{P_{ae}/R_e}$
Field intensity: $\epsilon \propto \sqrt{P_{ae}} \propto I_{ae}$

where E_0 = r.m.s. voltage across tuned load circuit ($= I_0/\omega C = \omega L I_0$);
I_0 = r.m.s. current in tuned load circuit;
I_{ae} = r.m.s. value of aerial current;
R_e = effective resistance of aerial (ohmic + radiation resistance), ohms;
R_L = resistance of load circuit, ohms.

EXAMPLE 4.—*What percentage increase of aerial current and field intensity will occur with an amplitude-modulated transmitter when the depth of modulation is increased from 75 to 100 per cent? By what percentage must the current be reduced to halve the radiated power?*

(a) R.m.s. aerial current: $I_{ae} = I_c \sqrt{(1 + m^2/2)}$

At 75 per cent modulation: $I_{ae} = I_c \sqrt{(1 + 0{\cdot}75^2/2)} = 1{\cdot}132\ I_c$

At 100 per cent modulation: $I_{ae} = I_c \sqrt{(1 + 1^2/2)} = 1{\cdot}225\ I_c$

Ratio of increase of aerial current and field intensity

$$= \frac{1{\cdot}225 - 1{\cdot}132}{1{\cdot}132} = 0{\cdot}082 \text{ or } \underline{8{\cdot}2 \text{ per cent increase}}$$

(b) Since $I_{ae} \propto \sqrt{P_{ae}}$, in order to halve the power

$I'_{ae}/I_{ae} = \sqrt{P'_{ae}/P_{ae}}$

$I'_{ae} = I_{ae}/\sqrt{2} = 0{\cdot}707\ I_{ae}$ or a reduction of $\underline{29{\cdot}3 \text{ per cent}}$

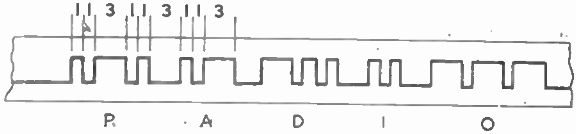
SIGNALLING AND MODULATION

Telegraph Signalling

Number of elementary units (dots + spaces) per letter

Morse	8
Cable Morse . . .	5
Five-unit teleprinter code .	7 (including start and stop impulses)

Average Morse letter = 8 units or 4 rectangular keying c/s.
Average word = 5 letters + space = 6 letters.



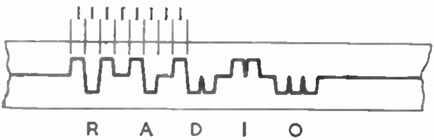


Fig. 77.—Relative Lengths of Elements in Morse and Cable Morse Keying.

Speed in Bauds is the number of elementary impulses or alternations of dots and spaces per second, and is equal to twice the rectangular keying cycles per second.

$$S = sN/60$$

where S = speed in Bauds;
s = speed in words/minute;
N = number of elements per word (dots + spaces per second × average number of letters per word).

Example 1.—*Calculate the signalling speed in Bauds corresponding to:* (a) *hand Morse at* 20 *w.p.m.;* (b) *teleprinter transmission in* 5 + 2 *unit code at* 60 *w.p.m.;* (c) *cable Morse at* 200 *w.p.m. Take the average word as equivalent to* 6 *letters.*

Speed in Bauds:

$$S = sN/60$$

(a) $$S = \frac{20 \times 8 \times 6}{60} = \underline{16 \text{ Bauds}}$$

(b) $$S = \frac{60 \times 7 \times 6}{60} = \underline{42 \text{ Bauds}}$$

(c) $$S = \frac{200 \times 5 \times 6}{60} = \underline{100 \text{ Bauds}}$$

Amplitude Modulation

TABLE 8.—AMPLITUDE MODULATION FACTORS

	D.S.B. and Carrier	*S.S.B. and Carrier*
Depth of modulation (modulation index), m	I_m/I_c	I_m/I_c
Ratio peak modulated current/unmodulated current	$1 + m$	$1 + m/2$
R.m.s. value of total modulated current, I_t	$I_c\sqrt{1 + m^2/2}$	$I_c\sqrt{1 + m^2/4}$
Ratio modulated power/unmodulated power, P_t/P_c	$1 + m^2/2$	$1 + m^2/4$
Total modulated power, P_t	$P_c(1 + m^2/2)$	$P_c(1 + m^2/4)$
Total sideband power, P_s	$P_c m^2/2$ $= P_t/(1 + 2/m^2)$	$P_c m^2/4$ $= P_t/(1 + 4/m^2)$

where m = depth of modulation;
I_c = r.m.s. value of unmodulated carrier current, amperes;
I_m = r.m.s. value of modulating current, amperes;
I_t = r.m.s. value of total modulated current, amperes;
P_c = unmodulated power, watts;
P_t = total modulated power, watts.

EXAMPLE 2.—*The measured power output from an amplitude modulated transmitter is 20 kW when the carrier is unmodulated and 28 kW when*

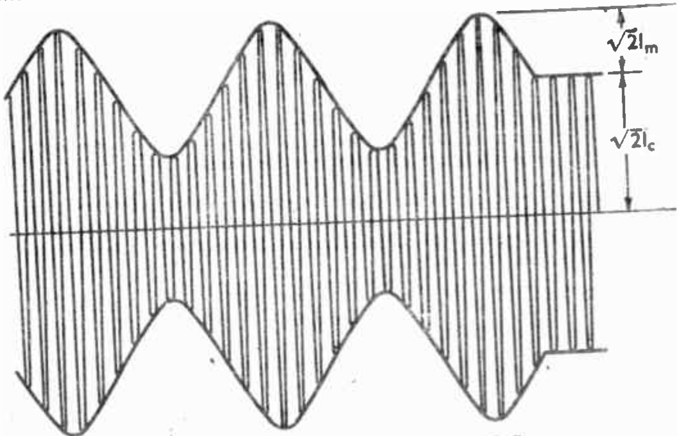


FIG. 78.—CARRIER ENVELOPE SHOWING AMPLITUDE MODULATION.

modulated by a steady sinusoidal tone. What is the total power in the sidebands and the percentage modulation? If the modulation is now reduced to 50 per cent what is the percentage reduction of sideband power?

Total power in sidebands:

$$P_s = P_t - P_c = 28 - 20 = \underline{8\ \text{kW}}$$

Percentage modulation:

$$P_t/P_c = 1 + m^2/2$$

$$m = \sqrt{2(P_t/P_c) - 1} = \sqrt{2(28/20 - 1)}$$

$$= 0{\cdot}895 \text{ or } \underline{89{\cdot}5 \text{ per cent}}$$

Since sideband power $\propto m^2$, the reduction of sideband power at 50 per cent modulation:

$$P_s'/P_t = (m'/m)^2 = \left(\frac{0{\cdot}5}{0{\cdot}895}\right)^2 = 0{\cdot}313$$

i.e., a reduction of $1 - 0{\cdot}313 = 0{\cdot}687$ or $\underline{68{\cdot}7 \text{ per cent}}$

Frequency Modulation

Ratio modulated current/unmodulated current, I_t/I_c . .	1
R.m.s. value of total modulated current, I_t	I_c
Ratio modulated power/unmodulated power, P_t/P_c . .	1
Total modulated power, P	P_c

Transmission Bandwidth

C.W. telegraphy, on–off keyed	$6f_k$ *
C.W. telegraphy, frequency shift keyed . .	$2(f_d + 3f_k)$ *
M.C.W. telegraphy, on–off keyed . . .	$2(f_m + 3f_k)$ *
Amplitude modulation, D.S.B. and carrier .	$2f_{m.\,max.}$
Amplitude modulation, S.S.B. and carrier .	$f_{m.\,max.}$
Amplitude modulation, S.S.B. only . . .	$f_{m.\,max.} - f_{m.\,min.}$
Wide-band frequency modulation ($f_m < f_d$) .	$2f_d$
Facsimile, keyed carrier	$1{\cdot}5f_e$ †
Facsimile, tone modulated	$1{\cdot}5f_e + 2f_m$ †
Television	$1{\cdot}5f_f f_e f_s$

* For good signal formation at least the third harmonic of the keying frequency, f_k, must be transmitted. Hence the total bandwidth due to keying $= 2 \times 3f_k$.

† 1·5 is a factor that allows for signal shaping and synchronization.

where
f_k = keying frequency, rectangular c/s;
f_d = deviation frequency, c/s;
f_m = modulation frequency, c/s;
f_e = no. of picture elements per line;
f_f = no. of picture frames per second;
f_s = no. of scanning lines per frame.

EXAMPLE 3.—*Determine:* (a) *the number of Morse telegraph transmitters sending at 250 w.p.m.; and* (b) *the number of double-sideband*

D

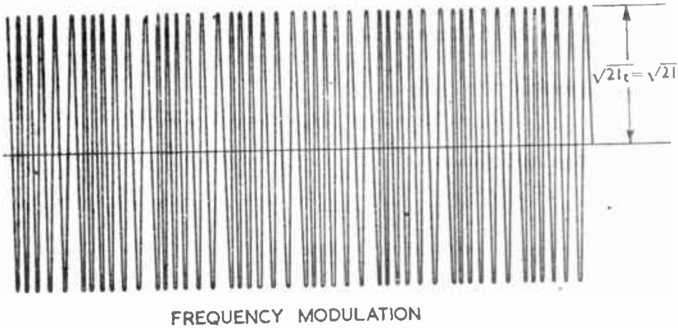


FIG. 79.—CARRIER ENVELOPE SHOWING FREQUENCY MODULATION

radiotelephone stations using a maximum modulation frequency of 3,000 c/s which can operate at the same time within the high-frequency band of 10–16 Mc/s.

(*a*) Speed in Bauds:

$$S = sN/60 = \frac{250 \times 8 \times 6}{60}$$

$$= 200 \text{ Bauds or } 100 \text{ c/s fundamental}$$

Total bandwidth per channel is twice the third-harmonic frequency

$$= 2 \times 3 \times 100 = 600 \text{ c/s}$$

Total number of 600-c/s channels between 10 and 16 Mc/s

$$= \frac{(16 - 10) \times 10^6}{600} = \underline{10,000}$$

(*b*) Total bandwidth per channel

$$= 2 \times 3,000 = 6,000 \text{ c/s}$$

Total number of 6,000-c/s channels between 10 and 16 Mc/s

$$= \frac{(16 - 10) \times 10^6}{6,000} = \underline{1,000}$$

EXAMPLE 4.—*A transmitter is controlled by a master oscillator oscillating at a frequency of 2,750 kc/s with a stability of* ± 10 *parts in* 10^6. *The sixth harmonic is selected, amplified and modulated at the final stage. What are the maximum and minimum limits of the radiated frequency band produced by modulating at broadcasting frequencies up to 10 kc/s?*

Fundamental carrier frequency:

$$f_c = 6f_0 = 6 \times 2,750 = 16,500 \text{ kc/s}$$

Maximum and minimum frequency products of modulation :

$$f_{max.} = f_c + f_m = 16{,}500 + 10 = 16{,}510 \text{ kc/s}$$
$$f_{min.} = f_c - f_m = 16{,}500 - 10 = 16{,}490 \text{ kc/s}$$

Maximum and minimum frequencies due to instability :

$$f'_{max.} = f_{max.} + \Delta f = 16{,}510 + \frac{10 \times 16{,}500}{10^6} = \underline{16{,}510 \cdot 17 \text{ kc/s}}$$

$$f_{min.} = f_{min.} - \Delta f = 16{,}490 - \frac{10 \times 16{,}500}{10^6} = \underline{16{,}489 \cdot 84 \text{ kc/s}}$$

EXAMPLE 5.—*What is the highest modulation frequency and the overall bandwidth in a television system transmitting 25 frames per second, each built up of 460 lines and 460 elements per line ?*

Highest modulation frequency :

$$f_m = f_f f_l f_e = 25 \times 460 \times 460$$
$$= 5 \cdot 29 \times 10^6 \text{ c/s or } \underline{5 \cdot 29 \text{ Mc/s}}$$

The overall bandwidth is 1·5 times this :

$$= 1 \cdot 5 \times 5 \cdot 29 = \underline{7 \cdot 94 \text{ Mc/s}}$$

VALVE COOLING AND VENTILATION

Power Losses in Valves

The total power losses in an amplifying valve are made up of power dissipated as heat by electron bombardment : (*a*) at the anode and (*b*) at the grid or grids, and as heat (*c*) in the filament.

$$P_L = P_a + P_g + P_f$$

where
P_a = power loss at anode ;
P_g = power loss at grid ;
P_f = filament heating loss.

Water-cooling Systems

Rate of flow required to dissipate a given amount of power as heat with a given temperature rise of water :

$$Q = 3 \cdot 2\, P_L/T$$

where Q = rate of flow of water, gal./minute ;
P_L = power dissipated as heat, kW ;
T = temperature rise of water, ° C.

LOSSES IN WATER-INSULATING COLUMNS

Resistance of circular water column :

$$R = (\rho l/A) \times 10^6$$

D.C. loss in water column :

$$P = E_a^2/R = (E_a^2 A/\rho l) \times 10^{-6}$$

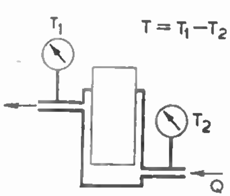


FIG. 80.—WATER COOLING.

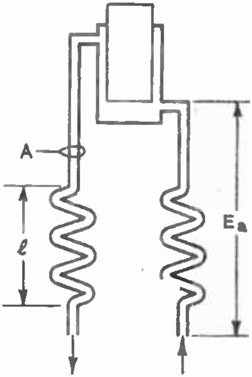


FIG. 81.—LOSSES IN WATER COLUMNS.

where P = power loss, watts;
E_a = D.C. anode voltage;
R = resistance of column, ohms;
ρ = resistivity of water, MΩ-cm. (0·15 for freshly distilled water, 0·002 or less for tap water);
A = cross-sectional area of bore, sq. cm.;
l = developed length of tube, cm.

EXAMPLE 1.—*The anode jacket of a water-cooled valve is fed with water via a spiral inlet and outlet insulating tube, the tubes being electrically in parallel between anode and earth. Each tube is 30 ft. long and 1 in. bore. The resistivity of the water is 0·05 MΩ-cm. What is the D.C. resistance between anode and earth and the D.C. leakage loss when the anode voltage is 10 kV?*

$$\rho = 0{\cdot}05 \text{ M}\Omega\text{-cm.} = 5 \times 10^4 \text{ ohms-cm.}$$

$$l = 30 \text{ ft.} = 30 \times 30{\cdot}5 = 915 \text{ cm.}$$

$$A = \pi d^2/4 = \frac{\pi \times 2{\cdot}54^2}{4} = 5{\cdot}07 \text{ sq. cm.}$$

Resistance of one column:

$$R = (\rho l/A) = \frac{5{\cdot}0 \times 10^4 \times 915}{5{\cdot}07}$$

$$\rho = 9{\cdot}05 \times 10^6 \text{ ohms or } 9{\cdot}05 \text{ M}\Omega$$

Total resistance to earth of both columns is half this, i.e., 4·52 MΩ.
D.C. loss:

$$P = E_a^2/R = \frac{10{,}000^2}{4{\cdot}52 \times 10^6} = \underline{22{\cdot}2 \text{ watts}}$$

Water-cooling Tanks

The area of the cooling surface required varies directly as the power dissipated and the thickness of the material, and inversely as the coefficient of thermal conductivity and the temperature rise.

COEFFICIENT OF THERMAL CONDUCTIVITY.—The rate at which heat is transmitted through unit thickness over unit area of material per degree difference in temperature between surfaces.

$$A = 238\ Pt/kT$$

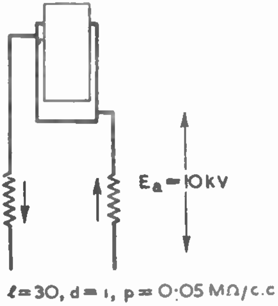


FIG. 82.—EXAMPLE 1.

where P = power to be dissipated, watts;
t = thickness of material, cm.;
A = surface area, sq. cm.;
T = temperature difference between surfaces, ° C.;
k = coefficient of thermal conductivity.

EXAMPLE 2.—*A 500-watt resistor is fitted in a cylindrical copper container $\frac{1}{8}$ in. thick and cooled by circulating water. Calculate the total cooling surface required and the dimensions of the container, if the length is to be twice the diameter and the temperature rise of the copper is not to exceed 10° C. Take the thermal conductivity of copper as 0·918.*

$$t = 1/8 \times 2{\cdot}54 = 0{\cdot}318 \text{ cm.}$$

Surface area required:

$$A = 238\, P_t/kT = \frac{238 \times 500 \times 0{\cdot}318}{0{\cdot}918 \times 10}$$

$$= \underline{4{,}130 \text{ sq. cm.}}$$

$$A = \pi d^2 l/4$$

But $l = 2d$,

and $A = 2\pi d^3/4$,

Hence $d = \sqrt[3]{2A/\pi} = \sqrt[3]{\frac{2 \times 4{,}130}{\pi}}$

$$= \underline{13{\cdot}8 \text{ cm.}}$$

$$l = 2d = \underline{27{\cdot}6 \text{ cm.}}$$

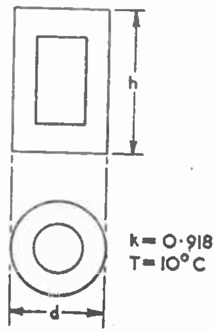


FIG. 83.—EXAMPLE 2.

Air-cooling Systems

Air flow and temperature:

$$Q = 5{\cdot}92\, P_L T_0/T_d$$

where Q = air flow, cu. ft./minute;
P_L = power dissipated as heat, kW;
T_0 = absolute temperature of incoming air (° C. + 273);
T = temperature difference between outgoing and incoming air, ° C.

EXAMPLE 3.—*An air-cooled transmitter valve requires 300 cu. ft. of air per minute. The measured temperature of the incoming air is 72° F. and the heated exhaust air 130° F. When the H.T. voltage is reduced, the temperature of the exhaust air falls to 95° F. Calculate the power in kilowatts dissipated as heat in both cases.*

$$T_0 = {}^\circ\text{C.} + 273 = (72 - 32)/1{\cdot}8 + 273 = 295{\cdot}2^\circ \text{ K.}$$

$$T_d = (130 - 72)/1{\cdot}8 = 32{\cdot}2^\circ \text{ C.}$$

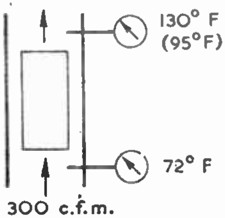


FIG. 84.—EXAMPLE 3.

Since $$Q = 5{\cdot}92\ PT_0/T_d,$$

Power dissipated at full H.T. voltage:

$$P = QT_d/5{\cdot}92T_0 = \frac{300 \times 32{\cdot}2}{5{\cdot}92 \times 295{\cdot}2}$$

$$= \underline{5{\cdot}53 \text{ kW}}$$

When H.T. voltage is reduced, the temperature rise :

$$T_d' = (95 - 72)/1{\cdot}8 = 12{\cdot}78° \text{ C.}$$

Since $$P \propto T_d,$$

$$p'/p = T_d'/T_d$$

Power dissipated at reduced H.T. voltage :

$$P' = PT_d'/T_d = \frac{5{\cdot}53 \times 12{\cdot}78}{32{\cdot}2} = \underline{2{\cdot}2 \text{ kW}}$$

Air Velocity in Ducts

$$v = Q/A$$

where v = velocity, ft./minute ;
Q = rate of flow, cu. ft./minute ;
A = cross-sectional area of duct, sq. ft.

Ventilation of Transmitters

The expression for air flow in air-cooling systems can be applied to problems on the ventilation of cubicles and rooms. A simpler empirical formula, sufficiently accurate for most practical purposes, is based on the fact that 1 kW dissipated will raise the air temperature 1° C. with an air displacement of 1,725 cu. ft./minute approximately, i.e.,

$$Q = 1{,}725\ P_L/T_d$$

where Q = air displacement, cu. ft./minute ;
P_L = power dissipated in heating air, kW ;
T = permissible temperature rise of air above ambient temperature, ° C.

AERIALS AND PROPAGATION

Effective Height of Aerial

Vertical aerial with sinusoidal distribution of current:

$$h = 2l/\pi$$
$$= \lambda/2\pi \text{ for a quarter-wave aerial}$$
$$= \lambda/\pi \text{ for a half-wave aerial}$$

where h = effective height, metres;
l = length of wire, metres;
λ = wavelength, metres.

Aerial Resistance

The effective resistance is made up of two components:

(a) Radiation resistance, R_r, a fictitious resistance which governs the proportion of the total energy radiated.

(b) Loss resistance, R_L, which is accounted for by losses in conductors, the earth, dielectrics of insulators and eddy currents in metal supports and stays.

Effective Resistance

$$R_e = R_r + R_L$$

The resistance (voltage/current ratio) varies with the sinusoidal distribution of voltage and current along the aerial, and is usually referred to the current anti-node at the base of a quarter-wave aerial or at the centre of a half-wave aerial.

Radiation Resistance

$$R_r = 160\ (\pi h/\lambda)^2$$
$$= 40 \text{ ohms for a quarter-wave aerial (referred to the base)}$$
$$= 160 \text{ ohms for a half-wave aerial (referred to the centre)}$$

Power Radiated

$$P_r = I_{ae}^2 R_r = 160\ (\pi h I_{ae}/\lambda)^2$$

Power Loss

$$P_L = I_{ae}^2 R_L$$

where h = effective height of aerial, metres;
I_{ae} = r.m.s. value of aerial current, amperes.

Example 1.—*Calculate* (a) *the radiation resistance, and* (b) *the power radiated by a vertical half-wave aerial if the aerial current is* 3 *amperes.*

(a) Radiation resistance:

$$R_r = 160(\pi h/\lambda)^2$$

Substituting λ/π for h,

$$R_r = 160(1)^2 = \underline{160 \text{ ohms}}$$

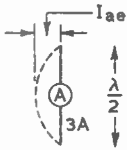


FIG. 85.—
EXAMPLE 1.

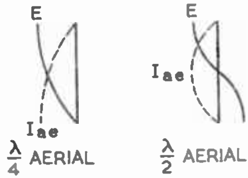


FIG. 86.—
VERTICAL AERIALS.

(*b*) Power radiated :

$$P_r \doteq I_{ae}^2 R_r$$
$$= 3^2 \times 160 = 1{,}440 \text{ watts or } \underline{1{\cdot}44 \text{ kW}}$$

Field Strength

Sommerfeld formula for ground wave at medium and low frequencies

$$\epsilon = \alpha\beta(300\sqrt{P} \cos \theta/d)$$

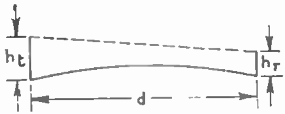


FIG. 87.—FIELD STRENGTH AT VERY HIGH FREQUENCIES.

where α = ground loss factor, dependent on earth conductivity, dielectric constant, frequency and distance;
β = directivity factor $\beta = 1$ for short aerials where field strength in vertical plane follows a cosine law;
P_r = radiated power, kW;
d = distance from transmitting aerial, km.;
ϵ = field strength, μV/metre;
θ = angle of elevation of ray.

Field strength of direct wave at very high frequencies, taking into account reflection from ground

$$\doteq 4\pi h_t h_r \epsilon_0/\lambda d^2 \text{ when } d \gg h$$

where H = field strength, μV/metre;
d = distance from transmitting aerial, metres;
h_t = elevation of transmitting aerial, metres;
h_r = elevation of receiving aerial, metres;
ϵ_0 = field strength at unity distance, mV/metre.

EXAMPLE 2.—*A broadcast transmitter delivers 20 kW to an aerial having a radiation efficiency of 65 per cent.* (a) *Estimate the field strength of the ground wave at a distance of 200 miles, if the ground loss factor, determined by means of a low-power mobile transmitter, is 0·08.* (b) *If the sky wave travels 50 per cent farther than the ground wave at an effective angle to the earth of 45°, determine its field strength, assuming the wave after reflection from the ionosphere to be reduced to 25 per cent of the incident wave.*

Power radiated :

$$P_r = 20 \times 0{\cdot}65 = 13 \text{ kW}$$

(a)

$$d = 200 \times 1{\cdot}609 = 321{\cdot}8 \text{ km.}$$
$$\alpha = 0{\cdot}08$$
$$\beta = 1$$
$$\cos\theta = \cos 0 = 1$$

Field strength of ground wave :

$$\epsilon = \alpha\beta\,(300\sqrt{P_r}\cos\theta/d)$$
$$= \frac{0{\cdot}08 \times 1 \times 300 \times \sqrt{13} \times 1}{321{\cdot}8} = \underline{0{\cdot}269\ \mu\text{V/metre}}$$

(b)

$$d = 1{\cdot}5 \times 321{\cdot}8 = 482 \text{ km.}$$
$$\cos\theta = \cos 45^\circ = 0{\cdot}707$$
$$\alpha = 1, \text{ as there are no ground losses}$$

Field strength of sky wave :

$$\epsilon = \frac{1 \times 300\sqrt{13} \times 0{\cdot}707 \times 0{\cdot}25}{482} = \underline{0{\cdot}396\ \mu\text{V/metre}}$$

Maximum Range of Direct Wave

Geometric (straight line) range :

$$d_{max.} = 3{\cdot}55(\sqrt{h_t} + \sqrt{h_r})$$

Actual range when refraction is taken into account :

$$d_{max.} = 3{\cdot}55c(\sqrt{h_t} + \sqrt{h_r})$$

where $d_{max.}$ = maximum range of reception, km. ;
c = a factor which allows for refraction of the wave over the earth's curvature. An average value is 1·3.

EXAMPLE 3.—*A V.H.F. transmitting aerial and a distant receiving aerial are both at the same height above earth. Assuming a straight-line ray path, what will be the gain in field strength at the receiver, expressed in decibels if :* (a) *the height of the transmitting aerial is increased* 50 *per cent ;* (b) *the height of both aerials is increased* 50 *per cent ?*

$$\epsilon \propto \sqrt{h_t} + \sqrt{h_r} = 2\sqrt{h_r}$$

(a) Here $h_t = 1{\cdot}5\ h_r$

Hence modified field strength :

$$\epsilon' \propto \sqrt{1{\cdot}5h_r} + \sqrt{h_r} = 2{\cdot}225\sqrt{h_r}$$
$$\therefore \quad \epsilon'/\epsilon = 2{\cdot}225\sqrt{h_r}/2\sqrt{h_r} = 1{\cdot}1125$$

Gain :

$$m = 20 \log_{10}(\epsilon'/\epsilon) = 20 \log_{10} 1{\cdot}1125 = \underline{0{\cdot}924 \text{ db}}$$

(b) Modified field strength :

$$\epsilon' \propto \sqrt{1{\cdot}5h_r} + \sqrt{1{\cdot}5h_r} = 2{\cdot}45\sqrt{h_r}$$
$$\therefore \quad \epsilon'/\epsilon = 2{\cdot}45\sqrt{h_r}/2\sqrt{h_r} = 1{\cdot}225$$

Gain :

$$m = 20 \log_{10} 1{\cdot}225 = \underline{1{\cdot}762 \text{ db}}$$

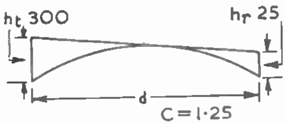


FIG. 88.—EXAMPLE 4.

EXAMPLE 4.—*A television transmitting aerial is elevated 300 ft. above earth. If the average receiving aerial is 25 ft. high, what is the geometric range for straight-line reception and the maximum range of the direct ray, if the refraction factor is 1·25 and there are no obstructions in the path?*

Geometric (straight line) range :

$$d = 3{\cdot}55(\sqrt{h_t} + \sqrt{h_r}) = 3{\cdot}55\left(\sqrt{\frac{300}{3{\cdot}281}} + \sqrt{\frac{25}{3{\cdot}281}}\right) = \underline{43{\cdot}7 \text{ km.}}$$

Maximum range for direct ray reception :

$$d_{max.} = cd = 1{\cdot}25 \times 43{\cdot}5 = \underline{54{\cdot}7 \text{ km.}}$$

Half-wave Dipole

Resonates when length is about 95 per cent of the half wavelength

$$l = 0{\cdot}475\lambda = 142{\cdot}5/f$$

where l = length of wire, metres;
λ = wavelength, metres;
f = frequency, Mc/s.

Directional Arrays

The energy gain of a directional array is the ratio of the power which would be required by a single aerial element to the power taken by the array to produce a given field strength in a required direction. Gain in decibels

$$m = 10 \log_{10} P_e/P_{ae}$$

where P_e = power which would be required by a single element;
P_{ae} = power required by array.

Field strength produced by array : $\epsilon \propto \sqrt{P_{ae}} \propto I_{ae}$

Field strength produced by each element : $\epsilon \propto \sqrt{P_{ae}/n}$

where I_{ae} = total aerial current of array;
n = number of elements in array.

EXAMPLE 5.—*What is the directional gain in field strength of an array of ten aerial elements over a single element, neglecting mutual reactionary effects between elements?*

Let power supplied to array $= P$
Then power supplied to each element $= P/10$
Field strength due to array : $\epsilon' \propto \sqrt{P}$
Field strength due to each element : $\epsilon \propto \sqrt{(P/10)}$

Gain: $$\epsilon'/\epsilon = \frac{\sqrt{P}}{\sqrt{(P/10)}} = \sqrt{10} \text{ or } 3{\cdot}162 \text{ times}$$

Gain in db:

$$m = 20 \log_{10} (\epsilon'/\epsilon) = 20 \log_{10} 3{\cdot}162 = \underline{10 \text{ db}}$$

Receiving Aerials

Voltage Developed in a Receiving Aerial

Vertical wire aerial: $E = \epsilon h$
Frame aerial: $E = 2\pi\epsilon AN/\lambda$

where
E = e.m.f. mV;
ϵ = field strength, mV/metre;
h = effective height of aerial, metres;
A = area of frame, sq. metres;
N = number of turns in frame.

Example 6.—*What voltage will be produced by a field of intensity 10 μV/metre at a wavelength of 300 metres in:* (a) *a vertical quarter-wave receiving aerial*; (b) *a frame aerial of ten turns, 2 metres square?*

(*a*) Effective height of $\lambda/4$ aerial:

$$h = \lambda/2\pi = \frac{300}{2\pi} = 47{\cdot}8 \text{ metres}$$

Voltage induced in wire:

$$E = \epsilon h = 10 \times 47{\cdot}8 = 478\ \mu\text{V or } \underline{0{\cdot}5 \text{ mV approx.}}$$

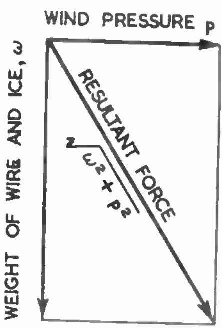
(*b*) Voltage induced in frame:

$$E = 2\pi\epsilon AN/\lambda = \frac{2\pi \times 10 \times 2^2 \times 10}{300} = \frac{8\pi}{3} = \underline{8{\cdot}38\ \mu\text{V}}$$

Mechanical Factors

The total load on a horizontal wire is made up of the weight of wire per unit length acting downwards and the wind load acting horizontally.

$$W = \sqrt{w^2 + p^2}$$



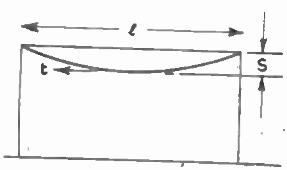


Fig. 89 (*left*).—Mechanical Factors.

Fig. 90 (*above*).—Determination of Total Load.

Wind pressure : $P = pd$

Sag : $s \doteq wl^2/8t$, neglecting effect of wind pressure

$S = Wl^2/8t$, including effect of wind pressure

Length of wire : $L = l + 8S^2/3$ approximately

where W = total load on wire, lb./ft.;
w = weight of wire, lb./ft.; *
p = wind load, lb./sq. ft. of projected area;
d = diameter of wire, ft.;
s = sag, neglecting effect of wind pressure, ft.;
S = sag, including effect of wind pressure, ft.; †
l = span, ft.;
L = length of wire, ft.; †
t = horizontal tension, lb.†

* If ice is present, w includes the weight of ice. A safety factor on the breaking load of 2 at 22° F. is normally allowed in Britain under normal loading conditions imposed by a layer of ice ⅜ in. thick radially and a wind pressure of 8 lb/sq. ft. acting horizontally.

† Length and sag vary directly and tension inversely as the ambient air temperature. They are therefore determined for the lowest temperature likely to be encountered, when the length and sag are a minimum and tension is a maximum.

Example 7.—*An aerial curtain weighing* 300 *lb. is supported from a steel wire rope* ½ *in. diameter weighing* 0·6 *lb./ft. between two masts* 400 *ft. apart, the load being uniformly distributed over the span. What must be the minimum sag at the centre of the rope if the safe working stress is not to exceed* 3 *tons under a horizontal wind load of* 30 *lb./sq. ft. of projected area?*

Weight per foot run of aerial and rope :

$$w = w_1 + w_2 = \frac{300}{400} + 0{\cdot}6 = 1{\cdot}35 \text{ lb./ft.}$$

Wind load per foot run :

$$P = pd = \frac{1/2}{12} \times 30 = 1{\cdot}25 \text{ lb./ft.}$$

Total load :

$$W = \sqrt{w^2 + P^2} = \sqrt{1{\cdot}35^2 + 1{\cdot}25^2} = 1{\cdot}84 \text{ lb./ft.}$$

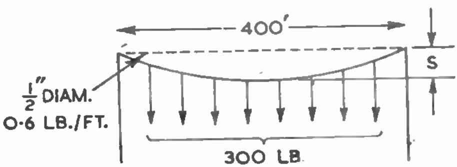


Fig. 91.—Example 7.

Minimum sag :

$$S = Wl^2/8t = \frac{1{\cdot}84 \times 400^2}{8 \times 3 \times 2{,}240} = \underline{5{\cdot}48 \text{ ft.}}$$

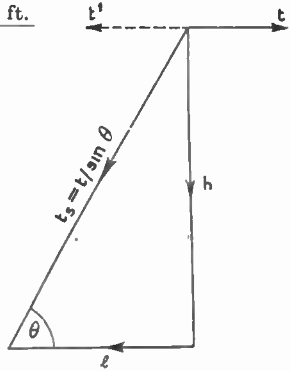


FIG. 92.—TENSION IN STAYS.

EXAMPLE 8.—*The horizontal tension in an aerial wire is 10 lb. when the sag is 7 ft. To what value can the sag be reduced if:* (a) *the maximum allowable tension is 25 lb.; and* (b) *the maximum tension is 25 lb. and the span is reduced by one quarter?* (c) *How will the sag be affected if a heavier wire of twice the diameter is used?*

(*a*) Sag varies inversely as tension

$$s \propto 1/t$$

Hence

$$s_2 = s_1(t_1/t_2) = \frac{7 \times 10}{25} = \underline{2{\cdot}8 \text{ ft.}}$$

(*b*) Sag varies as (span)^2

$$s \propto l^2$$

$$s_3 = S_2(l_2/l_1)^2 = 2{\cdot}8 \times (\tfrac{3}{4})^2 = \underline{1{\cdot}58 \text{ ft.}}$$

(*c*) Both w and t vary directly as the cross-sectional area or as the (diameter)^2. But sag varies directly as w/t. So the minimum sag will remain substantially the same.

Tension in Stays

$$\begin{aligned} t_s &= t/\sin\theta \\ &= t \operatorname{cosec}\theta \\ &= t\sqrt{h^2 + l^2}/h \end{aligned}$$

where t_s = tension in stay, lb;
t = horizontal tension in aerial wire, lb.;
θ = angle between stay and ground;
h = height of stay, ft.;
l = distance between mast and anchor point of stay, ft.

EXAMPLE 9.—*A twin wire aerial transmission line is supported on 15-ft. poles spaced 150 ft. apart. A stay is attached to the terminal pole and anchored to the ground 9 ft. from the base. The sag at the centre of the span is 9 in., and the weight of the wire is 262 lb./mile. Calculate the tension in the stay.*

Weight of wires per foot length

$$w = \frac{2 \times 262}{5{,}280} = 0{\cdot}0992 \text{ lb./ft.}$$

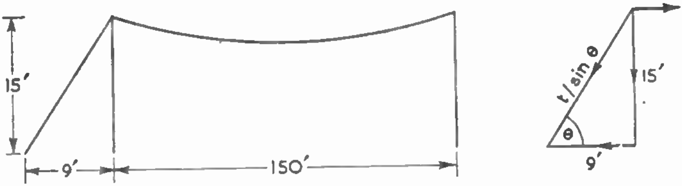


FIG. 93.—EXAMPLE 9.

Horizontal tension in wires:

$$t = wl^2/8s = \frac{0{\cdot}0992 \times 150^2}{8 \times 0{\cdot}75} = 372{\cdot}3 \text{ lb.}$$

Tension in stay:

$$t_s = t/\sin\theta$$

where $\theta = \text{arc tan }(15/9) = \text{arc tan } 1{\cdot}666;$

$$\sin\theta = 0{\cdot}8572.$$

Hence $$t = \frac{372{\cdot}3}{0{\cdot}8572} = \underline{435 \text{ lb.}}$$

RADIO-FREQUENCY TRANSMISSION LINES

Line Constants

The constants which determine the electrical characteristics of a transmission line are classified as primary and secondary (derived) constants.

PRIMARY CONSTANTS

L_0 = inductance, henrys/metre
C_0 = capacitance, farads/metre
R_0 = resistance, ohms/metre
G_0 = leakance, mhos/metre

SECONDARY CONSTANTS

Z_0 = characteristic impedance, ohms $= \sqrt{L_0/C_0}$

a = attenuation constant, nepers/metre * $= \frac{1}{2}R/Z_0 + \frac{1}{2}GZ_0$

B = phase constant, radians/metre $= \omega\sqrt{L_0C_0}$

v = velocity of propagation, metres/second $= 1/\sqrt{L_0C_0}$

* 1 neper = 8·68591 db.

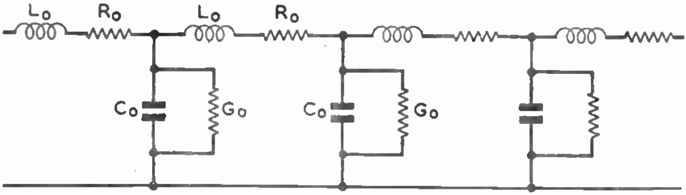


FIG. 94.—EQUIVALENT LINE

Characteristic Impedance

The characteristic impedance is the voltage/current ratio at any position along a line which is long enough for reflections to be negligible, or which is terminated at the far end in an impedance equal to the characteristic impedance so as to prevent reflection.

At Radio Frequencies

Twin-wire line (Fig. 95(a)):

$$Z_0 = 276 \log_{10} (d/r)$$

Four-wire symmetrical line (Fig. 95(b)):

$$Z_0 = 138 \log_{10} (d/r) - 20{\cdot}8$$

Co-axial line (Fig. 95(c)):

$$Z_0 = (138/\sqrt{k}) \log_{10} (r_2/r_1)$$

where d = distance between centres of conductors;
r = radius of wire;
r_1 = outer radius of inner conductor;
r_2 = inner radius of outer conductor;
k = average dielectric constant of insulating medium between conductors (= 1 for air spacing).

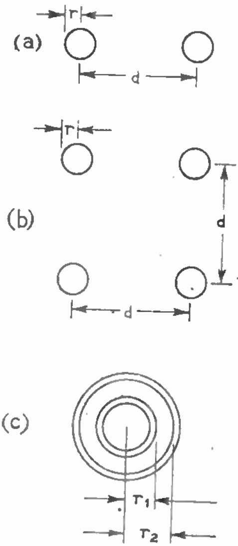


FIG. 95.—CHARACTERISTIC IMPEDANCE.

Sending-end Impedance

The impedance Z_1 at the sending end of a line terminated in any impedance Z_2:

$$Z_1 = Z_0 \frac{Z_2 + jZ_0 \tan (2\pi l/\lambda)}{Z_0 + jZ_2 \tan (2\pi l/\lambda)}$$

If the line is short-circuited at the far end:

$$Z_1 = jZ_0 \tan (2\pi l/\lambda)$$

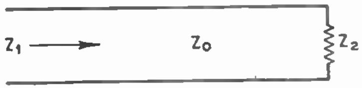


FIG. 96 (*above*).—EXAMPLE 1.

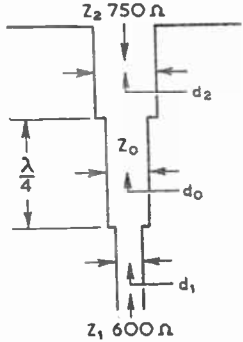


FIG. 97 (*right*).—MATCHING BY QUARTER WAVE LINE.

If the line is open-circuited at the far end :

$$Z_1 = -j\frac{Z_0}{\tan(2\pi l/\lambda)}$$

If the line is a quarter wavelength long : $2\pi l/\lambda = 90°$.

Two different impedances Z_1 and Z_2, may be matched by a quarter-wave line of impedance Z_0 connected between them, when

$$Z_1 = Z_0^2/Z_2$$

$$Z_0 = \sqrt{Z_1 Z_2}$$

EXAMPLE 1.—*A 600-ohm twin wire line with conductors spaced 10 in. apart is to be matched to a dipole having a terminal resistance of 750 ohms by means of a quarter-wave matching section at a frequency of 20 Mc/s. Calculate the length of the matching section and the spacing between conductors if the same gauge wire is used throughout.*

(*a*) Wavelength at 20 Mc/s :

$$\lambda = \frac{3 \times 10^8}{20 \times 10^6} = 15 \text{ metres}$$

$$\lambda/4 = \frac{15}{4} = 3{\cdot}75 \text{ metres}$$

(*b*) Characteristic impedance of matching section :

$$Z_0 = \sqrt{Z_1 Z_2} = \sqrt{600 \times 750}$$

$$= 100\sqrt{45} = \underline{670} \text{ ohms}$$

(*c*) Spacing of matching section :

$$Z_0 = 276 \log_{10}(d_0/r)$$

$$d_0/r = \text{antilog}(Z_0/276) \quad \ldots\ldots \quad (1)$$

Also $$d_1/r = \text{antilog}(Z_1/276) \quad \ldots\ldots \quad (2)$$

Dividing (1) by (2),

$$d_0/d_1 = \frac{\text{antilog}\,(Z_0/276)}{\text{antilog}\,(Z_1/276)}$$

$$\frac{d_0}{10} = \frac{\text{antilog}\,(670/276)}{\text{antilog}\,(600/276)} = \frac{269 \cdot 2}{149 \cdot 6}$$

$$d_0 = \underline{18 \text{ in.}}$$

Current, Voltage and Power

When the line is correctly terminated in Z_0:

R.m.s. current of carrier wave $I_c = \sqrt{P_c/Z_0}$

R.m.s. voltage of carrier wave $E_c = \sqrt{P_c Z_0}$

Peak voltage of amplitude-modulated wave $E_{mp} = \sqrt{2}E_c(1 + m)$

Maximum voltage on line with a standing wave $E_{max.} = \sqrt{n}E_{mp}$

Standing wave ratio

$$\begin{aligned} n &= I_{max.}/I_{min.} &&= Z_0/R_{min.} \\ &= E_{max.}/E_{min.} &&= R_{max.}/Z_0 \\ &= \sqrt{R_{max.}/R_{min.}} &&= R_L/Z_0 \text{ if } R_L/Z_0 > 1 \\ & &&= Z_0/R_L \text{ if } R_L/Z_0 < 1 \end{aligned}$$

where P_c = carrier power;
m = modulation factor (= 1 for 100 per cent modulation);
$E_{max.}$ = maximum voltage on line at point of maximum resistance;
$E_{min.}$ = minimum voltage on line at point of minimum resistance;
$R_{max.}$ = resistance of line at point of minimum current;
$R_{min.}$ = resistance of line at point of maximum current;
$I_{max.}$ = maximum current in line at point of minimum resistance;
$I_{min.}$ = minimum current in line at point of maximum resistance;
R_L = load resistance;
n = standing wave ratio.

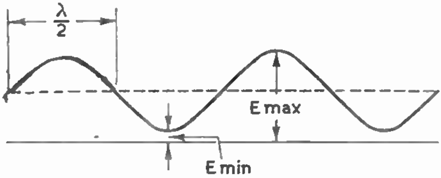
EXAMPLE 2.—*An amplitude-modulated transmitter of* 10 *kW carrier power feeds a balanced twin-wire transmission line having a characteristic impedance of* 600 *ohms. What is the maximum peak voltage occurring on the line when the transmitter is fully modulated:* (a) *when the line is correctly terminated; and* (b) *if the standing-wave ratio is* 1·4/1*?*

R.m.s. carrier voltage on line when terminated in Z_0:

$$E_c = \sqrt{P_c Z_0} = \sqrt{10{,}000 \times 600} = 2{,}450 \text{ volts}$$

Peak voltage of amplitude-modulated wave:

$$\begin{aligned} E_{mp} &= \sqrt{2}E_c(1 + m) = \sqrt{2} \times 2{,}450 \times (1 + 1) \\ &= 6{,}926 \text{ volts or } \underline{6 \cdot 93 \text{ kV}} \end{aligned}$$



$n = E_{max}/E_{min} = 1{\cdot}4$

FIG. 98.—EXAMPLE 2.

When the standing wave ratio is 1·4/1

$$E'_{mp} = \sqrt{n}E_{mp} = \sqrt{1{\cdot}4} \times 6{\cdot}93 = \underline{8{\cdot}2 \text{ kV}}$$

EXAMPLE 3.—*A co-axial cable having a characteristic impedance of 100 ohms was subjected to voltage tests at radio frequency and found to flash over at the following peak voltages: 5·2 kV at 1 Mc/s; 3·95 kV at 10 Mc/s; 3·5 kV at 20 Mc/s. If the cable is to be rated at a safety factor of 2/1, what is the maximum r.m.s. current and power it can transmit safely at each frequency?*

Maximum safe r.m.s. voltage with a safety factor of 2/1:

$$E_{max.} = E_p/2\sqrt{2}$$

Maximum safe r.m.s. current:

$$I_{max.} = E_p/2\sqrt{2}Z_0$$

At 1 Mc/s $= \dfrac{5{,}200}{2\sqrt{2} \times 100} = \underline{18{\cdot}4 \text{ amperes}}$

At 10 Mc/s $= 18{\cdot}4 \times \dfrac{3{\cdot}95}{5{\cdot}2} = \underline{14{\cdot}0 \text{ amperes}}$

At 20 Mc/s $= 18{\cdot}4 \times \dfrac{3{\cdot}5}{5{\cdot}2} = \underline{12{\cdot}4 \text{ amperes}}$

Max. safe power:

$$P_{max.} = I_{max.}^2 Z_0$$

At 1 Mc/s $= 18{\cdot}4^2 \times 100 = 33{,}790$ watts or $\underline{33{\cdot}8 \text{ kW}}$

At 10 Mc/s $= 14{\cdot}0^2 \times 100 = 19{,}530$ watts or $\underline{19{\cdot}5 \text{ kW}}$

At 20 Mc/s $= 12{\cdot}4^2 \times 100 = 15{,}320$ watts or $\underline{15{\cdot}3 \text{ kW}}$

EXAMPLE 4.—*A 320-ohm, five-wire cage transmission line consists of four earthed outer conductors in square formation and a central live conductor. For what peak voltage should the line be insulated if it is to transmit 100 kW carrier power at 100 per cent modulation, allowing for a standing wave ratio of 1·5? Compare this with the peak voltage to earth on a 600-ohm twin-wire line carrying the same amount of power.*

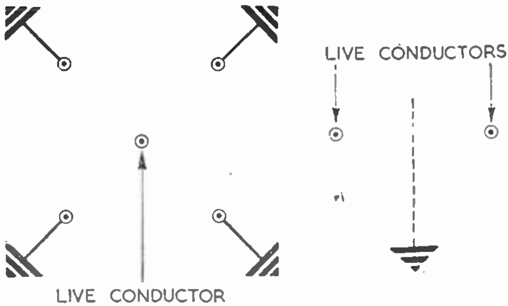


FIG. 99.—EXAMPLE 4.

Peak modulation voltage to earth on a five-wire line :

$$E_{mp} = \sqrt{2}E_c(1 + m)$$

But $\qquad E_c = \sqrt{P_c Z_0}$

$$E_{mp} = \sqrt{2P_c Z_0}(1 + m) = \sqrt{2 \times 10^5 \times 320}(1 + 1)$$
$$= 16{,}000 \text{ volts}$$
$$E_{\max.} = \sqrt{n}E_{mp} = \sqrt{1{\cdot}5} \times 16{,}000$$
$$= 19{,}600 \text{ volts or } \underline{19{\cdot}6 \text{ kV}}$$

Peak modulation voltage between conductors of a two-wire line :

$$E_{\max.} = 19{,}600 \times \sqrt{\frac{600}{320}} = \underline{26{,}840 \text{ volts}}$$

The voltage between each conductor and earth is half this, i.e.,

$$13{,}420 \text{ volts or } \underline{13{\cdot}42 \text{ kV}}$$

Attenuation

Attenuation in decibels :

$$\alpha = 10 \log_{10} (P_1/P_2)$$
$$= 20 \log_{10} (E_1/E_2) \text{ if input and output impedances are equal}$$

EXAMPLE 5.—*Two co-axial cables suitable for television reception have respectively attenuations of 6·0 db/100 ft. and 2·5 db/100 ft. at 45 Mc/s. What will be the percentage gain in power delivered by the aerial to the receiver in a length of 25 ft. if the smaller cable is replaced by the larger ?*

The smaller cable has an attenuation of 6·0 db/100 ft.

Attenuation in 25 ft. of the smaller cable :

$$\alpha = \frac{6{\cdot}0 \times 25}{100} = 1{\cdot}5 \text{ db}$$

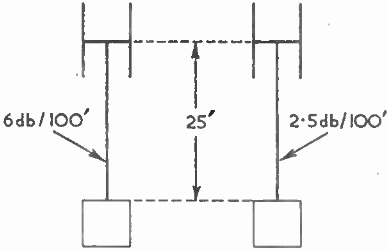


FIG. 0.—EXAMPLE 5.

Ratio power input/power output :

$$P_1/P_2 = \text{antilog } (1{\cdot}5/10) = 1{\cdot}413$$
$$P_2/P_1 = 0{\cdot}708$$

Attenuation in 25 ft. of the larger cable

$$= \frac{2{\cdot}5 \times 15}{100} = 0{\cdot}625 \text{ db}$$

$$P_1/P_2 = \text{antilog } (0{\cdot}625/10) = 1{\cdot}154$$
$$P_2/P_1 = 0{\cdot}866$$

Gain in power

$$= \frac{0{\cdot}866}{0{\cdot}708} = 1{\cdot}223 \text{ times or } \underline{22{\cdot}3 \text{ per cent gain}}$$

Variation of Attenuation with Frequency

Twin open-wire line :

$$\alpha = c_1\sqrt{f}$$

Co-axial line :

$$\alpha = c_2\sqrt{f/r_2}$$

where P_1 = power input;
P_2 = power output;
E_1 = input voltage;
E_2 = output voltage;
c_1c_2 = attenuation factors, depending on conductor resistance, insulation resistance and radiation losses. Average values are $c_1 = 0{\cdot}33$ and $c_2 = 0{\cdot}27$ when α is in db/km. and r_2 = inner radius of outer conductor, cm.;
f = frequency, Mc/s.

EXAMPLE 6.—*A balanced twin-wire line has an attenuation of* 1·5 *db/kilometre at a frequency of* 10 *Mc/s. Calculate the attenuation in a line* 400 *metres long at a frequency of* 30 *Mc/s.*

Attenuation at a frequency f_1: $\alpha_1 \doteq c\sqrt{f_1}$

Attenuation at a frequency f_2: $\alpha_2 = c\sqrt{f_2}$

Hence $\alpha_2/\alpha_1 = \sqrt{f_2/f_1}$

Attenuation per kilometre at 30 Mc/s: $\dfrac{\alpha_2}{1 \cdot 5} = \sqrt{\dfrac{30}{10}} = 2 \cdot 6$ db/km.

Attenuation in a length of 400 metres: $\alpha_{400} = \dfrac{400 \times 2 \cdot 6}{1{,}000} = \underline{1 \cdot 04 \text{ db}}$

Example 7.—*A master oscillator delivers a stabilized radio-frequency output of* 1*-watt to a transmitter through a co-axial cable* 50 *ft. long. What will be the power output to the transmitter at* 5 *and* 20 *Mc/s if the cable has an attenuation of* 1·3 *db/*100 *ft. at* 10 *Mc/s?*

Since $\alpha \propto \sqrt{f}$, attenuation at 5 Mc/s:

$$\alpha_2/\alpha_1 = \sqrt{f_2/f_1}$$

$$\alpha_2 = 1 \cdot 3\sqrt{5/10} = 0 \cdot 92 \text{ db/100 ft. or } 0 \cdot 46 \text{ db/50 ft.}$$

Power output at 5 Mc/s:

Since $\alpha = 10 \log_{10} (P_1/P_2)$

$$P_2 = P_1/\text{antilog } (\alpha/10) = 1/\text{antilog } 0 \cdot 046 = \underline{0 \cdot 9 \text{ watt}}$$

Attenuation at 20 Mc/s

$$= 1 \cdot 3\sqrt{20/10} = 1 \cdot 3 \times 1 \cdot 414$$

$$= 1 \cdot 83 \text{ db/100 ft. or } 0 \cdot 915 \text{ db/50 ft.}$$

Power output at 20 Mc/s:

$$P_2 = P_1/\text{antilog } (\alpha/10) = 1/\text{antilog } 0 \cdot 0915 = \underline{0 \cdot 81 \text{ watt}}$$

POWER SUPPLY

A.C. Supplies

The power, voltage and current relationships in the load on a single-phase or three-phase supply are :

	SINGLE-PHASE SUPPLY	BALANCED THREE-PHASE SUPPLY
Load power, kW	$EI \cos \phi$	$\sqrt{3}EI \cos \phi$
Load, VA	EI	$\sqrt{3}EI$
Line current	$P/E \cos \phi$	$P/\sqrt{3}E \cos \phi$
Line-to-neutral voltage . .	—	$E/\sqrt{3}$

where

E = line voltage ;
I = line current ;
P = power, watts ;
$\cos \phi$ = power factor.

EXAMPLE 1.—*A transmitter takes* 10 *kW with a line current of* 16 *amperes from a* 400*-volt three-phase transformer having an efficiency of*

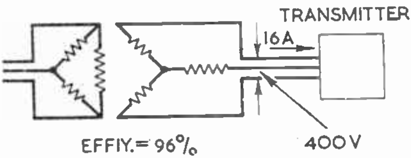


FIG. 101.—EXAMPLE 1.

96 *per cent. Find the power factor of the load and the kVA input to the transformer, assuming the phases to be equally loaded.*

Power factor

$$\cos \phi = P/\sqrt{3}EI = \frac{10{,}000}{\sqrt{3} \times 400 \times 16} = \underline{0{\cdot}904}$$

kVA input to transformer :

$$\text{kVA} = P/\eta \cos \phi = \frac{10}{0{\cdot}96 \times 0{\cdot}904} = \underline{11{\cdot}5 \text{ kVA}}$$

Measurement of A.C. Power and Power Factor

The power taken by an installation can be measured by the three-voltmeter or three-ammeter method without the use of a watt-meter. In the voltmeter method a non-inductive resistance is connected in series, and in the ammeter method in parallel with the load. Three voltmeters or ammeters are connected as shown in Fig. 102 (*a*) and (*b*) and their readings taken.

	THREE-VOLTMETER METHOD	THREE-AMMETER METHOD
Power, P . .	$(V_3^2 - V_1^2 - V_2^2)/2R$	$R(I_3^2 - I_1^2 - I_2^2)/2$
Power factor, $\cos \phi$	$(V_3^2 - V_1^2 - V_2^2)/2V_1V_2$	$(I_3^2 - I_1^2 - I_2^2)/2I_1I_2$
Current, I . .	V_1/R	—

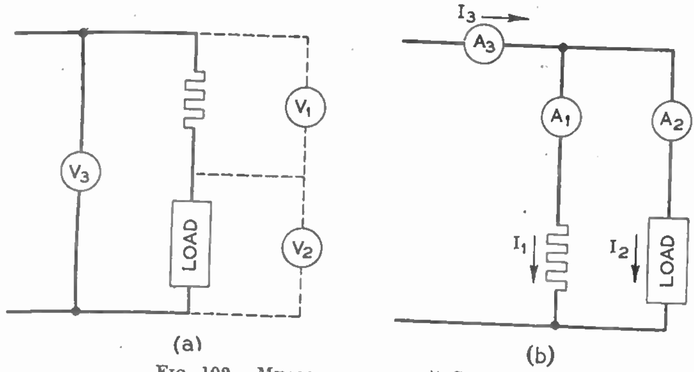


FIG. 102.—MEASUREMENT OF A.C. POWER.

where V_1 = voltage across non-inductive resistance;
V_2 = voltage across load;
V_3 = voltage across load and resistance together;
I_1 = current in non-inductive resistance;
I_2 = current in load;
I_3 = current in load and resistance together;
R = resistance in series or parallel, ohms.

EXAMPLE 2.—*The A.C. power input to a transmitter was measured by the three-voltmeter method, which gave the following readings: 81 volts across a non-inductive resistance of 3 ohms in series, 152 volts across the load and 230 volts across the load and resistance together. Calculate:* (a) *the power of the transmitter;* (b) *the load current; and* (c) *the power factor of the load.*

(*a*) Power input to transmitter:

$$P = (V_3^2 - V_1^2 - V_2^2)/2R = \frac{230^2 - 81^2 - 152^2}{2 \times 3} = \underline{3{,}873 \text{ watts}}$$

(*b*) Load current:

$$I = V_1/R = 81/3 = \underline{27 \text{ amperes}}$$

(*c*) Power factor of load:

$$\cos\phi = P/(\text{kVA}) = P/V_2I$$

$$= \frac{3{,}873}{152 \times 27} = \underline{0{\cdot}944}$$

Power-factor Correction

The usual method of improving a lagging power factor is to add to the existing load a capacitor which will take a leading current to neutralize the lagging current.

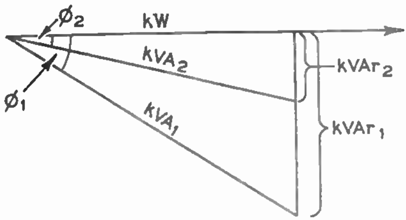


FIG. 103.—POWER FACTOR CORRECTION.

Capacitor kVA required to raise the power factor from a value $\cos \phi_1$ to $\cos \phi_2$:

$$\begin{aligned} \text{kVA} &= (\text{kVAr})_1 - (\text{kVAr})_2 \\ &= (\text{kVA})_1 \sin \phi_1 - (\text{kVA})_2 \sin \phi_2 \end{aligned}$$

where
ϕ_1 = angle of lag before correction;
ϕ_2 = angle of lag after correction;
kVA_1 = kVA of load before correction;
kVA_2 = kVA of load after correction;
kVAr_1 = reactive kVA of load before correction;
kVAr_2 = reactive kVA of load after correction.

EXAMPLE 3.—*Capacitors are to be installed to improve the power factor of a station taking 200 kW at 0·87 power factor lagging. Find:* (a) *the kVA rating of capacitors necessary to raise the power factor to 0·95 lagging; and* (b) *the reduction of load kVA resulting from the correction.*

Present kVA of station:

$$\text{kVA}_1 = P/\cos \phi_1 = \frac{200}{0{\cdot}87} = 230 \text{ kVA}$$

Reactive kVA:

$$\text{kVAr}_1 = (\text{kVA}_1) \sin \phi_1 = 230 \times 0{\cdot}4924 = 113{\cdot}3 \text{ kVA}$$

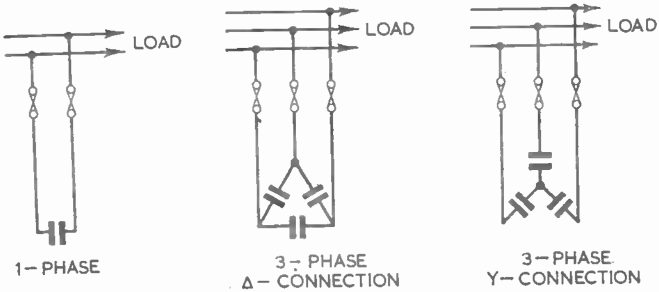
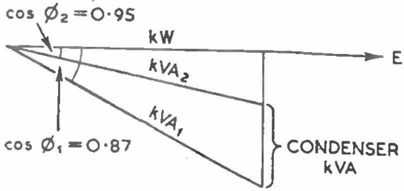


FIG. 104.—CAPACITORS FOR POWER FACTOR CORRECTION.

FIG. 105.—EXAMPLE 3.



kVA after correction :

$$kVA_2 = P/\cos\phi_2 = \frac{200}{0{\cdot}95} = 210{\cdot}5 \text{ kVA}$$

Reactive kVA after correction :

$$kVAr_2 = (kVA_2)\sin\phi_2 = 210{\cdot}8 \times 0{\cdot}3123 = 65{\cdot}8 \text{ kVA}$$

Capacitor kVA

$$= (kVAr_1) - (kVAr_2) = 113{\cdot}3 - 65{\cdot}8 = \underline{47{\cdot}5 \text{ kVA}}$$

Load reduction

$$= (kVA_1) - (kVA_2) = 230 - 210{\cdot}5 = \underline{19{\cdot}5 \text{ kVA}}$$

Engine Generators

Overall efficiency :

$$\eta = \eta_e \eta_g$$

where η_e = thermodynamic efficiency of engine;
η_g = electromechanical efficiency of generator.

Average thermodynamic efficiencies of engines :

Diesels of 10–50 b.h.p.	28–30 per cent
Diesels of 50 b.h.p. per cylinder	30–33 per cent
Diesels of 100 b.h.p. per cylinder	33–36 per cent
Paraffin engines	17–22 per cent
Four-stroke petrol engines	15–20 per cent

Electromechanical efficiencies of generators :

Alternators of 5–50 kVA	81–90 per cent
Alternators of 50–500 kVA	90–94·5 per cent
D.C. generators of 5–50 kW	82–91 per cent
D.C. generators of 50–500 kW	91–94 per cent

Equivalent Power

The energy value of fuel is specified by its calorific value in B.Th.U./lb. or B.Th.U./gal., and the equivalent mechanical h.p. or electrical power in kW is determined from the thermal power in B.Th.U./sec.

1 B.Th.U./sec. = 1,058 watts

Average calorific values of liquid fuel :

Diesel oil	18,400–19,500 B.Th.U./lb.
Paraffin	22,000 B.Th.U./lb.
Petrol	18,000 B.Th.U./lb.

EXAMPLE 4.—*Find the generating cost per unit (kWH) of electricity of a Diesel-alternator, given the following data :*

Cost of fuel oil	1*s. per gal.*
Calorific value	19,000 *B.Th.U./lb.*
Specific gravity	0·9
Overall efficiency of engine . . .	32 *per cent*
Efficiency of generator	91 *per cent*

1 B.Th.U./second = 1,058 watts

$$\text{Hence 1 kWH} = \frac{3{,}600 \times 1{,}000}{1{,}058} = 3{,}400 \text{ B.Th.U.}$$

Number of B.Th.U. required to generate 1 kWH with engine and generator efficiencies η_e and η_a :

$$\text{B.Th.U.} = \frac{3{,}400}{\eta_e \eta_a} = \frac{3{,}400}{0{\cdot}32 \times 0{\cdot}95} = 11{,}190 \text{ B.Th.U.}$$

1 gal. water weighs 10 lb.
∴ 1 gal. fuel oil of specific gravity 0·9 weighs 10 × 0·9 = 9 lb.
Hence 1 gal. fuel oil produces 9 × 19,000 B.Th.U.
No. of gal. required to produce 11,190 B.Th.U.

$$= \frac{11{,}190}{9 \times 19{,}000} = 0{\cdot}0654 \text{ gal.}$$

Cost per kWH at 1*s*. per gal. = 0·0654 × 12 = 0·785*d*.

Total Power Requirement of a Transmitter

The total kVA input is the vector sum of the powers in kW and the reactive kVA (kVAr) of the individual pieces of apparatus.

Power input to each load : $P = (\text{kVA}) \cos \phi$

Reactive kVA of each load : $\text{kVAr} = P \tan \phi = (\text{kVA}) \sin \phi$

Total kVA of combined loads $= \sqrt{(\Sigma P)^2 + (\Sigma \text{kVAr})^2}$

Overall power factor $= \Sigma(\text{kw})/\Sigma(\text{kVA})$

A.C. RECTIFIERS AND SMOOTHING FILTERS

Rectification Factors

Table 9 gives the principal factors used in the design of polyphase rectifiers. They are given in terms of the average D.C. output voltage, neglecting the voltage drops in the rectifier and filter choke, which must be taken into account separately.

TABLE 9.—POLYPHASE RECTIFIER FACTORS

Circuit	*R.m.s. Secondary Voltage*	*Inverse Peak Voltage*	*Frequency of Ripple Voltage*	*R.m.s. Ripple Voltage*
Single-phase full wave	1·11 (half section)	3·14	$2f_0$	0·483
Single-phase, full-wave bridge	1·11 (whole section)	1·57	$2f_0$	0·483
Three-phase half wave	0·855	2·09	$3f_0$	0·188
Three-phase, half-wave interconnected	0·95 (0·493 half leg)	2·09	$3f_0$	0·183
Three-phase, full-wave double star	0·74 (phase neutral)	2·09	$6f_0$	0·042
Three-phase, full-wave single star	0·428 (phase neutral)	1·05	$6f_0$	0·042

If the circuit voltage drops are taken into account, the r.m.s. secondary voltage of the transformer

$$E_2 = k_c(E_0 + E_a + E_L)$$

where E_0 = average output D.C. voltage;
E_a = voltage drop in rectifier element;
E_L = voltage drop in filter choke;
k_c = voltage conversion factor given in column 2.

EXAMPLE 1.—*Calculate the secondary voltage between outers and the secondary kVA of a transformer for a three-phase, half-wave rectifier to deliver* 2·5 *amperes at* 5,000 *volts D.C. The voltage drop in the rectifier is* 15 *volts, and the resistance of the filter choke* 48 *ohms.*

Voltage drop in choke :

$$E_L = I_0R_L = 2{\cdot}5 \times 48 = 120 \text{ volts}$$

Voltage conversion factor :

$$k_c = 0{\cdot}855$$

Transformer secondary voltage between outers :

$$E_2 = k_c(E_0 + E_a + E_L) = 0{\cdot}855\ (5{,}000 + 15 + 120)$$
$$= \underline{4{,}390 \text{ volts r.m.s.}}$$

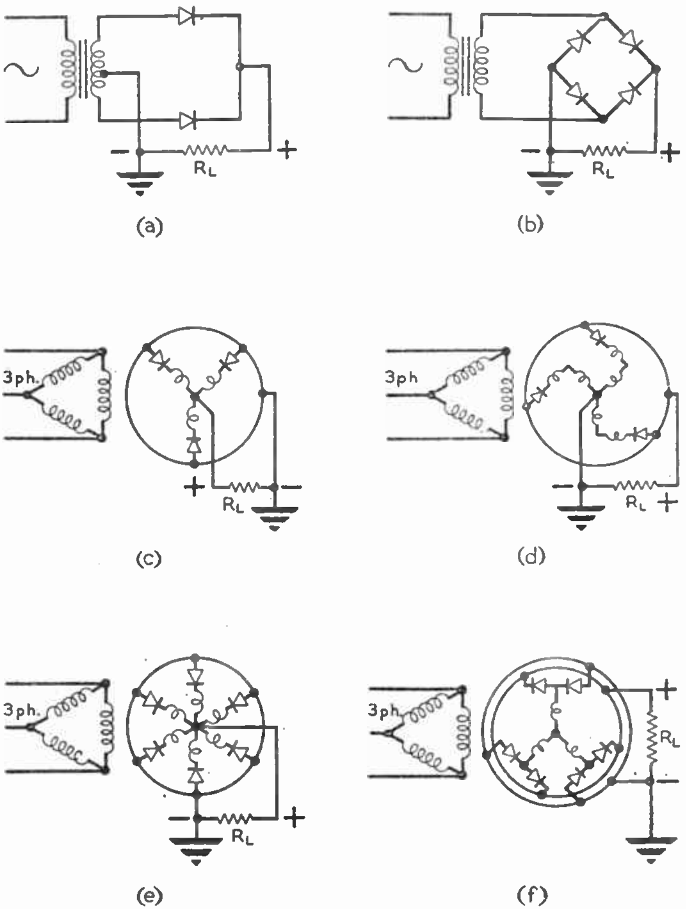


FIG. 106.—(a) SINGLE-PHASE, FULL WAVE; (b) SINGLE-PHASE, FULL-WAVE BRIDGE; (c) THREE-PHASE, HALF WAVE; (d) THREE-PHASE, HALF-WAVE INTERCONNECTED STAR; (e) THREE-PHASE, FULL-WAVE DOUBLE Y; (f) THREE-PHASE, FULL-WAVE SINGLE Y.

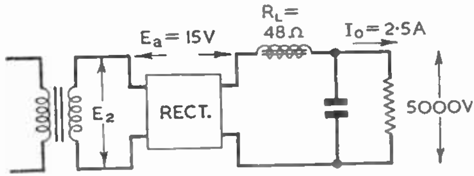


FIG. 107.—EXAMPLE 1.

Secondary kVA

$$= \sqrt{3}E_2I_0/1{,}000$$

$$= \frac{\sqrt{3} \times 4{,}390 \times 2{\cdot}5}{1{,}000} = \underline{19\ \text{kVA}}$$

Smoothing Filters

RIPPLE REDUCTION FACTOR

Input capacitor alone (Fig. 108 (*a*)):

$$\alpha_1 = 1/\sqrt{2}\pi f_r C_0 R_L$$

Choke input filter (L section) (Fig. 108 (*b*)):

$$\alpha_2 = 1/(\omega^2{}_r L_1 C_1 - 1)^n$$

Choke input filter multi-stage (Fig. 108 (*c*)):

$$\alpha_2 = 1/(\omega^2{}_r L_1 C_1 - 1)^n$$

Choke input filter with input capacitor (Fig. 108 (*d*)):

$$\alpha_3 = \alpha_1\alpha_2$$

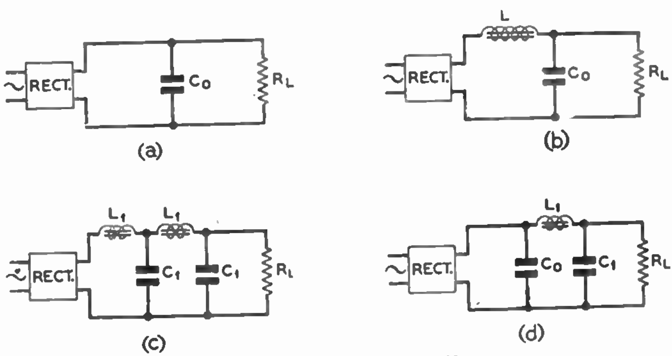


FIG. 108.—RIPPLE REDUCTION FACTOR.

where f_r = frequency of ripple voltage (see column 4 of rectification factors);
R_L = load resistance, ohms;
C_0 = capacitance of input capacitor, farads;
L_1 = inductance of choke, henrys;
C_1 = capacitance of output capacitor, farads;
n = number of similar filter sections in cascade;
ω = $2\pi \times$ ripple frequency.

EXAMPLE 2.—*A three-phase half-wave rectifier fed from 50-c/s mains delivers a D.C. output at 500 volts. The smoothing filter consists of a 30-henry choke followed by a 10-μF capacitor. Calculate the output ripple voltage.*

Ripple frequency:

$$f_r = 3 \times 50 \quad = 150 \text{ c/s}$$
$$\omega = 2\pi \times 150 = 942$$

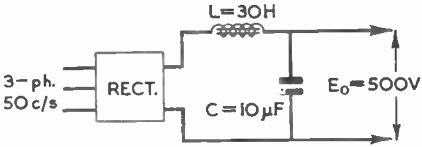


FIG. 109.—EXAMPLE 2.

Initial ripple-reduction factor:

$$\alpha_0 = 0{\cdot}188$$

Ripple-reduction factor of filter:

$$\alpha_1 = 1/(\omega_r^2 L_1 C_1 - 1) = 1/(\omega_r^2 L_1 C_1) \text{ approx.}$$
$$= \frac{1}{942^2 \times 30 \times 10 \times 10^{-6}} = 0{\cdot}00376$$

Total ripple reduction:

$$\alpha_t = \alpha_0 \alpha_1 = 0{\cdot}188 \times 3{\cdot}76 \times 10^{-3} = 7{\cdot}06 \times 10^{-4}$$

Output ripple voltage:

$$E = \alpha_t E_0 = 7{\cdot}06 \times 10^{-4} \times 500 = \underline{0{\cdot}353 \text{ volt}}$$

EXAMPLE 3.—*How much will the output ripple voltage of a single stage L-section filter be reduced by:* (a) *doubling the capacitance;* (b) *adding a similar filter section in cascade;* (c) *employing three-phase, full-wave rectification instead of single-phase, full-wave?*

(*a*) Since $\alpha \propto 1/C$,

$$\alpha_2/\alpha_1 = \frac{1/2C}{1/C} = \tfrac{1}{2}$$

∴ The ripple voltage will be halved.

(b) $$\alpha \propto (1/LC)^2$$

$\therefore$ a two-stage filter will square the ripple reduction factor.

(c) The ratio

$$\frac{\text{Initial reduction for single-phase full wave}}{\text{Initial reduction for three-phase full wave}} = \frac{0{\cdot}483}{0{\cdot}042} = 11{\cdot}5$$

i.e., the ripple voltage will be reduced 11·5 times.

Critical Filter Inductance

Minimum value of choke inductance to ensure that the output voltage does not fall below the average value of the A.C. voltage wave:

$$L_c = R_L(\sqrt{2k_c} - 1)/\omega_r$$

RECTIFIER EFFICIENCY:

$$\eta = 100\, E_0/(E_0 + E_a)$$

VOLTAGE REGULATION OF RECTIFIER:

$$v_r = 100(E_{NL} - E_L)/E_L$$

where
E_0 = output D.C. voltage;
E_a = voltage drop in rectifier;
E_{NL} = output voltage at no load;
E_L = output voltage at full load.

EXAMPLE 4.—*A* 1,000-*volt rectifier feeds an amplifier taking* 500 *mA at* 750 *volts through a series resistance. The voltage regulation of the rectifier alone is* 15 *per cent from full load to no load. Determine:* (a) *the series resistance;* (b) *the watts dissipated in the resistance;* (c) *the overall voltage regulation;* (d) *the efficiency of the rectifier at full load; and* (e) *the efficiency of the combination.*

(a) Series resistance:

$$R_s = \frac{\text{Voltage drop in } R}{I_0} = \frac{1{,}000 - 750}{0{\cdot}5} = \underline{500 \text{ ohms}}$$

(b) Watts dissipated in resistance:

$$P = I_0^2 R_s = 0{\cdot}5^2 \times 500 = \underline{125 \text{ watts}}$$

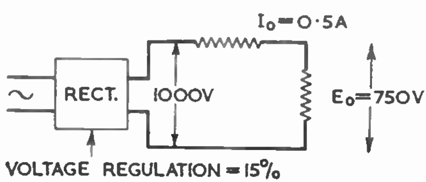


FIG. 110.—EXAMPLE 4.

Voltage rise across rectifier at no load	$= 0{\cdot}15 \times 1{,}000 =$	150 volts
Voltage rise across resistor at no load	$=$	250 volts
Total voltage rise at no load		400 volts

$$\text{Overall voltage regulation} = \frac{400 \times 100}{1{,}000} = \underline{40 \text{ per cent}}$$

(*d*) Efficiency of rectifier alone :

$$\eta_r = 100\, E_0/(E_0 + E_a) = \frac{100 \times 750}{750 \times 150} \doteq \underline{83{\cdot}3 \text{ per cent}}$$

(*e*) Efficiency of combination :

$$\eta_t = 100\, E_0/(E_0 + E_a + E_r) = \frac{100 \times 750}{750 + 150 + 250} = \underline{65{\cdot}2 \text{ per cent}}$$

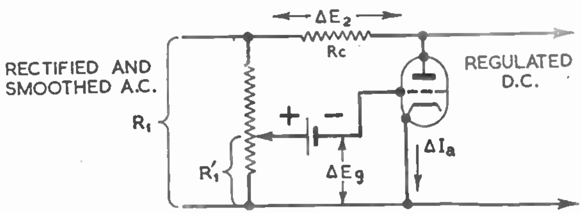


FIG. 111.—THERMIONIC VOLTAGE REGULATOR.

Thermionic Voltage Regulators

The general method of determining the condition for effective voltage regulation is illustrated by the following procedure for the circuit shown in Fig. 111.

Let $\Delta E_c =$ compensating voltage change;
$\Delta E_2 =$ supply voltage variation;
$\Delta I_a =$ anode current change in control valve;
$R_c =$ series compensating resistance;
$R_1 =$ total resistance of voltage divider;
$R_1' =$ resistance of tapped portion of divider;
$g =$ mutual conductance of valve, amperes/volt.

COMPENSATING VOLTAGE

$$\Delta E_c = \Delta I_a \,.\, R_c$$

But
$$\Delta I_a = \Delta E_g \,.\, g$$
$$= \Delta E_2(R_1'(R_1) \,.\, g$$

Hence
$$\Delta E_c = \Delta E_2 \,.\, g R_c R_1'/R_1$$

Effective regulation is achieved when $\Delta E_c = \Delta E_2$ or $\Delta E_c/\Delta E_2 = 1$; i.e., when
$$R_c = R_1/gR_1'$$

For a given anode current change ΔI_a obtainable, the range of voltage control

$$\Delta E_2 = \Delta I_a \,.\, R_c$$

EXAMPLE 5.—*Calculate the value of series compensating resistance required in the voltage compensating circuit of Fig.* 112, *and the maximum*

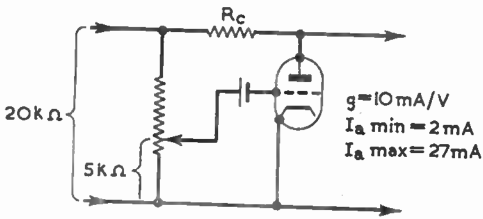


FIG. 112.—EXAMPLE 5.

variation of supply voltage over which control will be effective. The anode current change obtainable is 2–27 *mA*, g = 10 *mA/volt, and the resistance of the voltage divider is* 20 *k*Ω *tapped a quarter of the way up.*

At quarter tapping $R_1/R_1' = 20/5 = 4$ kΩ

Compensating resistance :

$$R_c = R_1/gR_1' = \frac{4}{0{\cdot}01} = \underline{400 \text{ ohms}}$$

Maximum range of control voltage :

$$\Delta E_2 = \Delta I_a R_c = (0{\cdot}027 - 0{\cdot}002)400$$
$$= \underline{10 \text{ volts or } \pm 5 \text{ volts}}$$

TRANSFORMERS AND REACTORS

Transformation Ratio

Turns ratio for an ideal transformer, 100 per cent efficient

$$k = N_2/N_1 = E_2/E_1 = I_1/I$$

where N_1N_2 = number of primary and secondary turns;
E_1E_2 = primary and secondary voltages;
I_1I_2 = primary and secondary currents.

In practice, the turns ratio for a transformer with a voltage regulation of v per cent from no load to full load, or of efficiency η

$$k = E_2(1 + v/100)/E_1$$
$$= E_2/\sqrt{\eta}E_1 \text{ if the efficiency} > 50 \text{ per cent}$$

Equivalent Impedance of Transformer

Equivalent impedance of load on secondary, referred to primary terminals:

Double wound, neglecting impedances of windings

$$Z_1 = Z_2/k^2$$

Double wound, including impedances of windings

$$Z_1 = Z_1' + (Z_2 + Z_2')/k^2$$

Double wound, with centre tapped primary, each half

$$Z_1 = Z_2/4k^2$$

where Z_1 = primary load impedance;
Z_2 = secondary load impedance;
Z_1' = impedance of primary winding;
Z_2' = impedance of secondary winding.

Multi-tapped primary, tapped section:

$$Z' = Z(N'/N)^2$$

where Z' = load impedance of tapped section;
Z = load impedance of whole winding;
N' = number of turns in tapped section;
N = number of turns in whole winding.

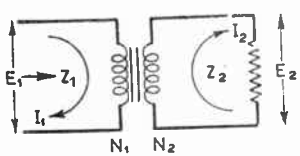


Fig. 113.—Equivalent Impedance of Transformer.

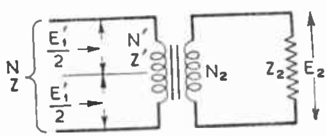


Fig. 114.—Tapped Primary.

EXAMPLE 1.—*Find the turns ratio of a transformer for use between an output stage having a rated load resistance of 5,000 ohms and an output load resistance of 20 ohms. At what points must the secondary be tapped for alternative load resistances of 15 ohms and 10 ohms?*

Turns ratio, primary to secondary:

$$k = \sqrt{Z_1/Z_2} = \sqrt{\frac{5{,}000}{20}} = \underline{15{\cdot}8/1}$$

Secondary tapping for 15 ohms output:

$$N_2'/N_2 = \sqrt{\frac{15}{20}} = \underline{0{\cdot}866 \text{ of the secondary turns}}$$

Secondary tapping or 10 ohms output:

$$N_2'/N_2 = \sqrt{\frac{10}{20}} = \underline{0{\cdot}707 \text{ of the secondary turns}}$$

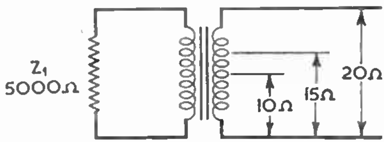


FIG. 115 (*left*).—EXAMPLE 1.

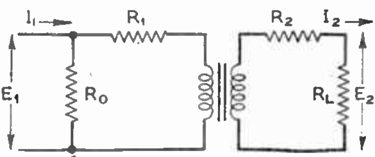


FIG. 116 (*right*).—TRANSFORMER EFFICIENCY FACTORS.

Transformer Efficiency and Power Losses

Efficiency	$\eta = P_2/(P_2 + P_c + P_i)$
Power input	$P_1 = P_2/\eta$
Total power losses	$P_L = P_2(1/\eta - 1)$
Copper losses *	$P_c = P_L/2$
Core losses *	$P_i = P_L/2$
Primary copper loss	$P_{c1} = I_1^2R_1$
Secondary copper loss . . .	$P_{c2} = I_1^2R_2$
Total copper losses	$P_c = I_2^2(k^2R_1 + R_2)$
Eddy current loss, ergs/cycle/c.c. of iron	$P_e = KB^{1{\cdot}6}$
Hysteresis loss, watts/c.c. of iron .	$P = fKB^{1{\cdot}6} \times 10^{-7}$

where P_2 = secondary volt-amperes;
R_1 = resistance of primary winding;
R_2 = resistance of secondary winding;
K = a factor, depending on the type of iron;
B = flux density, lines per c.c.;
f = frequency, c/s.

* For economy in design of power-supply transformers, primary and secondary copper losses are equalized, and core losses are made approximately equal to copper losses.

AVERAGE EFFICIENCIES OF SMALL TRANSFORMERS

50 VA .	. 75 per cent	300 VA .	. 93 per cent
100 VA .	. 85 per cent	500 VA .	. 94 per cent
200 VA .	. 90 per cent	1,000 VA .	. 96 per cent

EXAMPLE 2.—*A transformer has an output of 100 VA at 400 volts and a turns ratio of 1/1·75. The resistance of the primary winding is 20 ohms and the secondary 85 ohms. Find the efficiency, assuming that the copper and iron losses are equalized.*

Secondary current :

$$I_2 = P_2/E_2 = \frac{100}{400} = 0{\cdot}25 \text{ ampere}$$

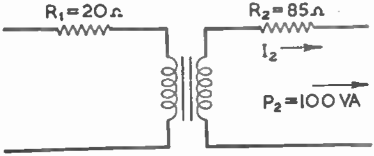


FIG. 117.—EXAMPLE 2.

Total copper loss :

$$P_c = I_2^2(k^2R_1 + R_2)$$
$$= 0{\cdot}25^2(1{\cdot}75^2 \times 20 + 85) = 9{\cdot}14 \text{ watts}$$

Iron losses = copper losses

∴ Total losses = 2 × 9·14 = 18·28 watts

Efficiency per cent :

$$\eta = 100P_2/(P_2 + P_L)$$
$$= \frac{100 \times 100}{100 + 18{\cdot}28} = \underline{84{\cdot}6 \text{ per cent}}$$

Voltage Regulation of Transformers

Percentage regulation :

$$v = 100(E_{NL} - E_{FL})/E_{NL}$$

where E_{FL} = output voltage at full load,
E_{NL} = output voltage at no load.

AVERAGE VOLTAGE REGULATION OF POWER TRANSFORMERS

1 kVA .	. 3·2 per cent	20 kVA .	. 1·8 per cent
5 kVA .	. 3·0 per cent	50 kVA .	. 1·5 per cent
10 kVA .	. 2·2 per cent	100 kVA .	. 1·25 per cent

EXAMPLE 3.—*A rectifier transformer has a secondary voltage of 400 at full load and a voltage regulation of 5 per cent. What is the peak voltage applied to the rectifier when the load is switched off?*

Regulation per cent :

$$v = 100(E_{NL} - E_{FL})/E_{NL} = 100(1 - E_{FL}/E_{NL})$$

$$E_{NL} = E_{FL}/(1 - v/100) = \frac{400}{1 - 5/100} = 421 \text{ volts}$$

Peak voltage at no load :

$$E_p = \sqrt{2}E_{NL} = \sqrt{2} \times 421 = \underline{596 \text{ volts peak}}$$

Transformer Design Factors

The various quantities are determined in the following order :

Secondary power output, watts	$P_2 = E_2 I_2$
Primary power input, watts	$P_1 = P_2/\eta$
Core area, sq. in.	$A = \sqrt{P_1}/5{\cdot}58$
Primary current, amperes	$I_1 = P_1/E_1$
Secondary current, amperes	$I_2 = P_2/E_2$
Gauge of wire for primary and secondary	Wire tables
Primary turns per volt	$N_1/E_1 = 10^8/4{\cdot}44fAB_{max.}$
For $B_{max.}$ = 60,000 lines/sq. in. and 50 c/s	$= 7{\cdot}5/A$
For $B_{max.}$ = 80,000 lines/sq. in. and 50 c/s	$= 5{\cdot}63/A$
Number of primary turns	$N_1 = (N_1/E_1)E_1$
Number of secondary turns	$N_2 = kN_1$
Size of core laminations	To suit winding space

EXAMPLE 4.—*Determine the following data for a transformer to operate on a 230-volt, 50-c/s supply and deliver 100 VA at 500 volts:* (a) *cross-sectional area of core;* (b) *current rating for the wire used in each winding;* (c) *number of primary and secondary turns, assuming a maximum flux density of 80,000 lines/sq. in.*

Assume an efficiency of 85 per cent for this rating.

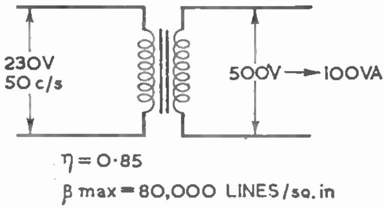


FIG. 118.—EXAMPLE 4.

Primary power input :

$$P_1 = P_2/\eta = \frac{100}{0{\cdot}85} = 117{\cdot}7 \text{ watts}$$

Core area :

$$A = \sqrt{P_1}/5{\cdot}58 = \frac{\sqrt{117{\cdot}7}}{5{\cdot}58} = \underline{1{\cdot}85 \text{ sq. in.}}$$

Primary current :

$$I_1 = P_1/E_1 = \frac{117{\cdot}7}{230} = \underline{0{\cdot}512 \text{ ampere}}$$

Secondary current :

$$I_2 = P_2/E_2 = \frac{100}{500} = \underline{0{\cdot}2 \text{ ampere}}$$

Turns per volt of primary :

$$N_1/E_1 = 5{\cdot}63/A = \frac{5{\cdot}63}{1{\cdot}85} = \underline{3{\cdot}04 \text{ turns/volt}}$$

Turns ratio :

$$k = E_2/\sqrt{\eta}E = \frac{500}{\sqrt{0{\cdot}85} \times 230} = 2{\cdot}36$$

Total number of primary turns :

$$N_1 = 3{\cdot}04 \times 230 = \underline{669 \text{ turns}}$$

Total number of secondary turns :

$$N_2 = kN_1 = 2{\cdot}36 \times 699 = \underline{1{,}647 \text{ turns}}$$

Auto Transformers

The saving in copper over a double-wound transformer is determined by the ratio

$$\frac{\text{Copper in auto transformer}}{\text{Copper in double-wound transformer}} = 1 - 1/k \text{ or } 1 - k \text{ if } k < 1$$

For ratios of 2/1 or 1/2 the quantity of copper is halved, and the current in the common portion of the winding is zero in an ideal transformer. Little economy is realized for higher transformation ratios.

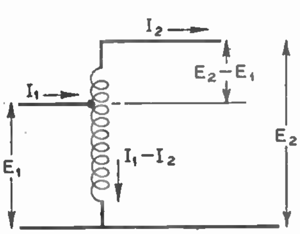


FIG. 119.—AUTO TRANSFORMER.

EXAMPLE 5.—*An auto transformer is required to deliver 250 VA at 50 volts with an input of 230 volts. Calculate the current rating of each section of the winding, assuming an efficiency of 90 per cent, and the percentage saving in copper over a double-wound transformer.*

Secondary current :

$$I_2 = P_2/E_2 = \frac{250}{50} = 5 \text{ amperes}$$

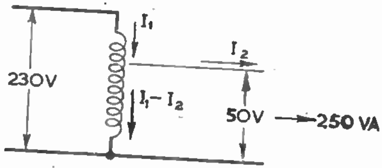


FIG. 120.—EXAMPLE 5.

Primary current :

$$I_1 = P_2/\eta E_1 = \frac{250}{0{\cdot}9 \times 230} = 1{\cdot}208 \text{ amperes}$$

Current rating of upper section of winding $= \underline{5 \text{ amperes}}$

Current rating of common section of winding $= I_2 - I_1$
$= 5 - 1{\cdot}21$
$= \underline{3{\cdot}79 \text{ amperes}}$

Turns ratio $k = E_2/\sqrt{n}E_1 = \dfrac{50}{0{\cdot}9 \times 230} = 0{\cdot}229$

Percentage saving in copper $= 1 - 1/k$ (or $1 - k$ when $k < 1$)
$= 1 - 0{\cdot}229$
$= 0{\cdot}771,$

or a saving of 77 per cent approx.

Iron-core Chokes

Audio-frequency chokes and transformers with A.C. only flowing :

$$L = (3{\cdot}2N^2\mu A/l) \times 10^{-8}$$

Ripple filter chokes with large air gap. D.C. component of current large compared with A.C. component :

$$L = (3{\cdot}2N^2A/l_g) \times 10^{-8}$$

where μ = A.C. permeability, dependent on flux density and iron alloy used — for audio-frequency chokes at very low flux density $\mu = 1{,}000$;
L = inductance, henrys;
N = number of turns in winding;
A = cross-sectional area of core, sq. in.;
l = length of magnetic path, in.;
l_g = length of air gap, in.

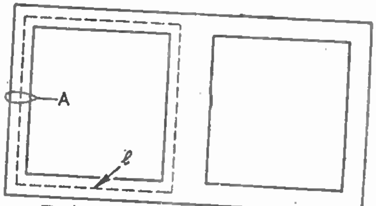


FIG. 121.—MAGNETIC PATH OF CHOKE.

CABLES AND LINES

Rating of Power Cables

The determination of the size of a cable is normally governed by its current-carrying capacity, but in low-voltage circuits carrying heavy current, the voltage drop and power loss in the cable must also be taken into consideration. The current rating of power cables of different sizes run under various conditions can be obtained from the *I.E.E. Regulations for the Electrical Equipment of Buildings.*

The current taken by ordinary power supply and conversion apparatus can be calculated from the formulæ given in Table 10.

TABLE 10.—FORMULÆ FOR CURRENT TAKEN

	R.m.s. Line Current
Single-phase transformer	$P_2/\eta E_1$
Three-phase transformer *	$P_2/\sqrt{3}\eta E_1$
Single-phase rectifier	$kI_0(E_0 + E_L)/\eta E_1$
Three-phase or six-phase rectifier *	$kI_0(E_0 + E_L)/\sqrt{3}\eta E_1$
D.C. motor	$746\ P_h/\eta E_1$
Single-phase A.C. motor	$746\ P_h/\eta E_1 \cos\phi$
Three-phase A.C. motor *	$746\ P_h/\sqrt{3}\eta E_1 \cos\phi$
Single-phase alternator	$P_0/E \cos\phi$
Three-phase alternator *	$P_0/\sqrt{3}E \cos\phi$

* On a three-phase system with earthed neutral the voltage rating of a cable is $1/\sqrt{3}$ or 0·578 of the line voltage.

where P_2 = secondary VA;
P_h = rated horse-power;
P_0 = power output, watts;
η = efficiency;
E_0 = D.C. output voltage;
E_1 = input r.m.s. voltage;
E_L = voltage drop across filter choke;
I_0 = D.C. output current, amperes;
k = a conversion factor, depending on the number of rectified phases.

Single-phase, full wave	$k = 1{\cdot}11$
Single-phase, full wave, bridge	$k = 1{\cdot}11$
Three-phase, half wave	$k = 1{\cdot}21$
Three-phase, full wave, star	$k = 1{\cdot}28$
Three-phase, full wave, bridge	$k = 1{\cdot}05$

Voltage Drop and Power Loss in Cables

Resistance per foot length, ohms	$r = c/A$
Voltage drop per foot length, ohms	$Ir = cI/A$
Power loss per foot length, watts	$I^2r = cI^2/A$

where A = cross-sectional area of conductor, sq. in.;
c = resistance per unit length (= $8{\cdot}75 \cdot 10^{-6}$ ohm/ft. for soft copper wire)

TESTS AND MEASUREMENTS

Calibration of Meters

VOLTMETER SERIES MULTIPLIER RESISTANCE

$$R_s = R_m(n - 1)$$

AMMETER SHUNT RESISTANCE

$$R_{sh} = R_m/(n - 1)$$

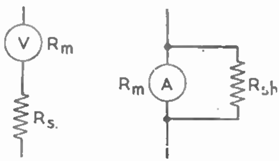


FIG. 122.—SERIES AND SHUNT MULTIPLIERS.

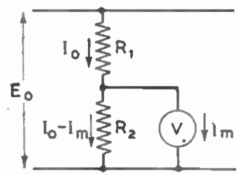


FIG. 123.—VOLTMETER WITH POTENTIAL DIVIDER.

VOLTMETER WITH POTENTIAL DIVIDER

When a moving-coil voltmeter is used with a potential divider for measuring high D.C. voltages, the resistances of the unshunted and shuntod sections are :

$$R_1 = (E_0 - I_m R_m)/I_0$$
$$R_2 = I_m R_m/(I_0 - I_m)$$

where R_m = resistance of meter (ohms/volt × range in the case of a voltmeter);
n = scale multiplier;
R_1 = resistance of unshunted section of divider;
R_2 = resistance of shunted section of divider;
I_m = meter current for full scale deflection, amperes;
I_0 = total current in potential divider ($I_0 > I_m$).

EXAMPLE 1.—*A voltmeter, scaled* 0–40 *volts, has a resistance of* 2,500 *ohms/volt and gives full-scale deflection with* 0·4 *mA.* (a) *What resistance will be required in series to read* 100 *volts full scale?* (b) *If the meter is tapped across a potential divider to read* 1,000 *volts full scale, what will be the resistance of each section if the total current is to be* 1 *mA?*

(*a*) Voltmeter resistance :

$$R_v = 2{,}500 \times 40 = 100{,}000 \text{ ohms}$$

Scale multiplier :

$$n = 100/40 = 2{\cdot}5$$

Resistance required in series :

$$R_s = R_m(n - 1)$$
$$= 100{,}000(2{\cdot}5 - 1) = \underline{150{,}000 \text{ ohms}}$$

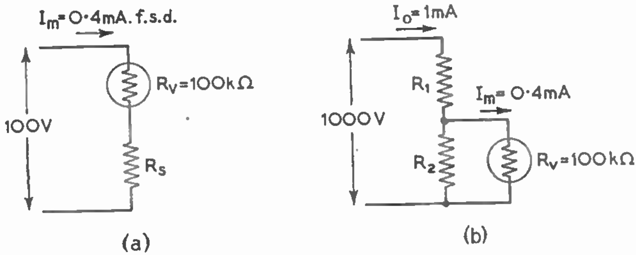


FIG. 124.—EXAMPLE 1.

(*b*) Resistance of unshunted section :

$$R_1 = (E_0 - I_m R_m) I_0$$
$$= \frac{1{,}000 - (0{\cdot}0004 \times 100{,}000)}{0{\cdot}001}$$
$$= 0{\cdot}96 \times 10^6 \text{ ohms} = 0{\cdot}96 \text{ M}\Omega \text{ say } \underline{1 \text{ M}\Omega}$$

Resistance of shunted section :

$$R_2 = I_m R_m / (I_0 - I_m)$$
$$= \frac{0{\cdot}0004 \times 100{,}000}{0{\cdot}001 - 0{\cdot}0004}$$
$$= 6{\cdot}67 \times 10^4 \text{ ohms or } \underline{66{\cdot}7 \text{ k}\Omega}$$

Bridge-measuring Instruments

A Wheatstone resistance bridge is balanced when

$$R_x = R_2 R_3 / R_1$$

EXAMPLE 2.—*The resistance of a co-axial cable having a norma resistance of 23·6 ohms/1,000 yd. was measured on a Wheatstone bridge to determine the position of an earth fault. The resistances of the two ratio*

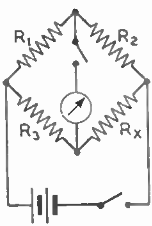


FIG. 125.—WHEATSTONE RESISTANCE BRIDGE.

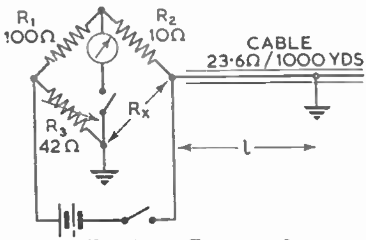


FIG. 126.—EXAMPLE 2.

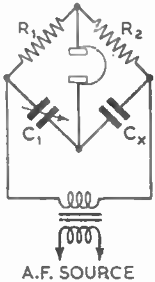


FIG. 127.—CAPACITANCE BRIDGE.

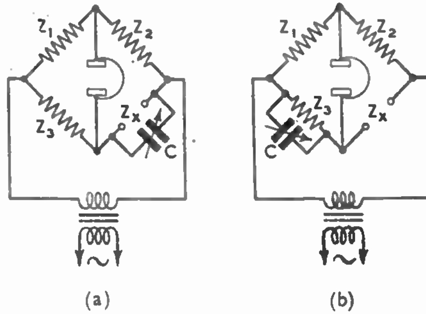


FIG. 128.—CONNECTIONS FOR (*a*) POSITIVE AND (*b*) NEGATIVE REACTANCES.

arms were respectively 100 *and* 10 *ohms, and balance was obtained when the variable arm was* 42 *ohms. Find the distance of the fault from the testing end.*

Resistance of cable :

$$R_x = R_2R_3/R_1 = \frac{42 \times 10}{100} = \underline{4{\cdot}2 \text{ ohms}}$$

Distance of fault from end of cable

$$= l'R_x/R' = \frac{1{,}000 \times 4{\cdot}2}{23{\cdot}6} = \underline{178 \text{ yd.}}$$

A simple capacitance bridge (Fig. 127) is balanced when

$$C = C_1R_2/R_1$$

where R_1R_2 = calibrated resistance arms;
R_3 = adjustable calibrated resistance;
R_x = resistance to be measured;
C_1 = adjustable calibrated capacitance,
C_x = capacitance to be measured.

An A.C. impedance bridge is balanced when

$$Z = Z_2Z_3/Z_1$$

Usually Z_1 and Z_2 are made equal, so that the impedances of the opposite arms Z_3 and Z_x must be equal to satisfy the balance. Then if the balancing arm Z_x is made a pure resistance R_3, $Z_x = R_3$.

For a positive (inductive) reactance (Fig. 128 (*a*)) :

$$Y_x = G_x + jB_x$$

$$R_x + jX_x = (R_3 + j\omega CR_3^2)/\{1 + (\omega CR_3)^2\}$$

$$R_x = R_3/\{1 + (\omega CR_3)^2\}$$
$$X_x = \omega CR^2{}_3/\{1 + (\omega CR_3)^2\}$$

For a negative (capacitive) reactance (Fig. 128 (*b*)):

$$R_x = R_3/\{1 + (\omega CR_3)^2\}$$
$$X_x = -\omega CR_3^2/\{1 + (\omega CR_3^2)\}$$

where R_x = resistance component of unknown impedance;
X_x = reactance component of unknown impedance;
R_3 = resistance of balancing arm;
C = added capacitance.

EXAMPLE 3.—*A cable of unknown impedance is connected to an A.C. impedance bridge. Balance is obtained at a frequency of 1,000 c/s when the impedances of the ratio arms are equal, the resistance of the balancing arm is 500 ohms and the shunt capacitance is 0·4 µF. Calculate the resistance and reactance components of the cable.*

$$\omega C = 2 \times 10^3 \times 4 \times 10^{-7} = 2{\cdot}513 \times 10^{-3}$$

Impedance of cable:

$$Z_x = Z_2 Z_3/Z_1$$

But since $Z_2 = Z_1$

$$Z_x = Z_3$$
$$= \frac{R_3(jX_3)}{R_3 + jX_3}$$
$$= \frac{-jR_3/\omega C}{R_3 - j/\omega C}$$
$$= \frac{(-jR_3/\omega C)(R_3 + j/\omega C)}{R_3^2 + (1/\omega C)^2}$$
$$= \frac{R_3 - j\omega CR_3^2}{1 + (\omega CR_3)^2}$$
$$= \frac{500 - j(2{\cdot}513 \times 10^{-3} \times 500^2)}{1 + (2{\cdot}513 \times 10^{-3} \times 500)^2} = 194 - j244$$

or 194 ohms resistance
244 ohms capacitive reactance

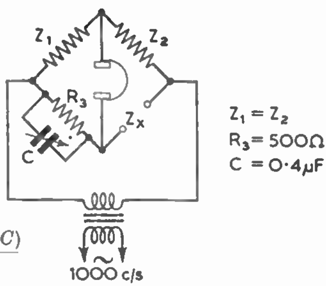


FIG. 129.—EXAMPLE 3.

Resistance, Inductance and Capacitance

Resistance, inductance and capacitance in low-frequency and radio-frequency circuits can often be determined from simple A.C. measurements of voltage and current.

Resistance	$R = E/I$
Inductance	$L = E/\omega I$
Capacitance	$C = I/\omega E$
Resistance with known resistance in series	$R = E/I - R_s$
Inductance with known resistance in series	$L = \{\sqrt{(E/I)^2 - R_s^2}\}/\omega$
Capacitance with known resistance in series	$C = 1/\{\omega\sqrt{(E/I)^2 - R_s^2}\}$

where E = r.m.s. value of voltage;
I = r.m.s. value of current, amperes;
$\omega = 2\pi f$;
R_s = known resistance in series, ohms.

EXAMPLE 4.—*A choke is connected in series with a capacitor across a 100-volt, 1,000-c/s supply. The capacitor is adjusted until a maximum current of 200 mA flows in the circuit. The voltage across the capacitor,*

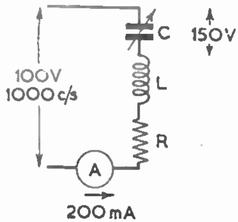


FIG. 130.—EXAMPLE 4.

measured with a valve voltmeter, is then 150 volts r.m.s. Calculate the resistance and inductance of the choke and the capacitance of the capacitor assuming that the capacitor has negligible resistance.

Maximum current flows at resonance, and effective reactance of circuit is zero. Then resistance of choke:

$$R = E/I = 100/0{\cdot}2 = \underline{500 \text{ ohms}}$$

Capacitance of condenser:

$$I = \omega CE$$

$$C_{\mu F} = 1/\omega E \times 10^6$$

$$= \frac{0{\cdot}2 \times 10^6}{2 \times 1{,}000 \times 150} = 1/1{\cdot}5\pi = \underline{0{\cdot}212\ \mu\text{F}}$$

Since inductive and capacitive reactances are equal at resonance,

$$\omega L = 1/\omega C$$

$$L_H = 1/\omega^2 C$$

$$= \frac{1}{(2 \times 1{,}000)^2 \times 0{\cdot}212 \times 10^{-6}} = \underline{0{\cdot}12 \text{ henry}}$$

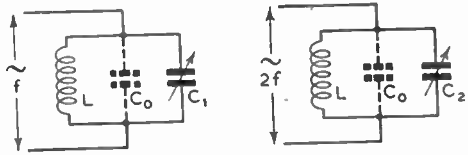


FIG. 131.—DETERMINATION OF SELF-CAPACITANCE OF INDUCTOR L.

Resonant Circuits

Apparent inductance of resonant circuit, capacitance known	$L_a = 1/\omega^2 C_1$
True inductance, (capacitance and self-capacitance of coil known)	$L_1 = L_a C_1/(C_1 + C_0)$

Self-capacitance of inductance coil. This is found as follows:

The coil is energized with a known capacitance in parallel by means of a calibrated test oscillator at the resonant frequency of the combination. The frequency is then doubled and the capacitance adjusted until resonance is again obtained. The self-capacitance:

$$C_0 = (C_1 - 4C_2)/3$$

where C_1 = capacitance required for resonance at frequency f;
C_2 = capacitance required for resonance at frequency $2f$.

EXAMPLE 5.—*A coil of unknown inductance with a capacitance of 250 pF in parallel tunes to a test oscillator at a frequency of 1·8 Mc/s. When the frequency is doubled, the circuit tunes with a capacitance of 45 pF. Find the self-capacitance of the coil and its true inductance. A capacitor of unknown capacitance is now inserted in parallel, and the circuit is found to tune to a frequency of 1·63 Mc/s. Find the capacitance of the capacitor.*

Self-capacitance of coil:

$$C_0 = (C_1 - 4C_2)/3 = \frac{250 - 180}{3} = \underline{23{\cdot}3 \text{ pF}}$$

$$\omega^2 = (2 \times 1{\cdot}8 \times 10^6)^2 = 1{\cdot}277 \times 10^{14}$$

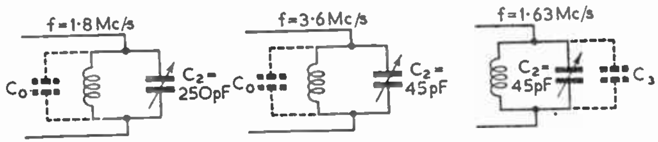


FIG. 132.—EXAMPLE 5.

Apparent inductance:

$$L_a = 1/\omega_1^2 C_1$$

$$L_a\,(\mu\text{H}) = 10^{18}/\omega_1^2 C_1 \text{ when } C_1 \text{ is in pF}$$

$$= \frac{10^{18}}{1{\cdot}277 \times 10^{14} \times 250} = 31{\cdot}3\,\mu\text{H}$$

True inductance :

$$L = \frac{L_a C_1}{(C_1 + C_0)} = \frac{31{\cdot}3 \times 250}{250 + 23{\cdot}3} = \underline{28{\cdot}6\,\mu\text{H}}$$

$$\omega_2^2 = (2\pi \times 1{\cdot}63 \times 10^6)^2 = 1{\cdot}046 \times 10^{14}$$

$$L_a\,(\mu\text{H}) = 10^{18}/\{\omega_2^2(C_2 + C_3)\}$$

Capacitance of added capacitor :

$$C_3\,(\text{pF}) = (10^{18}/\omega_2^2 L_a) - C_2$$

$$= \frac{10^{18}}{(1{\cdot}046 \times 10^{14} \times 31{\cdot}4)} - 45 = \underline{260\text{ pF}}$$

Radio-frequency Resistance

Resistance Variation Method

The apparatus of which the resistance is to be measured is inserted at X, and the circuit is coupled to a stable radio-frequency oscillator. A current reading is taken on a thermo-ammeter. A known non-inductive resistance is then inserted in series and a second reading is taken. Then

$$R = R_1/(I/I_1 - 1)$$

where I = current reading without added resistance;
I_1 = current reading with added resistance;
R_1 = added resistance, ohms.

Added Reactance Method

Suitable for measuring the total resistance of a circuit. The current is measured at resonance. Reactance is then zero. The circuit reactance is varied by a known amount and the current is read again. The reactance can be varied by adjusting a calibrated variable condenser or the number of turns of a known inductance or by changing the frequency. Then

$$R = X_1\sqrt{I_1^2/(I^2 - I_1^2)}$$

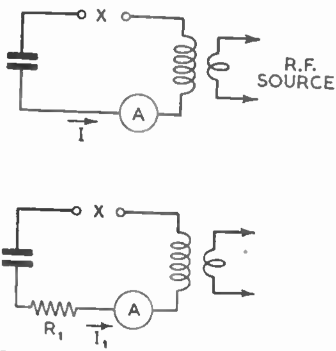


Fig. 133.—Measurement of Radio-Frequency Resistance

where I_r = current at resonance, amperes;
I_1 = current with added reactance, amperes;
X_1 = added reactance, ohms
$= \pm (C_r - C)/\omega C_r C$ for variation of capacitance
$= \pm \omega(L - L_r)$ for variation of inductance
$= \pm L(\omega^2 - \omega_r^2)/\omega$ for variation of frequency.

Valve Testing

The constants g, μ and R are determined from measurements of small relative changes of anode current (I), anode voltage (δE) and grid voltage (δE). They are usually specified for receiver valves at an anode voltage of 100, and zero grid bias. g should not be allowed to fall below 80 per cent of its specified value for satisfactory operation.

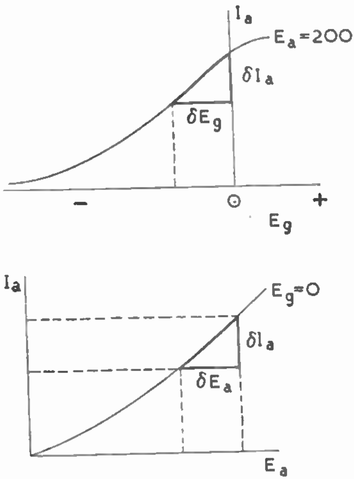


Fig. 134.—Determination of Valve Constants.

Mutual conductance : $g = \dfrac{\delta I_a}{\delta E_g}$

Amplification factor : $\mu = \dfrac{\delta E_a}{\delta E_g}$

A.C. resistance $R = \dfrac{\delta E_a}{\delta I_a}$

If two of the three constants are known, the third can be found from the relationship

$$\mu = gR_a$$

Example 6.—*Tests on a triode showed that when the anode voltage was reduced from* 200 *to* 180 *with zero grid bias, the anode current fell from* 30 *to* 23·3 *mA. When the anode voltage was kept constant at* 200 *and the bias voltage was altered from* 0 *to* −2 *volts, the anode current fell to* 22 *mA. Find the values of* R_a, g *and* μ.

A.C. resistance

$$R_a = \frac{\delta E_a}{\delta I_a} = \frac{200 - 180}{(30 - 23{\cdot}3)/1{,}000} = \underline{2{,}985 \text{ ohms}}$$

Mutual conductance :

$$g = \frac{\delta I_a}{\delta E_g} = \frac{(30 - 22)/1{,}000}{2}$$

$$= 0{\cdot}004 \text{ ampere/volt or } \underline{4 \text{ mA/volt}}$$

Amplification factor :

$$\mu = gR_a = 0{\cdot}004 \times 2{,}985 = 11{\cdot}9$$

W. E. P.

AIDS TO ENGINEERING CALCULATIONS

Accuracy

Nearly all engineering data is approximate to within the limits imposed by measuring instruments and no useful purpose is served by carrying the result of a calculation to more than the number of significant figures given in the data. As an example, the diameter of a wire can be measured accurately with a micrometer to three significant figures. In calculating the resistance from the measured diameter and length, the accuracy of the result is not increased by working to more than three significant figures.

Again, suppose it is required to convert a wavelength of 16·4 metres to frequency in megacycles. The result would be correctly stated as 18·3 Mc/s. If the result were given as 18·29 Mc/s, no reliance could be placed on the second decimal, as the original figure is given to only three significant figures.

In most practical applications the results are seldom required to a greater accuracy than three significant figures, which is obtainable on a 10-in. slide rule. Two significant figures is often sufficient for rough mental computations.

A meter reading or a measurement given as 4·3 is understood to have limits of accuracy of $4{\cdot}3 \pm 0{\cdot}05$. If the reading were given as 4·30, the limits would be $4{\cdot}30 \pm 0{\cdot}005$, and so on. The procedure for adding, subtracting, multiplying and dividing data of equal accuracy when the limits are required is illustrated by the following examples :

ADDITION

$$\begin{array}{r} 4{\cdot}3 \pm 0{\cdot}05 \\ 3{\cdot}2 \pm 0{\cdot}05 \\ \hline 7{\cdot}5 \pm 0{\cdot}1 \end{array}$$

SUBTRACTION

$$\begin{array}{r} 4{\cdot}3 \pm 0{\cdot}05 \\ 3{\cdot}2 \pm 0{\cdot}05 \\ \hline 1{\cdot}1 \pm 0 \end{array}$$

or $1{\cdot}1 \pm 0{\cdot}1$ if the errors are of the same sign.

MULTIPLICATION

$$\begin{aligned} &(4{\cdot}3 \pm 0{\cdot}05)\,(3{\cdot}2 \pm 0{\cdot}05) \\ &= 13{\cdot}76 \pm 0{\cdot}05\,(4{\cdot}3 + 3{\cdot}2) + 0{\cdot}0025 \text{ neglecting the last term.} \\ &= 13{\cdot}76 \pm 0{\cdot}37 \end{aligned}$$

DIVISION

$$\frac{4{\cdot}3 \pm 0{\cdot}05}{3{\cdot}2 \pm 0{\cdot}05} = \frac{4{\cdot}35}{3{\cdot}15} = 0{\cdot}1381$$

$$\text{or } \frac{4{\cdot}25}{3{\cdot}25} = 0{\cdot}1308$$

Find the mean of the two results to as many decimal places as the datum. Then

$$\begin{aligned} \text{Mean quotient} &= 0{\cdot}1344 \\ \text{Mean of difference} &= 0{\cdot}0036 \\ \text{Limits of quotient} &= 0{\cdot}1344 \pm 0{\cdot}0036 \end{aligned}$$

Approximations

An approximate calculation can be made to determine the position of the decimal point or to obtain a rough answer to an extended product or division by the following method. With practice the working can be performed mentally.

Write down each term of the expression as an approximate number of whole units to the left of the decimal point, multiplied or divided by the necessary number of tens written in index form, e.g.,

$$0{\cdot}00365 = 4 \,.\, 10^{-3}$$
$$692 = 7 \,.\, 10^{2}$$

Examples

(1)

$$\frac{0{\cdot}00365 \times 692 \times 8{\cdot}2}{77{\cdot}4 \times 0{\cdot}01}$$

$$\simeq \frac{4 \times 7 \times 8}{8 \times 1} \times \frac{10^{-3} \times 10^{2}}{10^{1} \times 10^{-2}} = \underline{28 \text{ approx.}}$$

(2)

$$\sqrt{\frac{2{\cdot}1 \times 8{,}074}{39{\cdot}4 \times 0{\cdot}93}}$$

$$\simeq \sqrt{\frac{2 \times 8}{4 \times 1} \times \frac{10^{3}}{10^{1}}}$$

$$= 10\sqrt{4} = \underline{20 \text{ approx.}}$$

Short Cuts in Calculations

Use of Indices. Numerical calculation is simplified by the use of indices or powers of ten. The index number, when positive, denotes the number of figures to the left of the decimal point; when negative to the right of the decimal point, e.g.,

$$127{,}000 = 1{\cdot}27 \times 10^{5}$$
$$0{\cdot}0038 = 3{\cdot}8 \times 10^{-3}$$
$$0{\cdot}054 = 5{\cdot}4 \times 10^{-2}$$

The rules for operating on indices are

$$x^{m} \times x^{n} = x^{m+n} \qquad 10^{3} \times 10^{2} = 10^{5}$$
$$x^{m}/x^{n} = x^{m-n} \qquad 10^{3}/10^{2} = 10$$
$$(x^{m})^{n} = x^{mn} \qquad (10^{3})^{2} = 10^{6}$$

Examples

(1)

$$\frac{3{\cdot}2 \times 0{\cdot}00625 \times 0{\cdot}73}{6{\cdot}29 \times 25{\cdot}64 \times 0{\cdot}01} = \frac{3{\cdot}2 \times 6{\cdot}25 \times 0{\cdot}73}{6{\cdot}29 \times 2{\cdot}564} \times \frac{10^{-3}}{10^{-1}}$$

$$= \underline{0{\cdot}00905}$$

(2) $\sqrt{0.003 \times 0.17} = \sqrt{3 \times 1.7 \times 10^{-4}}$

$= 10^{-2}\sqrt{5.1}$

$= \underline{0.0226}$

Squares and Square Roots

The calculation of square roots and cube roots is simplified by remembering that

$$\sqrt{x \times 10^n} = 10^{n/2}\sqrt{x}$$

$$\sqrt[3]{x \times 10^n} = 10^{n/3}\sqrt[3]{x}$$

Examples

(1) $\sqrt{0.0009} = \sqrt{9 \times 10^{-4}} = 10^{-2}\sqrt{9} = \underline{0.03}$

(2) $\sqrt[3]{27{,}000} = \sqrt[3]{27 \times 10^3} = 10\sqrt[3]{27} = \underline{30}$

In such expressions as a^2b^2, a^2/b^2, $\sqrt{a+b}/\sqrt{c+d}$ the operation of squaring or extracting the square root can be reduced to one operation.

Examples

(1) $\omega^2L^2 = (\omega L)^2$

(2) $a^2/n^2 = (a/n)^2$

(3) $\sqrt{R^2 + \omega^2L^2}/\sqrt{G^2 + 1/\omega^2C^2} = \sqrt{\{R^2 + (\omega L)^2\}/\{G^2 + 1/(\omega C)^2\}}$

Differences of Two Squares

The difference of two squares can be calculated by factorizing without having to square the two quantities.

$$a^2 - b^2 = (a+b)(a-b)$$

Other useful approximations obtained by squaring are:

$$(a+b)^2 = a^2 + 2ab + b^2$$
$$= a^2 + 2ab \qquad \text{when } a \gg b$$

$$(a-b)^2 = a^2 - 2ab + b^2$$
$$= a^2 - 2ab \qquad \text{when } a \gg b$$

Examples

(1) $14^2 - 6{\cdot}5^2 = (14 + 6{\cdot}5)(14 - 6{\cdot}5) = \underline{153{\cdot}75 \text{ approx.}}$

(2) $42^2 = 40^2 + 2(40 \times 2) = \underline{1{,}760 \text{ approx.}}$

(3) $4{\cdot}8^2 = 5^2 - 2(5{\cdot}0 \times 0{\cdot}2) = \underline{23 \text{ approx.}}$

Products and Quotients

The following approximations for the products and powers of binomial factors can often be applied to shorten certain calculations.

$$(1 + x)(1 + y) \simeq 1 + x + y$$
$$(1 + x)(1 + y)(1 + z) \simeq 1 + x + y + z$$
$$(1 \pm x)^n \simeq 1 \pm nx$$
when x, y and $z \ll 1$

Examples

(1) $1{\cdot}003 \times 0{\cdot}995 = (1 + 0{\cdot}003)(1 - 0{\cdot}005)$
$= 1 + 0{\cdot}003 - 0{\cdot}005$ approx.
$= \underline{0{\cdot}998}$

(2) $\dfrac{1}{3{\cdot}006} = \frac{1}{3}(1 + 0{\cdot}002)^{-1}$
$= \frac{1}{3}(1 + (-1 \times 0{\cdot}002))$ approx.
$= \underline{0{\cdot}333}$

(3) $\dfrac{1}{\sqrt{0{\cdot}994}} = (1 - 0{\cdot}006)^{-\frac{1}{2}}$
$= 1 - (-\frac{1}{2})(0{\cdot}006)$ approx.
$= \underline{1{\cdot}003}$

Calculating Successive Values from a Formula

When an equation contains several variables, only one of which is varied, successive calculations for each different value of the variable factor are simplified by first working out the constant part of the expression, and then applying the appropriate law of proportions to the variable factor.

Example

The frequency in kc/s of a tuned circuit is given by the formula $f = 159/\sqrt{LC}$, where L is the inductance in microhenrys and C the capacitance in microfarads. Find f when $L = 50$, 100, 150 and 200 μH respectively and $C = 0{\cdot}0002$ μF.

When $L = 50$, $f_1 = 159/\sqrt{LC}$

$$= \frac{159}{\sqrt{50 \times 2 \times 10^{-4}}} = \underline{1{,}590 \text{ kc/s}}$$

Note that, since $f \propto 1/\sqrt{LC}$

at any other frequency $f_2 = f_1/\sqrt{L_2/L_1}$
$= f_1\sqrt{L_1/L_2}$

Hence

when $L = 100$ $f_2 = f_1/\sqrt{2} = \underline{1{,}120 \text{ kc/s}}$

$L = 150$ $f_3 = f_1/\sqrt{3} = \underline{920 \text{ kc/s}}$

$L = 200$ $f_4 = f_1/2 = \underline{795 \text{ kc/s}}$

Expressions Containing ω or ω^2

Pulsatance ω ($= 2\pi \times$ frequency) can be evaluated quickly in many cases by remembering that at 100 c/s. $\omega = 628$ approximately, and multiplying by 10^3 when the frequency is given in kc/s and 10^6 when in Mc/s.

Examples

(1) $f =$ 1,000 c/s. $\quad \omega = \omega_{100} \times 10 = \underline{6{,}280}$

(2) $f =$ 5 kc/s. $\quad \omega = \frac{1}{2}\omega_{100} \times 10^3 = \underline{3{\cdot}142 \times 10^5}$

(3) $f =$ 2·5 Mc/s. $\quad \omega = \frac{1}{4}\omega_{100} \times 10^6 = \underline{1{\cdot}571 \times 10^8}$

(4) $f =$ 40 Mc/s. $\quad \omega = \frac{1}{4}\omega_{100} \times 10^7 = \underline{2{\cdot}51 \times 10^{10}}$

ω^2 commonly occurs in such expressions as $\omega^2 LC$. Calculation can be shortened by considering that

$$\omega^2 = 3{\cdot}94 f^2 \times 10^7 \text{ when } f \text{ is in kc/s}$$
$$= 3{\cdot}94 f^2 \times 10^{13} \text{ when } f \text{ is in Mc/s}$$

Examples

(1) $f =$ 5 kc/s. $\quad \omega^2 = 3{\cdot}94 \times 5^2 \times 10^7 = \underline{9{\cdot}85 \times 10^7}$

(2) $f =$ 8 Mc/s. $\quad \omega^2 = 3{\cdot}94 \times 8^2 \times 10^{13} = \underline{2{\cdot}52 \times 10^{15}}$

Conversion of Units

In radio technique quantities are frequently measured in sub-multiples of basic units, e.g., current in milliamperes, inductance in microhenrys and capacitance in microfarads; whereas fundamental formulæ are derived from basic units. The procedure in the following examples will often simplify the conversion of formulæ:

Examples

(1) Find the power in watts dissipated by a resistance of 2,000 ohms when carrying a current of 30 mA. Power in watts $P = I^2R$.

Instead of converting to amperes and squaring, it is simpler to divide mentally (milliamperes) by 1,000 and resistance by 1,000.

$$P = 0{\cdot}9 \times 2 = 1{\cdot}8 \text{ watts}$$

(2) The frequency in c/s is given by the formula $f = 1/2\pi\sqrt{LC}$, when L is in henrys and C in farads. What is the frequency in Mc/s when L is in microhenrys and C in microfarads?

$$\sqrt{L_{\mu H} C_{\mu F}} = \sqrt{L_H C_F \times 10^{12}}$$
$$= 10^6 \sqrt{L_H C_F}$$

Also

$$f_{Mc/s} = 10^6 f_{c/s}$$

Hence the formula remains unchanged.

Where the conversion involves several factors, it is advisable to set down each factor step by step to reduce risk of errors.

Example

Find the number of kilogram-calories per hour equivalent of 1 kW given that

1 kg.cal. = 3·968 B.Th.U.
1 B.Th.U./sec. = 1·058 kW

kW to B.Th.U./sec.	$\div$ 1·058
B.Th.U./sec. to kg.cal./sec.	$\div$ 3·968
kg.cal./sec. to kg.cal./hr.	$\times$ 3,600

$$1 \text{ kW} = \frac{3{,}600}{1{\cdot}058 \times 3{\cdot}968}$$

$$= \underline{860 \text{ kg.cal./hr.}}$$

Use of Logarithms

The laws of indices expressed in logarithmic form are:

Product of two numbers . . . . (Add the logs and take the antilog.)	$\log mn = \log m + \log n$
Quotient of two numbers . . . (Subtract the logs and take the antilog.)	$\log m/n = \log m - \log n$
nth power of a number . . . . (Multiply the log by the power and take the antilog.)	$\log m^n = n \log m$
nth root of a number . . . . (Divide the log by the power and take the antilog.)	$\log \sqrt[n]{m} = (1/n) \log m$

In applying these principles to calculations, note that the decimal part of a logarithm is positive, but the whole number (the index) to the left of the decimal point may be either positive or negative, according as the number corresponding to the logarithm is greater or less than unity. When the index is negative, convert the logarithm to a positive quantity by the device of adding +10 −10, as shown in the following examples.

Examples

(1) Evaluate $\dfrac{29{\cdot}32 \times 0{\cdot}0076}{3{\cdot}451}$.

log 29·32 .	= 1·4672
log 0·0076 = $\bar{3}$·8808 (+ 10 − 10)	= 7·8808 − 10
	9·3480 − 10
log 3·451	= 0·5379
	8·8101 − 10
antilog 8·8101 − 10	= antilog $\bar{2}$·8101
	= 0·06459

(2) Evaluate $\sqrt[3]{\dfrac{2{\cdot}303 \times 69{\cdot}5}{26{\cdot}42 \times 0{\cdot}007 \times 0{\cdot}124}}$

Denominator

$$\begin{aligned}
\log 26{\cdot}42 &= 1{\cdot}4219\\
\log 0{\cdot}007 &= \bar{3}{\cdot}8451\\
\log 0{\cdot}124 &= \bar{1}{\cdot}0934\\
\hline
&\ \bar{2}{\cdot}3604
\end{aligned}$$

Numerator

$$\begin{aligned}
\log 2{\cdot}303 &= 0{\cdot}3623\\
\log 69{\cdot}05 &= 1{\cdot}8420\\
\hline
&\ 2{\cdot}2043\\
&\ \bar{2}{\cdot}3604\\
\hline
\text{log quotient} &= 3{\cdot}8439
\end{aligned}$$

$$\begin{aligned}
\log \sqrt[3]{} &= 1/3\ (3{\cdot}8439)\\
&= 1{\cdot}2813\\
\text{antilog} \sqrt[3]{} &= \underline{19{\cdot}11}
\end{aligned}$$

Decibel Conversions

Tables 11 and 12 provide a quick means of converting any power ratio P_2/P_1 or voltage ratio E_2/E_1 into decibels N, and vice versa, without the use of logarithmic tables. These tables have been derived from the basic formulæ

$$N = 10 \log_{10} (P_2/P_1)$$

or

$$N = 20 \log_{10} (E_2/E_1)$$

and, conversely,

$$P_2/P_1 = \text{antilog } (N/10)$$

$$E_2/E_1 = \text{antilog } (N/20)$$

To convert a power ratio greater than the ratios given in the table, divide the ratio by 10 in succession, until the quotient falls within the ratio column. Read the corresponding number of decibels, and add +10 db for each division by 10.

Examples

(1) To express a power ratio of 580 in decibels.

$$\frac{580}{10 \times 10} = 5{\cdot}8$$

From the table

$$\begin{aligned}
\text{A power ratio of } 5{\cdot}8 &= 7{\cdot}634 \text{ db}\\
\text{A power ratio of } 580 &= 7{\cdot}634 + 10 + 10 \text{ db}\\
&= \underline{27{\cdot}63 \text{ db}}
\end{aligned}$$

To convert a power ratio smaller than the ratios given in the table, multiply by 10 in succession, until the product falls within the ratio column. Read the corresponding number of decibels, and add −10 db for each multiplication by 10.

(2) To express a power ratio of 0·036 in decibels.

$$0{\cdot}036 \times 10 \times 10 = 3{\cdot}6$$

$$\begin{aligned}
\text{A power ratio of } 3{\cdot}6 &= 5{\cdot}563 \text{ db}\\
\text{A power ratio of } 0{\cdot}036 &= 5{\cdot}563 - 10 - 10 \text{ db}\\
&= \underline{-14{\cdot}44 \text{ db}}
\end{aligned}$$

TABLE 11.—CONVERSION OF POWER, VOLTAGE OR CURRENT RATIOS TO DECIBELS

Power Ratio	db	Power Ratio	db	Power Ratio	db	Power Ratio	db
1·0	0·000	3·3	5·185	5·6	7·482	7·9	8·976
1·1	0·414	3·4	5·315	5·7	7·559	8·0	9·031
1·2	0·792	3·5	5·441	5·8	7·634	8·1	9·085
1·3	1·139	3·6	5·563	5·9	7·709	8·2	9·138
1·4	1·461	3·7	5·682	6·0	7·782	8·3	9·191
1·5	1·761	3·8	5·798	6·1	7·853	8·4	9·243
1·6	2·041	3·9	5·911	6·2	7·924	8·5	9·294
1·7	2·304	4·0	6·021	6·3	7·993	8·6	9·345
1·8	2·553	4·1	6·128	6·4	8·062	8·7	9·395
1·9	2·788	4·2	6·232	6·5	8·129	8·8	9·445
2·0	3·010	4·3	6·335	6·6	8·195	8·9	9·494
2·1	3·222	4·4	6·435	6·7	8·261	9·0	9·542
2·2	3·424	4·5	6·532	6·8	8·325	9·1	9·590
2·3	3·617	4·6	6·628	6·9	8·388	9·2	9·638
2·4	3·802	4·7	6·721	7·0	8·451	9·3	9·685
2·5	3·979	4·8	6·812	7·1	8·513	9·4	9·731
2·6	4·150	4·9	6·902	7·2	8·573	9·5	9·777
2·7	4·314	5·0	6·990	7·3	8·633	9·6	9·823
2·8	4·472	5·1	7·076	7·4	8·692	9·7	9·868
2·9	4·624	5·2	7·160	7·5	8·751	9·8	9·912
3·0	4·771	5·3	7·243	7·6	8·808	9·9	9·956
3·1	4·914	5·4	7·324	7·7	8·865	10·0	10·000
3·2	5·051	5·5	7·404	7·8	8·921		

To convert a voltage or current ratio, treat the value in the power ratio column as a voltage or current ratio, and double the corresponding number of decibels read in the decibel column. If the ratio is outside the range of the table, follow the procedure outlined above.

(3) To express a voltage ratio of 75 in decibels.

$$\frac{75}{10} = 7{\cdot}5$$

A power ratio of 7·5 = 8·751 db
A voltage ratio of 7·5 = 8·751 × 2 = 17·50 db
A voltage ratio of 75 = 17·5 + 10 = 27·5 db

To convert values intermediate between 10 and 100 db, not included in the table, proceed as follows:

(1) Select from the decibel column the nearest value below the given decibels, and note the corresponding ratio.
(2) Subtract the selected decibels from the given decibels, and note the corresponding ratio.
(3) Multiply the two ratios thus found.

TABLE 12.—CONVERSION OF DECIBELS TO POWER, VOLTAGE OR CURRENT RATIOS

Gain		+ *db.*	*Loss*	
Power Ratio	*Voltage or Current Ratio*		*Power Ratio*	*Voltage or Current Ratio*
1·000	1·000	0	1·0000	1·0000
1·259	1·122	1	0·7943	0·8193
1·585	1·259	2	0·6310	0·7943
1·995	1·413	3	0·5012	0·7079
2·512	1·585	4	0·3981	0·6310
3·162	1·778	5	0·3162	0·5623
3·981	1·995	6	0·2512	0·5012
5·012	2·239	7	0·1995	0·4467
6·310	2·512	8	0·1585	0·3981
7·943	2·818	9	0·1259	0·3548
10	3·162	10	10^{-1}	$3{\cdot}162 \times 10$
10^2	10	20	10^{-2}	10^{-1}
10^3	$3{\cdot}162 \times 10$	30	10^{-3}	$3{\cdot}162 \times 10^{-1}$
10^4	10^2	40	10^{-4}	10^{-2}
10^5	$3{\cdot}162 \times 10^2$	50	10^{-5}	$3{\cdot}162 \times 10^{-2}$
10^6	10^3	60	10^{-6}	10^{-3}
10^7	$3{\cdot}162 \times 10^3$	70	10^{-7}	$3{\cdot}162 \times 10^{-3}$
10^8	10^4	80	10^{-8}	10^{-4}
10^9	$3{\cdot}162 \times 10^4$	90	10^{-9}	$3{\cdot}162 \times 10^{-4}$
10^{10}	10^5	100	10^{-10}	10^{-5}

Examples

(1) To express a gain of 28 decibels as a power ratio.
From the table :

Power ratio for + 20 db = 100
Power ratio for + 8 db = 6·31
Power ratio for + 28 db = 100 × 6·31 = <u>631</u>

(2) To express a level of 15 decibels below zero level (1 mW) as power output in milliwatts.
From the table :

Power ratio for −10 db = 0·1
Power ratio for − 5 db = 0·3162
Power output (reference 1 mW) = 0·1 × 0·3162 × 1
= <u>0·0316 mW</u>

(3) To express a voltage gain of 44 decibels as output voltage when the input level is 0·5 volts. Input and output impedances are assumed to be equal.

From the table:

Voltage ratio for +40 db	= 100
Voltage ratio for + 4 db	= 1·585
Output voltage (reference 0·5 V)	= 100 × 1·585 × 0·5
	= 79·25 volts

Slide-rule Calculations

The principal scales on the face of a slide rule are divided into lengths proportional to the logarithms of the numbers engraved opposite the lengths. On a 10-in. rule the full scale from 1 to 10 corresponds to $\log_{10} 10$, and the length in inches corresponding to any number between 1 and 10 is equal to the logarithm of that number multiplied by 10.

Multiplication of two numbers by addition of their logarithms is performed directly by adding the length representing one number on the sliding scale to the length representing the other on the fixed scale. Fig. 135 (*a*), with the aid of an example, will illustrate the principle of this operation. To multiply 3·2 × 2·5 set 1 on the sliding scale opposite 3·2 on the fixed scale. Opposite 2·5 on the sliding scale, read off the product 8·0 on the fixed scale.

Similarly, to divide two numbers by subtracting their logarithms, the lengths are subtracted, as in Fig. 135 (*b*). To divide 8·0 by 3·2 we set 3·2 on the fixed scale opposite 8·0 on the sliding scale. Opposite 1 on the fixed scale read off the quotient 2·5 on the sliding scale. The method of approximating to determine the position of the decimal point has been described in an earlier section.

The ordinary straight slide rule has four scales on the front face which will be referred to in the sections that follow as A, B, C, D. A and D are fixed to the body of the scale; B and C are on the slide. A and B are similar; C and D are also similar. On the reverse of the slide are three more scales for calculating sines, tangents and logarithms, which will be referred to as S, T and L respectively.

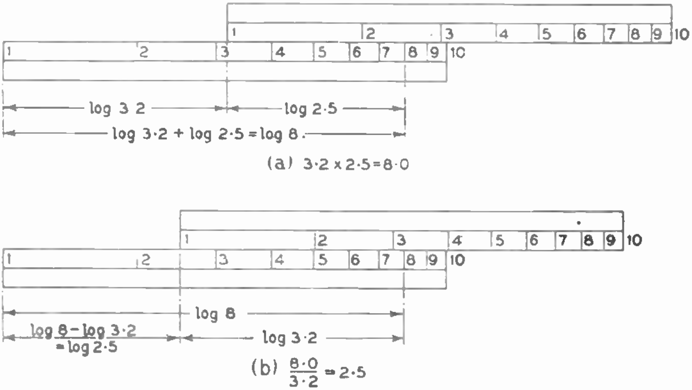


Fig. 135.—The Fixed and Sliding Scales of a Slide Rule, showing: (*a*) the Method of Multiplication, and (*b*) the Method of Division.

Multiplication and Division

When the numerator and denominator both contain two or more factors, the number of movements of the slide can be reduced by cross dividing and multiplying alternately.

Example

$$\frac{0{\cdot}038 \times 625}{89{\cdot}6 \times 0{\cdot}00011}$$

Approximation for the decimal point 2700.

(i) Set hair line on cursor over 3·8 on Scale D.
(ii) Slide Scale C until 8·96 appears under hair line.
(iii) Set hair line on cursor over 6·25 on Scale C.
(iv) Slide Scale C until 1·1 appears under hair line.
(v) Read off 241 under 1 on Scale D.

Answer to 3 significant figures = 2,410

Squares

Scales C and D are each twice the length of Scales A and B. Since $\log x^2 = 2 \log x$, numbers on Scale A are the squares of the numbers opposite them on Scale D.

Example

$$0{\cdot}0627^2$$

Approximation for the decimal point 0·0036.

(i) Set hair line on cursor over 6·27 on Scale D.
(ii) Read off 393 under hair line on Scale A.

Answer to 3 significant figures = 0·00393

Square Roots

Numbers on Scale D are the square roots of the numbers opposite them on Scale A. Use the left-hand half of Scale A if the number of figures to the left of the decimal point is odd, and the right-hand half if it is even. When the number is less than unity, use the left-hand half of Scale A if the number of 0s to the right of the decimal point is odd, and the right-hand half if it is even.

Examples

(1) $$\sqrt{0{\cdot}00548}$$

Approximation for the decimal point 0·07.

(i) Set hair line on cursor over 5·48 on right-hand half of Scale A.
(ii) Read off 74 under hair line on Scale D.

Answer to 3 significant figures 0·074

(2) $$\sqrt{39500}$$

Approximation for the decimal point 200.

(i) Set hair line on cursor over 395 on left-hand half of Scale A.
(ii) Read off 199 under hair line on Scale D.

Answer to 3 significant figures 199

Cubes

A cube is obtained on Scale A from the number on Scale D in the following manner:

Example

$$2{\cdot}56^3$$

(i) Move Scale C until 1 (or 10 if number >316 . . .) is opposite 2·56 on Scale D.
(ii) Set hair line on cursor over 2·56 on Scale B.
(iii) Read off answer on Scale A.

Answer 16·8

Cube Roots

The process is the reverse of cubing. The root is obtained on Scale D from the number of Scale A.

Example

$$\sqrt[3]{87}$$

(i) Set hair line on cursor over 87 on Scale A.
(ii) Move Scale C until 1 (or 10 if number >464 . . .) is opposite the same number on D as appears under cursor on B.
(iii) Read off answer on Scale D.

Answer 4·43

Sines

Use Scale S on back of slide for the angle, and read the sine on Scale B, and vice versa. Note that for angles from approximately ½° to 5½° the figures for the sine must be prefixed by 0·0, and for angles above about 5½° by 1.

Example

$$\sin 18^\circ 30'$$

(i) Withdraw slide to right and set 18° 30′ on Scale S below the datum mark on the body of the scale.
(ii) Reverse rule and read off 3·17 on Scale B below 10 on Scale A at right-hand end.

Answer $\dot{0}{\cdot}3\dot{1}7$

Cosines

The cosine of an angle is equal to the sine of the complementary angle θ. Use the procedure for the sine and find $\sin (90^\circ - \theta)$.

Example

$$\cos 62^\circ$$

(i) $\cos 62^\circ = \sin(90^\circ - 62^\circ) = \sin 28^\circ$.
(ii) Withdraw slide to the right and set 28 on Scale S below the datum mark on the body of the rule.
(iii) Reverse rule and read off 4·7 on Scale B below 10 on Scale A at right-hand end.

Answer 0·47

Tangents

Use Scale T on back of slide for the angle, and read the tangent on Scale C above 1 on Scale D, and vice versa. Note that for angles between approximately $\frac{1}{2}°$ and $5\frac{1}{2}°$ the figures for the tangent must be prefixed by 0·0, and for angles between $5\frac{1}{2}°$ and 45° by 0. For angles between 45° and 90° the tangent varies between 1 and infinity. When the angle exceeds 45°, subtract the angle from 90° and read the tangent on Scale D below 10 on Scale C.

Example

$$\tan 63° \, 35'$$

(i) Since the angle is greater than 45°, use the complementary angle. $90° - 63° \, 35' = 26° \, 25'$.

(ii) Withdraw slide to left and set 26° 25′ above datum mark on body of rule.

(iii) Reverse the rule and read off 2·02 on Scale D below 10 at right-hand end of Scale C.

Answer 2·02

Cotangents

The contangent of an angle θ is equal to the tangent of its complementary angle. When the angle is less than 45°, find $\tan \theta$ and read the tangent on Scale D below 10 on Scale C. When the angle is greater than 45°, find $\tan (90° - \theta)$ and read the tangent on Scale C above 1 on Scale D.

Example

$$\cot 21° \, 50$$

(i) Withdraw slide to left and set 21° 50′ above datum mark on body on rule.

(ii) Reverse rule and read off 2·5 on Scale D below 1 on Scale C.

Answer 2·5

Common Logarithms

Use Scale D for the number and Scale L on the back of the slide for the mantissa of the corresponding logarithm, and vice versa. The index of the logarithm is determined in the manner previously described for logarithms.

Examples

(1) $$\log_{10} 74{\cdot}6$$

(i) Since the number lies between 10 and 100, the index will be 1.

(ii) Set 1 on Scale C opposite 7·46 on Scale D.

(iii) Reverse rule and read off 8·73 on Scale L above datum mark on body of rule.

Answer 1·873

(2) $\log_{10} 0{\cdot}0382$

(i) Since the number lies between 0·01 and 0·1, the index will be 2.
(ii) Set 1 on Scale C opposite 3·82 on Scale D.
(iii) Reverse rule and read off 5·82 on Scale C above datum mark on body of rule.

Answer $\bar{2}{\cdot}582$

Complex Quantities

Complex quantities are readily resolved with the aid of the j operator, Just as the symbol $-$ prefixed to a quantity representing a vector denotes that the vector is displaced by an angle of 180°, so the symbol $\sqrt{-1}$, represented by the prefix j, denotes displacement through 90°. It follows that

$$\begin{array}{rrl}
j = & \sqrt{-1} & \text{rotation through } 90^\circ \\
j \times j = +j^2 = & -1 & \text{rotation through } 180^\circ \\
-j \times j = -j^2 = & +1 & \text{rotation through } 0^\circ \\
-j \times -j = +j^2 = & -1 & \text{rotation through } 180^\circ \\
j \times j \times j = +j^3 = & -1\sqrt{-1} & \text{rotation through } 270^\circ \\
j \times j \times j \times j = +j^4 = & +1 & \text{rotation through } 360^\circ
\end{array}$$

A vector displaced by an angle intermediate between 0° and 90° can be regarded as the resultant of two component vectors a and b at 90° with respect to each other, and is designated by the complex quantity $a + jb$, of which a is termed the real part and jb the imaginary part. For example, the vector representing the impedance Z of a resistance R and reactance X in series is the resultant of two vectors R and jX, and forms the hypotenuse of a right-angled triangle, of which R and jX are the two sides (see Fig. 136). Hence:

$$Z = R + jX$$

$$|Z| = \sqrt{R^2 + X^2}$$

where the phase angle

$$\phi = \text{arc tan } (X/R)$$

Complex expressions must be added vectorially by treating the real and imaginary components separately. Addition, subtraction, multiplication and division are performed according to the ordinary rules of algebra. The scalar value of the resultant can then be calculated by taking the square root of the sum of the squares of the real and imaginary quantities.

Addition

$$Z_1 + Z_2 = (R_1 + jX_1) + (R_2 + jX_2) = (R_1 + R_2) + j(X_1 + X_2)$$

Subtraction

$$Z_1 - Z_2 = (R_1 + jX_1) - (R_2 + jX_2) = (R_1 - R_2) + j(X_1 - X_2)$$

Multiplication

$$Z_1 Z_2 = (R_1 + jX_1)(R_2 + jX_2) = (R_1 R_2 - X_1 X_2) + j(X_1 R_2 + R_1 X_2)$$

Division

$$Z_1/Z_2 = (R_1 + jX_1)/(R_2 + jX_2) = (R_1 R_2 + X_1 X_2)/(R_2^2 + X_2^2) + j(X_1 R_2 - R_1 X_2)/(R_2^2 + X_2^2)$$

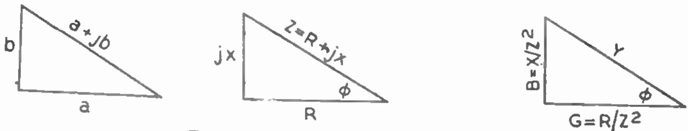


FIG. 136.—USE OF J OPERATOR.

Conjugate Numbers

To split a complex expression, such as $1/(R + jX)$, into real and imaginary components, multiply numerator and denominator by the conjugate of $R + jX$, i.e., $R - jX$.

Then since $(R + jX)(R - jX) = R^2 + X^2$

$$1/(R + jX) = R/(R^2 + X^2) - jX/(R^2 + X^2)$$

In dealing with parallel circuits involving complex quantities, it is easier to proceed as follows. The admittance of impedances in parallel is the algebraic sum of the separate admittances, $Y = 1/Z_1 + 1/Z_2 + \ldots$.

(1) Find the separate conductances $G = R/Z^2$ and susceptances $B = X/Z^2$ of each path.
(2) Find the combined admittance $Y = (G_1 + jB_1) + (G_2 + jB_2) + \ldots$
(3) Take the reciprocal of the combined admittance to find the effective impedance.

Example

What is the effective impedance and phase angle of two impedances $3 + j4$ and $2 - j1$ connected in parallel?

	Branch A	Branch B
(impedance)²	$Z^2 = 3^2 + 4^2 = 25$	$Z^2 = 2^2 - 1^2 = 3$
conductance	$G_1 = 3/25 = 0{\cdot}12$	$G_2 = 2/3 = 0{\cdot}67$
susceptance	$B_1 = 4/25 = 0{\cdot}16$	$B_2 = 1/3 = 0{\cdot}33$

Combined admittance of branches A and B

$$Y = (G_1 + G_2) + j(B_1 + B_2)$$
$$= 0{\cdot}79 - j0{\cdot}17$$

Effective impedance of branches A and B

$$Z = \frac{1}{0{\cdot}79 - j0{\cdot}17}$$

$$= \frac{0{\cdot}79 + j0{\cdot}17}{(0{\cdot}79 - j0{\cdot}17)(0{\cdot}79 + j0{\cdot}17)}$$

$$= \frac{0{\cdot}79 + j0{\cdot}17}{0{\cdot}62 + 0{\cdot}029} = \underline{1{\cdot}21 + j0{\cdot}26}$$

LOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	4	8	12	17	21	25	29	33	37
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	4	8	11	15	19	23	26	30	34
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	3	7	10	14	17	21	24	28	31
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	3	6	10	13	16	19	23	26	29
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	3	6	9	12	15	18	21	24	27
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	3	6	8	11	14	17	20	22	25
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	3	5	8	11	13	16	18	21	24
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	2	5	7	10	12	15	17	20	22
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	2	5	7	9	12	14	16	19	21
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	2	4	7	9	11	13	16	18	20
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	2	4	6	8	11	13	15	17	19
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	2	4	6	8	10	12	14	16	18
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	2	4	6	8	10	12	14	15	17
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	2	4	6	7	9	11	13	15	17
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	2	4	5	7	9	11	12	14	16
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	2	3	5	7	9	10	12	14	15
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	2	3	5	7	8	10	11	13	15
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	2	3	5	6	8	9	11	13	14
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	2	3	5	6	8	9	11	12	14
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	1	3	4	6	7	9	10	12	13
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	1	3	4	6	7	9	10	11	13
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	1	3	4	6	7	8	10	11	12
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	1	3	4	5	7	8	9	11	12
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	1	3	4	5	6	8	9	10	12
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	1	3	4	5	6	8	9	10	11
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	1	2	4	5	6	7	9	10	11
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	1	2	4	5	6	7	8	10	11
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	1	2	3	5	6	7	8	9	10
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	1	2	3	5	6	7	8	9	10
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	1	2	3	4	5	7	8	9	10
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	1	2	3	4	5	6	8	9	10
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	1	2	3	4	5	6	7	8	9
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	1	2	3	4	5	6	7	8	9
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	1	2	3	4	5	6	7	8	9
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	1	2	3	4	5	6	7	8	9
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	1	2	3	4	5	6	7	8	9
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	1	2	3	4	5	6	7	7	8
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	1	2	3	4	5	5	6	7	8
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	1	2	3	4	4	5	6	7	8
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	1	2	3	4	4	5	6	7	8
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	1	2	3	3	4	5	6	7	8
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	1	2	3	3	4	5	6	7	8
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	1	2	2	3	4	5	6	7	7
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	1	2	2	3	4	5	6	6	7
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	1	2	2	3	4	5	6	6	7

Hyperbolic Logarithms = Common Logarithms × 2·30258.

LOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	1	2	2	3	4	5	5	6	7
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	1	2	2	3	4	5	5	6	7
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	1	2	2	3	4	5	5	6	7
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	1	1	2	3	4	4	5	6	7
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	1	1	2	3	4	4	5	6	7
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	1	1	2	3	4	4	5	6	6
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	1	1	2	3	4	4	5	6	6
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	1	1	2	3	3	4	5	6	6
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	1	1	2	3	3	4	5	5	6
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	1	1	2	3	3	4	5	5	6
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	1	1	2	3	3	4	5	5	6
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	1	1	2	3	3	4	5	5	6
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	1	1	2	3	3	4	5	5	6
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	1	1	2	3	3	4	4	5	6
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	1	1	2	2	3	4	4	5	6
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	1	1	2	2	3	4	4	5	6
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	1	1	2	2	3	4	4	5	5
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	1	1	2	2	3	4	4	5	5
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	1	1	2	2	3	4	4	5	5
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	1	1	2	2	3	4	4	5	5
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	1	1	2	2	3	3	4	5	5
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	1	1	2	2	3	3	4	5	5
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	1	1	2	2	3	3	4	4	5
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	1	1	2	2	3	3	4	4	5
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	1	1	2	2	3	3	4	4	5
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	1	1	2	2	3	3	4	4	5
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	1	1	2	2	3	3	4	4	5
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	1	1	2	2	3	3	4	4	5
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	1	1	2	2	3	3	4	4	5
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	1	1	2	2	3	3	4	4	5
85	9294	9299	9304	9309	9315	9320	9325	9330	9535	9340	1	1	2	2	3	3	4	4	5
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	1	1	2	2	3	3	4	4	5
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	0	1	1	2	2	3	3	4	4
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	0	1	1	2	2	3	3	4	4
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	0	1	1	2	2	3	3	4	4
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	0	1	1	2	2	3	3	4	4
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	0	1	1	2	2	3	3	4	4
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	0	1	1	2	2	3	3	4	4
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	0	1	1	2	2	3	3	4	4
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	0	1	1	2	2	3	3	4	4
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	0	1	1	2	2	3	3	4	4
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	0	1	1	2	2	3	3	4	4
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	0	1	1	2	2	3	3	4	4
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	0	1	1	2	2	3	3	4	4
99	0056	9961	9965	9969	9974	9978	9983	9987	9991	9996	0	1	1	2	2	3	3	3	4

Common Logarithms = Hyperbolic Logarithms × 0·43429

F

ANTILOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
·00	1000	1002	1005	1007	1009	1012	1014	1016	1019	1021	0	0	1	1	1	1	2	2	2
·01	1023	1026	1028	1030	1033	1035	1038	1040	1042	1045	0	0	1	1	1	1	2	2	2
·02	1047	1050	1052	1054	1057	1059	1062	1064	1067	1069	0	0	1	1	1	1	2	2	2
·03	1972	1074	1076	1079	1081	1084	1086	1089	1091	1094	0	0	1	1	1	1	2	2	2
·04	1095	1099	1102	1104	1107	1109	1112	1114	1117	1119	0	1	1	1	1	2	2	2	2
·05	1122	1125	1127	1130	1132	1135	1138	1140	1143	1146	0	1	1	1	1	2	2	2	2
·06	1148	1151	1153	1156	1159	1161	1164	1167	1169	1172	0	1	1	1	1	2	2	2	2
·07	1175	1178	1180	1183	1186	1189	1191	1194	1197	1199	0	1	1	1	1	2	2	2	2
·08	1202	1205	1208	1211	1213	1116	1219	1222	1225	1227	0	1	1	1	1	2	2	2	3
·09	1230	1233	1236	1239	1242	1245	1247	1250	1253	1256	0	1	1	1	1	2	2	2	3
·10	1259	1262	1265	1268	1271	1274	1276	1279	1282	1285	0	1	1	1	1	2	2	2	3
·11	1288	1291	1294	1297	1300	1303	1306	1309	1312	1315	0	1	1	1	2	2	2	2	3
·12	1318	1321	1324	1327	1330	1334	1337	1340	1343	1346	0	1	1	1	2	2	2	2	3
·13	1349	1352	1355	1358	1361	1365	1368	1371	1374	1377	0	1	1	1	2	2	2	3	3
·14	1380	1384	1387	1390	1393	1396	1400	1403	1406	1409	0	1	1	1	2	2	2	3	3
·15	1413	1416	1419	1422	1426	1429	1432	1435	1439	1442	0	1	1	1	2	2	2	3	3
·16	1445	1449	1452	1455	1459	1462	1466	1469	1472	1476	0	1	1	1	2	2	2	3	3
·17	1479	1483	1486	1489	1493	1496	1500	1503	1507	1510	0	1	1	1	2	2	2	3	3
·18	1514	1517	1521	1524	1528	1531	1535	1538	1542	1545	0	1	1	1	2	2	2	3	3
·19	1549	1552	1556	1560	1563	1567	1570	1574	1578	1581	0	1	1	1	2	2	3	3	3
·20	1585	1589	1592	1506	1600	1603	1607	1611	1614	1618	0	1	1	1	2	2	3	3	3
·21	1622	1626	1629	1633	1637	1641	1644	1648	1652	1656	0	1	1	2	2	2	3	3	3
·22	1660	1663	1667	1671	1675	1679	1683	1687	1690	1694	0	1	1	2	2	2	3	3	3
·23	1698	1702	1706	1710	1714	1718	1722	1726	1730	1734	0	1	1	2	2	2	3	3	4
·24	1738	1742	1746	1750	1754	1758	1762	1766	1770	1774	0	1	1	2	2	2	3	3	4
·25	1778	1782	1786	1791	1795	1799	1803	1807	1811	1816	0	1	1	2	2	2	3	3	4
·26	1820	1824	1828	1832	1837	1841	1845	1849	1854	1858	0	1	1	2	2	3	3	3	4
·27	1862	1866	1871	1875	1879	1884	1888	1892	1897	1901	0	1	1	2	2	3	3	3	4
·28	1905	1910	1914	1919	1923	1928	1932	1936	1941	1945	0	1	1	2	2	3	3	4	4
·29	1950	1954	1959	1963	1968	1972	1977	1982	1986	1991	0	1	1	2	2	3	3	4	4
·30	1995	2000	2004	2009	2014	2108	2023	2028	2032	2037	0	1	1	2	2	3	3	4	4
·31	2042	2046	1051	2056	2061	2065	2070	2075	2080	2084	0	1	1	2	2	3	3	4	4
·32	2089	2094	2099	2104	2109	2113	2118	2123	2128	2133	0	1	1	2	2	3	3	4	4
·33	2138	2143	2148	2153	2158	2163	2168	2173	2178	2183	0	1	1	2	2	3	3	4	4
·34	2188	2193	2198	2203	2208	2213	2218	2223	2228	2234	1	1	2	2	3	3	4	4	5
·35	2239	2244	2249	2254	2259	2265	2270	2275	2280	2286	1	1	2	2	3	3	4	4	5
·36	2291	2296	2301	2307	2312	2317	2323	2328	2333	2339	1	1	2	2	3	3	4	4	5
·37	2344	2350	2355	2360	2366	2371	2377	2382	2388	2393	1	1	2	2	3	3	4	4	5
·38	2399	2404	2410	2415	2421	2427	2432	2438	2443	2449	1	1	2	2	3	3	4	4	5
·39	2455	2460	2466	2472	2477	2483	2489	2495	2500	2506	1	1	2	2	3	3	4	5	5
·40	2512	2518	2523	2529	2535	2541	2547	2553	2559	2564	1	1	2	2	3	4	4	5	5
·41	2570	2576	2582	2588	2594	2600	2606	2612	2618	2624	1	1	2	2	3	4	4	5	5
·42	2630	2636	2642	2649	2655	2661	2667	2673	2679	2685	1	1	2	2	3	4	4	5	6
·43	2692	2698	2704	2710	2716	2723	2729	2735	2742	2748	1	1	2	3	3	4	4	5	6
·44	2754	2761	2767	2773	2780	2786	2793	2799	2805	2812	1	1	2	3	3	4	4	5	6
·45	2818	2825	2831	2838	2844	2851	2858	2864	2871	2877	1	1	2	3	3	4	5	5	6
·46	2884	2891	2897	2904	2911	2917	2924	2931	2938	2944	1	1	2	3	3	4	5	5	6
·47	2951	2958	2965	2972	2979	2085	2992	2999	3006	3013	1	1	2	3	3	4	5	5	6
·48	3020	3027	3034	3041	3048	3055	3062	3069	3076	3083	1	1	2	3	4	4	5	6	6
·49	3090	3097	3105	3112	3119	3126	3133	3141	3148	3155	1	1	2	3	4	4	5	6	6

ANTILOGARITHMS

	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
·50	3162	3170	3177	3184	3192	3199	3206	3214	3221	3228	1	1	2	3	4	4	5	6	7
·51	3236	3243	3251	3258	3266	3273	3281	3289	3296	3304	1	2	2	3	4	5	5	6	7
·52	3311	3319	3327	3334	3342	3350	3357	3365	3373	3381	1	2	2	3	4	5	5	6	7
·53	3388	3396	3404	3412	3420	3428	3436	3443	3451	3459	1	2	2	3	4	5	6	6	7
·54	3467	3475	3483	3491	3499	3508	3516	3524	3532	3540	1	2	2	3	4	5	6	6	7
·55	3548	3556	3565	3573	3581	3589	3597	3606	3614	3622	1	2	2	3	4	5	6	7	7
·56	3631	3639	3648	3656	3664	3673	3681	3690	3698	3707	1	2	3	3	4	5	6	7	8
·57	3715	3724	3733	3741	3750	3758	3767	3776	3784	3793	1	2	3	3	4	5	6	7	8
·58	3802	3811	3819	3828	3837	3846	3855	3864	3873	3882	1	2	3	4	4	5	6	7	8
59	3890	3899	3908	3917	3926	3936	3945	3954	3963	3972	1	2	3	4	5	5	6	7	8
·60	3981	3990	3999	4009	4018	4027	4036	4046	4055	4064	1	2	3	4	5	6	6	7	8
·61	4074	4083	4093	4102	4111	4121	4130	4140	4150	4159	1	2	3	4	5	6	7	8	9
·62	4169	4178	4188	4198	4207	4217	4227	4236	4246	4256	1	2	3	4	5	6	7	8	9
·63	4266	4276	4285	4295	4305	4315	4325	4335	4345	4355	1	2	3	4	5	6	7	8	9
·64	4365	4375	4385	4395	4406	4416	4426	4436	4446	4457	1	2	3	4	5	6	7	8	9
·65	4467	4477	4487	4498	4508	4519	4529	4539	4550	4560	1	2	3	4	5	6	7	8	9
·66	4571	4581	4592	4603	4613	4624	4634	4645	4656	4667	1	2	3	4	5	6	7	9	10
·67	4677	4688	4699	4710	4721	4732	4742	4753	4764	4775	1	2	3	4	5	7	8	9	10
·68	4786	4797	4808	4819	4831	4842	4853	4864	4875	4887	1	2	3	4	6	7	8	9	10
·69	4898	4909	4920	4932	4943	4955	4966	4977	4989	5000	1	2	3	5	6	7	8	9	10
·70	5012	5023	5035	5047	5058	5070	5082	5093	5105	5117	1	2	4	5	6	7	8	9	11
·71	5129	5140	5152	5164	5176	5188	5200	5212	5224	5236	1	2	4	5	6	7	8	10	11
·72	5248	5260	5272	5284	5297	5309	5321	5333	5346	5358	1	2	4	5	6	7	9	10	11
·73	5370	5383	5395	5408	5420	5433	5445	5458	5470	5483	1	3	4	5	6	8	9	10	11
·74	5495	5508	5521	5534	5546	5559	5572	5585	5598	5610	1	3	4	5	6	8	9	10	12
·75	5623	5636	5649	5662	5675	5689	5702	5715	5728	5741	1	3	4	5	7	8	9	10	12
·76	5754	5768	5781	5794	5808	5821	5834	5848	5861	5875	1	3	4	5	7	8	9	11	12
·77	5888	5902	5916	5929	5943	5957	5970	5984	5998	6012	1	3	4	5	7	8	10	11	12
·78	6026	6039	6053	6067	6081	6095	6109	6124	6138	6152	1	3	4	6	7	8	10	11	13
·79	6166	6180	6194	6209	6223	6237	6252	6266	6281	6295	1	3	4	6	7	9	10	11	13
·80	6310	6324	6339	6353	6368	6383	6397	6412	6427	6442	1	3	4	6	7	9	10	12	13
·81	6457	6471	6486	6501	6516	6531	6546	6561	6577	6592	2	3	5	6	8	9	11	12	14
·82	6607	6622	6637	6653	6668	6683	6699	6714	6730	6745	2	3	5	6	8	9	11	12	14
·83	6761	6776	6792	6808	6823	6839	6855	6871	6887	6902	2	3	5	6	8	9	11	13	14
·84	6918	6934	6950	6966	6982	6998	7015	7031	7047	7063	2	3	5	6	8	10	11	13	15
·85	7079	7096	7112	7129	7145	7161	7178	7194	7211	7228	2	3	5	7	8	10	12	13	15
·86	7244	7261	7278	7295	7311	7328	7345	7362	7379	7396	2	3	5	7	8	10	12	13	15
·87	7413	7430	7447	7464	7482	7499	7516	7534	7551	7568	2	4	5	7	9	10	12	14	16
·88	7586	7603	7621	7638	7656	7674	7691	7709	7727	7745	2	4	5	7	9	11	12	14	16
·89	7762	7780	7798	7816	7834	7852	7870	7889	7907	7925	2	4	5	7	9	11	13	14	16
·90	7943	7962	7980	7998	8017	8035	8054	8072	8091	8110	2	4	6	7	9	11	13	15	17
·91	8128	8147	8166	8185	8204	8222	8241	8260	8279	8299	2	4	6	8	9	11	13	15	17
·92	8318	8337	8356	8375	8395	8414	8433	8453	8472	8492	2	4	6	8	10	12	14	15	17
·93	8511	8531	8551	8570	8590	8610	8630	8650	8670	8690	2	4	6	8	10	12	14	16	18
·94	8710	8730	8750	8770	8790	8810	8831	8851	8872	8892	2	4	6	8	10	12	14	16	18
·95	8913	8933	8954	8974	8995	9016	9036	9057	9078	9099	2	4	6	8	10	12	15	17	19
·96	9120	9141	9162	9183	9204	9226	9247	9268	9290	9311	2	4	6	8	11	13	15	17	19
·97	9333	9354	9376	9397	9419	9441	9462	9484	9506	9528	2	4	7	9	11	13	15	17	20
·98	9550	9572	9594	9616	9638	9661	9683	9705	9727	9750	2	4	7	9	11	13	16	18	20
·99	9772	9795	9817	9840	9863	9886	9908	9931	9954	9977	2	5	7	9	11	14	16	18	20

TABLE OF ORDINARY LOGARITHMS OF NUMBERS TO THE BASE 10
From 1 to 100 with Characteristics

No.	Log.	No.	Log.	No.	Log.	No.	Log.	No.	Log
1	0·000 000	21	1·322 219	41	1·612 784	61	1·785 330	81	1·908 485
2	0·301 030	22	1·342 423	42	1·623 249	62	1·792 392	82	1·913 814
3	0·477 121	23	1·361 728	43	1·633 468	63	1·799 341	83	1·919 078
4	0·602 060	24	1·380 211	44	1·643 453	64	1·806 180	84	1·924 279
5	0·698 970	25	1·397 940	45	1·653 213	65	1·812 913	85	1·929 419
6	0·778 151	26	1·414 973	46	1·662 758	66	1·819 544	86	1·934 498
7	0·845 098	27	1·431 364	47	1·672 098	67	1·826 075	87	1·939 519
8	0·903 090	28	1·447 158	48	1·681 241	68	1·832 509	88	1·944 483
9	0·954 243	29	1·462 398	49	1·690 196	69	1·838 849	89	1·949 390
10	1·000 000	30	1·477 121	50	1·698 970	70	1·845 098	90	1·954 243
11	1·041 393	31	1·491 362	51	1·707 570	71	1·851 258	91	1·959 041
12	1·079 181	32	1·505 150	52	1·716 003	72	1·857 332	92	1·963 788
13	1·113 943	33	1·518 514	53	1·724 276	73	1·863 323	93	1·968 483
14	1·146 128	34	1·531 479	54	1·732 394	74	1·869 232	94	1·973 128
15	1·176 091	35	1·544 068	55	1·740 363	75	1·875 061	95	1·977 724
16	1·204 120	36	1·556 303	56	1·748 188	76	1·880 814	96	1·982 271
17	1·230 449	37	1·568 202	57	1·755 875	77	1·886 491	97	1·986 772
18	1·255 273	38	1·579 784	58	1·763 428	78	1·892 095	98	1·991 226
19	1·278 754	39	1·591 065	59	1·770 852	79	1·897 627	99	1·905 635
20	1·301 030	40	1·602 060	60	1·778 151	80	1·903 099	100	2·000 000

TABLE OF HYPERBOLIC LOGARITHMS OF NUMBERS
From 1 to 100

No.	Log.	No.	Log.	No.	Log.	No.	Log.	No.	No.
1	0·000 000	21	3·044 522 44	41	3·713 572 07	61	4·110 873 86	81	4·394 449 15
2	0·693 147 18	22	3·091 042 45	42	3·737 669 62	62	4·127 134 39	82	4·406 719 25
3	1·098 612 29	23	3·135 494 22	43	3·761 200 12	63	4·143 134 73	83	4·418 840 61
4	1·386 294 36	24	3·178 053 83	44	3·784 189 63	64	4·158 883 08	84	4·430 816 80
5	1·609 437 91	25	3·218 875 82	45	3·806 662 49	65	4·174 387 27	85	4·442 651 26
6	1·791 759 47	26	3·258 096 54	46	3·828 641 40	66	4·189 654 74	86	4·454 347 30
7	1·945 910 15	27	3·295 836 87	47	3·850 147 60	67	4·204 692 62	87	4·465 908 12
8	2·079 441 54	28	3·332 204 51	48	3·871 201 01	68	4·219 507 71	88	4·477 336 81
9	2·197 224 58	29	3·367 295 53	49	3·891 820 30	69	4·234 410 65	89	4·488 636 37
10	2·302 585 09	30	3·401 197 38	50	3·912 023 01	70	4·248 495 24	90	4·499 809 67
11	2·397 895 27	31	3·433 987 20	51	3·931 825 63	71	4·262 679 88	91	4·510 859 51
12	2·484 906 65	32	3·465 735 90	52	3·951 243 72	72	4·276 666 12	92	4·521 788 58
13	2·564 949 36	33	3·496 507 56	53	3·970 291 91	73	4·290 459 44	93	4·532 599 49
14	2·639 057 33	34	3·526 360 52	54	3·988 984 05	74	4·304 065 09	94	4·543 294 78
15	2·708 050 20	35	3·555 348 06	55	4·007 333 19	75	4·317 488 11	95	4·553 876 89
16	2·772 588 72	36	3·583 518 94	56	4·025 351 69	76	4·330 733 34	96	4·564 348 19
17	2·833 213 34	37	3·610 917 91	57	4·043 051 27	77	4·343 805 42	97	4·574 710 98
18	2·890 371 76	38	3·637 586 16	58	4·060 443 01	78	4·356 708 [illegible]	98	4·584 967 48
19	2·944 438 98	39	3·663 561 65	59	4·077 537 44	79	4·369 447 85	99	4·595 119 85
20	2·995 732 27	40	3·688 879 45	60	4·094 344 56	80	4·382 026 63	100	4·605 170 19

TABLES OF NATURAL SINES, COSINES, TANGENTS, AND COTANGENTS FROM 0° TO 90°. READ THE TABLE DOWNWARDS FOR *SINES* AND UPWARDS FOR *COSINES*.

NATURAL SINES

′	0°	1°	2°	3°	4°	5°	6°	7°	8°	9°	′
0	000 000	017 452	034 899	052 336	069 756	087 156	104 528	121 869	139 173	156 434	60
5	1 454	8 907	6 353	3 788	071 207	8 605	5 975	3 313	140 613	7 871	55
10	2 909	020 361	7 806	5 241	2 658	090 053	7 421	4 756	2 053	9 307	50
15	4 363	1 815	9 260	6 693	4 108	1 502	8 867	6 199	3 493	160 743	45
20	5 818	3 269	040 713	8 145	5 559	2 950	110 313	7 642	4 932	2 178	40
25	7 272	4 723	2 166	9 597	7 009	4 398	1 758	9 084	6 371	3 613	35
30	8 727	6 177	3 619	061 049	8 459	5 846	3 203	130 526	7 809	5 048	30
35	010 181	7 631	5 072	2 500	9 909	7 293	4 648	1 068	9 248	6 482	25
40	1 635	9 085	6 525	3 952	081 359	8 741	6 093	3 410	150 686	7 916	20
45	3 090	030 539	7 978	5 403	2 808	100 188	7 537	4 851	2 123	9 350	15
50	4 544	1 992	9 431	6 854	4 258	1 635	8 982	6 292	3 561	170 783	10
55	5 998	3 446	050 883	8 306	5 707	3 082	120 426	7 733	4 998	2 216	5
60	7 452	4 899	2 336	9 756	7 156	4 528	1 869	9 173	6 434	3 648	0
Cos.	89°	88°	87°	86°	85°	84°	83°	82°	81°	80°	

Sin.	10°	11°	12°	13°	14°	15°	16°	17°	18°	19°	
0	173 648	190 809	207 912	224 951	241 922	258 819	275 637	292 372	309 017	325 568	60
5	5 080	2 237	9 334	6 368	3 333	260 224	7 035	3 762	310 400	6 943	55
10	6 512	3 664	210 756	7 784	4 743	1 628	8 432	5 152	1 782	8 317	50
15	7 944	5 090	2 178	9 200	6 153	3 031	9 829	6 542	3 164	9 691	45
20	9 375	6 517	3 599	230 616	7 563	4 434	281 225	7 930	4 545	331 063	40
25	180 805	7 942	5 019	2 031	8 972	5 837	2 620	9 318	5 925	2 435	35
30	2 236	9 368	6 440	3 445	250 380	7 238	4 015	300 706	7 305	3 807	30
35	3 665	200 793	7 859	4 859	1 788	8 640	5 410	2 093	8 684	5 178	25
40	5 095	2 218	9 279	6 273	3 195	270 040	6 083	3 479	320 062	6 547	20
45	6 524	3 642	220 697	7 686	4 602	1 440	8 196	4 864	1 439	7 917	15
50	7 953	5 065	2 116	9 098	6 008	2 840	9 589	6 249	2 816	9 285	10
55	9 381	6 489	3 534	240 510	7 414	4 239	290 981	7 633	4 193	340 653	5
60	190 809	7 912	4 951	1 922	8 819	5 637	2 372	9 017	5 568	2 020	0
Cos.	79°	78°	77°	76°	75°	74°	73°	72°	71°	70°	

Sin.	20°	21°	22°	23°	24°	25°	26°	27°	28°	29°	
0	342 020	358 368	374 607	390 731	406 737	422 618	438 371	453 990	469 472	484 810	60
5	3 387	9 725	5 955	2 070	8 065	3 936	9 678	5 286	470 755	6 081	55
10	4 752	361 082	7 302	3 407	9 392	5 253	440 984	6 580	2 038	7 352	50
15	6 117	2 438	8 649	4 744	410 719	6 569	2 289	7 874	3 320	8 621	45
20	7 481	3 793	9 994	6 080	2 045	7 884	3 593	9 160	4 600	9 890	40
25	8 845	5 148	381 339	7 415	3 369	9 198	4 896	460 458	5 880	491 157	35
30	350 207	6 501	2 683	8 749	4 693	430 511	6 198	1 749	7 159	2 424	30
35	1 569	7 854	4 027	400 082	6 016	1 823	7 499	3 038	8 436	3 689	25
40	2 931	9 206	5 369	1 415	7 338	3 135	8 799	4 327	9 713	4 953	20
45	4 291	370 557	6 711	2 747	8 660	4 445	450 098	5 615	480 989	6 217	15
50	5 651	1 908	8 052	4 078	9 980	5 755	1 397	6 901	2 263	7 479	10
55	7 010	3 258	9 392	5 408	421 300	7 063	2 694	8 187	3 537	8 740	5
60	8 368	4 607	390 731	6 737	2 618	8 371	3 990	9 472	4 810	500 000	0
Cos.	69°	68°	67°	66°	65°	64°	63°	62°	61°	60°	

Sin.	30°	31°	32°	33°	34°	35°	36°	37°	38°	39°	
0	500 000	515 038	529 919	544 639	559 193	573 576	587 785	601 815	615 661	620 320	60
5	1 259	6 284	531 152	5 858	560 398	4 767	8 961	2 976	6 807	630 450	55
10	2 517	7 529	2 384	7 076	1 602	5 957	590 136	4 136	7 951	1 578	50
15	3 774	8 773	3 615	8 293	2 805	7 145	1 310	5 294	9 094	2 705	45
20	5 030	520 016	4 844	9 509	4 007	8 332	2 482	6 451	620 235	3 831	40
25	6 285	1 258	6 072	550 724	5 207	9 518	3 653	7 607	1 376	4 955	35
30	7 538	2 499	7 300	1 937	6 406	580 703	4 823	8 761	2 515	6 078	30
35	8 791	3 738	8 256	3 149	7 604	1 886	5 991	9 915	3 652	7 200	25
40	510 043	4 977	9 751	4 360	8 801	3 069	7 159	611 067	4 789	8 320	20
45	1 293	6 214	540 974	5 570	9 997	4 250	8 325	2 217	5 923	9 439	15
50	2 543	7 450	2 197	6 779	571 191	5 429	9 489	3 367	7 057	640 557	10
55	3 791	8 685	3 419	7 987	2 384	6 608	600 653	4 515	8 129	1 673	5
60	5 038	9 919	4 639	9 193	2 576	7 785	1 815	5 661	9 320	2 788	0
	59°	58°	57°	56°	55°	54°	53°	52°	51°	50°	

NATURAL COSINES

NATURAL SINES

′	40°	41°	42°	43°	44°	45°	46°	47°	48°	49°	′
0	642 788	656 059	669 131	681 998	694 658	707 107	719 340	731 354	743 145	754 710	60
5	3 901	7 156	670 211	3 061	5 704	8 134	720 349	2 345	4 117	5 663	55
10	5 013	8 252	1 289	4 123	6 748	9 161	1 357	3 334	5 088	6 615	50
15	6 124	9 346	2 367	5 183	7 790	710 185	2 364	4 323	6 057	7 565	45
20	7 233	660 439	3 443	6 242	8 832	1 209	3 369	5 309	7 025	8 514	40
25	8 341	1 530	4 517	7 299	9 871	2 230	4 372	6 294	7 991	9 461	35
30	9 448	2 620	5 590	8 355	700 909	3 250	5 374	7 277	8 956	760 406	30
35	650 553	3 709	6 662	9 409	1 946	4 269	6 375	8 259	9 919	1 350	25
40	1 657	4 796	7 732	690 462	2 981	5 286	7 374	9 239	750 880	2 292	20
45	2 760	5 882	8 801	1 513	4 015	6 302	8 371	740 218	1 840	3 232	15
50	3 861	6 966	9 868	2 563	5 047	7 316	9 367	1 915	2 798	4 171	10
55	4 961	8 049	680 934	3 611	6 078	8 329	730 361	2 171	3 755	5 109	5
60	6 059	9 131	1 998	4 658	7 107	9 340	1 354	3 145	4 710	6 044	0
Cos.	49°	48°	47°	46°	45°	44°	43°	42°	41°	40°	
Sin.	50°	51°	52°	53°	54°	55°	56°	57°	58°	59°	
0	766 044	777 146	788 011	798 636	809 017	819 152	829 038	838 671	848 048	857 167	60
5	6 979	8 060	8 905	9 510	9 871	9 985	9 850	9 462	8 818	7 915	55
10	7 911	8 973	9 798	800 383	810 723	820 817	830 661	840 251	9 586	8 662	50
15	8 842	9 884	790 690	1 254	1 574	1 647	1 470	1 039	850 352	9 406	45
20	9 771	780 794	1 579	2 123	2 423	2 475	2 277	1 825	1 117	850 149	40
25	770 699	1 702	2 467	2 991	3 270	3 302	3 082	2 609	1 879	0 890	35
30	1 625	2 608	3 353	3 857	4 116	4 126	3 886	3 391	2 640	1 629	30
35	2 549	3 513	4 238	4 721	4 959	4 949	4 688	4 172	3 399	2 366	25
40	3 472	4 416	5 121	5 584	5 801	5 770	5 488	4 951	4 156	3 102	20
45	4 393	5 317	6 002	6 445	6 642	6 590	6 286	5 728	4 912	3 836	15
50	5 312	6 217	6 882	7 304	7 480	7 407	7 083	6 503	5 665	4 567	10
55	6 230	7 114	7 759	8 161	8 317	8 223	7 878	7 277	6 417	5 297	5
60	7 146	8 011	8 636	9 017	9 152	9 038	8 671	8 048	7 167	6 025	0
Cos.	39°	38°	37°	36°	35°	34°	33°	32°	31°	30°	
Sin.	60°	61°	62°	63°	64°	65°	66°	67°	68°	69°	
0	866 025	874 620	882 948	891 007	898 794	906 308	913 545	920 505	927 184	933 580	60
5	6 752	5 324	3 629	1 666	9 431	6 922	4 136	1 072	7 728	4 101	55
10	7 476	6 026	4 309	2 323	900 065	7 533	4 725	1 638	8 270	4 619	50
15	8 199	6 727	4 988	2 979	0 698	8 143	5 311	2 201	8 810	5 135	45
20	8 920	7 425	5 664	3 633	1 329	8 751	5 896	2 762	9 348	5 650	40
25	9 639	8 122	6 338	4 284	1 958	9 357	6 479	3 322	9 884	6 162	35
30	870 356	8 817	7 011	4 934	2 585	9 961	7 060	3 880	930 418	6 672	30
35	1 071	9 510	7 681	5 582	3 210	910 563	7 639	4 435	0 950	7 181	25
40	1 784	880 201	8 350	6 229	3 834	1 164	8 216	4 989	1 480	7 687	20
45	2 496	0 891	9 017	6 873	4 455	1 762	8 791	5 541	2 008	8 191	15
50	3 206	1 578	9 682	7 515	5 075	2 358	9 364	6 090	2 534	8 694	10
55	3 914	2 264	890 345	8 156	5 692	2 953	9 936	6 638	3 058	9 194	5
60	4 620	2 948	1 007	8 794	6 308	3 545	920 505	7 184	3 580	9 693	0
Cos.	29°	28°	27°	26°	25°	24°	23°	22°	21°	20°	
Sin.	70°	71°	72°	73°	74°	75°	76°	77°	78°	79°	
0	939 693	945 519	951 057	956 305	961 262	965 926	970 296	974 370	978 148	981 627	60
5	940 189	5 991	1 505	6 729	1 662	6 301	0 647	4 696	8 449	1 904	55
10	0 684	6 462	1 951	7 151	2 059	6 675	0 995	5 020	8 748	2 178	50
15	1 176	6 930	2 396	7 571	2 455	7 046	1 342	5 342	9 045	2 450	45
20	1 666	7 397	2 838	7 990	2 849	7 415	1 687	5 662	9 341	2 721	40
25	2 155	7 861	3 279	8 406	3 241	7 782	2 029	5 980	9 634	2 989	35
30	2 641	8 324	3 717	8 820	3 630	8 148	2 370	6 296	9 925	3 255	30
35	3 126	8 784	4 153	9 232	4 018	8 511	2 708	6 610	980 214	3 519	25
40	3 609	9 243	4 588	9 642	4 404	8 872	3 045	6 921	0 500	3 781	20
45	4 089	9 699	5 020	960 050	4 787	9 231	3 379	7 231	0 785	4 041	15
50	4 568	950 154	5 450	0 456	5 169	9 588	3 712	7 539	1 068	4 298	10
55	5 044	0 606	5 879	0 860	5 548	9 943	4 402	7 844	1 349	4 554	5
60	5 519	1 057	6 305	1 262	5 926	970 296	4 370	8 148	1 627	4 808	0
′	19°	18°	17°	16°	15°	14°	13°	12°	11°	10°	′

NATURAL COSINES

Natural Sines

′	80°	81°	82°	83°	84°	85°	86°	87°	88°	89°	′
0	984 808	987 688	990 268	992 546	994 522	996 195	997 564	998 630	999 391	999 848	60
5	5 059	7 915	0 469	2 722	4 673	6 320	7 664	8 705	9 441	9 872	55
10	5 309	8 139	0 669	2 896	4 822	6 444	7 763	8 778	9 488	9 894	50
15	5 556	8 362	0 866	3 068	4 969	6 566	7 859	8 848	9 534	9 914	45
20	5 801	8 582	1 061	3 238	5 113	6 685	7 953	8 917	9 577	9 932	40
25	6 045	8 800	1 254	3 406	5 256	6 802	8 045	8 984	9 618	9 948	35
30	6 286	9 016	1 445	3 572	5 396	6 917	8 135	9 048	9 657	9 962	30
35	6 525	9 230	1 634	3 735	5 535	7 030	8 223	9 111	9 694	9 974	25
40	6 762	9 442	1 820	3 897	5 671	7 141	8 308	9 171	9 720	9 983	20
45	6 996	9 651	2 005	4 056	5 805	7 250	8 392	9 229	9 762	9 990	15
50	7 220	9 859	2 187	4 214	5 937	7 357	8 473	9 295	9 793	9 996	10
55	7 460	990 065	2 368	4 360	6 067	7 462	8 552	9 339	9 821	9 999	5
60	7 688	0 268	2 546	4 522	6 195	7 564	8 630	9 391	9 848	1·00000	0
Cos.	9°	8°	7°	6°	5°	4°	3°	2°	1°	0°	

Natural Cosines

Natural Tangents

′	0°	1°	2°	3°	4°	5°	6°	7°	8°	9°	′
0	000 000	017 455	034 921	052 408	069 927	087 489	105 104	122 785	140 541	158 384	60
5	1 454	8 910	6 377	3 866	071 389	8 954	6 575	4 261	2 024	9 876	55
10	2 909	020 365	7 834	5 325	2 851	090 421	8 046	5 738	3 508	161 368	50
15	4 363	1 820	9 290	6 784	4 313	1 887	9 518	7 216	4 993	2 860	45
20	5 818	3 275	040 747	8 243	5 775	3 354	110 990	8 694	6 478	4 354	40
25	7 272	4 731	2 204	9 703	7 238	4 821	2 463	130 173	7 964	5 848	35
30	8 727	6 186	3 661	061 163	8 702	6 289	3 936	1 652	9 451	7 343	30
35	010 181	7 641	5 118	2 623	080 165	7 757	5 409	3 132	150 938	8 838	25
40	1 636	9 097	6 576	4 083	1 629	9 226	6 883	4 613	2 426	170 334	20
45	3 091	030 553	8 033	5 543	3 094	100 695	8 358	6 094	3 915	1 831	15
50	4 545	2 009	9 491	7 004	4 558	2 164	9 833	7 576	5 404	3 329	10
55	6 000	3 465	050 949	8 465	6 023	3 634	121 309	9 058	6 894	4 828	5
60	7 455	4 921	2 408	9 927	7 489	5 104	2 785	140 541	8 384	6 327	0
Cot.	89°	88°	87°	86°	85°	84°	83°	82°	81°	80°	

Tan.	10°	11°	12°	13°	14°	15°	16°	17°	18°	19°	
0	176 327	194 380	212 557	230 868	249 328	267 949	286 745	305 731	324 920	344 328	60
5	7 827	5 890	4 077	2 401	250 873	9 509	8 320	7 322	6 528	5 955	55
10	9 328	7 401	5 599	3 934	2 420	271 069	9 896	8 914	8 130	7 585	50
15	180 830	8 912	7 121	5 469	3 968	2 631	291 473	310 508	9 751	9 216	45
20	2 332	200 425	8 645	7 004	5 517	4 194	3 052	2 104	331 364	350 848	40
25	3 835	1 938	220 169	8 541	7 066	5 759	4 632	3 701	2 979	2 483	35
30	5 339	3 452	1 695	240 079	8 618	7 325	6 214	5 299	4 595	4 119	30
35	6 844	4 967	3 221	1 618	260 170	8 892	7 796	6 899	6 213	5 756	25
40	8 350	6 483	4 749	3 157	1 723	280 460	9 380	8 500	8 833	7 396	20
45	9 856	8 000	6 277	4 698	3 278	2 029	300 966	320 103	9 454	9 037	15
50	191 363	9 518	7 806	6 241	4 834	3 600	2 553	1 707	341 077	360 680	10
55	2 871	211 037	9 337	7 784	6 391	5 172	4 141	3 313	2 702	2 324	5
60	4 380	2 557	230 868	9 328	7 949	6 745	5 731	4 920	4 328	3 970	0
Cot.	79°	78°	77°	76°	75°	74°	73°	72°	71°	70°	

Tan.	20°	21°	22°	23°	24°	25°	26°	27°	28°	29°	
0	363 970	383 864	404 026	424 475	445 229	466 308	487 733	509 525	531 709	554 309	60
5	5 618	5 534	5 719	6 192	6 973	8 080	9 534	511 359	3 577	6 212	55
10	7 268	7 205	7 414	7 912	8 719	9 854	491 339	3 195	5 447	8 118	50
15	8 920	8 879	9 111	9 634	450 467	471 631	3 145	5 034	7 319	560 027	45
20	370 573	390 554	410 810	431 358	2 218	3 410	4 955	6 876	9 195	1 939	40
25	2 228	2 231	2 511	3 084	3 971	5 191	6 767	8 720	541 074	3 854	35
30	3 885	3 910	4 214	4 812	5 726	6 976	8 582	520 567	2 956	5 773	30
35	5 543	5 592	5 919	6 543	7 484	8 762	500 399	2 417	4 840	7 694	25
40	7 204	7 275	7 626	8 276	9 244	480 551	2 219	4 270	6 728	9 619	20
45	8 866	8 960	9 335	440 011	461 006	2 343	4 042	6 126	8 619	571 547	15
50	380 530	400 646	421 046	1 748	2 771	4 137	5 867	7 984	550 513	3 478	10
55	2 196	2 335	2 759	3 487	4 538	5 933	7 695	9 845	2 409	5 413	5
60	3 864	4 026	4 475	5 229	6 308	7 733	9 525	531 709	4 309	7 350	0
′	69°	68°	67°	66°	65°	64°	63°	62°	61°	60°	′

Natural Cotangents

NATURAL TANGENTS

′	30°	31°	32°	33°	34°	35°	36°	37°	′
0	577 350	600 861	624 869	649 408	674 509	700 208	726 543	753 554	60
5	9 291	2 842	6 894	651 477	6 627	2 377	8 767	5 837	55
10	581 235	4 827	8 921	3 551	8 749	4 551	730 996	8 125	50
15	3 183	6 815	630 953	5 629	680 876	6 730	3 230	760 418	45
20	5 134	8 807	2 988	7 710	3 007	8 913	5 469	2 716	40
25	7 088	610 802	5 027	9 796	5 142	711 101	7 713	5 019	35
30	9 045	2 801	7 070	661 886	7 281	3 293	9 961	7 327	30
35	591 006	4 803	9 117	3 979	9 425	5 430	742 214	9 640	25
40	2 970	6 809	641 167	6 077	691 572	7 691	4 472	771 959	20
45	4 938	8 819	3 222	8 179	3 725	9 897	6 735	4 283	15
50	6 908	620 832	5 280	670 285	5 881	722 108	9 003	6 612	10
55	8 883	2 849	7 342	2 394	8 042	4 323	751 276	8 946	5
60	600 861	4 869	9 408	4 509	700 208	6 543	3 554	781 286	0
Cot.	59°	58°	57°	56°	55°	54°	53°	52°	
Tan.	38°	39°	40°	41°	42°	43°	44°	45°	
0	781 826	809 784	839 100	869 287	900·404	932 515	965 689	1·00 000	60
5	3 631	812 195	841 581	871 844	3 041	5 238	8 504	002 913	55
10	5 981	4 612	4 069	4 407	5 685	7 968	971 326	005 835	50
15	8 336	7 034	6 563	6 977	8 336	940 706	4 157	008 765	45
20	790 698	9 463	9 062	9 553	910 994	3 451	6 996	011 704	40
25	3 064	821 897	851 568	882 136	3 659	6 204	9 842	014 651	35
30	5 436	4 336	4 081	4 725	6 331	8 965	982 697	017 607	30
35	7 813	6 782	6 599	7 322	9 010	951 733	5 560	020 572	25
40	800 196	9 234	9 124	9 924	921 697	4 508	8 432	023 546	20
45	2 585	831 691	861 655	892 534	4 391	7 292	991 311	026 529	15
50	4 979	4 155	4 193	5 151	7 091	960 083	4 199	029 520	10
55	7 379	6 624	6 736	7 774	9 800	2 882	7 095	032 521	5
60	9 784	9 100	9 287	900 404	932 515	5 689	1·000 000	035 530	0
Cot.	51°	50°	49°	48°	47°	46°	45°	44°	
Tan.	46°	47°	48°	49°	50°	51°	52°	53°	
0	1·035 530	1·072 369	1·110 613	1·150 368	1·191 754	1·234 897	1·279 942	1·327 045	60
5	038 549	075 501	113 866	153 753	195 280	238 576	283 786	331 068	55
10	041 577	078 642	117 131	157 150	198 818	242 269	287 645	335 108	50
15	044 614	081 794	120 405	160 557	202 360	245 974	291 518	339 162	45
20	047 660	084 955	123 691	163 976	205 933	249 693	294 406	343 233	40
25	050 715	088 127	126 987	167 407	209 509	253 426	299 308	347 320	35
30	053 780	091 309	130 294	170 850	213 097	257 172	303 225	351 422	30
35	056 854	094 500	133 612	174 304	216 698	260 932	307 158	355 541	25
40	059 938	097 702	136 941	177 770	220 312	264 706	311 105	359 676	20
45	063 031	100 914	140 282	181 248	223 939	268 494	315 067	363 828	15
50	066 134	104 137	143 633	184 738	227 579	272 296	319 044	367 996	10
55	069 247	107 369	146 995	188 240	231 231	276 112	323 037	372 181	5
60	072 369	110 613	150 368	191 754	234 897	279 942	327 045	376 382	0
Cot.	43°	42°	41°	40°	39°	38°	37°	36°	
Tan.	54°	55°	56°	57°	58°	59°	60°	61°	
0	1·376 382	1·428 148	1·482 561	1·539 865	1·600 335	1·664 280	1·732 051	1·804 048	60
5	380 600	432 578	487 222	544 779	605 526	669 776	737 883	810 252	55
10	384 835	437 027	491 904	549 716	610 742	675 299	743 745	816 489	50
15	389 088	441 494	496 606	554 674	615 982	680 849	749 637	822 759	45
20	393 357	445 980	501 328	559 655	621 247	686 426	755 559	829 063	40
25	397 644	450 485	506 071	564 659	626 537	692 031	761 511	835 400	35
30	401 948	455 009	510 835	569 686	631 852	697 663	767 494	841 771	30
35	406 270	459 552	515 620	574 735	637 192	703 323	773 508	848 176	25
40	410 610	464 115	520 426	579 808	642 558	709 012	779 552	854 616	20
45	414 967	468 697	525 254	584 904	647 949	714 728	785 629	861 091	15
50	419 343	473 298	530 102	590 024	653 366	720 474	791 736	867 600	10
55	423 736	477 920	534 973	595 167	658 810	726 248	797 876	874 146	5
60	428 148	482 561	593 865	600 335	664 280	732 051	804 048	880 727	0
′	35°	34°	33°	32°	31°	30°	29°	28°	

NATURAL COTANGENTS

NATURAL TANGENTS

′	62°	63°	64°	65°	66°	67°	68°	69°	′
0	1·880 727	1·962 611	2·050 304	2·144 507	2·246 037	2·355 852	2·475 087	2·605 089	60
5	887 344	969 687	057 895	152 676	254 857	365 412	485 489	616 457	55
10	893 997	976 805	065 532	160 896	263 736	375 037	495 966	627 912	50
15	900 687	983 964	073 215	169 168	272 673	384 729	506 520	639 455	45
20	907 415	991 164	080 944	177 492	281 669	394 489	517 151	651 087	40
25	914 179	998 406	088 720	185 869	290 726	404 317	527 860	662 809	35
30	920 982	2·005 690	096 544	194 300	299 843	414 214	538 648	674 622	30
35	927 823	013 016	104 415	202 784	309 021	424 180	549 516	686 527	25
40	934 702	020 386	112 335	211 323	318 261	424 217	560 465	698 525	20
45	941 620	027 799	120 303	219 918	327 563	444 326	571 496	710 619	15
50	948 577	035 257	128 321	228 568	336 929	454 506	582 609	722 808	10
55	955 574	042 758	136 389	237 274	346 358	464 760	593 807	735 093	5
60	962 611	050 304	144 507	246 037	355 852	475 087	605 089	747 477	0
Cot.	27°	26°	25°	24°	23°	22°	21°	20°	

Tan.	70°	71°	72°	73°	74°	75°	76°	77°	
0	2·747 477	2·904 211	3·077 684	3·270 853	3·487 414	3·732 051	4·010 781	4·331 476	60
5	759 961	917 991	092 983	287 949	506 656	753 882	035 778	360 400	55
10	772 545	931 889	108 421	305 209	526 094	775 952	061 070	389 694	50
15	785 231	945 905	123 999	322 636	545 733	798 266	086 663	419 364	45
20	798 020	960 042	139 719	340 233	565 575	820 828	112 561	449 418	40
25	810 913	974 302	155 584	358 001	585 624	843 642	138 772	479 864	35
30	823 913	988 685	171 595	375 943	605 884	866 713	165 300	510 709	30
35	837 020	3·003 194	187 754	394 063	626 357	890 045	192 151	541 961	25
40	850 235	017 830	204 064	412 363	647 047	913 642	219 332	573 629	20
45	863 560	032 595	220 526	430 845	667 958	937 509	246 84[illegible]	605 721	15
50	876 997	047 492	237 144	449 512	689 093	961 652	274 707	638 246	10
55	890 547	062 520	253 918	468 368	710 456	986 074	302 914	671 212	5
60	904 211	077 684	270 853	487 414	732 051	4·010 781	331 476	704 630	0
Cot.	19°	18°	17°	16°	15°	14°	13°	12°	

Tan.	78°	79°	80°	81°	82°	83°	84°	85°	
0	4·704 630	5·144 554	5·671 282	6·313 752	7·115 370	8·144 346	9·514 365	11·430 05	60
5	738 508	184 804	719 917	373 736	191 246	243 449	649 348	624 76	55
10	772 857	225 665	769 369	434 843	268 726	344 956	788 173	826 17	50
15	807 685	267 152	819 657	497 104	347 861	448 957	931 009	12·034 62	45
20	843 005	309 279	870 804	560 554	428 706	555 547	10·078 93	250 51	40
25	878 825	352 063	922 832	625 226	511 318	664 822	229 43	474 22	35
30	915 157	395 517	975 764	691 156	595 754	776 887	385 40	706 20	30
35	952 013	439 659	6·029 625	758 383	682 077	891 351	546 15	946 92	25
40	989 403	484 505	084 438	826 944	770 351	9·009 826	711 91	13·196 88	20
45	5·027 340	530 072	140 230	896 880	860 642	130 935	882 92	456 63	15
50	065 835	576 379	197 028	968 234	953 022	255 304	11·059 43	726 74	10
55	104 902	623 442	254 859	7·041 048	8·047 565	383 066	241 71	14·007 86	5
60	144 554	671 282	313 752	115 370	144 346	514 365	430 05	300 67	0
Cot.	11°	10°	9°	8°	7	6	5°	4°	

Tan.	86°	Diff.	87°	Diff.	88°	Diff.	89°	Diff.	
0	14·300 67		19·081 14		28·636 25		57·289 96		60
5	605 92	305 25	627 39	546 16	29·882 40	1·246 05	62·499 15	5·209 19	55
10	924 42	318 50	20·205 55	578 25	31·241 58	1·359 28	68·750 09	6·250 91	50
15	15·257 05	332 63	818 83	613 28	32·730 26	1·488 68	76·390 01	7·639 92	45
20	604 78	347 73	21·470 40	651 57	34·367 77	1·637 51	85·939 79	9·549 78	40
25	968 67	363 89	22·163 98	693 58	36·177 60	1·809 83	98·217 94	12·278 2	35
30	16·349 86	381 19	903 77	739 79	38·188 46	2·010 86	114·588 7	16·370 8	30
35	749 61	399 75	23·694 54	790 77	40·435 84	2·247 38	137·507 5	22·918 8	25
40	17·169 34	419 73	24·541 76	847 22	42·964 08	2·528 24	171·885 4	34·377 9	20
45	610 56	441 22	25·451 70	909 94	45·829 35	2·865 27	229·181 7	57·296 3	15
50	18·074 98	464 42	26·431 60	979 90	49·103 88	3·274 58	343·773 7	114·592 0	10
55	564 47	489 49	27·489 85	1·058 25	52·882 11	3·778 23	687·548 9	343·775 2	5
60	19·081 14	516 67	28·636 25	1·146 40	57·289 96	4·407 85	Infinite	Infinite	0
′	3°	Diff.	2°	Diff.	1°	Diff.	0°	Diff.	′

NATURAL COTANGENTS

Practical Hints on Solving a Problem

(1) Visual presentation is an invaluable aid to solving a problem. Where possible, always draw a rough diagram, inserting the data and the quantities required.

(2) Before commencing a lengthy calculation, see if a short cut can be used to advantage.

(3) It saves time in checking to set down each step of a calculation, rather than to rely too much on mental reckoning.

(4) Decimal points are sometimes poorly defined. To avoid ambiguity in reading quantities like ·635, always insert a 0 before the decimal point.

(5) Brackets omitted or inserted incorrectly can often cause confusion. Compare, for example, the following:

$\sin\theta^2$	sine of the square of θ
$(\sin\theta)^2$	square of sin θ
$\log P_2/P_1$	quotient of the log of P_2 and P_1
$\log (P_2/P_1)$	log of the quotient of P_2 and P_1

W. E. P.

2. COMMUNICATION THEORY, ELECTRON OPTICS AND COLOUR TELEVISION

2. COMMUNICATION THEORY, ELECTRON OPTICS AND COLOUR TELEVISION

Engineers who design systems for communication have hitherto done so without the aid of a means for quantitatively assessing the commodity being transmitted in those systems. Communication Theory aims at supplying such a means. Mathematical methods, particularly probability theory and mathematical analysis, are used to provide a measure of this commodity, "information", and to give a basis for comparison of the various communication systems. The study of information enables the theoretical limits, which an ideal system might reach under a given set of conditions, to be deduced, and with its aid better methods and more efficient systems may be evolved.

The history of communication is closely interwoven with the history of man himself. Primitive peoples showed a desire to convey information to one another, and used pictures or other symbolic representations to effect a record of events. These reduced, for example, to hieroglyphics, and civilizations built up agreed languages of selections from recognized sets of symbols. (The spoken word can also be considered as a combination of selections from a set of sounds.) In more recent times attempts have been made to devise ways of compressing the information contained in messages, or of simplifying the symbols themselves. One of the simplest means in use throughout history has been that employed in smoke signalling, i.e., the use of a two-valued code. Another such two-valued signalling code is that devised by Samuel Morse, in 1832: this was based on the probability of occurrence of letters in printed English as used in Morse's day. Thus the most frequently occurring letters, such as E, A, and T. were assigned the shortest code combinations, in order that messages might be as short as possible. The code was used in telegraphy, and came as a result of the wish to conserve time in the transmission of a given message. It showed also that for economy the statistics of the language must be taken into account.

The progress in communication from telegraphy to telephony, radio and television, with their concomitant techniques of multi-channel transmission by frequency-division-multiplex and pulse methods, threw emphasis on to the need for economy of band-width. It became clear that some means for assessing the merits of a system should be found and some relationship between the variables expressed.

It was Hartley,[1] who, in 1928, first formulated an equation relating the information in a communication channel to the band-width, transmission time, and distinguishable amplitude levels. This last factor, accounting for the amplitude levels, implies a lower limit on the power of the signal, and in fact can be simply related to the signal-to-noise power ratio for the channel.

The concepts, which have crystallized since those early years, have been the importance of signal band-width, time and signal power (as well as interfering noise power) and also the importance of the statistical nature of the language being used. These concepts have been correlated by Shannon [2] in his Mathematical Theory of Communication. Similarities exist between the expressions relating these variables and those used in modern thermodynamic theory for considering the entropy of systems with states of various probabilities. This has led to a recognition of the complementary nature of information and entropy. The subject has thus widened into an immense field of study embracing such diverse matters as linguistics, acoustics, radar, computation and psychology. Major contributions to wider aspects of the theory have been those of Shannon and Wiener,[3] whose powerful mathematical methods have been applied to many information-handling systems, such as pulse-modulation methods, secrecy codes, filters and automatic control systems.

Some classification of these theories is necessary. The widest of the categories considered has been the one concerned with systems in which action is controlled by the information received. These include control systems of all types, and the study of communication and control has been called " Cybernetics " by Wiener. " Information Theory " deals with the broad problem of conveying information from source to destination. This involves the recognition of the commodity as it is passed right through the system and provision of a means of measurement. The system may include channels other than those familiar to the engineer in that such components as parts of the nervous system may well be included. Information Theory is considered a part of modern physics and scientific method. " Communication Theory ", however, with which we are at present concerned, is confined to an understanding of communication processes and the improvement of methods for handling information which already exists in the spoken, written, visible or other clearly recognizable form.

Divisions of the Theory

Communication Theory is concerned with making a representation (in one space, at the receiving end) of a representation already existing (in another space, at the transmitting end). An example of the type of system concerned is shown in Fig. 1, and applies to such familiar equipments as radiotelephone links. Incoming information in the form of speech is processed in the transmitter, and may be multiplexed with other messages to produce an output signal which differs from the input in that filtering and distortion reduce the original information content and introduce errors. Over the radio link there will be an additional source of error in the form of noise. In this way the final output representation which goes to the destination of the message (the listener's ear) differs from the original by a certain amount. The message, however, is understood by virtue of the rules of the language and the context of the message. Information is gained at the receiver, when the state of existing knowledge (the *a priori* information) there is changed by the receipt of the message.

The Theory is first concerned with the analysis of the message into an alphabet of basic elements, or symbols. This is often termed the *Representation of Signals*. The communication of these elements within the system with or without the addition of noise can then be treated by means of the application of Probability Theory, and this is the *Statistical Theory of Communication*.

Further subdivision of the Statistical Theory may be made :

(*a*) Discrete signals in a noiseless channel.
(*b*) Discrete signals in a noisy channel.
(*c*) Continuous signals in a noisy channel.

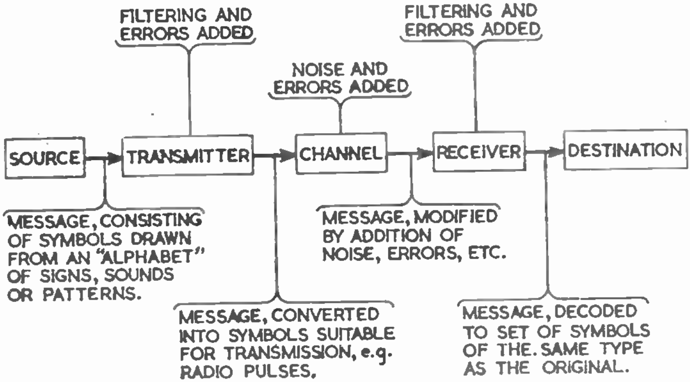


FIG. 1.—BLOCK DIAGRAM OF A COMMUNICATION SYSTEM.

The message is modified during transfer by the equipment characteristics, which reduce its true information content by filtering and by adding errors to it.

Definitions and Theory

Discrete Symbols in a Noiseless Channel

Information is conveyed only if the data received cannot be predicted by the receiver from *a priori* information received in the past.

A measure of information may be based on the number of choices that are necessary to select a given symbol from a set of possible symbols. Thus the logarithmic measure of information of Hartley is defined as

$$I = \log_2 n \text{ binary digits (bits)} \quad . \quad . \quad . \quad . \quad (1)$$

where I binary choices (two alternatives at each choice) are required to select one out of n equiprobable alternatives.

Practical sources of messages, however, do not produce symbols or even groups of symbols with equal probability of occurrence, and averages over large samples must be taken in order that a measure of the difficulty of transmission may be obtained. If a source produces symbols independently in a long series, x, of different representations (ergodic series), of which the ith has a prior probability of p_i, then the average selective information per symbol from that source is given by :

$$H(x) = -\sum_1^n p_i \log_2 p_i \text{ bits/symbol} \quad . \quad . \quad . \quad (2)$$

H is then the *selective entropy of the source.*

Shannon's Fundamental Coding Theorem, for a Noiseless Channel

The idea of the source supplying a communication channel of limited capacity has been incorporated in a theorem known as the Fundamental Coding Theorem. This states that :

If H is the entropy of the source (bits/symbol), and C is the channel capacity (bits/second), then it is possible to encode the message in such a way as to transmit it at an average rate of up to, but not exceeding $\frac{C}{H}$ symbols/second. The code which actually gives this rate is said to be "ideal", and requires the messages to be independent, and coded in groups of virtually infinite length.

Sampling Theorem

It is possible to determine signals by discrete samples of the amplitude taken at a finite number of times. This has been expressed as a result of work by Whittaker,[4] in Shannon's Sampling Theorem:

If a signal is confined to a time T and to a frequency band-width W; it can be completely determined by samples taken at $2TW$ equally spaced interpolation points.

A mathematical model of the signal can be built up by thinking of these $2TW$ data as defining a point in a reference space having $2TW$ dimensions. In consequence, algebraic operations may be performed on the co-ordinates of the point, to represent operations performed on the signal itself. This is only strictly correct if the number of dimensions is infinite, but $2TW$ is usually large enough.

The signal can also be represented by an area in the time-frequency plane. Gabor[5] has shown that this area must be at least unity in the inequality:

$$\Delta t \,.\, \Delta f \geqslant 1 \qquad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

The similarity of this to the uncertainty relation of modern quantum physics is obvious.

Types of Entropy

The selective entropy, $H(x)$, defined in equation (2), was a measure of the information conveyed by the occurrence of a single symbol, the ith of a set, x. The entropy to be ascribed to the joint occurrence of two symbols, symbol i of set x, and symbol j of set y, can be written:

$$\mathrm{H}(x, y) = -\sum_{i}\sum_{j} p(i, j) \log p(i, j) \qquad . \quad . \quad . \quad (4)$$

This is the *entropy of the joint event*, as opposed to the *conditional entropy* of the set y, when set x is already known. This is:

$$H_x(y) = -\sum_{i}\sum_{j} p(i, j) \log p_i(j) \qquad . \quad . \quad . \quad . \quad (5)$$

The *relative entropy* is defined as the ratio of the actual entropy of the source to the maximum entropy it could have, still using the same symbols.

The *redundancy* of the source is then expressed as the difference between unity and the relative entropy.

To give an example, consider English, as used in a communication system, to consist of a series of selections from a thirty-two letter alphabet. Were these equally likely, the information per symbol, calculated from equation (2) or more simply, equation (1), would be five bits. From analysis of the statistical structure of English, however, Shannon has estimated the average information per letter to be only about one bit. From the definitions above, then, the relative entropy of a source producing English is 0·2, and the redundancy some 0·8. The conclusion to be drawn from this is that in a noiseless channel, 80 per cent of the letters of a long message could be lost and the information still be conveyed.

Communication by Discrete Signals in Noise

In the previous section the entropy of a joint event was defined. It is easy to see how this can be applied to the reception of a message, which though transmitted as a set of events, x, is received as a set of events, y, differing from x because of errors introduced by noise in the channel. As a result there exists uncertainty at the receiver as to the exact form of x, and this is the *ambiguity* or *equivocation*. It is given by equation (5), the conditional entropy.

The *Channel Capacity* can then be expressed as the maximum difference between the entropy of the source and this equivocation. Thus the source is varied until it is matched to the channel and then:

$$\text{Channel Capacity, } C = [H(x) - H_y(x)]_{\max.} \qquad (6)$$

Shannon's Fundamental Coding Theorem for this noisy channel can be stated:

If H is the entropy of the source, and C is the channel capacity, then it is possible to encode the message in such a way that:

(a) when H is less than C, there is an arbitrarily small frequency of errors, and

(b) when H is greater than C, the equivocation is as little as $(H-C)$.

The two ways in which the transmission rate may be improved can be described as: (a) coding, in which the message symbols are taken in large groups, and (b) matching, in which the transmission symbols are chosen to reduce uncertainty due to noise to a minimum.

The Continuous Noisy Channel

It was mentioned that signals can be represented by points or vectors in a signal space of $2TW$ dimensions, where T is the time interval and W the frequency band confining the signal. If P is the mean power of the signal, and N the mean power of the noise, the resultant signal plus noise can be represented by the random resultant vector PN. As the dimensions increase in number, the vectors tend to terminate on a hypersphere and the number of distinguishable signals can be shown to be:

$$\left[\frac{P+N}{N}\right]^{TW}.$$

The capacity is then found by taking the logarithm to base two:

Capacity of Noisy Continuous Channel

$$C = TW \log_2\left(\frac{P+N}{N}\right) \text{ bits} \qquad (7)$$

This is known as Shannon's Mean Power Theorem and holds for noise which is "white" gaussian, and signals of the same amplitude distribution.

These are the principal relations which can be applied in assessing practical communications. Certain modifications must, however, be made to account for the deviations from the idealized assumptions that were made to derive them.

Practical Applications of Communication Theory

Practical systems with which engineers are concerned are usually designed to transmit data, speech or pictures. Communication Theory can be used to calculate the efficiency with which these can transmit their information by comparing with theoretically ideal systems.

The waveform to be transmitted by the equipment may be far from ideal. This is particularly true of speech, which has an amplitude and power spectrum making it unsuitable for efficient use of the available transmitter power and band-width. Shannon's Mean Power Theorem (Eqn. (7)) indicates that maximum channel capacity is realized when the signal assumes an amplitude distribution resembling that of noise. This may be approached by multiplexing large numbers of speech channels for transmission over a single link. The multiplexed signals not only have the required amplitude distribution but also make better use of the power, since peaks do not occur infrequently as in single channel speech.

The possibility of reducing the band-width necessary to transmit spoken intelligence has long been contemplated. Telephone circuits transmit only some 3 kc/s of the full fifteen associated with the voice and, even with a noisy line, this is usually perfectly intelligible. On the basis of one bit per letter of English, the average spoken sentence should convey information at an overall rate of some ten to twenty bits per second only. Were it possible to use an ideal code, equation (7) indicates that even with the poor signal-to-noise ratio of three to one it would be possible to transmit this information over a frequency band of only 10 c/s. That the three kilocycles can be greatly reduced has been demonstrated by the "vocoder" with which Dudley [6] was able to transmit speech using ten channels of 25 c/s to give the frequency range, and another for operating an artificial larynx, a total of 275 c/s. An even narrower band has been found necessary for synthesizing speech, using the six vocal parameters.

These reductions are made as a result of investigations into the characteristics of the voice. Equivalent reductions in band-width of television signals might be made by the review of eye characteristics. In both cases coding methods are not known and theoretical channel capacities cannot be calculated.

Reducing Band-width of the Communication Channel

The reductions so far considered have applied to the coding of the message input. The question arises: with a given form of input signal (e.g., speech) by what means may it be transmitted to make the most use of the available band-width ?

Various modulation methods, such as A.M., F.M., P.A.M., P.P.M., P.C.M., etc., have been devised, and for each of these the channel capacity may be derived. These must then be compared with the theoretically ideal system capacity given in equation (7). Frequency modulation is found to enable band-width and signal-to-noise-ratio to be exchanged linearly (the familiar property of improving the signal-to-noise-ratio with increased deviation ratio). On the other hand, Pulse Code Modulation allows exchange on a logarithmic basis, and this is similar to the ideal of equation (7). For a detailed comparison of the theoretical capacities of the various systems the reader is referred to the work of Jelonek.[7]

Applications to Television Systems

The possibility of reducing the band-width of television signals has been considered by Cherry and Gouriet.[8] Their proposal was to change the statistical structure of the signal by recoding, just as Morse did with the code described earlier. Information about the rate of change of video signal, and therefore about boundaries between light and shade in the picture rather than about intensities, is transmitted.

Attempts have been made to reduce the band-width necessary for the transmission of colour television signals. This is important since higher frequencies at present indicated would prove a costly proposition. Methods for coding the video signal and thus compressing the band-width necessary have been suggested, one simple method being that of Valensi.[9] The range of colour can be described by means of Maxwell's Colour Triangle, which would be divided into thirty segments. Information supplied in the transmitted signal would merely consist of the code to select particular segments and the corresponding brightness.

Filters

Application of mathematical theory to the design of optimum filters has been effected by Wiener. Complicated filters for separating messages from a background of noise have been the centre of much interest in this field of application of the theory. It is possible that this work may ultimately lead to a solution of the problem of speech recognition.

Conclusions

Communication Theory has concentrated attention on the essential features of communication. It has provided a quantitative means for evaluating the efficiency of our systems and for comparing them with theoretically ideal systems. It clearly indicates limits to the performance of equipment.

One desirable result of the theory has been the way in which it has shown just how close some present methods are to the ideal. Although in many cases reductions in band-width are possible, they will be great only if flexibility is sacrificed. Methods necessary to achieve large signal-band-width savings in general require complex equipment with means for storing large amounts of information. Decisions as to the value of such methods must necessarily be based on economics.

Perhaps the most useful result of the development of the theory has been the way in which it has focused the attention of many fields of thought on to the possibility of applying statistical methods to communication of all types.

B. Z. de F.

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ELECTRON OPTICS

Behaviour of an Electron in an Electric Field

An electron situated in an electric field, such as that between the parallel plates of Fig. 2, is urged towards the positively-charged plate. The electric field is usually represented by arrows pointing in the

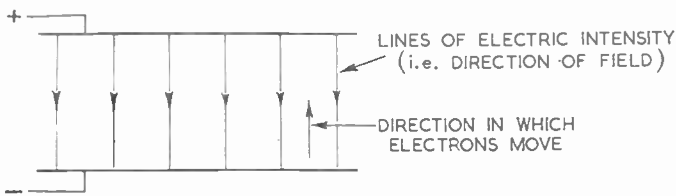

Fig. 2.—Electric Field between Two Parallel Charged Plates.

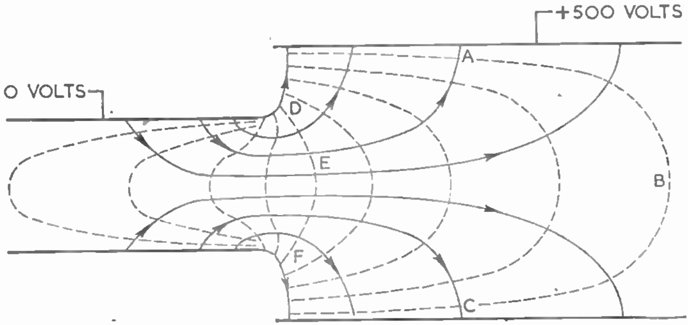

Fig. 3.—Electric Field (Solid Lines) and Equipotential Surfaces (Dotted Lines) between Two Co-axial Charged Cylinders.

direction in which a *positive* charge would move; thus an electron, by virtue of its negative charge, tends to move in the opposite direction to the lines of electric intensity representing the field.

Fig. 3 shows in solid lines the field pattern between two charged co-axial cylinders; this is a form of construction frequently used in electrostatic electron lenses. It is significant that the lines of intensity in Figs. 2 and 3 all leave or arrive at the charged conductors at right angles. In Fig. 3 a number of dotted lines are drawn; these have the property that they are, at all points, at right angles to the lines of intensity. The most important property of these is that for each of them the potential acquired by a conductor placed at any point in the line is constant. These are, in fact, lines of constant potential, and are usually termed *equipotential lines.*

Equipotential lines are sometimes labelled with their value of potential; for example, line *ABC*, which is very near the outer cylinder,

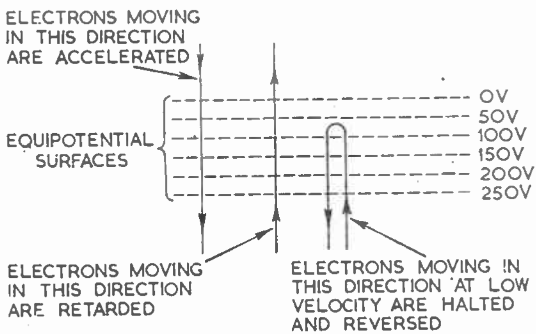

Fig. 4.—Behaviour of an Electron Beam Entering an Electric Field Normal to the Equipotential Surfaces.

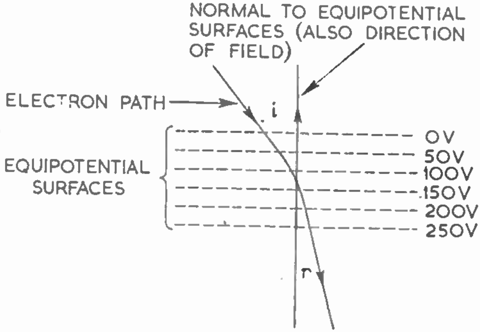

FIG. 5.—PATH TRACED BY AN ELECTRON BEAM ENTERING A PARALLEL ELECTRIC FIELD OBLIQUELY AND TRAVELLING UP THE POTENTIAL GRADIENT.

has a potential of nearly 500 volts (that of the cylinder), whereas *DEF*, approximately equidistant from both cylinders, probably has a potential of 250 volts, a value mid-way between those of the two cylinders. Fig. 3 is, of course, a sectional drawing, and if it were rotated about its axis of symmetry the dotted lines would generate *equipotential surfaces*. The shape of these surfaces determines the behaviour of electrostatic focusing systems, and has a bearing of the deflection of electron beams passing through them, just as the shape of an optical lens determines the deflection experienced by a light beam passing through the lens.

An electron moving in an electric field so as to cross equipotential surfaces at right angles is, of course, travelling parallel to the lines of electric intensity (which are always at right angles to the surfaces). Such electrons experience no lateral deflecting forces, and continue to move in the same straight line, being accelerated if they are moving up the potential gradient (i.e., towards a positively-charged conductor), but retarded if moving down the potential gradient. If a retarding field is sufficiently strong it can bring a moving electron to a temporary halt and then reverse its motion. Such conditions exist near the target of all low-velocity television camera tubes. Where an electron beam is required to penetrate a retarding field it must be given an initial velocity

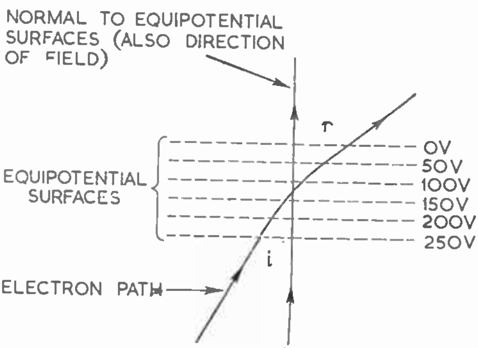

FIG. 6.—PATH TRACED BY AN ELECTRON BEAM ENTERING A PARALLEL ELECTRIC FIELD OBLIQUELY AND TRAVELLING DOWN THE POTENTIAL GRADIENT.

adequate for the purpose. Fig. 4 illustrates the three effects which can occur when an electron stream enters a field at right angles to the equipotential surfaces.

If an electron enters an electric field at an angle to the equipotential surfaces, its velocity is not parallel to the lines of intensity, and the electron experiences a lateral deflecting force which tends to align it with the lines of intensity (if the electron is travelling *up* the potential gradient) or tends to make it take up a direction at right angles to the lines of intensity (if it is moving *down* the potential gradient). These two modes of behaviour are illustrated in Figs. 5 and 6. The effect on the beam is perhaps easier to understand when it is realized that the beam is accelerated downwards in Fig. 5, tending to make the velocity more nearly at right angles to the equipotential surfaces, whereas in Fig. 6 the beam is retarded and tends to become parallel to the equipotential surfaces. There is a relationship between the angles of incidence and refraction for electron beams analogous to Snell's law in Optics. The relationship is

$$\frac{\sin i}{\sin r} = \sqrt{\left(\frac{V_i}{V_r}\right)} \quad . \quad . \quad . \quad . \quad (1)$$

where V_i is the potential of the medium above the uppermost equipotential surface and V_r is that of the medium below the lowest equipotential surface. Both media are assumed to be of constant potential.

The potential of a medium can readily be varied, but the refractive index of an optical medium cannot; in this respect electron-optical systems are more flexible than their optical counterparts.

Figs. 5 and 6 show another feature of electron lenses which is not shared by optical lenses; in an optical system regions of different refractive indices are separated by surfaces, and light beams suffer an abrupt change in direction as they cross these surfaces. In an electron-optical system there is a boundary zone of appreciable thickness between two media of different potential, and an electron beam, crossing from one medium to the other, follows a curved path in such zones.

Electrostatic Immersion Lens

Fig. 7 illustrates a typical electrostatic lens of the type commonly used near the cathode of a cathode-ray tube to concentrate emission from the cathode into a beam in the direction of the screen. The lens consists essentially of two conducting plates, each containing a circular hole, so arranged with respect to the cathode that the three electrodes form a system with an axis of symmetry. The electrode nearest the cathode is the grid, sometimes termed a modulator, and the second is the first anode. The shape of the electric field is also indicated, and from this we can deduce the shape of the paths traced out by electrons released from the cathode. To do this we apply the rules implicit in Figs. 5 and 6, namely, that electrons travelling in the direction of *increasing* positive potential are deflected *towards* the normal, and those travelling in the direction of *decreasing* positive potential are deflected *away from* the normal.

Electrons entering the field and travelling up the potential gradient meet equipotential surfaces of convex form; they are therefore deflected towards the normals to these surfaces, and the divergent beam leaving

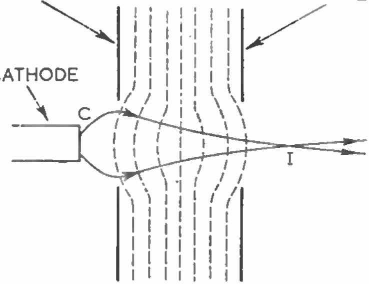

FIG. 7.—ELECTROSTATIC IMMERSION LENS USED TO CONCENTRATE EMISSION FROM THE CATHODE OF A CATHODE-RAY TUBE INTO A CONVERGENT BEAM.

the cathode is made convergent immediately inside the lens. As the beam approaches the anode, however, it meets equipotential surfaces of concave form, and electrons are here deflected away from the normal. In this region of the lens the beam is made more divergent, but, because of the increased electron velocity in this region, the effect is less than that of the convex surfaces, and the beam leaving the lens is still convergent. This is therefore an example of a positive electron lens, and it has a focus at I which can be regarded as an image of the cathode at C.

Fig. 7 may thus be regarded as an electron–optical equivalent of a convex lens. The simple convex lens is, however, not a perfect analogy, because the media on the two sides of the lens have the same refractive index (both being air presumably), but in Fig. 7 the potential near the anode is considerably higher than that near the grid. In spite of this, the analogy is useful, and it is helpful to remember that the shape of the equipotential surfaces, encountered by an electron travelling up the potential gradient, is also the shape of the front surface of the equivalent optical system (in which it is assumed that light is travelling into a medium of greater refractive index).

Electrostatic Two-cylinder Lens

There are usually two lenses of the type shown in Fig. 7 in an electrostatically-focused cathode-ray tube. The first is formed between the grid and anode as described, and the second is between the first and second anodes, as shown in Fig. 8. The space between the second anode and the screen is made equipotential as far as possible by a third anode, consisting of a conductive coating deposited on the inside walls of the tube and extending to within a small distance of the screen. Such an electrode is known as a wall anode, and is usually connected internally to the second anode (to avoid the formation of a third lens between the second and third anodes). The focusing of the electron beam can be achieved by variation of the first or second anode potential.

Symmetrical or Univoltage Electrostatic Lenses

Another type of electron lens is that illustrated in Fig. 9, which consists of three parallel conducting plates with circular apertures

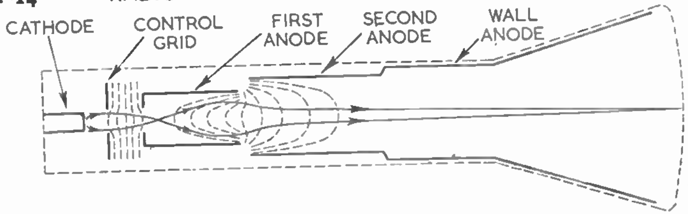

FIG. 8.—ELECTROSTATIC FOCUSING SYSTEM OF A CATHODE-RAY TUBE.

having a common axis. The outer plates have the same potential, and the centre conductor has a positive or negative bias which can be varied to control the focal length. The diagram gives the shape of the equipotential surfaces for a positively-biased centre plate, and also indicates two typical electron paths through the lens, showing it to be convergent. Univoltage lenses are chiefly employed in electron microscopes, but they are mentioned here as an example of an electron lens in which both initial and final media have the same potential. Such a lens corresponds very closely with a single convex optical lens in air.

Electric Deflection of Electron Beams

Fig. 10 illustrates an electrostatic deflecting system; such systems have been used in television camera and picture tubes, but are now chiefly employed in cathode-ray tubes for oscilloscopes.

The electron beam is passed between two parallel plates situated between the electron gun and the screen. If there is no potential difference between the plates, the beam passes between them without deflection and strikes the centre of the screen (as shown by the dotted line), but if one plate is biased positively with respect to the other, the beam experiences an accelerating force in the direction of the positive plate (i.e., opposite to the field) and is deflected in the direction of that plate, as shown by the solid line. The deflection D on the screen is related to the voltage V between the plates according to the approximate expression

$$\frac{D}{V} = \frac{lL}{2dV_0} \qquad (2)$$

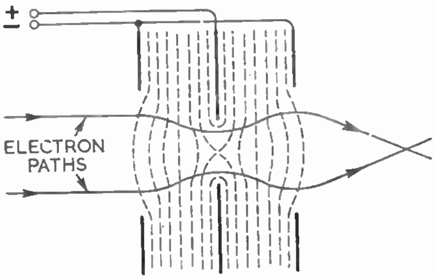

FIG. 9.—SYMMETRICAL OR UNIVOLTAGE ELECTRON LENS.

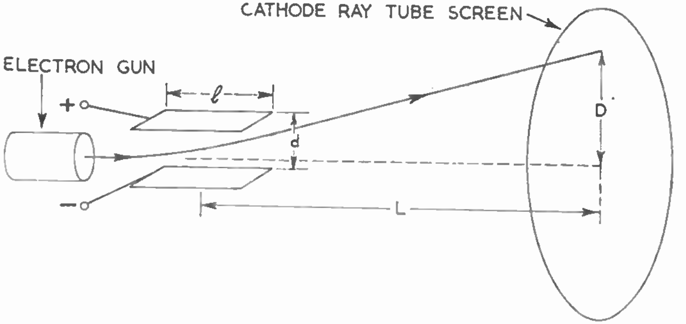

FIG. 10.—ELECTROSTATIC DEFLECTION OF AN ELECTRON BEAM.

where V_0 is the final accelerating potential of the electron gun, d is the spacing between the plates, l is the length of the plates and L is the distance of the plates from the screen. The expression shows that the deflection per volt applied between the plates (usually known as the *deflection sensitivity*) is inversely proportional to the voltage on the final anode of the electron gun. For this reason deflection sensitivity is usually given in a form expressing the dependence on V_0. For example, the deflection sensitivity may be quoted as $450/V_0$ mm./volt; if the final anode potential is 1,250 volts, the deflection is $450/1{,}250 = 0{\cdot}36$ mm./volt.

As shown in expression (2), the deflection sensitivity can be increased by increasing the length of the deflecting plates or decreasing the spacing between them, but both methods limit the maximum angular deflection of the beam. This limitation can be avoided to some extent by setting the plates at a small angle to one another, as shown in Fig. 11, in which the beam is shown deflected through the maximum angle by the first pair of plates.

The two plates illustrated in Fig. 10 deflect the beam in the vertical direction only, and to scan a target or screen, another pair of plates is required to give horizontal deflection. To avoid interaction between the two fields, the two pairs of plates are generally mounted in tandem, as shown in Fig. 11. To give a rectangular scanning pattern as required in a television camera or picture tube, the horizontal deflecting plates

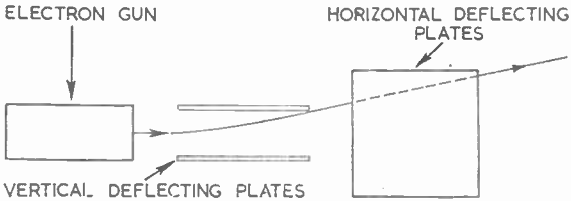

FIG. 11.—ARRANGEMENT OF VERTICAL AND HORIZONTAL DEFLECTING PLATES IN A CATHODE-RAY TUBE.

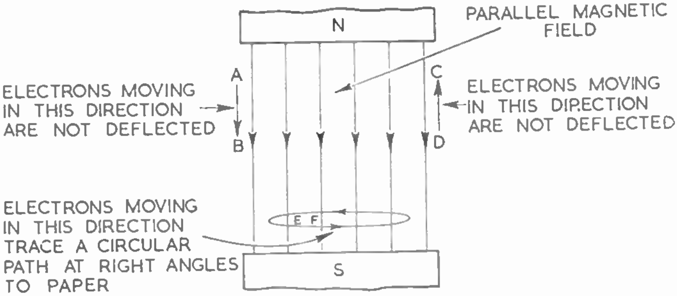

FIG. 12.—ILLUSTRATING THE FORCES ACTING ON ELECTRONS MOVING IN A PARALLEL MAGNETIC FIELD.

are supplied with a saw-tooth voltage at line frequency and the vertical deflecting plates with a saw-tooth voltage at frame frequency.

Behaviour of an Electron in a Magnetic Field

The forces acting on an electron travelling in a magnetic field can be illustrated by reference to a diagram such as that shown in Fig. 12. This shows a parallel magnetic field between two pole pieces, the field being represented by arrowed lines; these indicate, at any point in them, the direction in which a north magnetic pole would be urged, i.e., the direction of a compass needle situated at that point.

An electron moving parallel to the field as in the direction of the arrows *AB* or *CD* experiences no force due to the field, and continues to move in the same straight line. If the electron is moving in a direction at right angles to the magnetic field, as shown by arrow *EF*, it experiences a deflecting force at right angles to *EF* and to the field. The force, in this instance, acts downwards into the paper, and tends to make the electron follow a curved path in a plane at right angles to the paper. As the direction of the electron velocity alters, that of the force changes also, because the force is, at every instant, at right angles to field and velocity. Under the action of this ever-changing force the electron is forced to move in a circular path lying in a plane at right

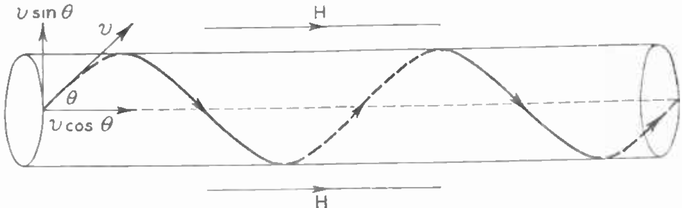

FIG 13.—PATH TRACED BY AN ELECTRON ENTERING A PARALLEL MAGNETIC FIELD AT AN ANGLE Θ.

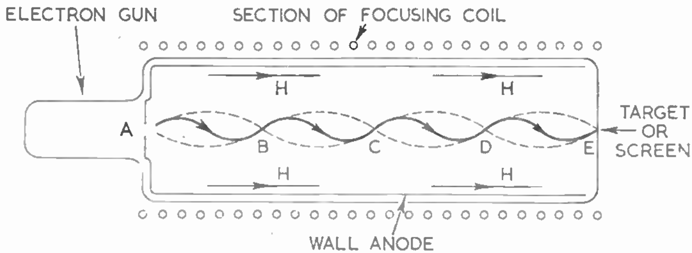

FIG. 14.—ACTION OF A LONG MAGNETIC LENS.

angles to the paper; an attempt has been made to illustrate the path in Fig. 12.

From the behaviour of the electrons entering the field as indicated in Fig. 12, it is possible to deduce the path traced by an electron entering a parallel field at a small angle θ, as shown in Fig. 13. The electron velocity v can be resolved into two components $v \cos \theta$ parallel to the field and $v \sin \theta$ at right angles to it. The first-mentioned component produces a uniform motion of the electron parallel to the field, and the second produces motion in a circular path in a plane at right angles to the field. Under the combined effect of these two motions, the electron follows a corkscrew path similar to that shown in Fig. 13. The path is represented by a line drawn on the surface of a cylinder and which, after making each complete revolution of the cylinder, returns to the level of the starting point. The cross-section of the cylinder on a plane at right angles to the field is, of course, the circle due to interaction of $v \sin \theta$ and H; the projection of the electron motion on a plane parallel to the magnetic field is a sine wave.

Long Magnetic Lens

One method of focusing an electron beam by a magnetic field is shown in Fig. 14. This shows a tube similar in shape to some television camera tubes and surrounded, for most of its length, by a coil (the turns of which are shown in section) which carries D.C. and produces an axial magnetic field represented by H. The electron gun, the details of which are not shown, includes a cathode, a control grid and a number of anodes consisting of plates with apertures. A wall anode is connected to the final anode of the gun to form a field-free space in the body of the tube through which electrons may travel at uniform velocity. The potential on the control grid controls the density of the electron beam, and the anodes have the effect of intercepting all electrons leaving the cathode except those which make small angles with the tube axis (and hence with the magnetic field). The anodes are not focusing electrodes, and the electrons leave the gun in the form of a divergent beam of small angle as shown. Each electron in this beam, except those parallel to the magnetic field, follows a path of the form shown in Fig. 13, and each returns to its initial level at the same distance from the gun. Thus the cross-section of the beam narrows to a small value, approximately equal to the area of the final aperture in the electron gun, at regular intervals along the tube axis. The envelope of all the electron paths is shown by the dotted line in Fig. 14, and one particular path is indicated

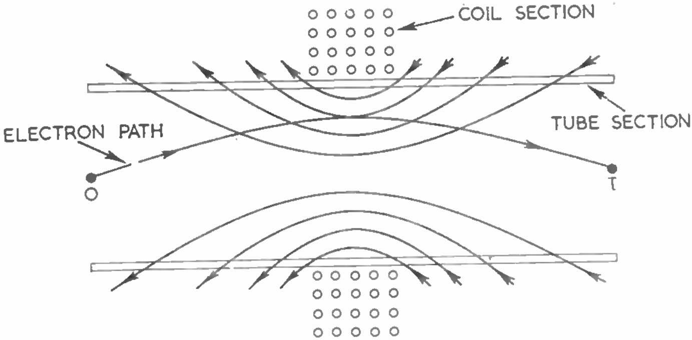

FIG. 15.—LONGITUDINAL SECTION THROUGH A TUBE WITH A SHORT MAGNETIC LENS SHOWING FIELD AND A TYPICAL ELECTRON PATH.

by the solid line. At the points B, C, D, etc., the electron beam can be said to form images of the initial opening A, and by adjustment of the final anode potential (i.e., the electron velocity) or the current in the focusing coil surrounding the tube, the intervals between the foci can be varied and one of the foci can be made to coincide with the target or screen. This is the type of magnetic electron lens used in low-velocity camera tubes of the orthicon, image–orthicon and c.p.s. Emitron types.

Short Magnetic Lens

Another type of magnetic lens commonly used to focus the beam in picture tubes consists of a short solenoid carrying D.C., as shown in Fig. 15. The magnetic field due to such a coil has the form shown: although it is parallel to the coil axis at the centre of the coil, it has marked radial components (at right angles to the axis) elsewhere. In this type of lens the focusing action is principally due to the radial components (not the axial component as in the long magnetic lens). The mechanism of the focusing effect is somewhat complex; briefly, electrons entering the lens and meeting the radial field at the entrance are deflected at right angles to the paper and begin to spin around the lens axis. At the centre of the lens this motion interacts with the axial magnetic field to produce a force which deflects the electron towards the lens axis. At the exit of the lens the radial field components again predominate, and interaction of these with the electron velocity gives rise to deflecting forces, again causing a spinning motion around the lens axis. The radial field is, however, in the opposite direction to that at the lens entrance, and the spinning motion is also in the reverse direction, tending to cancel that produced earlier. In practice, the cancellation is not perfect, and the image is usually rotated through a small angle with respect to the inverted position, which is normal for a positive lens. The spinning motion is ignored in the typical electron path shown in Fig. 15.

The focal length of the lens depends on the radial field components, and can be reduced (implying increased lens power) by enclosing the winding in a shroud of highly permeable material (such as soft iron),

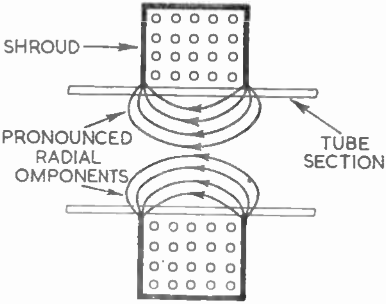

Fig. 16.—Increase in Radial Field Components Due to Use of a Magnetic Shroud.

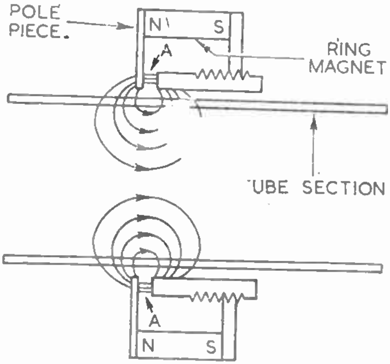

Fig. 17.—Cross-section of One Type of Short Magnetic Lens Employing a Permanent Magnet.

as shown in Fig. 16. The focal length is proportional to the square of the current in the coil, and can thus be varied by adjustment of the current. Lenses of this type are used in the image section of image-iconoscope camera tubes.

Short magnetic lenses often employ a permanent magnet instead of the electromagnet illustrated in Figs. 15 and 16, and a section through one type of permanent-magnet lens is given in Fig. 17. The magnet is in the form of a ring surrounding the tube neck, and is magnetized so that there is a North pole on one side of the ring and a South pole on the opposite side as shown. The flux is led to the tube by pole pieces in the form of annular rings. One of these has a larger inner diameter than the other, and is threaded on the inside edge so as to engage with a soft-iron tube which can be rotated so as to increase or decrease the air gap *A*. This gap behaves as a magnetic shunt, and by varying it the flux density in the tube can be controlled and the focal length of the lens adjusted.

Magnetic Deflection of Electron Beams

Fig. 18 (*a*) gives a theoretical diagram for a magnetic deflection system. Electrons are fired down the centre of the tube by an electron gun, and meet a vertical magnetic field due to the two series-connected coils carrying D.C. These coils are arranged one above and the other

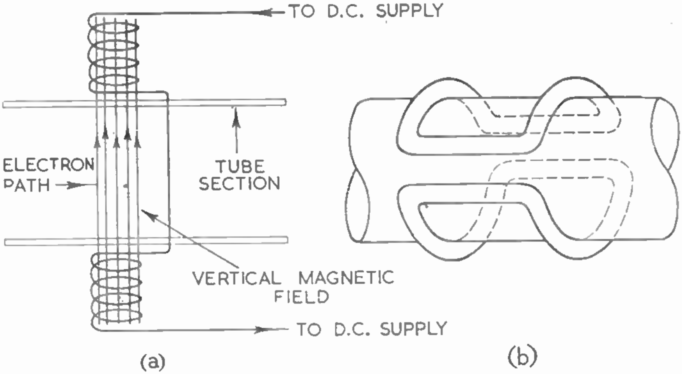

FIG. 18.—THEORETICAL DIAGRAM ILLUSTRATING THE MECHANISM OF MAGNETIC DEFLECTION IS SHOWN IN (a) AND A PRACTICAL FORM FOR SCANNING COILS FOR HORIZONTAL DEFLECTION IS SHOWN IN (b).

below the tube, and their common axis is at right angles to the axis of the tube. The deflecting force is at right angles to the electron motion and the magnetic field, and is thus at right angles to the paper; in other words, this arrangement of coils gives horizontal deflection of the beam.

The deflection D produced on the screen of a cathode-ray tube is given by

$$\frac{D}{H} = \frac{lL}{\sqrt{(kV_0)}}$$

where H is the magnetic field in the coils, l is the length of the magnetic field along the tube axis, k is a constant dependent on the ratio of the charge to the mass of an electron and V_0 is the final anode potential of the electron gun. The deflection is thus linearly related to the deflecting field, but is inversely proportional to the *square root* of the final

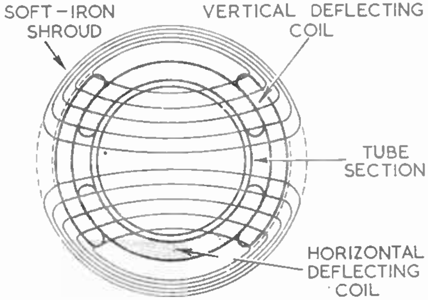

FIG. 19.—CROSS-SECTION OF TUBE FITTED WITH SADDLE COILS, FOR VERTICAL AND HORIZONTAL DEFLECTION, AND A SHROUD. THE FIELD ILLUSTRATED IS DUE TO THE VERTICAL DEFLECTION COILS.

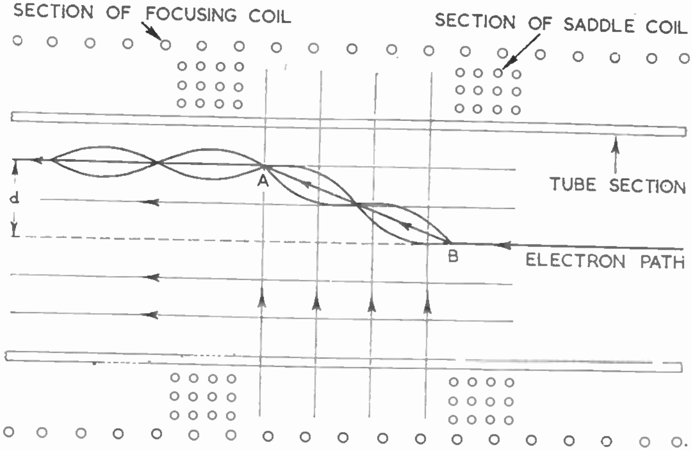

FIG. 20.—LONGITUDINAL SECTION OF TUBE WITH SADDLE AND FOCUSING COIL ILLUSTRATING THE PRINCIPLE OF ORTHOGONAL SCANNING.

anode potential. In electric deflection the beam displacement is inversely proportional to the final anode potential itself.

Fig. 18 (*b*) shows a practical arrangement of a magnetic deflecting system; the coils are of approximately square or rectangular shape, and are moulded to the contour of the tube to form "saddle coils". Another pair of coils with their common axis horizontal is needed to give vertical deflection, and these coils can overlap the horizontal-deflecting coils, as shown in the sectional view of Fig. 19. There is no need for the two deflecting systems to be mounted in tandem, as in an electrostatic deflecting system, and thus a tube with magnetic deflection can be appreciably shorter than one with electrostatic deflection. The efficiency of a magnetic deflection system can be approximately doubled by the use of a tube of highly-permeable material surrounding the coils, as shown in Fig. 19. This tube provides a low-reluctance path for the magnetic flux external to the scanning coils, and the flux path for the vertical deflecting coils is shown in Fig. 19.

To give a rectangular scanning pattern as required in a television camera or picture tube, the horizontal deflecting coils are supplied with a saw-tooth current at line frequency and the vertical deflecting coils with a saw-tooth current at frame frequency.

Magnetic Deflection in the Presence of an Axial Magnetic Field (Orthogonal Scanning)

As mentioned earlier, low-velocity camera tubes have long magnetic lenses. They also have magnetic deflection systems, the saddle coils being inside the focusing coils, and thus the electron beam in such a tube is subjected to axial and lateral magnetic fields. The operation of such a magnetic deflecting system is quite different from that just described.

Consider the behaviour of an electron beam entering the combined field due to the focusing coil and a pair of saddle coils (giving a vertical field), as shown in Fig. 20, in which both coils are shown in section. The two fields can be combined to produce a single field inclined at an angle to the tube axis as shown by arrow AB. Electrons entering this field along the axis of the tube perform spirals around the lines of force, as in a long magnetic lens (Fig. 14). After they have passed through the lateral field, however, the electrons are subjected to the axial field only, and are again deflected, emerging parallel to the lens axis but laterally displaced from it by a distance d (Fig. 20). A similar behaviour occurs with respect to the horizontal deflecting coils, which can be mounted on top of or underneath the vertical deflecting coils, as in Fig. 19.

The presence of the axial magnetic field brings about two significant differences in the effect of the deflecting coils :

(1) Coils giving a vertical magnetic field cause vertical deflection (such coils give horizontal deflection in the absence of an axial field).

(2) Coils cause electrons to suffer a double deflection, first on entering and then on leaving the lateral field, and as a result electrons approach the target at normal incidence, no matter what the extent of the deflection. This is known as *orthogonal scanning*, and is an essential requirement in any camera tube employing a low-velocity beam.

S. W. A.

COLOUR TELEVISION

Display Devices

Due to the fact that human colour vision is substantially trichromatic, it is possible to produce an illuminated patch of light of practically any desired colour by the superimposition of three suitably chosen beams of different coloured lights having the required intensities. This is an "additive" colour process, and there are at least three methods of synthesizing any given coloured illumination :

(i) Superimposition of the three chosen primary coloured beams, each having its own required intensity. This method is employed in three-tube receiver display systems in which two semi-reflecting or dichroic mirrors, which may, for example, be either crossed or in the shape of a circumflex, combine the simultaneous rasters of three separate cathode-ray picture tubes with phosphors which fluoresce red, green and blue respectively.

(ii) Juxtaposition of the three beams if the area which encloses the patches of light formed by the three juxtaposed beams subtends such a small angle at the observer's eye that each beam is not separately distinguishable. This method, Fig. 21, is used in three-beam shadow-mask colour picture tubes.[1,2] The 21-in. R.C.A. version (tricolour kinescope) of these contains a screen of 400,000 groups of three colour-fluorescing dots of phosphor. The three dots in each group fluoresce red, green and blue respectively and three electron beams, each emerging from its own individual gun inside

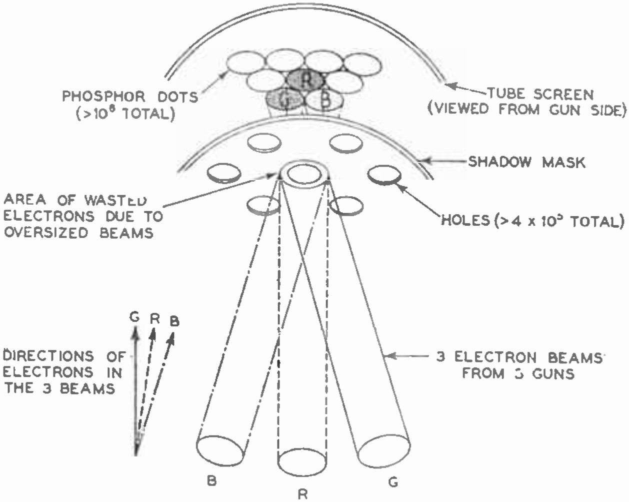

Fig. 21.—Three-gun Display Tube Using Shadow or Aperture Mask.

the tube, are so directed that each beam impinges upon only one coloured dot in each group. There are thus effectively a red, a blue and a green gun operating simultaneously. The modulating electrodes of the guns may be supplied with primary-colour information just as would be done in the case of the three-separate-display-tube system described in (i). What may be an improved version, with curved rather than flat shadow-mask, is known as the colortron.[2]

(iii) Sequential presentation of the three primary beams of coloured light at a rate sufficient to overcome brightness flicker. With present-day phosphors the minimum acceptable presentation frequency for brightnesses up to about 12 ft.-lamberts is 50 c/s. If consideration be given to a field-sequential system of colour television, then the transmission of an object having a colour requiring the transmission of one only of the three chosen primaries will give rise to a signal which will be presented at a third of the total field frequency; thus each primary-colour signal must be presented at a frequency of at least 50 c/s, and if three different primary-colour signals are presented in successive fields, then a total field frequency of 150 c/s is required. It might be thought that, with a field sequential three-tube display, it would be legitimate to use long-persistence phosphors to abate the flicker, but this is not so because of the "colour poisoning" which would result from the afterglow of one colour phosphor being superimposed upon the colour being presented in the succeeding field by another phosphor. The classic field-sequential display is the single tube having a "white" phosphor emitting an adequate quantity of light at all wavelengths in the visible spectrum (400–700 mμ). Immediately in front of the picture-tube screen is interposed a disc having a suitable number, say six, of gelatine colour filters of suitable density/light wavelength characteristics. If the disc rotates at 1,500 r.p.m. and is synchronized in phase as well as frequency with the television field-synchronizing pulses, then correct colour sequence will be obtained.

Combination Methods

Display systems which make use of more than one of the above methods exist. The single-gun horizontal-line-screen post-deflection-focusing tube known as the chromatron or Lawrence tube [3] can employ a rapid sequential presentation of coloured fluorescence from closely packed horizontal lines of three phosphors, Fig. 22. Here, juxtaposition can be combined with rapid (more than 1 Mc/s) sequence. Beam-indexing picture tubes are the subject of research at the present time. To avoid the power required to deflect electrostatically a single beam on to an appropriate colour-fluorescing line of phosphor when the beam electrons are close to the screen, it is possible to arrange the lines of phosphor vertically, i.e., at right angles to the scanning lines. By the use of a signal generator built into the screen, e.g., a colour-sensitive photo device [4] which emits a pulse each time one of the colour phosphor strips is being bombarded by the beam, it is possible to switch the control electrode of the single gun to the appropriate primary colour-signal channel. The availability of a pulse which denotes the time at which the scanning beam is in a given position in any one tricolour group of phosphor dots or lines opens a wide range of possibilities for feedback or servo-control of beam position : extreme line-scan linearity ceases to have such vital importance. Present receiver practice in the

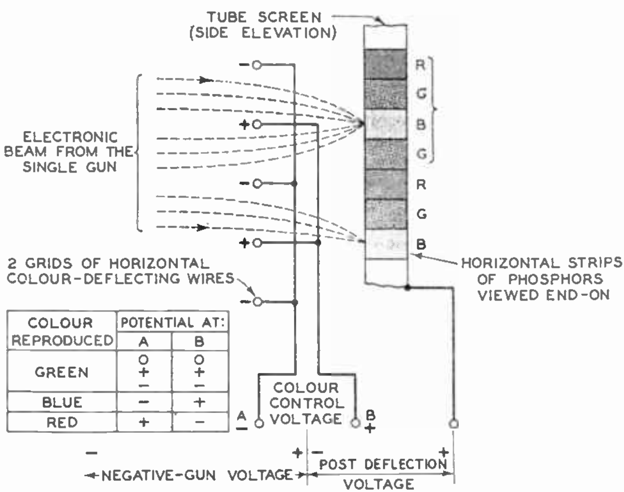

FIG. 22.—ONE-GUN DISPLAY TUBE USING COLOUR SELECTION BY DEFLECTION SWITCHING.

U.S.A. uses exclusively the three-gun shadow-mask type of tube, but this may well not be indicative of future trend.

All the above display means may be used whether the transmitted signals are simultaneous, in rapid sequence, or in slow sequence such as field sequential, so that the type of transmission is not of necessity tied to the display means. This is not to say that greater efficiency and economy cannot be achieved by so doing.[5]

Transmission Systems

The question of "compatibility" is one to which much importance is attached. A colour television system is said to be fully compatible when colour signals transmitted by it through existing black-and-white transmission equipment can be received not only by colour receivers in colour but also by existing black-and-white sets in black-and-white. It will be noted that this definition requires the colour transmission to occupy the same band-width as the black-and-white transmission with which it is compatible as well as to have the same scanning standards. There are many forms of partially compatible systems, e.g., one in which a signal containing information about colour is sent outside the black-and-white band-width.

Of completely incompatible systems, the classic one uses the field-sequential method. Here, the scanning standards are quite different from black-and-white; field and line-scan frequencies being three times the black-and-white standards. Band-width for equivalent definition is likewise three times the normal. Such a system would have to be built up in the U.H.F. band for lack of adequate band-width in Bands I and III. Although it would appear that this system is now unsuitable for broadcasting, but not necessarily for closed-circuit or industrial or medical colour television, it has the advantage of being time multiplexed in contradistinction to frequency multiplex on the one hand, or separate-channel working on the other: each of the three colour fields passes through the same channel, so that exact matching of three channels often containing non-linear circuits is not required, as it is in parts of some simultaneous systems. The very considerable difficulties associated with precise alignment of cameras which are required for three-tube simultaneous operation has led to a scheme known as the chromacoder, which is described later.

The only nationally adopted colour television system is the American National Television System Committee (N.T.S.C.) system, which is, at least nominally, fully compatible. This will be described with reference to tentative British figures scaled from the U.S.A. standards adopted by the Federal Communications Commission (F.C.C.). If it is assumed that present black-and-white pictures are panchromatic, then they represent the brightness or luminance component of the light reflected from the scene to be televised; thus a compatible colour system must, as its prime condition, transmit a luminance signal.

Colorimetry

The coloured light which illuminates the television pick-up device can be adequately described, as has been said, by specifying the amounts, fluxes (in lumens) of three suitable primary coloured lights which would be required to synthesize it, and in fact any colour-television system must start and finish with three primary beams—after analysis at the pick-up device and before synthesis at the display device. In between analysis and synthesis, however, the three primary signals may be coded and decoded in any suitable manner, and it is at this stage that an examination of some of the properties of human colour vision becomes most profitable. If we assume that the colour-addition properties of the eye are linear,[6] then any linear transformation of the, say, red, green and blue primary signals will give rise to three new signals of equal validity. The idea of panchromaticity necessitates the concept of brightness or luminance as distinct from colour or chrominance. The luminance of a patch of colour light is simply a measure of the radiant power coming from it after passage through a transmission filter having a transmission/light-wavelength characteristic equal to that of the "standard eye". This is called the photopic curve. Luminance is thus what remains of a quantity of light after chrominance has been removed. Chrominance can be separated into two components; conveniently, hue and saturation. Hue is measured by the dominant wavelength or the wavelength of a monochromatic radiation which when added to a suitable amount of a standard achromatic "white" will match the coloured light. Saturation is the ratio of the photometric flux (in lumens) which would be required from a source at the dominant wavelength to the sum of the fluxes required from a neutral or achromatic source and the source at the dominant wavelength for matching the given sample of light. A 100 per cent saturated colour has, therefore, no grey in it (although its constituent light has luminance) and is, in consequence, a dominant hue. A colour of 0 per cent saturation is grey, and is thus achromatic.

Thus a quantity of coloured light can be described either in terms of three quantities of primary coloured lights of suitable spectral distribution or in terms of luminance, hue and saturation or luminance and two chrominance components. Now the advantage of the latter descriptions is that they correspond with certain subjective properties of colour vision. First, the eye is far more sensitive to luminance flicker (at constant chrominance) than it is to chrominance flicker (at constant luminance). Unfortunately, no method of taking advantage of this has yet been devised. Secondly, the eye is far more sensitive to small-area luminance changes than it is to small-area chrominance changes. This means that fine detail (viewed at a certain distance) consisting of, say, black and white bars will be clearly discernible, whereas the same-sized pattern will not be visible at all if it consists of, say, green and red bars of the same brightness or luminance. This leads to the important conclusion that chrominance information can be sent in a narrower band-width (coarser detail) than luminance without any visible effects.[7] It has also been found that any medium-small patch of coloured light can be matched by two primary coloured lights of an orange and a cyan (blue-green) hue respectively.[8,9] Thus a system which derived the most benefit from the foregoing knowledge would transmit as a full three-colour system in large areas, a two-colour system in areas of medium size and a no-colour black-and-white system in small areas. Translating this into practical terms, we might choose to send full three-colour information in a video band-width of 340 kc/s, two-colour information in a 1-Mc/s band and luminance information only, in the full 3-Mc/s band in accordance with present practice.

It has also been shown[10] that interference from random noise with chrominance, at constant luminance, is considerably less visible than interference with the luminance channel. It is, therefore, to be expected that colour television, if the system is a constant-luminance one, will suffer scarcely any more than black-and-white television from noise disturbances.

The N.T.S.C. System Scaled to United Kingdom Standards

The result of the processes described above is termed "band saving", because if the signals representing the luminance and the two chrominance components are each sent through separate channels the total band-width will be $3 + 1 + 0{\cdot}34 = 4{\cdot}34$ Mc/s instead of the 9 Mc/s which would be required for a non-coded three-channel system or a sequential system.

The luminance signal may be obtained from the three primary signals by adding them together in the proportions in which each one of them contributes to the brightness of a panchromatic, achromatic or grey signal. Thus, if one lumen of grey (at a colour temperature of, say, 6,500° Kelvin) can be synthesized by the sum of 0·3 lumens of the chosen red primary, 0·59 lumens of the green and 0·11 lumens of the blue at the display screen, then the luminance signal can be constituted as

$$Y = 0{\cdot}3R + 0{\cdot}59G + 0{\cdot}11B \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where R, G and B are the voltages or currents, functions of time, obtained when a coloured scene is scanned by three simultaneous pick-up tubes. If we subtract the luminance signal from any one of the primary signals we obtain a colour-difference or chrominance signal from which brightness has been removed. Thus, two suitable chrominance signals are

$$R - Y = 0{\cdot}7R - 0{\cdot}59G - 0{\cdot}11B \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

$$B - Y = -0{\cdot}3R - 0{\cdot}59G + 0{\cdot}89B \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

As we have already seen, a third chrominance signal is unnecessary because it is necessary to transmit only three signals: R, G and B or linear transformations of them. Nevertheless, the colour-difference signal

$$G - Y = -0{\cdot}2\,(B - Y) - 0{\cdot}51\,(R - Y) \quad . \quad . \quad . \quad . \quad (4)$$

is useful in some receivers where the three colour-difference signals given by equations (2), (3) and (4) are put on to the grids of the three electron guns whilst a $(-Y)$ signal is put on to the three cathodes. In this way the transformation from colour-difference signals to colour-primary signals takes place in the electron beams of the three-gun tube. The processes of combining and decombining the three primary signals into luminance and chrominance signals, and vice versa, are sometimes called matrixing, since they correspond to the operation of three-line and column linear mathematical matrices. We shall have made maximum use of band saving if we transmit Y, $R - Y$ and $B - Y$ in band-widths of 3 Mc/s, 340 kc/s and 340 kc/s respectively, but we shall not have allowed the eye the benefit of a two-colour (orange–cyan) system in medium-fine detail. This feature can be achieved by changing the two chrominance signals into linear transformations of two of the colour-difference signals in such a manner that each chrominance signal represents colour at constant luminance along an axis in the chromaticity diagram [6,9] joining a dominant hue with its complement and ensuring that one of these axes is the orange–cyan line. This implies that the amplitude of a chrominance signal is a direct measure of the saturation of a transformed primary colour having a constant dominant hue and variable saturation. For instance, the chrominance signal lying on the orange–cyan axis, which will be referred to as the I component of chrominance (in place of $R - Y$), is positive when an orange scene is being transmitted, negative for a cyan scene and zero for a grey scene. Similarly, the other chrominance component, known as the Q signal (in place of $B - Y$), is positive for a magenta scene, negative for a green scene and zero for a grey scene. Now, if we transmit the Q signal through a 340-kc/s band-width and the I signal through a 1-Mc/s band-width, we shall have a full three-colour system down to a detail size of nine picture elements corresponding to 340 kc/s, and a two-colour system for smaller detail down to three picture elements in size corresponding to 1 Mc/s, whilst for detail corresponding to between one and three picture elements only the luminance signal will remain and so such detail will be reproduced in grey. Thus, an abrupt transi-

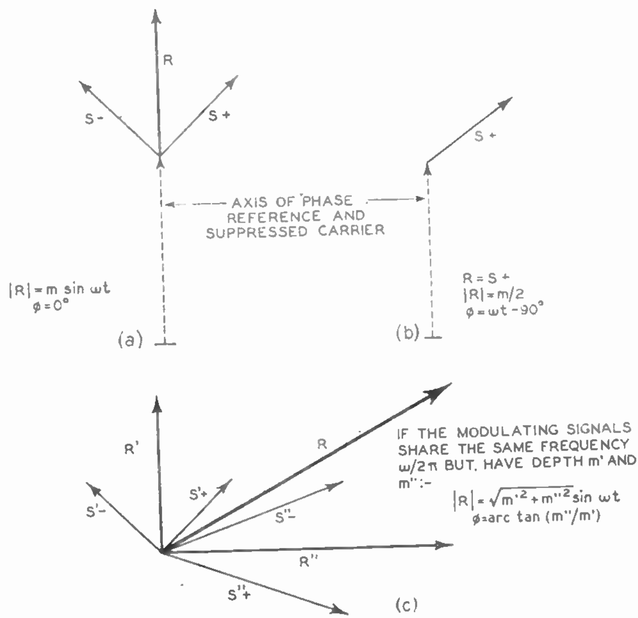

FIG. 23.—METHODS OF MODULATION.

$S-$ and $S+$ are sidebands resulting from the modulation of a carrier wave $\sin \omega_0 t$ by a modulating signal $m \sin \omega t$. R is the resultant wave of the entire process and has an amplitude $|R|$, a frequency $\omega_0/2\pi$ and a phase angle ϕ.

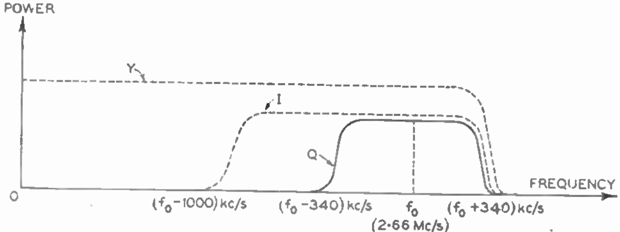

FIG. 24.—SPECTRUM OF THE COMPOSITE VIDEO SIGNAL.

tion in the scene from, say, a lemon yellow to a blue at constant luminance would commence with the correct yellow, but as the Q chrominance component has a limited band-width, the colour would veer towards the desaturated orange until the I component band-width attenuated it and then a grey interspace would be visible upon close inspection. This would give way to a desaturated cyan as the I component returned, and this would finally be replaced by the correct blue as the Q component returned to complete the three-colour display. It should be noted that because all the band-width limitation is effected upon chrominance signals at constant luminance instead of upon primary signals carrying luminance information, there is no defect comparable with chromatic aberration.

The I and Q components, being linear transformations of $B - Y$ and $R - Y$, may be expressed in terms of the primary signals thus :

$$I = 0{\cdot}6R - 0{\cdot}27G - 0{\cdot}32B \qquad (5)$$

$$Q = 0{\cdot}21R - 0{\cdot}52G + 0{\cdot}31B \qquad (6)$$

We have seen how the maximum band-saving consistent with the colour-vision properties of the eye can be achieved. We now examine the method of modulation used for the chrominance components in order that "band sharing" shall be accomplished with minimum interference, thus enabling more or less complete compatibility to be achieved.

The method adopted, Fig. 23, is to modulate the I and Q signals on to two suppressed sub-carriers in quadrature. This results in two double-sideband suppressed-carrier systems in quadrature, with no cross-talk between the two sets of sidebands. However, although the Q sidebands are transmitted without asymmetric attenuation, the I suppressed-carrier system is transmitted in an asymmetric-sideband fashion, only the lower set of sidebands extending beyond 340 kc/s to 1 Mc/s. This would cause cross-talk, Fig. 23(b), into the Q channel, were it not for the fact that the Q signal does not exist in the 340 kc/s to 1 Mc/s cross-talk region and can therefore be protected by selectivity. It will be seen that the sub-carrier carrying the chrominance components is situated within the normal black-and-white or luminance video band, Fig. 24, and, of course, interference results. This is the result of band sharing. The interference can be eliminated in new receivers, whether monochrome or colour, by the insertion of a narrow-band elimination or notch filter in the receiver luminance channel. In receivers already in use, the interference cannot be affected, but it is abated in the transmission by arranging that the sub-carrier frequency is locked to an odd multiple of half the line-scan frequency. This effects a partial cancellation of the beat-frequency pattern arising from the chrominance signal on homologous lines of alternate fields.

This method of quadrature modulation requires synchronous detection in the receiver: ordinary envelope detection would not be capable of separating the two quadrature components. The process, Fig. 48, of extracting the chrominance components consists of, first, selecting the chrominance signal from the composite video signal by means of bandpass filters, and then feeding it into two separate channels, in one of which it is multiplied by a reference signal (obtained by means to be described) having the same frequency and phase as those of one of the quadrature components of the suppressed sub-carrier, and in the other it is similarly multiplied by a reference signal having the same frequency and phase as those of the other of the quadrature components of the sub-carrier. Thus the I or "in-phase" chrominance component will be obtained from one channel whilst the Q or "quadrature" component will be found in the other. A matrix unit which is basically a set of potentiometers then reverses the processes described by equations (1), (5) and (6) and the original primary signals R, G and B are finally obtained.

There are several methods of locking the receiver synchronous detection reference oscillator in frequency and phase to a reference signal. The crystal and automatic phase-control feedback loop is perhaps the best example. The reference signal consists of a "burst" of eight cycles of a sine-wave of sub-carrier frequency and phase transmitted in each post line-synchronizing suppression period. The amplitude of the "burst" sine-wave is half the magnitude of the synchronizing pulses, and its mean value is situated upon suppression or electrical black level. It has been shown [11] that eight cycles of reference "burst" repeated in every line-suppression interval is adequate to lock a local oscillator to within $\pm 5°$ r.m.s. when the peak-white signal to r.m.s. noise ratio is 10 db. An effective band-width in the automatic-phase control-loop circuit of the local oscillator of about 100 c/s is required if phase lock to within $\pm 5°$ is required in conditions of such high noise level.

We can now write introductory expressions for the composite [12] video signal:

$$M = Y + I \cos(\omega_0 t + 33°) + Q \sin(\omega_0 t + 33°) \quad . \quad . \quad . \quad (7)$$

remembering that the video components of Q only exist for frequencies below 340 kc/s whilst those of I exist for frequencies below 1 Mc/s. $\omega_0/2\pi$ is the chrominance sub-carrier frequency. Below 340 kc/s, where a full three-colour system is available, we may replace I and Q by their linear transforms $R - Y$ and $B - Y$, thus obtaining an equivalent formula

$$M = Y + 0{\cdot}88\,(R - Y)\cos\omega_0 t + 0{\cdot}49\,(B - Y)\sin\omega_0 t \quad . \quad . \quad (8)$$

In fact, the 33° in equation (7) is the angle through which it was necessary to rotate the vector $R - Y$ in order to superimpose its chromaticity equivalent upon the orange–cyan chromaticity axis.

So far, the N.T.S.C. system has been explained as if luminance and chrominance were transmitted without error, but this is not, in fact, entirely the case. It will be remembered that all picture-tube brightness/control-electrode characteristics approximate to a simple power

TABLE 2.—MAGNETIC POWDER CORE MATERIALS

Material	Relative Permeability (μ_e)	Loss Factors			Applications
		$a \times 10^6$	$c \times 10^6$	$e \times 10^{10}$	
Electrolytic iron, <325 mesh	16	250	1,200	200	(Little used.)
Carbonyl-iron, E type	12	3	200	2	Radio frequencies: first-grade cores.
Carbonyl-iron, F type (special fine particle size)	10	2	100	<2	Radio frequencies (above 10 Mc/s).
Hydrogen-reduced iron	20	50	750	30	Radio frequencies, where permeability more important than losses, permeability tuning, etc.

facture. Since they are made from finely divided oxides mixed with a binding agent, compressed and baked, the mechanical properties depend to a large extent upon the nature of the binder, the amount of compression and the subsequent heat treatment. In general, however, the ferrites are hard and brittle, and are frequently classed as ceramics.

The high electrical resistivity of the ferrite magnets and their consequent low eddy-current losses in alternating fields make them very suitable for providing a polarizing field in inductors and transformers. Examples are unidirectional pulse transformers, polarized relays, and microphone and telephone units. A further advantage is the comparatively short length of magnet required, leading to a reduction in the effective air gap introduced into the transformer.

Because of their resistance to demagnetization by external fields, ferrite magnets can be used in opposing pairs to provide a field of strength variable with their relative position. This principle has been used in the focusing of television receivers, and could be applied also to the control of inductance in variable reactors.

Magnetic Powder Cores

The original commercial application of magnetic powder cores was for telephone loading coils. Powder cores in radio receivers can supply the need for high-Q inductors for tuning circuits, but their use is based more on economic considerations in providing a given selectivity at less expense or in a smaller space.

For broadcast frequencies and above, small cores of a variety of shapes and sizes are used, giving similar Q-values, reduction in size being possible because of the preponderance of eddy-current losses at high frequencies, for which the effective resistance is independent of core dimensions. For the lower radio (i.e., broadcast-receiver intermediate) frequencies or for high-frequency carrier telephony, a compromise is often made in the form of cores completely or partially shrouding a winding.

Besides these main functions of powder cores, subsidiary but important uses have developed, mainly in the radio-frequency fields. These depend on the possibilities of inductance variation by the relative motion of core and coil. There are inherent limitations on the range of possible adjustment; because the toroidal form is impossible, only a fraction of the effective permeability of the core materials can be utilized. Where the maximum range is desired, it is necessary to use a cylindrical core with a large length/diameter ratio, which restricts the Q-value. However, cores giving an inductance ratio of 10/1 have been successfully made for radio-receiver tuning.

High-permeability Powder Cores

Attention has recently been given to the development of powder cores of specially high permeability, at the expense of increased losses. Such materials may be made from inexpensive but readily compressible and relatively coarse iron powder, sometimes in the form of flake. They have uses at power frequencies as inexpensive replacements for silicon iron (Stalloy) stampings in small transformers, reactors, etc., in which power loss is unimportant. They also bridge the gap between dust cores and normal laminations. In general, their hysteresis losses are high compared with those of silicon iron and similar laminations, but eddy-current losses are reduced. An important application in television receivers is to line-frequency transformers for providing the E.H.T. voltage. A high permeability is required to obtain adequate coupling, and under these circumstances lower losses can be obtained than with silicon–iron stampings. Another use is for the moulded pole pieces for electromagnetic deflectors for television cathode-ray tubes.

A typical selection of coils and cores is shown in Fig. 4.

Ferro-ceramic Beads

There are a number of applications in radio and television receivers for small beads, in the form of short tubes about 3 mm. in length with an outside diameter of about 3·5 mm., of ferro-ceramic material, e.g., Ferroxcube. The beads are slipped over the appropriate lead, and held in place, if necessary, by a small piece of sleeving. The effect of the bead is to increase the inductance and H.F. resistance offered by the wire on which it is threaded. The increased H.F. resistance is useful for decoupling purposes, and, when the bead is threaded on the grid lead, for suppressing parasitic oscillation. A common use is for decoupling heater wiring in Band II radio tuners and Band III television tuners, H.F. reactance being added at no increase in the D.C. resistance.

Microwave Unidirectional Ferrite Elements

Several unidirectional waveguides and isolators which form a convenient means of isolating micro-wave circuit elements to reduce the effect of reflection caused by mismatch have been marketed. These isolators depend for their operation on the non-reciprocal behaviour of ferrite materials at micro-wave frequencies. This property is caused by the phenomena of gyro-magnetic resonance exhibited when a sliver of ferrite is placed in a waveguide and subjected to a transverse magnetic field. This has the property of attenuating a wave propagating in one direction but not one propagating in the opposite direction. A typical X-band isolator provides an isolation of about 20 dB for an insertion loss of less than 1 dB.

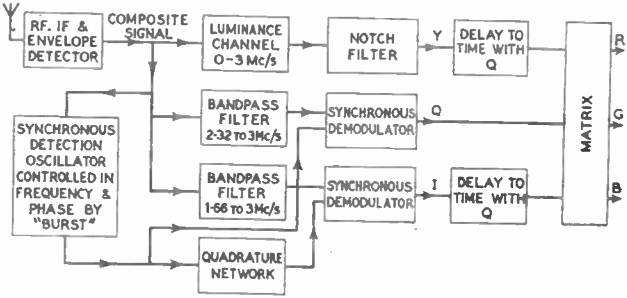

FIG. 25.—RECEIVER BLOCK DIAGRAM.

law of exponent between 2 and 3, and to ensure a linear overall scene-to-viewing-screen contrast law, the transmitted signal is "gamma corrected" by being raised to about the 1/2·5 power in all black-and-white systems. An ideal compatible colour system would send a gamma-corrected luminance or Y signal just as in monochrome practice, but to avoid considerable extra complication in colour receivers, the N.T.S.C. luminance signal, instead of itself being gamma-corrected, is composed of gamma-corrected primary signals: $R^{1/\gamma}$, $B^{1/\gamma}$, $G^{1/\gamma}$. This results in the transmission of saturated colours with inadequate luminance components, the missing luminance information finding its way into the chrominance signal which is relatively narrow band. Black-and-white receivers would thus show a compatible picture which would be non-panchromatic in that the more saturated the colour at the transmitting end of the broadcast the lower would the brightness allotted to it. Colour receivers would regain the missing luminance information from their chrominance channels, except in fine and medium-fine detail requiring a band-width greater than those possessed by the chrominance channels. These effects are quite visible for really saturated blues and reds; so much so, that for saturated colours it is scarcely possible to distinguish between a transmission in which the I signal is given its full 1-Mc/s band-width and one in which the I band-width is reduced so that it is the same as that for the Q signal (340 kc/s). This failure of the application of the constant-luminance principle rapidly ceases to have much significance when colours of normal saturations are reproduced.

To indicate that gamma-correction is applied to the primary signals before they are coded or matrixed into luminance and chrominance signals, the Americans use primes; thus primes would be added to the letters R, G, B, I, Q, Y and M in equations (1) to (8).

Pick-up Devices

In the present state of the art these can be subdivided into slow-sequential and simultaneous.

(i) The slow-sequential method of pick-up has been used for some time for experimental field-sequential systems of colour television. It consists of interposing between the optical lens and the photo-cathode or photo-sensitive surface of a camera tube a rotating disc carrying translucent colour filters in the same way as described under "Display Devices" method (iii). Care must be taken to avoid colour poisoning resulting from any persistence or memory of one colour field being carried over into the next due to the storage properties of the pick-up tube. The clumsiness of the rotating-disc method of colour synthesis is much less when the same method is applied to analysis at the transmitting end of the broadcast. This is because of the much smaller size of the photo-sensitive surface of a pick-up tube in comparison with the size of a phosphor screen in a viewing or picture tube: much smaller rotating discs can thus be used for cameras than is possible for receivers.

So convenient is the sequential method of analysis when compared with simultaneous means that, in spite of the degradation caused to any television signal by present techniques of "standards conversion",[13] a device known as the chromacoder, mentioned under "Transmission System", is being tried out in the U.S.A. in place of a three-tube camera for simultaneous colour television (in fact the N.T.S.C. system). The method is to analyse sequentially the scene to be televised and to "gate" each of the three sequential fields on to the control electrode of each of three standards-converter display tubes. Three low-velocity pick-up tubes, for preference orthicons (on account of their stable "black" level), are used to convert the field-sequential displays to simultaneous signals: a red, a green and a blue. These standards-converted simultaneous signals can then be used for simultaneous colour television.

(ii) Three-tube cameras[14] employing three image-orthicons are, at present, the usual live pick-up devices used. Owing to the programme requirement for a multi-lens turret, a light-relay[15] system is employed, and this contributes to a loss of sensitivity which, when coupled to the need for operating the camera tubes on the linear portions of their transfer characteristics, results in the need for a tenfold increase in studio lighting—400 ft.-candles. Dichroic-mirror filters are used for splitting the incident beam of light representing the scene to be televised into the required three primary beams which in turn feed the three camera tubes.

A bright future is claimed for cameras using vidicon pick-up tubes based on photo-conductivity rather than on photo-emission. It has been suggested that cameras using less than three tubes may be developed.

For film-scanner use both the flying spot system with three photo-multiplier tubes and dichroic beam-splitting mirror-filters and the three-vidicon[16] pulsed-light methods are used. Considerable competition between these two methods seems likely in the future.

Light-analysis Filters

The need to analyse the light incident upon the image pick-up device, be it camera or film scanner, into three primary-coloured beams is common to all television picture sources.

The spectral transmission characteristics of the analysis filters must, to take a film scanner for example, be such that, in combination with the spectral distribution of the light from the flying spot tube and the spectral sensitivity curve of each of the pick-up photocells, the correct signals emerge from each photocell head amplifier. The variations of these signals with scene brightness and colour must depend directly upon the choice of primary phosphors (red, green and blue) in the display system. Once these receiver-tube screen phosphors have been chosen and their chromaticities determined, the analysis filters at the picture source become likewise fixed. All colour-television receiver display chromaticities must be identical to those upon which the picture-source spectral sensitivities were based. The method of finding what spectral distribution characteristics the flying-spot-tube-cum-analysis-filters-cum-photocell combination should present is, at least theoretically, quite simple. It consists of finding, in principle by observer tests, what quantities (in lumens) of each of the three receiver primary lights are required to match (visually) given quantities of light at each wavelength in the visible spectrum. Usually an equal-energy spectrum is assumed, that is, one in which the energy density (per unit wavelength) is uniform. The spectral distribution for the red channel (including flying-spot tube, analysis filter and photocell sensitivity) will therefore be proportional to the curve showing the amount of red receiver primary which, when added to suitable proportions of the green and the blue, will match each visible wavelength; similarly for the green and the blue channels.

The application of the above description to a colour camera as distinct from a colour-film scanner will be evident.

R. D. A. M.

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3. MATERIALS

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MAGNETIC MATERIALS

The improved performances achieved by the use of anisotropic alloys and the reduction in size and scaling down of other parts which this has made possible have led to their almost universal adoption for magnets for loudspeakers, microphones and for the focusing of television tubes.

Loudspeakers of the moving-coil type, which have a diaphragm to which is attached a light coil arranged to move freely in the annular air-gap of the magnet, usually employ a permanent magnet, generally of one of the types shown in Fig. 1.

The first of these uses a ring of magnet alloy magnetized axially, clamped between mild-steel pole pieces as shown. The efficiency reckoned as the ratio of flux in the gap to total magnet flux is about 40 per cent, falling as the gap field exceeds 1 Wb/cu. in. (10,000 gauss). The slug or central-pole type is shown in Fig. 1 (*b*), and comprises a magnet block surmounted by a cylindrical soft-iron tip, the whole being fixed into a cup or yoke with a front plate to form the outer gap face and return circuit for the flux: the maximum efficiency is around 55 per cent. A modification of this type, Fig. 1 (*c*), uses a centre pole which is completely magnet alloy. Efficiencies as high as 65 per cent are obtained within a very limited range of gap flux density, using Alcomax or Ticonal, which is directionally magnetized so that the flux lines, normal to the axis throughout the greater part of the centre pole, are turned through 90° in the vicinity of the air-gap. These last two types are widely used in television receivers, for which they are particularly suitable, owing to their almost complete lack of external leakage field, due to the shielding effect of the soft-iron cup or yoke.

Various assembly methods are used: ring types are invariably clamped with screws as shown; centre poles may be screwed, soldered brazed or held solely by magnetic attraction. In this latter case a positioning device for centralizing the centre pole is necessary. Microphone magnets of the moving-coil type utilize a magnet essentially

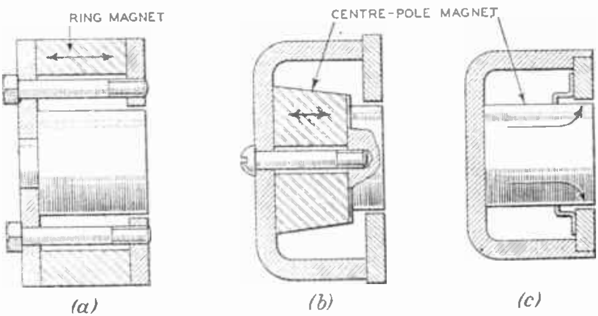

FIG. 1.—TYPICAL FORMS OF LOUDSPEAKER MAGNETS.

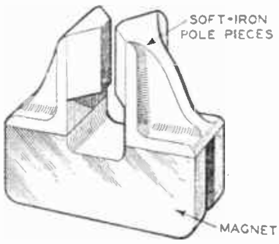

FIG. 2.—TYPICAL RIBBON-TYPE MAGNET FOR LOUDSPEAKERS.

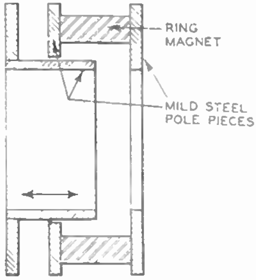

FIG. 3.—TYPICAL MAGNETIC FOCUSING ASSEMBLY FOR A TELEVISION CATHODE-RAY TUBE.

similar to that of a loudspeaker. Ribbon types use magnets similar to those shown in Fig. 2.

Focusing magnets for cathode-ray tubes use magnets basically of the form shown in Fig. 3. Flux densities at the centre of the annulus of the order of 0·02–0·03 Wb/cu. m. (200–300 gauss) are usual. Magnets of short length are required because of space considerations, and since they usually work on almost open magnetic-circuit conditions, and consequently have a low permeance, the alloys having very high coercivity, such as Alcomax III, IV or Hycomax, are most suitable.

Permanent magnets are also used for the purpose of picture centring in television receivers. In one common device, a small Alnico magnet is attached to mild-steel pole-pieces which are formed to surround the neck of the picture tube. Rotating the whole assembly adjusts the direction of the field of the magnet. Rotating the magnet in relation to the mild-steel pole-pieces, from a position at which the poles of the magnet are in line with the mild-steel pole-pieces (minimum strength) to one where they bridge the mild-steel pole-pieces (maximum strength), adjusts the field strength. The electron beam in the tube is centred by a combination of the effects of field strength and direction. An alternative system, especially suitable for 110-degree picture tubes, is to use two sheet steel magnets. These are magnetized diametrically, each being the equivalent of two C-shaped magnets in parallel. The field strength in the region of the electron beam can be varied from zero, when the two magnets have unlike poles adjacent, to a maximum when they have like poles adjacent. Rotating the two magnets together provides the means of adjusting the direction of the field.

Ferrites

The ferrites used for permanent magnets consist of mixed oxides of iron and one or more other metals, the heat treatment of the mixed oxides producing complex crystals with the required magnetic properties.

From the point of view of mechanical properties, ferrite magnets vary considerably, according to their particular method of manu-

TABLE 1.—COMPARATIVE TABLE OF BRITISH PERMANENT MAGNETS

Material	*(BH) Max.* $\times 10^6$	B_r *Gauss*	H_c *Oersted*	B_d	H_d	*Recommended Saturation Values* *		*Normal Composition (Balance Fe)*	*Approximate Heat Treatment* *
						B_{sat} *Gauss*	H_{sat} *Gauss*		
Ticonal GX †	7·5	13,500	720	12,000	625	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu	Heat approx. 1250° C. Cool in 2–15 min. to to 600° C. in magnetic field Temper approximate 550–650° C. between 4 and 50 hr. depending upon type
Ticonal G	5·7	13,480	583	11,000	520	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu	
Ticonal C	5·0	12,500	680	9,620	520	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu	
Ticonal L	5·5	14,000	575	12,000	460	17,000	3,000	7 Al, 14·5 Ni, 19·5 Co, 1·5 Cu	
Alcomax III	5·0	12,500	670	9,680	516	17,000	3,000	8 Al, 13 Ni, 24 Co, 3 Cu, 1 Cb	
Ticonal F (44/44)	4·8	12,400	600	10,000	480	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu, 0·5 Ti	
Ticonal H	4·5	11,800	770	8,270	545	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu	
Alcomax IV	4·3	11,200	750	8,000	537	17,000	3,000	8 Al, 11·5 Ni, 24 Co, 6 Cu, 2 Cb	
Alcomax II	4·3	12,400	575	9,570	450	17,000	3,000	8 Al, 11·7 Ni, 24 Co, 6 Cu	
Ticonal S	4·2	11,070	620	8,830	470	17,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu	
Ticonal E (42/50)	4·1	11,070	710	7,500	550	16,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu, 1·5 Ti	
Ticonal D (2·8)	3·8	12,000	600	9,000	420	16,000	3,000	8 Al, 14 Ni, 24 Co, 3 Cu, 1·0 Ti	
Ticonal E †	3·6	8,500	1,150	4,650	775	15,000	6,000	—	
Alcomax I	3·5	12,000	475	9,500	370	16,000	2,000	8 Al, 11·7 Ni, 24 Co, 3 Cu	
Hycomax	2·8	8,500	790	5,500	500	15,000	3,000	8·5 Al, 21 Ni, 20 Co, 15 Cu, 1·5 Ti	
Magnadur 2 (blue/red)	2·5	3,600	1,400	2,200	1,135	16,800	14,000	$BaFe_{12}O_{19}$	—
Magnadur 3 (blue/orange)	2·25	3,000	1,900	1,600	1,400	18,400	14,000	$BaFe_{12}O_{19}$	—

(*Mullard Ltd.*)

Table 1 (*contd.*)

Material	*(BH) Max.* × 10^6	B_r *Gauss*	H_c *Oersted*	B_d	H_d	*Recommended Saturation Values* * B_{sat} *Gauss*	H_{sat} *Gauss*	*Normal Composition (Balance Fe)*	*Approximate Heat Treatment* *
Reco 2A	1·92	5,500	1,000	3,300	600	13,000	4,000	7 Al, 20 Ni, 20 Co, 7 Cu, 6·5 Ti	
Reco 3A	1·7	7.200	645	4,350	390	13,500	3,000	—	
Alnico (high Br)	1·7	8,000	500	5,200	327	13,500	3,000	—	
Alnico	1·7	7,250	560	4,700	362	13,500	3,000	10 Al, 17 Ni, 12 Co, 6 Cu	Heat 1200–1250° C. Force cool to 600° C. between 1 and 4 min., depending upon type
Alnico (high Hc)	1·7	6,500	620	4,250	400	13,500	3,000	—	
Hynico	1·63	7,250	628	4,660	350	12,500	3,000	10 Al, 20 Ni, 13:5 Co, 6 Cu	
Alni (high Br)	1·25	6,200	490	4,000	312	12,000	2,000	—	
Alni	1·25	5,800	550	3,650	340	12,000	2,000	12 Al, 24 Ni	
Alni (high Hc)	1·25	4,700	700	2,840	440	11,500	3,000	—	
Hynical	1·15	5,250	674	3,290	350	11,500	3,000	12 Al, 32 Ni	
Magnadur 1 (blue/brown)	0·95	2,000	1,750	950	1,000	17,800	14,000	$BaFe_{12}O_{19}$	—
35% cobalt steel	0·95	9,060	250	5,930	160	15,500	1,000	—	Multiple heat-treatment usually involving at least one quench
15% cobalt steel	0·62	8,200	180	5,250	118	15,000	600	—	
9% cobalt steel	0·5	7,800	160	5,000	100	15,000	600	—	
6% cobalt steel	0·44	7,500	145	4,680	94	15,000	500	—	
3% cobalt steel	0·35	7,200	130	4,200	83	15,000	500	—	
6% tungsten steel	0·30	10,500	65	6,980	43	14,500	300	—	Water or oil quench from approx. 850° C.
3% chrome steel	0·285	9,800	70	6,200	46	13,500	300	—	

Columns 1–6 are based on manufacturers' published data. * Approximate figures not taken from manufacturers' data.

† These materials can only be supplied in certain shapes.

(*Mullard Ltd.*)

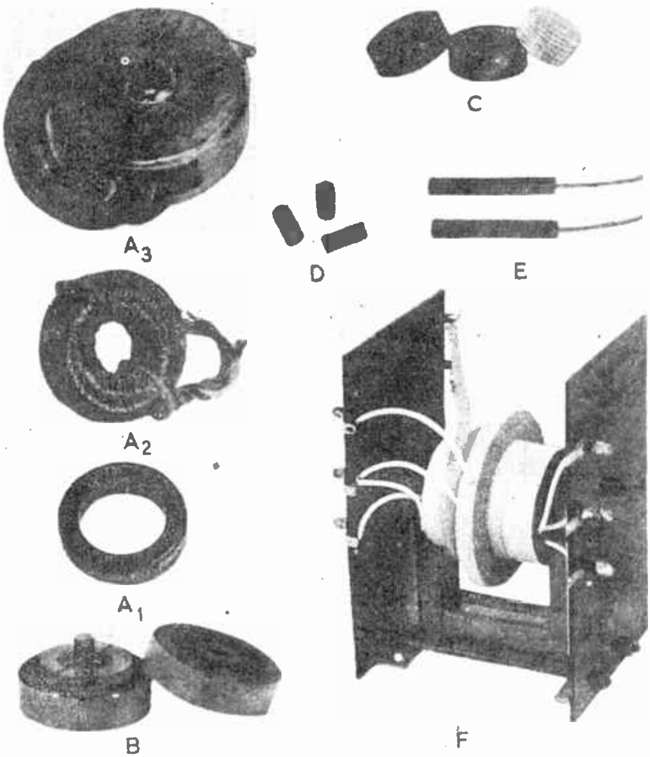

FIG. 4.—VARIOUS TYPES OF MAGNETIC POWDER CORES.

A_1. Toroidal core of high-permeability alloy powder for small loading-coil.
A_2. Wound coil on A_1.
A_3. Completed coil in screening can; Q-factor of about 100 at 1,500 c/s.
B. Large "pot" core showing winding former for carrier-frequency applications; $Q > 300$ from 50 kc/s to 2 Mc/s.
C. Small "pot" core for carrier and radio applications; $Q > 200$ from 80 kc/s to 3 Mc/s.
D. Small screw cores for radio applications: effective permeability about 2; $Q > 150$ from 500 kc/s to 20 Mc/s, > 100 up to 50 Mc/s.
E. Permeability-tuning radio cores: effective permeability about 10.
F. Television line-frequency transformer, wound on core of high-permeability iron-powder pressings.

(Salford Electrical Instruments, Ltd.)

INSULATING MATERIALS

The materials used for insulation are numerous and of the most diverse nature, including the majority of non-metallic substances.

There is no definite dividing-line between conducting and insulating materials, as the majority of the latter allow some current to flow, though usually only a very small amount. The definition of an insulating material (also called "insulator" or "dielectric"), given in the B.S.I. glossary of electrical terms, is "material which offers relatively high resistance to the passage of an electric current".

In Fig. 5 some typical materials, including metals and insulators, are arranged in order of electrical resistivity or "specific resistance" (i.e., the resistance measured across opposite faces of a unit cube of the material at a given temperature) on a logarithmic scale, somewhat similar to a spectrum, in order to give a picture of the relationships between conductors, semi-conductors, poor insulators and good insulating materials.

Whilst the conductors are practically all metals, capable of being produced and used in many different forms (e.g., wire, strip, sheet, bar and castings), and are easily formed to shape and machined, insulating materials have widely different characteristics: gaseous, liquid and solid; organic and inorganic; natural, refined, manufactured or synthesized. They vary from air—universally abundant and available at no cost—to expensive and somewhat rare minerals, such as mica (very limited in size and form).

No single insulating material has the adaptability of manufacture or the combination of properties required to enable its use to be extensive, such as occurs with metals—particularly copper and iron. Hence, the engineer has to employ different materials for insulation in various applications, according to the shape, properties, etc., required, and to the availability, cost, adaptability and often the reliability of the insulating materials otherwise suitable. For example, good electrical properties may have to be sacrificed to some extent to obtain the requisite mechanical strength, and certain materials having good heat resistance may not be appropriate because they cannot be applied as desired, e.g., as wrappings or tubes.

The chemical and relevant properties of insulating materials which are of importance in regard to their use as such, may conveniently be grouped as follows:

(*a*) resistance to external chemical effects;
(*b*) effects on other materials;
(*c*) chemical changes of the insulating material itself.

Under (*a*) there are such properties as resistance to:

(i) the effect of oil on materials liable to be used in oil (in transformers and switchgear), or to be splashed with lubricating oil;

(ii) effects of solvents used with varnishes employed for impregnating, bonding and finishing;

(iii) attack by acids and alkalis, e.g., nitric acid resulting from corona, acid and alkali vapours and sprays, and deposits of salts from sea spray;

(iv) oxidation, hydrolysis and other influences of atmospheric conditions, especially under damp conditions and in direct sunlight.

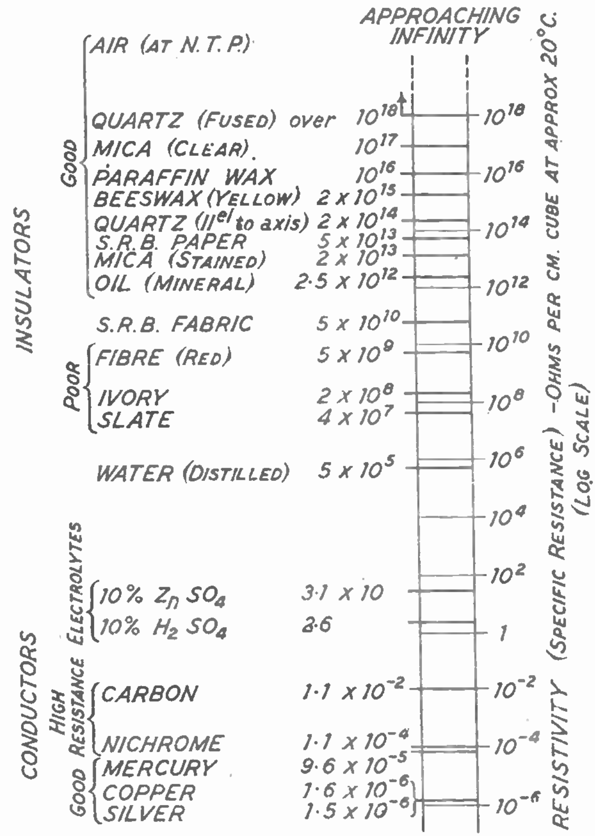

FIG. 5.—"Spectrum" of Electrical Resistivity of Materials.

In group (*b*) typical effects of the insulating materials on other substances with which they may be used are:

(i) direct solvent action, e.g., of oils and of spirits contained in varnishes, on bitumen and rubber; corrosion of metals in contact with the insulation; and attack on other materials by acids and alkalis contained in the insulating materials in a free state;
(ii) effects of impurities contained in the insulation;
(iii) effects resulting from changes in the material, for example, acids and other products of decomposition and oxidation affecting adjacent materials.

Group (*c*) includes such features as:

(i) oxidation resulting from driers included in varnishes;
(ii) deterioration due to acidity (e.g., in oils, papers and cotton products);
(iii) chemical instability of synthetic resins;
(iv) self-polymerization of synthetic compounds;
(v) vulcanization of rubber–sulphur mixtures.

Most of these chemical properties are determined by well-known methods of chemical analysis and test. The principal tests, and examples of B.S. specifications in which they are detailed, are as follows: acidity and alkalinity (by titration) and *p*H value (see B.S. 119 and B.S. 698), chloride content in vulcanized fibre (B.S. 934), presence of injurious sulphur (B.S. 688) and conductivity of aqueous extract (B.S. 698) for presence of electrolytes.

CLASSIFICATION OF INSULATING MATERIALS

Class	*Temperature* (° C.)	*Material*
Y	90	Paper, pressboard, cotton, rayon, silk, nylon, without impregnation
A	105	As Class Y but impregnated with bituminous and synthetic resin oil varnishes
E	120	Materials which can operate at 15° higher than Class A. Examples: varnished Terylene, cellulose triacetate film, Melinex film
B	130	Mica, glass fibre, asbestos, with suitable bonding or impregnating substances (e.g., synthetic resin varnishes, alkyds and polyesters)
F	155	As Class B but operating at 25° higher, e.g., alkyds, polyesters and silicone alkyds
H	180	Mica, etc., with suitable bonding or impregnating substances, e.g., silicone elastomers and resins
C	Above 180	Mica, glass, porcelain, quartz, with or without an inorganic binder

Materials are classified under temperature limits that will give an acceptable life under usual industrial conditions of service.

Epoxide Resins

Epoxide resins (e.g., Araldite, Epikote) are prepared by combining two complex organic chemicals into another yet more complex, and are generally made useful by the further addition of a hardener. The manufacturing process and choice of hardener are varied to produce resins suitable for various purposes. Further variation can be achieved by the addition of solvents, fillers, plasticizers, etc. The main uses in the electronic field are for impregnating coils and other components, for potting, and for laminating. They possess good electrical and mechanical properties with excellent resistance to ageing, and have good insulation properties. They are of particular use to meet Service requirements for potting components in equipment for use under adverse conditions.

TABLE 3.—RESISTIVITY (VOLUME), PERMITTIVITY AND POWER FACTOR OF TYPICAL INSULATING MATERIALS AT NORMAL TEMPERATURE (APPROX. 20° C.)

(*Note:* Figures given in this table are derived from various sources and are representative. In many cases the several properties of one material were determined on different samples by separate investigators.)

Material	*Resistivity (Volume) (MΩ/cm cube)*	*Permittivity (S.I.C.)*	*Power Factor*		
			f * = 50	f = 800–1,000	$f = 10^6$
Vacuum	Infinity	1·0000		Zero	
GASES					
Air (at N.T.P.)	Approaching Infinity	1·0006		Approaching zero	
Carbon dioxide	”	1·0010		” ”	
LIQUIDS					
Mineral Insulating Oil (B.S. 148)	10^7–10^8	2·0–2·5	0·0002	0·0001	—
Chlorinated dipbenyl	10^6–10^8	4·5–5·0	0·003	0·01	—
FUSIBLE COMPOUNDS, WAXES, ETC.					
Paraffin wax (refined)	10^{10}	2·2	—	0·0003	0·0001
Chlorinated naphthalene	10^8	5·0–5·4	0·002	0·001	0·006
Shellac	10^8	2·3–3·8	0·008	—	—
Bitumen compound	10^8	2·4–2·8	0·008	—	—
SHEET MATERIALS					
American white wood (dry)	10^6–10^8	3·0–3·8	—	0·07	0·04
Wood—Maple (dry)	10^6–10^7	4·4	—	—	—
Wood—Maple, paraffin-wax treated	3×10^4	3·2–4·1	—	—	—
Pressboard (dry)	10^4	3·1	0·013	—	—
Slate	10^2–5×10^4	6·0–7·5	—	—	—
Bitumen-asbestos compound	10^6–10^7	—	—	—	0·08
Ebonite (not loaded)	10^{16}	2·8	0·01	0·006	0·009
Hard rubber (loaded)	10^9–10^{12}	3·5–4·5	0·016	0·012	0·01
Paper (dry)	10^6	1·9–2·9	0·005	0·007	—
Paper (oil treated)	—	2·8–4·0	0·005	0·005	—
Varnished cotton cloth	3–5×10^8	4·5–5·5	0·2	0·15	—
Varnished silk cloth	3–5×10^8	3·2–4·5	—	0·06	—
Cellulose acetate film	6×10^8	4·0–5·5	0·023	0·04	—
Cellulose acetate moulding	10^8	4·0–6·5	0·016	0·03	0·06
Ethyl cellulose	10^7	2·5–3·7	0·02	0·03	0·01–0·03

Laminated Sheets					
Synthetic Resin Bonded paper, type I (B.S. 1137) †	10^7–10^8	4–6	0·02	0·03	0·04
Synthetic Resin Bonded paper ("Panilax") ‡	10^6–10^7	4·5–6·5	0·02–0·06	—	0·05
Synthetic Resin Bonded fabric (cotton) †	10^5–10^6	5–11	0·2–0·4	—	0·06
Synthetic Resin Bonded "Cotopa" fabric †	10^6	4·5	0·01–0·025	—	0·01
Synthetic Resin Bonded wood ("Permali") †	$>2 \times 10^6$	4·5–5·4	—	0·04	0·05
SYNTHETIC RESINS (UNFILLED)					
Phenol–formaldehyde	10^8	4–7	0·05	0·03	0·02
Phenol–formaldehyde (cast)	10^6	7–11	0·1	0·2	0·25
Aniline–formaldehyde	10^9	3·7–4·0	0·02	0·012	0·011
Polystyrene	10^{11}	2·5–2·7	0·0002	0·0002	0·0002
Polythene	10^{11}	2·3	0·0001	0·0001	0·0001
Methyl methacrylate	10^9	2·8	0·06	0·03	0·02
SYNTHETIC RESIN COMPOUNDS					
Phenol-formaldehyde—Wood filled	10^6 10^8	4–9	0·04–0·25	0·04–0·03	0·03–0·15
Phenol-formaldehyde—Mineral filled	10^6 10^8	5	0·01–0·02	0·01–0·02	0·007–0·02
Polystyrene—Mineral filled	—	3·2		—	0·0015
Urea-formaldehyde—Wood filled	10^6	5–8	0·06–0·1	0·04–0·1	0·038
Polyvinyl chloride compound	3×10^7	5–7	0·07–0·14	0·07	—
INORGANIC COMPOUNDS					
Porcelain	10^6–10^9	5–7	—	—	0·008
Steatite	10^6–10^9	4·1–6·5	0·0012	0·005	0·001
Special (steatite, etc.) ceramics for H.F.	10^{10} §	6–16	0·0005	—	0·0002
Rutile ceramics for H.F.	10^6–10^9 §	40–90	—	0·001	0·0004
Mycalex LDS. (Sheet and rod) grade	$>10^9$	7 average	—	0·007	0·002
Mycalex MLD. (Injection moulded) grade	$>10^9$	10·5	—	—	0·00125
Mycalon	$>10^{14}$	9·7	—	—	0·0013
VARIOUS (INORGANIC)					
Mica—Muscovite	10^7–10^{11}	4·5–7·0	0·0003	0·0002	0·0002
Mica—Phlogopite	—	5–6	0·005	0·002	—
Glass (plate)	2×10^7	6–7	—	0·0006	0·004
Glass special for H.F.	$>10^9$	5·25	—	—	0·0008
Fused quartz	$>10^{13}$	3·9	—	—	0·0002

* f is frequency in c/s. † Phenol-formaldehyde type. ‡ Aniline-formaldehyde type. § Resistivity at 300° C.

PLASTICS

There are two main groups of plastic materials: thermosetting and thermoplastic. The main disadvantage of the thermoplastic compared with the thermosetting materials is the relatively low heat resistance. The highest temperature at which a moulded or fabricated thermoplastic application will not lose over 10 per cent of its mechanical strength, or over which the appearance becomes affected or discoloured, is 140–190° F. at no load and 120–170° F. under mechanical load. The corresponding figures are 250–300° F. with phenolic materials.

Material Compounds

The following material compounds are used in the radio industry:

Thermosetting: phenolic plastics; cast phenolics; laminated phenolics; urea-formaldehyde; melamine-formaldehyde.

Thermoplastic: cellulose acetate and butyrate; polystyrene; ethyl cellulose; methyl methacrylate; vinyl material; polyethylene.

The mechanical, electrical, chemical and physical properties of the different materials are given in the manufacturers' catalogues. The *Plastics Comparator* (first published by the Bakelite Corporation, U.S.A.) shows the relative standing of various plastic materials for qualitative study. It should be noted that the numerical figures of the material properties given in Table 4 are established on standard test bars or pieces, and values of the finished articles or mouldings are usually different. In particular, the strength of the moulded articles is influenced by the flow of the plastics material during moulding, and also by the shape. The chemical, electrical and thermal properties are less affected by the manufacturing method or design.

Properties of Plastic Materials

For the radio industry, the power factor and dielectric properties of the different materials used are of special interest. The power factors of the principal materials are listed in Table 5. The general properties of thermoplastics and their applications to radio equipment are given in Table 3.

Design of Plastic Radio Components

The proper function of the plastic radio component depends to a great extent upon proper design, operating temperatures and the humidity of the surroundings. Material properties and the method of manufacture also have a great influence. Whilst it would be impracticable here to establish the rules to be considered when designing radio components, the following suggestions will be of service.

PROPERTIES	RECOMMENDATION AND REMARKS
(1) Mechanical properties	Even wall thickness and mass distribution. Openings well spaced and rounded to assure even flow. Avoid sharp edges.

TABLE 4.—PLASTICS COMPARATOR SHOWING MECHANICAL, ELECTRICAL, CHEMICAL AND PHYSICAL PROPERTIES OF DIFFERENT MATERIALS USED IN THE RADIO INDUSTRY

PLASTICS COMPARATOR

NUMBER 1 INDICATES THE MOST SUITABLE AND THE HIGHER NUMBERS THE ORDER OF SUITABILITY

	Mouldability around inserts	Dielectric strength	Resistivity	Loss factor	Hardness	Specific gravity	Heat insulation	Flammability	Continuous heat resistance	No dimensional change on heat ageing	Solvent resistance	Caustic resistance	Acid resistance	Water resistance	Cold flow	Colour stability	Tensile strength	Bending strength	Impact strength toughness
Phenolic general purpose	2	4	7	10	5	8	2	3	2	4	1	4	3	6	1	7	3	3	10
Phenolic low loss	2	3	3	4	3	12	7	1	3	2	1	4	4	3	1	7	7	3	11
Phenolic heat resistant	2	8	8		2	13	7	1	1	1	1	4	4	3	1	7	8	4	9
Phenolic acid & alkali resist.	3	7			4	5	2	2	3	5	1	3	2	4	1	7	8	6	10
Phenolic shock resistant	1	8	9		5	10	3	4	3	6	1	5	4	7	1	7	5	1	2
Phenolic transparent	3	6	5	7	4	6	2	2	3	5	1	3	3	4	1	7	3	1	7
Urea	4	2	4	9	1	11	5	5	7	7	1	4	4	9	2	1	1	1	8
Polystyrene	6	1	1	1	6	1	1	6	6	3	3	1	1	1	4	4	7	2	7
Cellulose-acetate	5	5	6	8	9	7	4	6	5	9	3	6	4	11	8	3	9	6	4
Aceto-butyrate	5			3	8	4	6	6	4	8	3	4	4	8	6	3	10	5	1
Ethyl-cellulose	5	1	2	2	8	2	4	6	5	8	3	2	4	10	7	6	6	2	3
Methyl-methacrylate	6	2		5	7	3	2	6	9	8	3	2	2	5	5	2	4	1	6
Vinyl no filler	5	1	3	6	7	9	2	6	8	3	2	1	1	2	3	5	2	1	5

TABLE 5.—POWER FACTORS

Material	*Power Factor*	
	At 60 cycles	*At 10^6 cycles*
Phenolic general-purpose powder	0·17	0·05–0·10
Phenolic low-loss powder	0·015–0·025	0·005
Phenolic medium shock resistors	0·10–0·25	0·19
Phenolic cast resin	0·113	0·027
Phenolic paper laminated	0·02–0·06	0·02–0·05
Phenolic fabric laminated	0·2–0·4	0·02–0·08
Urea	0·048	0·03
Melamine (cellulose filler)	0·037	0·029
Polystyrene	0·0002	0·0001
Methyl-methacrylate	0·05–0·07	0·02
Cellulose acetate	0·01–0·10	0·043
Vinyl chloride	0·02–0·15	0·043
Vinylidene chloride	0·02–0·03	0·04
Polyethylene	0·0003	0·0002
Porcelain	0·017	0·006
Steatite normal	0·003	0·002
Steatite H.F.	0·001	0·0001
Glass	0·03–0·005	0·002–0·006
Mica	0·02–0·003	0·006–0·002
Nylon	0·010	0·022

PROPERTIES	RECOMMENDATION AND REMARKS
(2) Wall thickness	Minimum for average parts $\frac{3}{32}$ in. for powders with fine fillers. For small mouldings, minimum 0·040 in. for thermosetting and 0·020 in. for thermoplastic materials.
(3) Shrinkage	Uniform cross-section for even shrinkage. Immediate moulding shrinkage can be kept within 0·002 in. per in. Urea materials shrink subsequently up to 0·015 in. per in. Large flat areas show shrink marks.
(4) Tolerances	Use the scheme of the British Plastics Federation, issued April 1945, and state tolerances on drawings.
(5) Threads	Avoid threads finer than 26 T.P.I. Standard metal fits and tolerances cannot be kept. Avoid thread cutting in mineral- and fabric-filled material. Consider self-tapping screws for permanent assembly.
(6) Assembly devices	Consider shrink fits, glued designs, spring hinges, self-tapping screws, speed nuts.

TABLE 6.—PROPERTIES OF THERMOPLASTIC MATERIALS

	Polythene	*Polystyrene*	*Polyvinyl Chloride*	*Methyl Methacrylate*	*Cellulose Acetate*	*Polyvinyl Acetate*	*Melamime*	*Ethyl Cellulose*
Specific gravity	0·96	1·05	1·2–1·6	1·18	1·3	1·3	1·5	1·10
Electrical strength breakdown (V/mil.)	1,000	500–700	250–450	400–450	200–300	500–700	700	400–700
Volume resistivity (50% H. 25° C.)	10^{17}	10^{17}	10^{15}	10^{11}	10^{10}–10^{13}	10^{15}	10^{11}–10^{13}	10^{13}–10^{14}
Power factor	0·0005–0·0006 At all frequencies	0·0001–0·0004	0·08–0·16	0·06–0·019	0·01–0·09	0·008–0·1	0·05–0·01	0·005–0·030
Dielectric constant	2·2 *	2·5–2·6	6·5–12	2·8–3·0	3·5–7·5	3·0–9·0	6·0–8·0	2·5–4·0
Moisture absorption (24 hr.) (%)	Nil	Nil	0·2	0·4	1·5–3·0	1·0	—	1·5–2·5
Softening temperature (°C.)	65–90	70–90	75–120	60–70	60–110	60–65	125–180	90–130
Coefficient of expansion per °C.	7×10^{-5}	7×10^{-5}	3×10^{-5}	9×10^{-5}	15×10^{-5}	3×10^{-5}	—	13×10^{-5}
Tensile strength (lb./sq. in.)	2,000	5,000–8,000	1,000–9,000	1,000–9,000	4,000–7,000	1,000–9,000	5,000–7,000	2,000–9,000
Hardness (Brinell)	1·0–2·0	20–30	2–50	18–20	8–15	12–15	16	4·0–8·5
Compression strength (lb./sq. in.)	Flows	13,000–13,500	—	11,000–13,000	—	—	—	10,000–12,000
Modulus of elasticity (lb./sq. in.)	Low	5×10^{5}	5×10^{5}	$4·6 \times 10^{5}$	$1·4 \times 10^{5}$	5×10^{5}	—	$2·5 \times 10^{5}$
Effect of acids	None	None	None	Very slight (oxidizing acids only)	Decomposes	Slight	Decomposes	Decomposes
Effect of alkalis	None	None	None	Very slight	Decomposes	Slight	Depends on filler: alpha-cell decomposes	Slight
Moulding or working temperature (°C.)	115–250	190–250	130–140	160–250	150–230	140–180	None	185–230
Applications	High freq. insulations; plugs; sockets; cable insulation; protective coating for coils and condensers	High freq. insulations; plugs; sockets; valve-holders; coil formers	Cable insulations; protective coatings; varnishes, sleevings and covered wire	Artistic parts; electrical optical systems	Conduits; boxes; formers; switch covers; housings; protective films	Similar to polyvinyl chloride	Varnishes special moulded parts	As cellulose acetate, also wire coverings

* At all frequencies falls slightly with increase in temperature.

Table 7.—Methods and Materials Used in Radio Industry

Methods	*Materials Used*	*Accuracy of Products*	*Remarks*
(1) Compression moulding	Thermosetting and thermoplastics with cooling	Fairly good—of standard quality	Standard commercial technique to produce mouldings
(2) Transfer moulding	Thermosetting materials	Improved closer tolerances	A variation of the compression method. Requires more pressure
(3) Lamination	Thermosetting (exceptionally, thermoplastic resins as binder)	Fairly good—of standard quality	Standard commercial technique to produce sheets, tubes or simple forms
(4) Pulp moulding	Pulp impregnated with thermosetting resins	Coarse	A process as used in the papier mâché industry is in course of development
(5) Injection moulding	Thermoplastic	Fairly close, better than compression moulding	Standard commercial technique for thermoplastic mouldings
(6) Jet moulding	Thermosetting	Fairly close, better than compression moulding	In development (injection moulding of thermosetting materials)
(7) Extrusion	Thermosetting or thermoplastic	Very good	To produce continuous rods, tubes, strips of different cross-section
(8) Casting	(Certain) thermosetting	Coarse	To produce advantageously larger mouldings of simple shape of various cross-sections

TABLE 8.—TYPICAL APPLICATIONS OF RADIO PLASTIC COMPONENTS

	Group	Plastics, Radio Part	Properties Required: Mechanical Strength	Appearance	H.F. Power Loss	Heat Resist.	Insulation	Water Resist.	Dimens. Stability	Accuracy	Material: Phenolic: Moulded	Phenolic: Cast	Phenolic: Laminated	Urea	Melamine	Polythene	Cellulose Ac-But.	Polystyrene	Meth. Methacrylate	Vinyl Mat.	Vinylidene Chl.	Manufacturing Method: Moulding: Compression	Moulding: Transfer	Moulding: Injection	Lamination	Casting	Extrusion
1	External	Radio Cabinet	●	●							●	●		●			●					●		●			
2	External	Knob		●						●	●			●			●					●		●			
3	External	Escutcheon	●	●							●			●	●		●		●			●		●			
4	External	Back Cover											●												●		
5	High Frequency	Valve Holder & Base	●		●	●	●	●	●		●											●	●				
6	High Frequency	H.F. Coil Former			●	●	●	●	●	●	●		●			●		●				●	●	●	●		
7	High Frequency	Disc Switch			●			●			●		●									●	●		●		
8	High Frequency	H.F. Cable Plug	●		●		●	●			●					●		●				●	●	●			
9	High Frequency	Crystal Holder			●		●	●			●											●	●				
10	High Frequency	Sleevings, Tubes, Rods			●								●			●			●	●	●				●	●	●
11	Low Freq.	L.F. Coil Former	●			●	●	●	●		●		●			●	●					●	●	●	●		
12	Low Freq.	Tag Panel	●			●	●	●	●		●		●				●					●	●	●	●		
13	Main Voltage	Main Plug & Socket	●				●	●		●	●		●									●	●		●		
14	Main Voltage	Control Panel & Term.	●	●			●	●		●	●		●									●	●		●		
15	Main Voltage	Insulated Control Spindle & Coupling	●				●	●	●	●	●											●	●				
16	Main Voltage	Main Switch Body	●		●	●	●	●	●	●	●			●	●							●	●				

Properties	Recommendation and Remarks
(7) Inserts	Use simply shaped, preferably round inserts, within correct manufacturing tolerances in order to fit mating mould parts. Provide for proper anchorage. Bolt-type inserts should have shoulder.
(8) Heat resistance	Actual service test should be made under load. Apply ribs or holes for air cooling and evenly distributed walls.
(9) Chemical properties.	Actual service tests should be made. Consider effect of ageing and effect of sunlight.
(10) Moisture absorption.	Detrimental results in loss of electrical properties. Choose the proper material for the particular job.
(11) Dimensional stability	Important, especially under tropical or humid conditions. Influenced mainly by water resistance of material, heat distortion and expansion, shrinkage or ageing. Cold flow.
(12) Electrical properties	State specific tests for moulding. Material specifications are correct to B.S.S. 771. At radio frequencies the values of dielectric strength are lower. Consider influence of moisture absorption, especially with machined surfaces.

Cabinets

The following moulding items will have to be considered when designing radio cabinets from plastics :

(1) Large and flat plane surfaces have to be avoided, curved or bowed portions, besides being structurally stronger, emphasize the lustre and depth of colour and by permitting an easier flow of the powder, waves and flow marks are reduced.

(2) Practically even wall-thickness at which the bottom thickness should be slightly above that of the side walls. Recommended thickness for the side walls 0·125–0·180 in., and for the bottom 0·170–0·200 in.

(3) Strong ribs inside result in shrinkage marks on the corresponding outside surface. It is advisable to cover this outside surface with ribs or by some surface treatment in the mould, such as sand blasting or planned patterns, or decorative fluting.

(4) Ribs are to be carefully designed and located, as by misapplication ribs give rise to stresses. Steps in the design may help.

(5) The after-shrinkage of the finished moulding should be considered, especially when using urea materials. Design in two or more parts should be considered as an alternative.

Low-loss Polymers

Examples of this group of materials are: polythene, polystyrene, polytetrafluorethylene (P.T.F.E.), polychlortrifluorethylene (Kel-F) and Terylene (polyethylene terephthalate). Information on capacitor applications is given in Section 29.

Polythene. This is the generic name of a synthetic, thermoplastic polyethylene with the following properties: low power factor at high frequencies; chemical inertness except in hot oil; flexibility and toughness. Its rather low softening point and ease of moulding prevent its use in applications where high temperatures are likely to be encountered. Treatment by electronic bombardment (irradiation), and other processing, can increase its resistance to oil and raise its softening point. The irradiated material is rubbery and elastic. Polythene does not resist discharge well, and has high contraction on cooling, leaving voids in the dielectric.

Polystyrene. This material is more rigid when cold, softening at 75° C. It is liable to crack, but has high intrinsic resistance at normal temperatures, falling only relatively slowly even up to 250° C. It is extensively used in the manufacture of small capacitors. It is supplied in films down to 10 μ thickness. It does not wrap as easily as paper, wrinkling being troublesome and static charges accumulating on the surface.

P.T.F.E. Examples of this are Teflon and Fluon. The material has many of the properties of polystyrene, but with a much wider temperature range (−60° to 250° C.). The power factor is very low (less than 0·0002), and the breakdown strength 1,000–2,000 volts/mil. As it cannot be melted or moulded, it is cold pressed as a powder and then sintered, resulting in a porous structure extremely susceptible to discharge. For wire covering, films down to 6 μ are available. As tape, the material is slippery and cannot be tied, and, because of its chemical inertness, adhesives cannot be used.

Kel-F. Also known as P.C.T.F.E., this is similar to P.T.F.E., but can be moulded and, as a wire covering, is less liable to plastic flow. Its mechanical properties are good, and it can be fabricated, extruded or moulded.

ELECTRICAL PROPERTIES OF POLYTHENE AND P.T.F.E.

Electrical Property	*Polythene*	*P.T.F.E.*
Specific insulation resistance	10^{17} ohms/cm.	10^{17} ohms/cm.
Dielectric loss angle	0·0004	0·0002
Permittivity	2·3	2·0
Breakdown voltage ($\frac{1}{16}$-in. sheet)	1,000 V/mil	500 V/mil

Terylene. This has a high melting point and is satisfactory for operation up to 150° C. It is a good insulator, of high electric strength, good heat resistance and low absorption properties. It can be made into fibres so that the woven material offers possibilities for use as a high-temperature-resistant material with good electrical properties and better mechanical properties than glass fibre. It has a common base with Melinex and Mylar, which have not the high electrical properties of Terylene, and is for that reason in insulation Class E.

SEMI-CONDUCTORS

Materials which are neither good conductors nor good insulators, i.e., the majority of substances, are known as "semi-conductors". There are, however, some which exhibit a peculiar behaviour, notably in the non-linearity of the relationship between some of the variables in the electrical circuit of which they may form a part.

Further discussion of "semi-conductors" will be found in Section 26 (Transistors). In this article information will be given of two further types of semi-conductors of particular importance to the radio and television engineer. These are:

(1) Thermistors (thermally-sensitive resistors) which have a large variation in resistance with temperature. They are synthetically prepared semi-conductors with a large negative temperature coefficient, i.e., their resistance decreases rapidly with rise of temperature. Thermistors are applicable to A.C. or D.C. circuits, and are generally made of mixtures of the oxides of various elements, including manganese, nickel, cobalt and copper, with a suitable binder in varying proportions according to the characteristics required. The mixture is moulded into the shape of a rod, disc or bead, and suitable leads are attached. In the form of a bead the thermistor may be mounted in a sealed glass bulb evacuated or filled with an inert gas according to the performance desired.

In indirectly heated thermistors, i.e., bead types, in which the bead is surrounded by a 100-ohm heater, the time constant may be defined as the time taken for the log of the resistance to change by $(e - I)/e$ (i.e., about 63 per cent when the heating power is suddenly changed. It is usual to measure the time constant for a change of heating power which will halve or double the resistance. It is not possible to obtain time constants of directly heated thermistors in the above manner, as these are largely dependent on the circuit in which a thermistor is used.

In vacuum-enclosed thermistors the time constant is larger for cooling than for heating; in gas-filled units the two periods are more nearly equal but shorter than for the evacuated types. The temperature change which brings about the variation of resistance may be conducted to the element from the surroundings, or may be developed internally by the passage of current through it.

(2) Non-linear elements, having a curvature in their current/voltage characteristic. Resistor elements are manufactured in the form of a disc or rod, usually containing silicon carbide. In these elements the relation between current and voltage is of the form $I = KV^x$, where K is a constant depending on composition and dimensions, and x lies between 3 and 5; so that doubling the voltage applied across the element may increase the current about twenty-fold, and trebling it about 100-fold.

Thermistors

The most widespread use of thermistors in radio and television practice is as current surge suppressors; suitable unpolarized resistive elements are available made in rod form. A typical range of such elements is manufactured and the following information on characteristics and applications is included by permission of the manufacturers.

The resistance of an element of this type decreases with rising temperature in such a manner that at room temperature an increase in temperature of the thermistor of approximately 20° C. will halve the value of the resistance. The temperature coefficient decreases as the temperature rises, and at 250° C. an increase in temperature of approximately 50° C. is required to halve the resistance value.

Voltage–Current Characteristic

When the voltage across the element is increased the current increases with it until a certain maximum voltage ($E_{max.}$) is reached. Thereafter the current will increase rapidly with a decrease of applied voltage, so that over this portion of the curve a negative resistance characteristic is obtained. The approximate value of $E_{max.}$ for a thermistor can be calculated by means of the following formula:

$$E_{max.} = \frac{\sqrt{R_c}}{k}$$

where R_c is the cold resistance and k is the maximum voltage factor for the type concerned. Typical values are given in Table 9.

Owing to the form of the voltage–current characteristic, a series-limiting resistor must always be employed to prevent excessive current flow when a low-impedance supply source is used.

The maximum operating current is a design centre rating which allows for normal supply voltage variation and an ambient temperature of 50° C. The maximum instantaneous current must not be exceeded, otherwise the component will be impaired permanently. A surge of this order may be experienced for a brief period soon after switch-on in certain valve-heater circuits. Should this surge exceed the rating, a suitable resistor must be shunted across the thermistor to ensure a slower and steady rate of increase of current during the warm-up period.

After a protracted period of storage under conditions of high humidity, some increase of initial resistance may be experienced, but this will return to its normal value once the component has been used.

At the maximum current ratings the temperature of the surge resistor may reach 250° C., so that the unit should be positioned carefully to prevent damage to other components; they should not be subjected to excessive mechanical stress, or fracture may occur.

Applications

Typical applications are shown in Fig. 6.

Fig. 6 (*a*) shows the use of a thermistor for surge suppression in A.C./D.C. receivers. Due to the low cold resistance of a series chain of valves and the fact that the hot resistance of the valve chain is an appreciable proportion of the total circuit resistance, an excessive current surge will flow at switch-on which may cause permanent damage to the valves. The thermistor should be connected in the circuit as shown, between the mains resistor and the heater of the rectifier valve; it must not be fitted between valve heaters or between the last valve heater and the chassis. In some equipment, especially television receivers, there is an appreciable variation in valve heating time within any

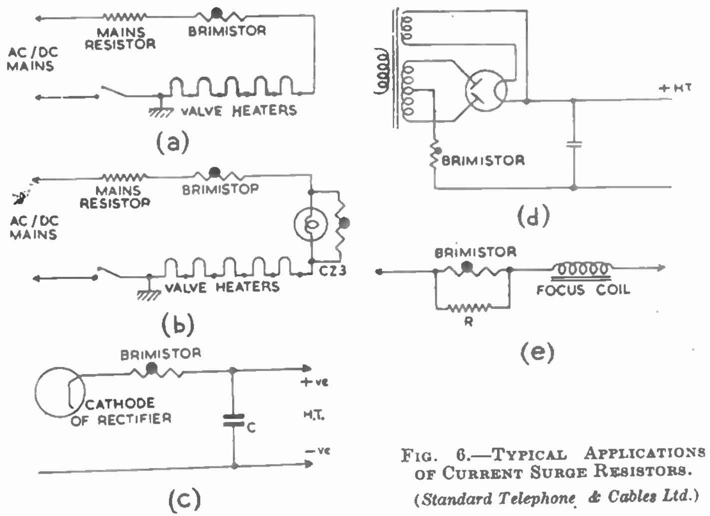

FIG. 6.—TYPICAL APPLICATIONS OF CURRENT SURGE RESISTORS.
(Standard Telephone & Cables Ltd.)

chain of valves, so that it is necessary to use a shunt resistor to prevent excessive voltage surges. It should also be remembered that the rate of heating of valves produced by different manufacturers may vary.

When a dial lamp is connected in the chain it will normally have to be under-run to allow for its short heating time and the consequent high voltage applied at switch-on. If a thermistor has been fitted as above for surge suppression, the dial lamp may be run at a higher temperature. An additional thermistor, as in Fig. 6 (*b*), may be wired across the lamp in place of a normal shunt resistor; should the dial

TABLE 9.—TYPICAL RANGE OF CURRENT-SURGE RESISTORS

Type	*Initial Resistance (Ohms)*			*Max. Voltage Factor k*	E_{max}. *Volts*			*Max. Operating Current (Amp.)*	*Resistance at Max. Operating Current (Ohms)*	*Max. Instantaneous Current (Amp.)*
	0° C.	*20° C.*	*50° C.*		*0° C.*	*20° C.*	*50° C.*			
CZ1, CZ1A	8,300	3,800	1,400	2·36	29·5	23	17	0·3	44	0·6
CZ2	12,500	5,500	1,850	2·47	37·5	30	20	0·3	38	0·4
CZ3	3,500	1,500	560	2·9	17	13·5	9	0·2	35	0·3
C4	1,700	800	320	1·2	28·5	23	15·5	1·25	5·5	2·0
CZ6	6,000	3,000	1,120	2·4	29	23	15	0·45	27	0·7
CZ8A	3,700	1,600	620	2·48	23·8	15·6	9·4	0·3	30	0·6
CZ9A	800	350	130	2·53	11·2	7·4	4·3	1·0	3·7	1·3

(Standard Telephones & Cables Ltd.)

lamp fail, the thermistor will warm up quickly and the set continue to function at full efficiency.

In order to reduce the switch-on current surge obtained with large reservoir capacitors associated with capacitor-input smoothing filters, a thermistor may be connected between the rectifier cathode and the leads to the reservoir capacitor, as shown in Fig. 6 (*c*).

The application of the full H.T. voltage to equipment when a directly-heated full-wave rectifier is used, may be delayed by connecting a thermistor in the centre-tap of the mains transformer (Fig. 6 (*d*)). Due allowance must be made for the fact that the r.m.s. current value must be used for selecting the correct current rating required. The r.m.s. current in the centre-tap will be 1·6 times the D.C. output current of the rectifier.

In order to compensate for the increase of resistance of a focus coil due to its rise of temperature while operating, the circuit shown in Fig. 6 (*e*) may be used. The thermistor should be connected in series, in close proximity to the coil, to ensure that it reaches a similar temperature. A shunt resistor may also be required for exact compensation of resistance change.

A common use of thermistors in television receivers is for picture-height shrinkage compensation. Frame time-base circuits tend to produce a constant voltage across the deflection coils, irrespective of changes of coil resistance due to self-heating, and under certain conditions this causes shrinkage of the picture height. This may be overcome by increasing the input drive to the frame output valve as the temperature of the deflection coils rises. The thermistor is connected in the H.T. supply lead to the frame oscillator (e.g., in the lead from the slider of the height control), and is mounted in close proximity to a hot component, such as a dropping resistor. By suitable lagging or spacing, correct compensation is obtained.

Thermistors may be used in resistor/capacitor smoothing circuits (replacing the resistor) to provide voltage stabilization under varying load conditions, provided that some increase of ripple with increasing current can be tolerated. As increased current is drawn from the power pack, the thermistor resistance decreases, thereby compensating the voltage drop. A thermistor in this application functions analogously to a swinging choke in an H.T. supply circuit for a Class B amplifier.

Temperature affects the operating current of transistor amplifiers (particularly in Class B circuits), and a thermistor is often connected across the transistor-base bias resistor to provide stabilization.

Non-linear Elements

The use in television receivers of non-linear resistive elements for the regulation of E.H.T. and for the provision of first-anode voltage for tetrode cathode-ray tubes has become widespread. Such elements also have a wide application for the protection of components, instruments and relays.

A typical element of this type is the " Metrosil " manufactured by a well-known electrical firm. Metrosil elements which contain silicon carbide are usually made in disc form, the faces of the discs being sprayed with brass. Electrical contact is then made by pressure on the sprayed surfaces or by soldering.

These elements have a negative temperature coefficient, and this property can be used for compensating purposes as already described under thermistors. The peculiar feature of this type of element, however, is its ability to form an "overflow" to prevent the build-up of high peak voltages. It can be arranged that under normal conditions comparatively small current flows through the element, but when a surge occurs this is conducted through the element. This characteristic may be used for the protection of instruments, relays, etc., under fault conditions, or for the protection of electrolytic smoothing capacitors under surge conditions, or for the protection of field coils, relay coils, etc., from the momentary rise in voltage across the coil that occurs when the current through the coil is stopped.

Television Applications

In television receivers suitable Metrosils may be connected directly across the E.H.T. provided by line-flyback-systems and the like in order to provide a degree of voltage regulation. When the output rises above normal, the current flowing through the Metrosil increases rapidly, and the extra load brings the voltage back to its normal level.

The provision of a first-anode voltage for tetrode cathode-ray tubes affords a problem, particularly for A.C./D.C. receivers. In many cases

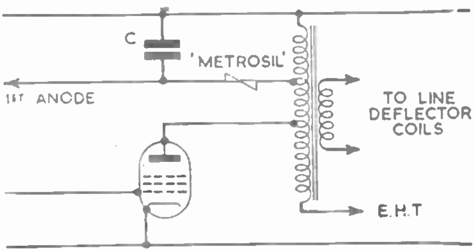

Fig. 7.—Use of a Non-linear Element for the Provision of a Cathode-ray-tube First Anode Voltage.

the cathode of the tube is modulated, and it is necessary to have the first anode about 150 volts or more above the cathode. This may require a supply of about 250 volts, whereas, unless some means of boost is incorporated, the H.T. available in an A.C./D.C. receiver may be of the order of only 190 volts. The non-resistive element provides a convenient means of producing a boost voltage, as seen in **Fig.** 7. A flyback pulse, so arranged that it is positive-going, charges C, and during the scan stroke the charge flows out, but conditions are such that the charge flowing out is less than that created by the pulse. The difference is about 150 volts, varying according to design, and is available as a positive potential for application to the first anode of the cathode-ray tube.

WIRES

Textile Coverings on Wires

Textile coverings on wires serve various purposes, depending upon the type of coil wound and the manner of winding. They act first as a protection, especially when the conductor is insulated by a thin enamel film. Secondly, they supply an absorbent layer between turns which, after impregnation with suitable varnishes or waxes, provides or augments the insulation of the conductor. At the same time, if the impregnation has been satisfactorily carried out, the coil is rendered impervious to moisture.

The protection of the wire is of the utmost importance in those cases where winding is necessarily severe, for instance, in machine-wound armatures, more especially when the turns are subsequently hammered or manipulated into armature slots.

Flexible lead-out wires and bunched enamelled copper conductors also need a textile covering. In the former case a variety of identification colours or colour combinations is more easily provided by a dyed textile; in the latter, a protection of the ultra-fine enamelled wires is a vital necessity, while in both cases it is necessary to bind the strands together to form as near round a cross-section as possible. Moreover, bunched enamelled copper conductors are invariably wave-wound, and a textile covering assists by its comparative roughness in building up the coil. It may be mentioned that B.S. 1258 : 1946 deals with the requirements for textile-covered bunched enamelled-copper wire conductors.

Choice of Textile

The choice of the most suitable textile is governed by the nature and function of the coil and the winding methods employed.

For field coils, single or double cotton coverings are used, since they most successfully withstand the hazards of the severe handling usually suffered. Rayon may be used, but does not generally provide adequate protection in these instances unless the thickness of the covering is increased.

For small solenoids, where generally space factor is important, pure silk and rayon are used. The high price of silk prevents its wide use, except on specialist apparatus, and rayon has almost entirely taken its place.

Medium and fine winding wires today are frequently covered with Regenerated Cellulose. There are two types :

(1) Cuprammonium.

(2) Fortisan.

Being cellulose, these rayons have dielectric properties similar to cotton, and can be treated with the same impregnating varnishes and waxes. In form they bear no comparison to cotton.

Cotton is made into thread by twisting together (spinning) fibres of a very short length. A lapping thread is made by winding parallel a determined number of such threads in tape form.

Rayon is produced in continuous filaments of extreme fineness which are collected in a determined number and wound without twist. The resulting single thread can be spread flat round the wire, and so produce a covering considerably finer than that permitted by the twisted threads of cotton.

Present production of the two rayons gives the advantage in elongation to Cuprammonium, while Fortisan enjoys the advantage in strength. High-speed winding demands wire with a reasonable elongation, which is best provided by Cuprammonium, but at the same time wave-winding is found easier with a wire on which the covering is comparatively rough. This feature is provided more readily by Fortisan, since the filaments tend to break up during the covering operation.

There are two other textiles used in wire covering, but with a more specialist application. These are "Celanese", a cellulose acetate yarn and "Cotopa", an acetylated cotton.

Both of these have outstanding insulating properties, a low moisture content and are resistant to attack by micro-organisms and insects. These characteristics make them especially suitable for coverings on switchboard wires and the like.

Designers should bear in mind that they are not bound rigidly to the dimensions and performances of conventional types of textile-covered wire It should also be remembered that many coil-winding problems

TABLE 10.—DATA FOR ROUND COPPER WIRES

S.W.G.	*Diameter (in.)*	*Area (sq. in.)*	*Lb. per 1,000 yd.*	*Ohms per 1,000 yd.*	*Ohms per lb.*
10	0·128	0·0129	148·22	1·866	0·0125
11	0·116	0·0106	122·22	2·272	0·0186
12	0·104	0·0085	98·24	2·826	0·0288
*	0·100	0·0078	90·83	3·057	0·0336
13	0·092	0·0066	76·88	3·612	0·0470
*	0·084	0·0055	64·09	4·332	0·0676
14	0·080	0·0050	58·13	4·776	0·0822
*	0·076	0·0045	52·46	5·292	0·1009
15	0·072	0·0040	47·09	5·897	0·1252
16	0·064	0·0032	42·00	6·611	0·1574
17	0·056	0·0024	28·48	9·747	0·3422
18	0·048	0·0018	20·93	13·267	0·6340
19	0·040	0·0012	14·53	19·105	1·314
20	0·036	0·0010	11·77	23·59	2·004
21	0·032	0·00080	9·30	29·85	3·209
22	0·028	0·000616	7·121	38·99	5·475
23	0·024	0·000452	5·232	53·07	10·144
24	0·022	0·000380	4·396	63·16	14·36
25	0·020	0·000314	3·633	76·42	21·03
26	0·018	0·000254	2·943	94·35	32·06
27	0·0164	0·000211	2·443	133·65	46·52
28	0·0148	0·000172	1·990	139·55	70·14
29	0·0136	0·000145	1·680	165·27	98·37
30	0·0124	0·000120	1·397	198·80	142·35
31	0·0116	0·000105	1·222	227·2	185·87
32	0·0108	0·0000961	1·059	262·1	247·4
33	0·0100	0·0000785	0·9083	305·7	336·5
34	0·0092	0·0000664	0·7688	361·2	469·8
35	0·0084	0·0000554	0·6409	433·2	676·0
36	0·0076	0·0000453	0·5246	529·2	1,008·7
37	0·0068	0·0000363	0·4200	661·1	1,574·0
38	0·0060	0·0000283	0·3270	849·1	2,597
39	0·0052	0·0000212	0·2456	1,130·5	4,603
40	0·0048	0·0000181	0·2093	1,326·7	6,340

* Denotes non-standard sizes.

arise through wrongly specified types. Of all covering media textiles are the most versatile, and can be modified in many cases to meet particular requirements or to overcome problems in the winding shop.

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Table 11.—Turns per Square Inch of Winding Space for Various Types of Wires

		Enamelled		S.C.C.		D.C.C.	
S.W.G.	*Bare Diam. (In.)*	*Covered Diam.*	*Turns per Sq. In.*	*Covered Diam.*	*Turns per Sq. In.*	*Covered Diam.*	*Turns per Sq. In.*
10	0·128	0·134	55	0·136	53	0·142	49
11	0·116	0·122	66	0·124	64	0·130	58
12	0·104	0·110	82	0·112	78	0·118	70
*	0·100	0·106	88	0·108	85	0·114	76
13	0·092	0·098	103	0·100	99	0·106	88
*	0·084	0·0895	124	0·092	116	0·098	103
14	0·080	0·0850	136	0·088	128	0·094	112
*	0·076	0·0805	152	0·084	140	0·090	122
15	0·072	0·0760	170	0·079	158	0·084	140
16	0·064	0·0675	217	0·071	196	0·076	170
17	0·056	0·0590	284	0·063	250	0·068	214
18	0·048	0·0508	384	0·055	326	0·059	284
19	0·040	0·0425	550	0·047	445	0·051	380
20	0·036	0·0384	670	0·042	560	0·047	450
21	0·032	0·0343	840	0·038	680	0·043	535
22	0·028	0·0302	1,080	0·034	850	0·039	650
23	0·024	0·0261	1,460	0·029	1,180	0·034	850
24	0·022	0·0240	1,720	0·027	1,360	0·032	968
25	0·020	0·0220	2,040	0·025	1,580	0·030	1,100
26	0·018	0·0198	2,520	0·023	1,870	0·028	1,260
27	0·0164	0·0181	3,020	0·0214	2,160	0·0264	1,420
28	0·0148	0·0164	3,670	0·0198	2,520	0·0248	1,600
29	0·0136	0·0151	4,350	0·0186	2,850	0·0236	1,770
30	0·0124	0·0138	5,200	0·0174	3,250	0·0224	1,970
31	0·0116	0·0129	5,920	0·0166	3,600	0·0216	2,120
32	0·0108	0·0121	6,750	0·0158	3,960	0·0208	2,270
33	0·0100	0·0112	7,850	0·0150	4,400	0·0200	2,470
34	0·0092	0·0103	9,330	0·0142	4,900	0·0192	2,670
35	0·0084	0·0095	11,000	0·0124	6,420	0·0174	3,250
36	0·0076	0·0086	13,400	0·0116	7,320	0·0166	3,600
37	0·0068	0·0078	16,200	0·0108	8,500	0·0158	3,950
38	0·0060	0·0069	20,700	0·0100	9,900	0·0150	4,400
39	0·0052	0·0061	26,500	0·0092	11,600	0·0142	4,900
40	0·0048	0·0056	31,500	0·0088	12,800	0·0138	5,200

* Denotes non-standard sizes.

TABLE 12.—TURNS PER INCH OF

S.W.G.	*S.C.C. Ord.*	*D.C.C. Ord.*	*S.S.C.*	*D.S.C.*	*Enam. and S.C.C.*	*Enam. and D.C.C.*	*Enam. and S.S.C.*	*Enam. and D.S.C.*	*Enam. and Paper*
6	4·954	4·810	—	—	—	—	—	—	—
7	5·382	5·215	—	—	—	—	—	—	—
8	5·895	5·694	—	—	—	—	—	—	—
9	6·519	6·272	—	—	—	—	—	—	—
10	7·283	6·979	—	—	7·105	6·811	—	—	—
11	7·987	7·623	—	—	7·787	7·441	—	—	—
12	8·850	8·403	—	—	8·621	8·197	—	—	—
13	9·918	9·352	—	—	9·636	9·101	—	—	—
14	11·26	10·55	—	—	10·93	10·26	—	—	11·70
15	12·55	11·81	—	—	12·15	11·45	—	—	12·94
16	13·96	13·04	14·97	14·68	13·50	12·64	14·43	14·16	14·47
17	15·74	14·59	17·01	16·64	15·17	14·10	16·37	16·03	16·42
18	18·02	16·82	19·72	19·23	17·33	16·21	18·90	18·45	18·98
19	21·10	19·46	23·47	22·78	20·20	18·69	22·37	21·74	22·47
20	23·61	21·11	25·91	25·06	22·52	20·24	24·63	23·87	24·75
21	26·10	23·08	28·99	27·93	24·81	22·08	27·40	26·46	27·55
22	29·17	25·46	32·79	31·45	27·62	24·27	30·86	29·67	31·06
23	34·20	29·21	37·88	36·76	32·15	27·70	35·34	34·36	35·59
24	36·73	31·04	42·19	39·68	34·48	29·41	39·22	37·04	38·46
25	39·68	33·11	46·08	43·10	37·17	31·35	42·74	40·16	41·84
26	43·15	35·49	50·76	48·31	40·32	33·56	46·95	44·84	45·87
27	46·38	37·66	55·25	52·36	43·29	35·59	51·02	48·54	49·75
28	50·12	40·08	61·35	57·47	46·73	37·88	56·18	52·91	54·35
29	53·36	42·12	66·23	61·73	49·75	39·84	60·61	56·82	58·48
30	57·08	44·40	71·94	66·67	53·19	42·02	65·79	61·35	63·29
31	59·81	46·04	76·34	70·42	55·87	43·67	69·93	64·94	67·11
32	62·85	47·83	81·30	74·63	58·48	45·25	74·07	68·49	70·92
33	66·23	49·75	86·96	79·37	61·73	47·17	79·37	72·99	75·76
34	69·93	51·81	93·46	84·75	65·36	49·26	85·47	78·13	81·30
35	80·00	57·14	101·0	90·91	74·07	54·05	91·74	83·33	86·96
36	85·47	59·88	109·9	98·04	79·37	56·82	100·0	90·09	94·34
37	91·74	62·89	120·5	106·4	84·75	59·52	108·7	97·09	—
38	99·01	66·23	133·3	116·3	91·74	62·89	120·5	106·4	—
39	107·5	69·93	149·3	128·2	99·01	66·23	133·3	116·3	—
40	112·4	71·94	158·7	135·1	104·2	68·49	142·9	123·5	—
41	—	—	166·7	140·8	—	—	149·3	128·2	—
42	—	—	178·6	149·3	—	—	158·7	135·1	—
43	—	—	192·3	158·7	—	—	169·5	142·9	—
44	—	—	208·3	169·5	—	—	185·2	153·8	—
45	—	—	227·3	181·8	—	—	204·1	166·7	—
46	—	—	250·0	196·1	—	—	222·2	178·6	—
47	—	—	277·8	212·8	—	—	250·0	196·1	—

Various Covered Conductors

Lewco-glass, Single	*Lewco-glass, Double*	*Lew-bestos*	*Enam.*	*Enam., Thick*	*Lewmex*			
					F	*HF*	*TF*	*QF*
—	—	—	—	—	—	—	—	—
—	—	—	—	—	—	—	—	—
—	—	—	—	—	—	—	—	—
—	—	—	—	—	—	—	—	—
—	—	6·930	7·532	7·342	—	—	—	—
—	—	7·565	8·307	8·085	—	—	—	—
—	—	8·333	9·259	8·994	—	—	—	—
—	—	9·264	10·44	10·12	—	—	—	—
11·66	11·26	10·44	11·98	11·60	11·96	11·81	11·64	11·36
12·87	12·39	11·81	13·28	12·87	13·26	13·07	12·89	12·56
14·37	13·77	13·05	14·90	14·43	14·88	14·64	14·41	14·01
16·23	15·48	14·58	16·98	16·47	16·92	16·61	16·31	15·80
18·69	17·70	16·53	19·72	19·08	19·65	19·27	18·87	18·21
22·03	20·66	19·08	23·53	22·73	23·42	22·88	22·37	21·51
24·15	22·52	20·66	26·04	25·13	25·91	25·32	24·69	23·64
26·81	24·81	22·57	29·15	28·09	29·07	28·33	27·62	26·39
30·03	27·55	24·81	33·11	31·85	33·00	32·05	31·15	29·67
34·25	31·06	27·62	38·31	36·76	38·31	37·17	35·97	34·01
36·76	33·11	29·24	41·67	39·84	41·67	40·32	38·91	36·63
39·68	35·46	31·06	45·66	43·48	45·45	43·86	42·37	39·84
43·10	38·17	34·25	50·51	48·08	50·25	48·31	46·51	43·67
46·30	40·65	36·23	55·25	52·36	54·95	52·63	50·51	47·17
50·25	43·67	38·61	60·98	57·80	60·98	58·48	56·18	52·36
53·48	46·08	40·49	66·23	62·50	73·53	62·89	60·24	55·87
57·14	48·78	42·55	72·46	68·03	71·94	68·97	65·79	60·98
59·88	50·76	44·05	77·52	72·99	76·92	73·53	69·93	64·52
62·89	52·91	45·66	82·64	78·13	81·97	78·13	74·07	68·03
66·23	55·25	47·39	89·29	84·03	88·50	84·03	79·37	72·99
—	—	—	97·09	90·91	96·15	90·09	84·75	77·52
—	—	—	105·3	99·01	104·2	98·04	92·59	84·03
—	—	—	116·3	108·7	114·9	107·5	101·0	91·74
—	—	—	128·2	119·0	126·6	119·0	111·1	101·0
—	—	—	144·9	133·3	142·9	135·1	126·6	113·6
—	—	—	163·9	151·5	163·9	153·8	142·9	128·2
—	—	—	178·6	163·9	178·6	166·7	153·8	138·9
—	—	—	192·3	175·4	192·3	181·8	169·5	151·5
—	—	—	208·3	192·3	212·8	200·0	185·2	166·7
—	—	—	227·3	208·3	232·6	222·2	204·1	185·2
—	—	—	256·4	232·6	256·4	243·9	227·3	208·3
—	—	—	294·1	263·2	—	277·8	256·4	232·6
—	—	—	333·3	303·0	—	322·6	294·1	263·2
—	—	—	400·0	370·4	—	384·6	344·8	303·0

Table 13.—Wire Gauges in Common Use

No.	*S.W.G.*		*B.W.G.*		*B. & S.*		*No.*	*S.W.G.*		*B.W.G.*		*B. & S.*	
	In.	*Mm.*	*In.*	*Mm.*	*In.*	*Mm.*		*In.*	*Mm.*	*In.*	*Mm.*	*In.*	*Mm.*
4/0	0·400	10·160	0·454	11·532	0·4600	11·684	24	0·022	0·559	0·022	0·559	0·0201	0·511
3/0	0·372	9·449	0·425	10·795	0·4096	10·404	25	0·020	0·508	0·020	0·508	0·0179	0·455
2/0	0·348	8·839	0·380	9·652	0·3648	9·266	26	0·018	0·457	0·018	0·457	0·0159	0·404
0	0·324	8·230	0·340	8·636	0·3249	8·252	27	0·0164	0·417	0·016	0·406	0·0142	0·361
1	0·300	7·620	0·300	7·620	0·2893	7·348	28	0·0148	0·376	0·014	0·356	0·0126	0·320
2	0·276	7·010	0·284	7·214	0·2576	6·543	29	0·0136	0·345	0·013	0·330	0·0113	0·287
3	0·252	6·401	0·259	6·579	0·2294	5·827	30	0·0124	0·315	0·012	0·305	0·0100	0·254
4	0·232	5·893	0·238	6·045	0·2043	5·189	31	0·0116	0·295	0·010	0·254	0·0089	0·226
5	0·212	5·385	0·220	5·588	0·1819	4·620	32	0·0108	0·274	0·009	0·229	0·0079	0·203
6	0·192	4·877	0·203	5·156	0·1620	4·115	33	0·0100	0·254	0·008	0·203	0·0071	0·180
7	0·176	4·470	0·180	4·572	0·1443	3·665	34	0·0092	0·234	0·007	0·178	0·0063	0·160
8	0·160	4·064	0·165	4·191	0·1285	3·264	35	0·0084	0·213	0·005	0·127	0·0056	0·142
9	0·144	3·658	0·148	3·759	0·1144	2·906	36	0·0076	0·193	0·004	0·102	0·0050	0·127
10	0·128	3·251	0·134	3·404	0·1019	2·588	37	0·0068	0·173	—	—	0·0045	0·114
11	0·116	2·946	0·120	3·048	0·0907	2·304	38	0·0060	0·152	—	—	0·0040	0·102
12	0·104	2·642	0·109	2·769	0·0808	2·052	39	0·0052	0·132	—	—	0·0035	0·090
13	0·092	2·337	0·095	2·413	0·0720	1·829	40	0·0048	0·122	—	—	0·0031	0·079
14	0·080	2·032	0·083	2·108	0·0641	1·628	41	0·0044	0·112	—	—	0·0028	0·071
15	0·072	1·829	0·072	1·829	0·0571	1·450	42	0·0040	0·102	—	—	0·0025	0·063
16	0·064	1·626	0·065	1·651	0·0508	1·290	43	0·0036	0·091	—	—	0·0022	0·056
17	0·056	1·422	0·058	1·473	0·0453	1·151	44	0·0032	0·081	—	—	0·0020	0·051
18	0·048	1·219	0·049	1·245	0·0403	1·024	45	0·0028	0·071	—	—	0·0018	0·046
19	0·040	1·016	0·042	1·067	0·0359	0·912	46	0·0024	0·061	—	—	0·00157	0·040
20	0·036	0·914	0·035	0·889	0·0320	0·813	47	0·0020	0·051	—	—	0·00140	0·036
21	0·032	0·813	0·032	0·813	0·0285	0·724	48	0·0016	0·041	—	—	0·00124	0·031
22	0·028	0·711	0·028	0·711	0·0253	0·643	49	0·0012	0·030	—	—	0·00099	0·025
23	0·024	0·610	0·025	0·635	0·0226	0·574	50	0·0010	0·025	—	—	0·00088	0·022
												—	—

Table 14.—Wire Abbreviations

The following abbreviations are the recognized trade descriptions and should therefore be used when ordering or specifying :

S.C.C.	Single Cotton Covered
D.C.C.	Double Cotton Covered
T.C.C.	Triple Cotton Covered
Lam.	Laminated
S.W.S.	Single White Silk
D.W.S.	Double White Silk
S.S.C.	Single Silk Covered
D.S.C.	Double Silk Covered
Enam.	Enamelled
Enam. & S.S.C.	Enamelled and Single Silk Covered
Enam. & D.S.C.	Enamelled and Double Silk Covered
Enam. & S.C.C.	Enamelled and Single Cotton Covered
Enam. & D.C.C.	Enamelled and Double Cotton Covered
S.P.C.	Single Paper Covered
D.P.C.	Double Paper Covered
T.P.C.	Triple Paper Covered
Standard	Standard Covering
Fine	Fine Covering
B/D or Brd.	Braided
Compd. strand	Compressed strand
H.D.	Hard Drawn
S.D.	Soft Drawn
H.C.	High Conductivity
Pl. cu.	Plain Copper
T/d. cu.	Tinned Copper
S.I.R. or S.P.R.	Single lapping of Pure Rubber
D.I.R. or D.P.R.	Double lapping of Pure Rubber
Pfd.	Paraffined
S.W.G.	Standard Wire Gauge
B.W.G.	Birmingham Wire Gauge
B. & S.	Brown & Sharp's Gauge
V.C. tape	Varnished cambric tape (also known as "Empire" or "Lino" tape)

Litz Wires

The term "Litz Wire" is derived from the German "Litzendraht" (*litze*, strand; *draht*, wire), and is generally understood to apply to conductors which are built up by successive stranding of wires or groups of wires in groups of three, each individual wire being insulated with enamel or silk.

Litz wires were developed for use in high-frequency work, in which "skin effect" (the tendency of the current to flow along the surface of the conductor only) is encountered. This tendency causes the resistance of a conductor at high frequencies to be much greater than its normal D.C. resistance. In order to reduce the effect the required cross-section of conductor is obtained by using a large number of small wires, each completely insulated from all the rest, twisted together in such a manner that throughout the whole length of the multiple conductor an individual wire occupies in turn all possible positions in the winding section. The necessary interweaving of the wires is usually brought about by constructing the stranded conductor on the " three " system, in which three wires are first twisted together, then three of the three-wire strands are twisted to form a nine-wire strand, and so on, the strands containing 3, 9, 27, 81, etc., wires. A slight variation of this construction is sometimes used, in which the number of wires in the original strands is other than three, but the further stranding conforms to the " three " system.

The accompanying tables give details of some of the standard sizes of Litz Wires, but the range can, of course, be extended considerably by using larger wires and by multiplying the stranding process to give 243, 729, etc., wires. If a copper area slightly different from that of the standards is required, this can be achieved by altering the number of wires in the first groups, e.g., 3/3/4 wires can be used instead of 3/3/3. The length of lay can also be adjusted, if necessary, to suit particular requirements. In the case of heavy Litz strands, it is sometimes the practice to compress the finished conductor into a rectangular section convenient for the particular application.

It is usual to apply an overall insulation of silk, glace-cotton or other textile covering to the stranded conductor, or alternatively to treat it with a coating of wax or varnish which bonds the individual wires sufficiently to avoid displacement in wiring.

Attention is drawn to the fact that B.S. 1258 : 1946 for " Textile-covered bunched enamelled-copper wire conductors " does not cover Litz wire (see Clause 1—" Scope ").

TABLE 15.—LITZ WIRES. LENGTH OF LAY (INCHES)

Diameter of Bare Wire		*Length of Lay (Inches)*								
S.W.G.	*Inch*	3/3 = 9 *Wire Conductors*		3/3/3 = 27 *Wire Conductors*			3/3/3/3 = 81 *Wire Conductors*			
48	0·0016	1/4	1/2	1/4	1/2	1/2	1/4	1/2	1/2	3/4
47	0·0020	1/4	1/2	1/4	1/2	1/2	1/4	1/2	1/2	3/4
46	0·0024	1/4	1/2	1/4	1/2	1/2	1/4	1/2	1/2	3/4
45	0·0028	1/4	1/2	1/4	1/2	1/2	1/4	1/2	1/2	3/4
44	0·0032	5/16	5/8	5/16	5/8	5/8	5/16	5/8	5/8	1
43	0·0036	5/16	5/8	5/16	5/8	5/8	5/16	5/8	5/8	1
42	0·0040	5/16	5/8	5/16	5/8	5/8	5/16	5/8	5/8	1
41	0·0044	3/8	3/4	3/8	3/4	3/4	3/8	3/4	3/4	1 1/4
40	0·0048	3/8	3/4	3/8	3/4	3/4	3/8	3/4	3/4	1 1/4
39	0·0052	3/8	3/4	3/8	3/4	3/4	3/8	3/4	3/4	1 1/4
38	0·0060	9/16	1	9/16	1	1	9/16	1	1	2
37	0·0068	9/16	1	9/16	1	1	9/16	1	1	2
36	0·0076	9/16	1	9/16	1	1	9/16	1	1	2
35	0·0084	5/8	1 1/4	5/8	1 1/4	1 1/4	5/8	1 1/4	1 1/4	2 1/4
34	0·0092	5/8	1 1/4	5/8	1 1/4	1 1/4	5/8	1 1/4	1 1/4	2 1/4
33	0·0100	5/8	1 1/4	5/8	1 1/4	1 1/4	5/8	1 1/4	1 1/4	2 1/4

TABLE 16.—LITZ WIRES. ENAMELLED WIRES, STRANDED AND D.S.C. OVERALL.

Diameter of Single Wires (Bare)		Diameter of Single Enamelled Wires (Inch)	Approximate Overall Diameter (Inch)			
Inch	S.W.G.		3 Wires 3	9 Wires 3/3	27 Wires 3/3/3	81 Wires 3/3/3/3
0·0016	48	0·0020	0·0062	0·0103	0·0178	0·0330
0·0020	47	0·0025	0·0075	0·0123	0·0222	0·0407
0·0024	46	0·0029	0·0083	0·0139	0·0253	0·0467
0·0028	45	0·0034	0·0093	0·0158	0·0291	0·0552
0·0032	44	0·0038	0·0101	0·0174	0·0322	0·0614
0·0036	43	0·0043	0·0111	0·0193	0·0360	0·0686
0·0040	42	0·0047	0·0119	0·0209	0·0390	0·0756
0·0044	41	0·0052	0·0129	0·0234	0·0430	0·0834
0·0048	40	0·0056	0·0137	0·0250	0·0461	0·0895
0·0052	39	0·0060	0·0145	0·0265	0·0491	0·0954
0·0060	38	0·0069	0·0163	0·0300	0·0570	0·1090
0·0068	37	0·0078	0·0181	0·0336	0·0639	0·1223
0·0076	36	0·0086	0·0197	0·0367	0·0711	0·1342
0·0084	35	0·0095	0·0220	0·0402	0·0780	0·1480
0·0092	34	0·0104	0·0238	0·0438	0·0850	0·1625
0·0100	33	0·0113	0·0256	0·0473	0·0918	0·1750

In order to avoid any possibility of confusion when ordering Litz wire, the method of stranding should be given, not only the number and size of wire: e.g., 81/0·0024 in. should be ordered as 3/3/3/3/0·0024 in.

The direction of twisting is usually as follows:

First process	Left hand
Second process . . .	Right hand
Third process	Right hand
Fourth process . . .	Left hand
Further processes . . .	Alternate directions

Self-bonding Wires

Self-bonding wires have recently been developed in order to meet the demand for a wire capable of being used to make rigid, self-supporting coils without formers or taping. An example of this type of wire is "Lewmexbond", which consists of standard "Lewmex" wire with an additional coating of thermo-bonding material. The increase in diameter due to the additional covering is usually of the order of from ½ to 1 mil, according to the size of the wire. The other conductor sizes to which this treatment is applied are limited to the range of 0·030 in. and finer.

Self-supporting coils are produced by heating the wire whilst it is on the former, and removing the coil when the wire has cooled. The bonding occurs through the welding of the outer coatings of adjacent turns of the wire. Any form of heating may be employed, but the best method is to pass a current through the coil sufficient to raise it momentarily to 130–150° C. The heat dissipation throughout the coil is almost uniform.

Such coils can be used for many purposes, including coils of matched inductance, such as those used as deflection coils for television, small frame aerials and tuning coils.

RARE AND PRECIOUS METALS

Fine silver, standard silver, iridium–platinum, palladium–silver, platinum–silver-gold, silver on copper bi-metal are largely used for metal contacts in scientific and precision instruments. Iridium, platinum and tungsten contacts find their chief applications in magnetos and coil-ignition equipment. Iridium–platinum and rhodium–platinum are utilized for the electrodes of cathode-ray tubes. Cæsium salts are employed chiefly in the manufacture of photo-electric cells. Platinum and rhodium–platinum provide two excellent materials for forming thermocouples, particularly where stability of operation is the prime consideration.

The table gives the properties of the rare and precious metals which are used for contact materials.

TABLE 17.—RARE AND PRECIOUS METALS

Metal	*Density*	*Vickers Hardness (Annealed)*	*Resistivity (MΩ/cm. cube)*	*Melting Point (° C.)*
Silver	10·5	26	1·6	961
Platinum	21·4	65	11·6	1,766
Palladium	11·9	40	10·7	1,552
Gold	19·3	20	2·4	1,063
Iridium	22·4	220	5·3	2,454
10% Gold-silver	11·4	29	3·6	965
5% Palladium-silver	10·5	33	3·8	975
10% Palladium–silver	10·6	40	5·8	1,000
20% Palladium–silver	10·7	55	10·1	1,070
Standard silver (7½% Copper)	10·3	56	1·9	778
10% Copper–silver	10·3	62	2·0	778
20% Copper-silver	10·2	85	·2·1	778
50% Copper–silver	9·7	95	2·1	778
Cadmium–silver (Elkonium 17)	10·3	55	5·8	850
Cadmium–copper-silver (Elkonium 18)	10·1	65	4·2	800
10% Iridium-platinum	21·6	120	24·5	1,780
20% Iridium-platinum	21·7	200	30·0	1,815
25% Iridium-platinum	21·7	240	32·0	1,845
30% Iridium-platinum	21·8	285	32·3	1,885
10% Ruthenium-platinum	19·9	200	42·2	1,780
14% Ruthenium-platinum	19·5	240	45·8	1,800
Iridium–ruthenium-platinum (Irru)	20·8	310	39·0	1,890
Iridium–osmium-platinum	22·0	540	—	2,400
Molybdenum-platinum	20·5	195	58·5	1,700
40% Silver-palladium	11·0	95	35·8	1,290
40% Copper–palladium	10·4	145	35·0	1,200
30% Silver–gold	16·6	32	10·4	1,025
Platinum-silver-gold	17·1	60	16·8	1,100

SOLDERS AND FLUXES

The economy of the soldering operation and the quality of joints depend on the choice of correct grades of solder and flux and the most suitable method of soldering.

Soft solders are basically lead–tin alloys, with sometimes a small proportion of antimony, which strengthens the metal. British standard grades of solders are given in the following table.

TABLE 18.—STANDARD GRADES OF SOLDERS

Grade	*Tin* (%)	*Antimony* (%)	*Melting Temperature (Approx.)* ° *C.*	° *F.*	*Uses*
A K	64–65 59–60	0·6 * 0·5 *	183–185 183–188	361–365 361–370	Components liable to damage by heat or requiring free-running solder, e.g., electrical, radio and instrument assemblies, machine soldering of can end-seams.
B F M	49–50 49–50 44–45	2·5–3·0 0·5 * 2·3–2·7	185–204 183–212 185–215	365–399 361–414 365–419	Coppersmiths' and tinsmiths' bit soldering; general machine soldering (e.g., can end-seams).
C G	39–40 39–40	2·0–2·4 0·4 *	185–227 183–234	365–441 361–453	Blowpipe soldering; soldering of side seams on high-speed, body-forming machines.
L D J	31–32 29–30 29–30	1·7–1·8 1·0–1·7 0·3 *	185–243 185–248 183–255	365–469 365–478 361–491	Wiping of cable and lead pipe joints; dipping baths.

* These solders are classified as non-antimonial types, the percentage antimony figure being the maximum percentage of antimony as an impurity.

Soldering Fluxes

ACTIVE FLUXES. These contain chlorides, are rapid in action and very effective in cleaning the metal surface and promoting spread and penetration of the solder. They are suitable for use on all common metals, with the exception of aluminium and its alloys.

SAFETY FLUXES. Resin-base fluxes are not so rapid or effective, but have the advantage that they are completely non-corrosive, so that the joint need not be cleaned after soldering. Such fluxes should always be used in the soldering of electrical conductors.

CORED SOLDER. Solder wire with a continuous core of flux is useful for "spot soldering", particularly where the joint is not readily accessible.

SOLDER PAINT. This is a creamy mixture of powdered solder with an active flux. It is used for tinning and for sweat soldering; in the latter application the joint members are brushed over with the paint, assembled and then heated to soldering temperature by any convenient method.

Fusible Solders

For delicate work which might be damaged by heat when a normal solder is used, special fusible solders are available.

There is a series of alloys covering a range of melting points down to 70° C. (158° F.).

Soldering of Radio Components

For general repair and construction of radio and television assemblies, the following points should be observed:

(1) Only non-corrosive fluxes should be used. Cored solders with resin-based fluxes are generally convenient. For miniaturized assemblies—where the lengthy application of heat may have harmful effects—a liquid flux of resin-base permits joints to be made rapidly. Acid-based fluxes, such as killed spirits, should not be used.

(2) All wires, tags and solder should be clean and free from oxidation. Since the cleansing action of non-corrosive fluxes is less effective than with active fluxes, greater care is needed in pre-cleaning of wires and tags, and it is often advisable to tin these parts before assembly.

(3) The iron should be clean, adequately tinned and should be about 50° C. above the liquefying point of the solder. With cored solders, "spitting" indicates too high a temperature of the bit, while "plasticizing" indicates too low a temperature.

(4) When using cored solders, the iron should never be used to carry the solder to the joint: whenever possible apply the iron beneath the joint and then apply the solder from above.

(5) Fine-gauge enamelled wires should not be scraped with a knife or with emery cloth, as this may damage the wire. A suggested method of cleaning them is to heat such wires in the flame of a spirit lamp, plunge them into methylated spirits and finally to wipe them dry.

(6) Dry joints, which may introduce considerable resistance into joints, are formed by layers of undiffused resin; this may be caused by too brief an application of the soldering-iron or by the bit being at too low a temperature.

Soldering to Aluminium

Soldering of connections to aluminium and its alloys by normal methods is seldom effective. In recent years, however, equipment has been marketed specifically for this purpose using an iron whose bit is subjected to a supersonic frequency produced by magneto-striction. This supersonic frequency prevents oxidation by disturbing the fluxing metal. A tin/zinc-based solder is used.

TABLE 19.—THE ELEMENTS IN ORDER OF ATOMIC NUMBER

Period	Atomic Number	Name	Symbol	Atomic Weight
I	1	Hydrogen	H	1·008
	2	Helium	He	4·002
II	3	Lithium	Li	6·94
	4	Beryllium	Be	9·02
	5	Boron	B	10·82
	6	Carbon	C	12·00
	7	Nitrogen	N	14·008
	8	Oxygen	O	16·00
	9	Fluorine	F	19·00
	10	Neon	Ne	20·18
III	11	Sodium	Na	22·99
	12	Magnesium	Mg	24·32
	13	Aluminium	Al	26·97
	14	Silicon	Si	28·06
	15	Phosphorus	P	31·02
	16	Sulphur	S	32·06
	17	Chlorine	Cl	35·46
	18	Argon	A	39·94
IV	19	Potassium	K	39·09
	20	Calcium	Ca	40·08
	21	Scandium	Sc	45·10
	22	Titanium	Ti	47·90
	23	Vanadium	V	50·95
	24	Chromium	Cr	52·01
	25	Manganese	Mn	54·93
	26	Iron	Fe	55·84
	27	Cobalt	Co	58·94
	28	Nickel	Ni	58·69
	29	Copper	Cu	63·57
	30	Zinc	Zn	65·38
	31	Gallium	Ga	69·72
	32	Germanium	Ge	72·60
	33	Arsenic	As	74·91
	34	Selenium	Se	78·96
	35	Bromine	Br	79·91
	36	Krypton	Kr	83·7
V	37	Rubidium	Rb	85·44
	38	Strontium	Sr	87·63
	39	Yttrium	Y	88·92
	40	Zirconium	Zr	91·22
	41	Niobium	Nb	92·91
	42	Molybdenum	Mo	96·0
	43	Technetium	Tc	99
	44	Ruthenium	Ru	101·7
	45	Rhodium	Rh	102·9
	46	Palladium	Pd	106·7
	47	Silver	Ag	107·9
	48	Cadmium	Cd	112·4
V	49	Indium	In	114·8
	50	Tin	Sn	118·7
	51	Antimony	Sb	121·8
	52	Tellurium	Te	127·6
	53	Iodine	I	126·9
	54	Xenon	Xe	131·3
VI	55	Cæsium	Cs	132·9
	56	Barium	Ba	137·4
	57	Lanthanum	La	138·9
	58	Cerium	Ce	140·1
	59	Praseodymium	Pr	140·9
	60	Neodymium	Na	144·3
	61	Promethium	Pm	147
	62	Samarium	Sm	150·4
	63	Europium	Eu	152·0
	64	Gadolinium	Gd	157·3
	65	Terbium	Tb	159·2
	66	Dysprosium	Dy	162·5
	67	Holmium	Ho	163·5
	68	Erbium	Er	167·6
	69	Thulium	Tm	169·4
	70	Ytterbium	Yb	173·0
	71	Lutecium	Lu	175·0
	72	Hafnium	Hf	178·6
	73	Tantalum	Ta	181·4
	74	Tungsten	W	184·0
	75	Rhenium	Re	186·3
	76	Osmium	Os	191·5
	77	Iridium	Ir	193·1
	78	Platinum	Pt	195·2
	79	Gold	Au	197·2
	80	Mercury	Hg	200·6
	81	Thallium	Tl	204·4
	82	Lead	Pb	207·2
	83	Bismuth	Bi	209·9
	84	Polonium	Po	210
	85	Astatine	At	211
	86	Radon	Rn	222
VII	87	Francium	Fr	223
	88	Radium	Ra	226·0
	89	Actinium	Ac	227
	90	Thorium	Th	232·1
	91	Protoactinium	Pa	231
	92	Uranium	U	238·1
	93	Neptunium	Np	237
	94	Plutonium	Pu	239
	95	Americium	Am	241
	96	Curium	Cm	242
	97	Berkelium	Bk	243
	98	Californium	Cf	244

Table 20.—Mechanical and Physical Properties of Electrical Structural Materials

S = specific gravity.
w' = weight, lb./cu. ft.
w'' = weight, lb./cu. in.
f_1 = elastic stress limit, tons/sq. in.
f_2 = ultimate tensile strength, tons/sq. in.
E = Young's modulus, thousands of tons/sq. in.
f_t, f_c, f_s = working stresses, tensile, compressive, shear, tons/sq. in.
α = coefficient of linear expansion $\times 10^{-6}/°$ C.
c = specific heat (15–100° C.).
k = thermal conductivity, W/c.c./° C.
θ_m = melting point, ° C.

Material	S	w'	w''	f_1	f_2	E	f_t	f_c	f_s	α	c	k	θ_m
Aluminium :													
Cast	2·67	166	0·097	3–4	4–6	4·3	1·5–2	—	—	24	0·214	2·0	660
Annealed	2·7	169	0·098	3·5	5	—	2	—	—	24	0·214	2·0	660
Wrought	2·7	169	0·098	8–9	12	4·3	7	—	—	24	0·214	—	660
Brass (60 Cu, 40 Zn) :													
Cast	8·4	525	0·304	9	20	5·5	2–3	1·5	—	18·9	0·094	1·1	890
Cold-rolled	8·4	525	0·304	20	30	5·5	10–12	—	—	—	—	—	—
Bronze :													
Hard-drawn	8·89	555	0·32	—	45	6–7	22	—	—	18	—	—	950
Cadmium-copper :													
Hard-drawn, 93% cond.	8·9	555	0·32	20–30	30–36	8·5	16	—	—	16·6	0·095	—	1,080
Copper :													
Cast 99·6% Cu.	8·7	545	0·31	3–4	8–11	5	1·75	2·75	—	17	0·093	3·8	—
Annealed, 99·9% Cu	8·9	555	0·32	3·5	12–18	7	2·5	—	1·0	17	0·093	—	1,100
Hard-drawn	8·9	555	0·32	—	22–29	8	8–12	—	—	16·6	0·093	—	1,100
Copper-clad steel :													
30–40% conductivity	8·25	514	0·298	—	45–60	10	18–35	—	—	12·8	—	—	—
Duralumin (94 Al, 4 Cu, 0·5 Mg) :													
Annealed	2·79	173	0·10	16	25–29	4·7	8–9	8–9	—	22·6	—	1·3	650
Rolled	2·79	173	0·10	33	35	—	20–25	—	—	—	—	—	—

Iron:														
Wrought (electric)		7·85	490	0·284	14–16	16–20	11–13	5–6	6	5	12	0·115	0·58	1,600
Cast		7·8	486	0·281	—	8–15	6–7	2·5	10	1·5	10·5	0·119	0·67	1,375
Malleable cast		7·4	461	0·266	14	25	11	8–10	10	7	—	0·110	0·66	—
Nomag cast		7·2	450	0·26	—	9–14	—	2·5	10	1·5	11·9	0·13	—	1,200
Phosphor-bronze:														
Cast		8·7	542	0·314	10–11	16–18	5–6	4·5	4·5	—	17	—	0·67	900
Hard-drawn		8·7	542	0·314	—	30–70	7·5	20–45	—	—	—	—	—	900
Steel:														
Cast (0·3–0·5% C)		7·7	480	0·278	12–16	25–35	13·5	8–9	8–9	5–6	11–13	0·117	0·46	1,400
Armature (2% Si)		7·75	483	0·280	13	20	—	—	—	—	—	0·117	—	1,400
Transformer (2% Si)		7·7	480	0·278	14–18	22	—	—	—	—	—	0·117	—	1,400
Transformer (4% Si)		7·5	468	0·271	30	40	—	—	—	—	—	0·117	—	1,400
Galvanized wire		7·84	490	0·284	15–50	23–80	13	11–40	—	—	11–13	0·117	—	—
Resistance alloys *:														
80 Ni, 20 Cr	(*a*)	8·34	522	0·302	—	47·5	12·5	—	—	—	12·5	0·103	0·038	1,420
59 Ni, 16 Cr, 25 Fe	(*b*)	8·28	517	0·299	—	44·5	12·5	—	—	—	12·5	0·112	—	1,420
37 Ni, 18 Cr, 2 Si, Fe Bal.	(*c*)	8·53	532	0·308	—	47	9	—	—	—	14·2	—	—	1,350
45 Ni, 54 Cu	(*d*)	8·88	555	0·321	—	32·5	12	—	—	—	14·9	0·098	0·054	1,250
80 Cu, 20 Ni	(*e*)	8·97	558	0·324	—	23	—	—	—	—	—	0·097	—	1,175
62 Cu, 15 Ni, 22 Zn	(*f*)	8·95	557	0·323	—	30	—	—	—	—	18	0·095	0·25	—
84 Cu, 12 Mn, 4 Ni	(*g*)	8·5	528	0·306	—	60	—	—	—	—	18	—	0·215	—

* *Alloys:* (*a*) Brightray C and S (working temperature up to 1,150° C.); (*b*) Brightray B (950° C.); (*c*) Brightray F (certain types of furnace up to 1,100° C.); (*d*) Ferry (200° C.); (*e*) Cupro (350° C.); (*f*) German Silver, Platinoid; (*g*) Manganin (Working temperature, 450° C.).

Table 21.—Electrical Properties of Conductors

$\rho_{cm.}$ = resistivity, $\mu\Omega$-cm. at 20° C.
$\rho_{in.}$ = resistivity, $\mu\Omega$-in., at 20° C.
γ/γ_{Cu} = relative conductivity, % of standard annealed copper.
α = resistance-temperature coefficient per ° C. × 10^{-4} at 20° C.

Material	$\rho_{cm.}$	$\rho_{in.}$	γ/γ_{Cu}	α
International standard annealed copper	1·72	0·678	100	39·3
Aluminium and alloy :				
Aluminium, cast	2·6	1·02	66	39·0
Aluminium, hard-drawn	2·8	1·20	62	39·0
Duralumin	4·7	1·85	36·5	—
Copper and alloys :				
Copper, annealed	1·69–1·74	0·665–0·685	102–99	39·3
Copper, hard-drawn	1·74–1·81	0·685–0·712	99–95	38·1–39
Brass (60/40), cast	7·5	2·95	23	16
Brass (60/40), rolled	9·0	3·55	19	16
Bronze	3·6	1·41	48	16·5
Phosphor-bronze, cast	6–12	2·4–4·8	29–14	10
Cadmium-copper, hard-drawn, 93% cond.	1·85	0·727	93	—
Cadmium–copper, hard-drawn, 82% cond.	2·1	0·825	82	40
Copper-clad steel, hard-drawn, 40% cond.	4·31	1·70	40	37·5
Copper-clad steel, hard-drawn, 30% cond.	5·75	2·27	30	37·5
Iron and steel :				
Iron, wrought (electric)	10·7	4·23	16	55
Iron, cast, grey	70	27·5	2·45	—
Iron, cast, white	100	39·3	1·72	20
Iron, cast, malleable	29·5	11·6	5·85	—
Iron, cast, nomag	160	63	1·12	4·5
Steel, 0·1% carbon	20	7·9	8·6	42
Steel, 0·3–0·5% carbon	12–19	4·7–7·5	14·3–9	42
Steel armature, <2% Si	15–20	5·9–7·9	11·5–8·6	—
Steel armature, 2% Si	35	13·8	4·9	—
Steel armature, 4% Si	55	21·6	3·1	—
Steel wire, galvanized	13–14·5	5·1–5·7	13·3–11·8	44
Steel wire, galvanized, 45-ton	17	6·7	10	34
Steel wire, galvanized, 80-ton	21·5	8·5	8	34
Resistance alloys * :				
80 Ni, 20 Cr (*a*)	109	40·6	1·65	1·0
59 Ni, 16 Cr, 25 Fe (*b*)	110	41·8	1·62	2·0
37 Ni, 18 Cr, 2 Si, Fe bal. (*c*)	108	35·8	1·89	2·6
45 Ni, 54 Cu (*d*)	49	18·9	3·6	0·4
80 Cu, 20 Ni (*e*)	26	10·2	6·6	2·9
62 Cu, 15 Ni, 22 Zn (*f*)	34·4	13·5	5·0	2·5
84 Cu, 12 Mn, 4 Ni (*g*)	48	18·9	3·6	0
Other Conductors :				
Carbon, graphitic	4,600	1,810	0·037	−0·2
Carbon, arc-lamp	5,080	2,000	0·034	−5
Gold	2·36	0·93	73	30
Lead	22·0	8·65	7·8	40
Mercury	95·5	37·6	1·8	7
Molybdenum	5·7	2·24	30	40
Nickel	13·6	5·34	12·6	50
Platinum	11·7	4·61	14·7	39
Silver, annealed	1·58	0·625	109	40
Silver, hard-drawn	1·75	0·685	98·5	40
Tantalum	15·5	6·1	11·1	31
Tungsten	5·6	2·2	31	45
Zinc	6·2	2·43	28	40

* *Alloys :* (*a*) Brightray C and S for furnaces, fires, heater elements; (*b*) Brightray B, for irons, tubular heaters; (*c*) Brightray F for certain types of furnace element; (*d*) Ferry for control resistances; (*e*) Cupro; (*f*) German silver, Platinoid, for instruments; (*g*) Manganin, for instrument shunts and resistance standards.

PRINTED CIRCUITS AND PRINTED COMPONENTS

The idea of "printing" the complicated wiring of radio equipment was first proposed by John Sargrove in 1947. At that time it seemed that the new process would have widespread application, but for several years only sporadic use was made of the idea. Recently, however, many cases have arisen where the printing methods are regarded as fully acceptable not only from an economic point of view but also on account of their reliability and assistance in miniaturization. With the complex circuitry of modern electronic equipment every means possible is being sought to reduce labour costs, especially those associated with wiring and maintenance.

Two major trends are evident:

(*a*) The use of proprietary sub-assemblies.
(*b*) The use of printed wiring and components.

Proprietary sub-assemblies are growing in number. A typical example of this type of component is the multiple ceramic capacitor, which enables all decoupling capacitors having a common earth and associated with one particular valve-holder to be embodied in a single unit; their use not only reduces cost but also provides a greater degree of control over "wiring-variables" often so troublesome in V.H.F. work. Another component of this type is a complete mains-interference filter.

The use of printed circuits and printed components now appears in many pieces of equipment such as hearing aids, computer circuits, amplifier strips, television deflection coils, transceivers, guided missiles, etc., where much of the circuitry is complete, apart from the simple addition of the valves.

Advantages and Disadvantages of Printed Circuitry

Printed circuits start off with the advantage that excellent conductors (i.e., silver or copper) can be used; further, such circuits are generally printed on a base-material of very good insulating value—high-grade laminated sheet to R.C.S. 1000 Grade 1 specification being a most suitable plastics base. Reliability is, however, perhaps one of the main advantages, in so far as the many hand-made joints of conventional constructional techniques are eliminated, each a potential "dry-joint" with its attendant troubles. The chances of a human error are reduced, and less inspection is necessary.

Other advantages are: mass-production methods can easily be introduced with the minimum of complicated tooling; the accuracy of printed components is high, being of the order of 1 per cent and the Q of circuits are also satisfactory; little screening is necessary owing to the close coupling of components, and connections, being short, ensure a high efficiency; reduction in size also encourages economy and promotes the trend towards miniaturization, so necessary in many items, such as deaf-aid equipment or in guided-missile work.

"Printing" is not without its disadvantages, for instance an advanced production unit can involve a heavy capital outlay, and it is difficult to change design at short notice.

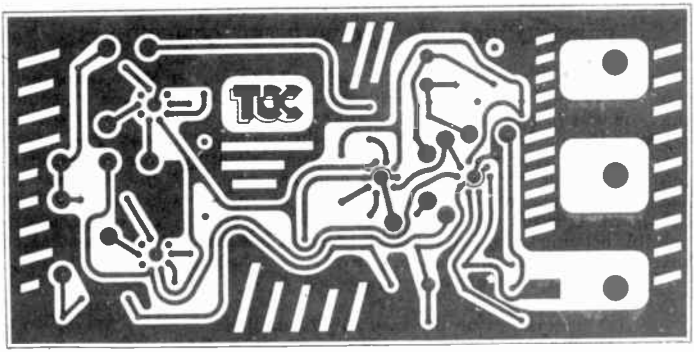

Fig. 8.—Stencil for a Printed Circuit.

Printed-circuit Methods

Several methods are available for "printing" circuits and components, the chief of which are :

The "Stencil" or "Masking" Technique, in which a stencil of the required outline is first prepared, by photographic reduction from a master drawing or by the use in simple cases of masking tape. The conducting material is then applied by one of a number of possible methods, e.g., by brush painting, by a dusting technique, by spraying, by chemical deposition or by cathodic sputtering in a high vacuum.

The Printing Technique, in which silver ink is printed on to a glass or ceramic base and then fired. This form gives particularly good adhesion, as required for rotary switches.

Photo-etching, in which an image of the circuit required is printed on to a copper-bonded laminate-base material (this is described in more detail later).

Die-stamping Process, in which an inlay of the correct shape is stamped out and applied to the ends of all the components and then automatically soldered in one dip operation or, alternatively, bonded to a sheet of insulating material.

Preparation of Master Drawings

An enlarged (two or three times) reproduction of the wiring pattern required is first prepared, the conductors being drawn in black on a dimensionally constant white paper mounted on Bristol Board ; any alterations are made in white ink—erasing and fold marks being avoided. Centre holes are shown by circles, and the line width will depend on the current to be carried—say $\frac{1}{8}$–$\frac{1}{4}$ in. for heavy current and from $\frac{1}{16}$ to $\frac{1}{8}$ in. only, for average work. The line spacing is made not less than $\frac{1}{32}$–$\frac{1}{16}$ in. to avoid difficulty when dip-soldering.

One body of opinion is that large areas of metal should preferably have their overall area reduced by the "picket-fence" method or by cross-hatching in such a manner that the lines in the newly broken-up areas have similar width to the circuit—this ensures uniform heating and consequent even coating with solder. Other manufacturers prefer to leave intact as much of the foil as possible and to leave the minimum thickness of insulation between conductors. The advantages of this are that less costly etching away is required, wide conductors of high current-carrying capacity and good heat dissipation are formed, and good screening results; furthermore, the presence of foil enables capacitors to be formed easily. The foil also presents a non-hygroscopic surface, eliminating fluctuations in electrical values with atmosphere changes. Where possible double-clad foil is preferable, as it is much stronger and warping is considerably reduced.

Photographic reduction from large dimensions enables accurate spirals for inductors to be produced with very small turn-width, so that, due to the physical characteristics of the flat spiral, its self-capacitance is almost negligible.

Photo-etching Process

This process is frequently used. The raw materials are the so-called copper-clad laminates, which are roughly of two types: (*a*) the simple insulation-backed metal-foil type; this may be *single-clad* or *double-clad*, depending on whether it is intended to print a circuit on one side only or on both sides of the card; (*b*) alternatively a multi-layer material may be used. For complicated circuit work, such multi-layer materials may comprise, for example, a copper film/nickel film/insulator or a copper film/carbon film/insulator.

The first stage in the photo-etching process is to print an acid-resisting image of the circuit on to the copper, so that the area of foil covered by the image will be protected against attack when immersed in acid, whereas the areas of foil not protected will be removed by the acid. This may be done in at least four different ways:

The Ordinary Printer's Procedure.—With this method the usual "block" is made from the master drawing of the required circuit. This is then used in the ordinary printing machine but using acid-proof ink and its impression imparted to the copper-clad laminate. The method is particularly applicable to long runs of mass-production work.

The Photo-mechanical Method.—In this method the copper laminate is uniformally coated with a material sensitive to ultra-violet light. It is then exposed through a negative of the required circuit and developed. The action of the ultra-violet light is to harden the coating into an insoluble film, whereas the unexposed parts are soluble in the developer. The method gives great accuracy and is particularly applicable to intricate work. The film may be heat toughened if required for further acid resistance.

Either chemical or electrolytic etching can be used to remove the foil not protected by the film depicting the required printed circuit. In the case of copper, ferric chloride is used as the mordant, and continuous agitation and close time and temperature control are necessary for consistent results. Washing in hot and cold water, followed by drying and removal of the photo-resist image, completes the operation.

"*Fusible-foil De-printing Method.*"—This is a method of limited application in which a low-melting-point foil of zinc, lead or tin on paper or similar support is used. The "block" of the circuit has its surface tinned, the foil is heated in the printing machine and, when the impression is made, the two metallic parts unite and the fused metal is lifted away, and subsequently removed by a "wiper". The main advantage of the process is that it is rapid and cheap.

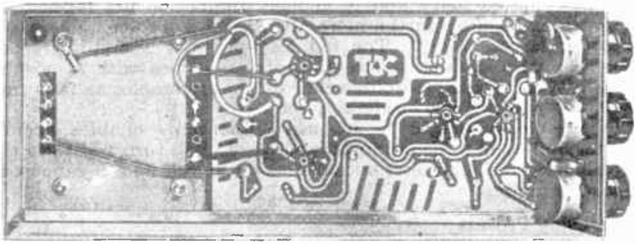

FIG. 9.—A TYPICAL PRINTED CIRCUIT, PREPARED WITH THE STENCIL SHOWN IN FIG. 8.

Stamping or Mechanical Printing.—In this system several variations are possible; but in general it may be said that the object is to first semi-punch-out the foil circuit, leaving temporary bridge pieces for support. This is then bonded to its insulating support, and the bridge pieces and the accompanying insulation are punched away. In one variation a Veé tool presses the " wanted " foil into the insulator, thereby, at one and the same time, effecting the bonding operation; the remainder is easily peeled away. Another variation removes the unwanted foil by " abrasion ", and this method is particularly applicable to " carbon-resistance materials ", which are unaffected by etching.

The Rating of Printed Wiring

Printed circuits can be used at both " ends " of a radio or television receiver, namely, for the very small currents associated with the radio-frequency input stages or, equally well, for the audio-frequency output and even for the " mains " connections. With the former, the " wiring " is frequently very much associated with the printed components themselves, but at the audio-frequency end the output transformers and other components are usually quite large and require mounting on a metal chassis. In this case the printed wiring may conveniently be carried out on a sub-panel which is slipped over the tags of the large components, the connections being completed by hand soldering. A cross connection becomes virtually impossiblé, and semi-skilled workers may be engaged on such assembly work.

The copper used in printed circuitry has a large surface area in comparison to its thickness and, being clean and bright, has a high radiation efficiency, so that it can safely carry comparatively heavy currents. The exact carrying capacity or temperature rise depends on ventilation, proximity of other components and other factors. As a guide, and for a copper thickness of $\frac{15}{1000}$ in. a $\frac{1}{16}$-in.-wide line could carry 2 amps. for a 5° C. rise or 4 amps. for 20° C. rise, while a $\frac{1}{4}$-in. line could deal with 6 amps. and 10 amps. respectively. For silver-plated circuits an additional 20 per cent could be expected. The high heat dissipation of foil resistors can largely overcome the usual cooling difficulties associated with heavy-duty resistors, and by suitable bonding much of the heat can be conducted to the chassis.

Eddy Current Loss.—Silvered-copper flexible-foil conductors can be twisted into such a shape that their width can lie in the direction of any magnetic field present, i.e., when carrying radio-frequency currents the skin effect causes it to flow in the wide silver surface, so reducing eddy current losses, and making the new medium comparable with litz-wire. An embossing operation on an experimental coil causes the conductors to stand up, and this raises the Q factor.

The *insulation resistance of printed wiring*, like that of all other wiring, depends on three main factors:

(*a*) the quality of the base or supporting material;
(*b*) the conditions of the exposed surface, i.e., its cleanliness, temperature and humidity; and,
(*c*) the length of the printed circuit and the distance apart of the component parts; thus two printed conductors 6 in. long and separated by $\frac{1}{16}$ in., and printed on material of surface resistivity of 5×10^4 MΩ/sq. cm., would have an insulation resistance of 50,000 MΩ.

Protective Coatings.—The variable factor in the insulation resistance of printed circuits is the surface condition, and three means may be used to combat ill-effects:

(*a*) the ink or photo-resist where used during manufacture may be left on except at junction points;
(*b*) a protective material may be applied: or
(*c*) the circuit may be folded and embedded in "casting-resin", in which case most of the difficulties disappear.

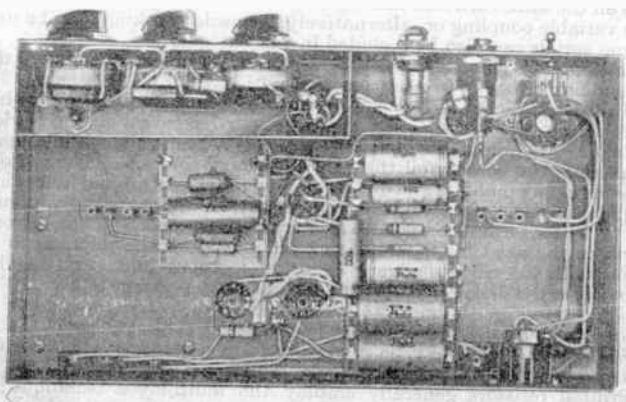

Fig. 10.—The Unit Shown in Fig. 9 Before Printed-circuit Techniques were Introduced.

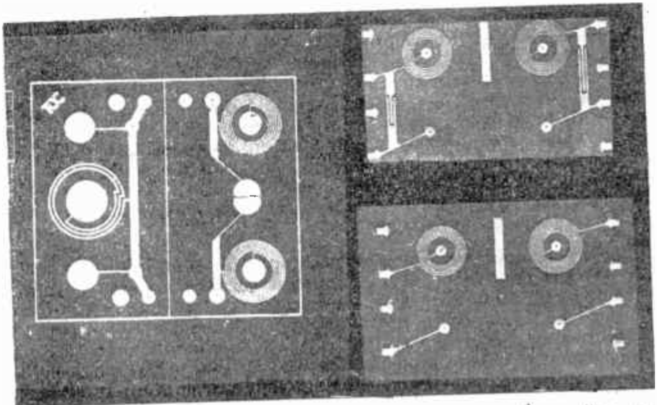

FIG. 11.—PRINTED COMPONENTS—NOTE PRINTED CAPACITOR ON RIGHT-HAND SIDE.

Printed Components

In addition to printing the wiring, it is also possible to "print" a number of components.

Inductors.—Inductors of low inductance value of the flat spiral type, i.e., those for television frequencies and for their corresponding intermediate frequencies, lend themselves to the printing technique. Variable inductors are also easily constructed either by printing the two coils on the same card and then folding it and varying the angle of vee to give variable coupling or, alternatively, a powder-backing may be used. Frame aerials can also be included in this category.

Capacitors.—These also lend themselves to printing, indeed they may be made by either of two processes: multi-layer or photo-etching.

Multi-layer technique capacitors start by coating a thin foil on both sides with a film of dielectric, the thickness depending on the working voltage. Any pinholes which may be present are ingeniously eliminated by passing the coated foil through an etching solution. The mordant penetrates any pinholes and dissolves away any metal around it, leaving a perfect dielectric after washing. Metal is then deposited or evaporated on to it to form the second electrode. The photo-etching process obviates the need for a sandwich technique. All the metal is in one plane with an edge-to-edge gap, often tortuous and curved, which constitutes the dielectric. With air as dielectric, a range of about 0–100 pF can be obtained, but by using a special dielectric this can be extended to about 0–3,000 pF.

Resistors.—These may be made by at least two methods: printed resistor or metal-foil resistors.

Printed resistors generally employ the multi-layer technique and consist of a layer of resistive material between an insulating layer and a highly conductive layer. A typical mode of manufacture is to coat a foil with varnish containing carbon particles and a resin binder, which is then stoved and cured. An insulator is bonded or coated on to the carbon resin film and the resistance per unit area checked; the metal foil is printed with the pattern of all the conductive and resistive areas, and then etched. The mordant exposes the areas not covered by ink, and presents the resistance material, which, together with the remaining ink, is removed by abrasion. The sheet is then over-printed, and again etched to remove metal where resistance material only is required, after which it is washed and coated with a protective phenolic material. The value of the resistor is determined by altering the composition of the coating and also its "aspect-ratio", i.e., its length to breadth dimensions. Alternatively, the resistance area may be formed by a shot-blasting process. The area required depends on the wattage rating and on the thermal conductivity of the supporting material; typical values are ½ watt/sq. cm. for laminated plastics, and 1 watt/sq. cm. for ceramics.

Metal-foil resistors are also used, especially for low-resistance applications such as strain gauges; in this type a continuous grid is either cut out or etched out, leaving a long, tortuous path of the right resistive value.

Foil Fuses.—Fuses can be made of values ranging from a few micro-amperes to several hundred amperes to quite close tolerances, and for either instantaneous or delayed action; they may be made self-indicating if required. Their fusing behaviour is controlled by shaping the conductor down to a few thousandths of an inch wide and by bonding it or by leaving it free, the effect of surface tension is most marked.

Potentiometers.—These may be produced by printing an accurate row of very closely spaced metal contacts and superimposing on them a resistance track by a photographic technique.

Switches.—Rotary switches can be printed easily using a suitable metal foil patterned to suit. Switches for automatic telephony have also been produced.

Valve-holders.—The conventional miniature valve-holder can be eliminated by using a flexible printed circuit whose conductor is made to contact directly with the valve pins.

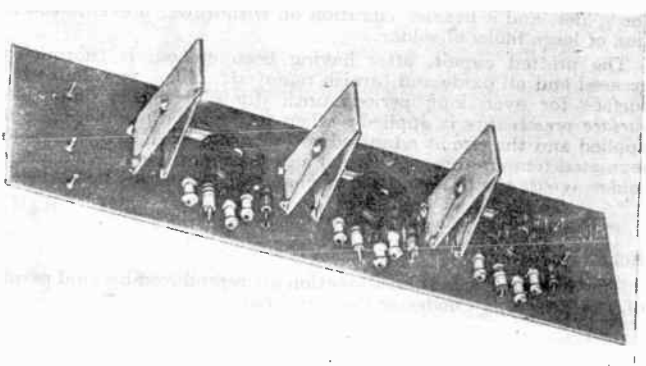

FIG. 12.—AN I.F. STRIP USING PRINTED CIRCUITRY AND PRINTED COILS.

Semi-conductors.—These can be incorporated in a similar manner to the process for carbon resistors, by forming a semi-conducting layer on the metal foil instead of the carbon layer; this considerably extends the field of application, and many interesting cases are now being developed.

Folding of Printed Circuits

With the continued trend towards miniaturization of equipment and the "printing" of components, it is not surprising to find that printed circuits are now appearing in folded form. The great advantage is that the circuit can be printed on a flexible flat film or on paper, and then its "space-factor" improved by bending or folding it into a three-dimensional form. It is found in a number of forms, i.e., as a "box" form when the angle between the films is 90°, in the "superimposed" type when the angle is 180° and as a "concertina" or "wound" version when the folding is repeated a number of times. Another reason for folding is to obtain a simple method of tuning, for example, of intermediate-frequency transformers.

Dip-soldering of Printed Circuits

One of the many advantages of the printed circuit is the ease of manufacture; another great advantage is the facility with which connections to valve-pins, other components and, of course, inter-connections between wafers can be made. By proper planning consistently good joints can be obtained automatically and all chance of mis-connection be eliminated.

The method of dip-soldering may vary from a simple hand-dip to a more automatic arrangement. A typical arrangement for the automatic dip-soldering of components consists essentially of a small trolley running on rails or guides fitted to the sides of the solder bath, and so bent that the printed-circuit panel is lowered into the solder and removed in a standard time. It is preferable that one side enter first at an angle of say 30°. The motion should then continue horizontally to ensure that no air bubbles are entrapped, and also that any solvent gas is released. A light vibration during the solder-dip assists in uniformally spreading the solder, and a heavier vibration on withdrawal prevents the formation of large blobs of solder.

The printed circuit, after having been etched, is thoroughly degreased and all oxide and tarnish removed. Then, to retain the clean surface for even long periods until dip-soldering is undertaken, a *surface preservative* is applied; when ready for dip-soldering a flux is applied and the circuit admitted to the solder bath, which is held at a regulated temperature. A film of anti-oxidant oil on the surface of the solder avoids the necessity for mechanical clearing of the surface.

R. H. G.

Acknowledgement

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4. STUDIOS AND STUDIO EQUIPMENT

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BROADCASTING STUDIOS

Broadcasting studios range in size from those suitable to accommodate one person delivering a talk up to concert halls capable of holding a symphony orchestra, chorus, organ and audience. Between these two extremes, the space and acoustic environment of small orchestras, discussion groups and dramatic personæ have to be met. The opportunity to build new premises occurs so rarely that when new studios are built by any broadcasting organization, interest in their construction is widespread. Before discussing the variations in acoustical quality and in equipment needed for these diverse needs, it is probably worth while considering the common requirements for all studios, that is, good sound insulation against external noise.

External Noise Sources

The most obvious source of interference, street traffic noise, is likely to be kept well in mind, but other sources of noise, such as low-flying aircraft, underground-train services or ships or works sirens, may have to be taken into account. Nor must it be forgotten that an organ or similar recital in one studio may represent a source of interference to an adjacent studio.

Sound Insulation

The particular need for good sound insulation between studios and between the studios and the outside world is extremely important, as any failing in this direction is almost impossible to correct. A quiet site on which to build, situated near the centre of a city for the convenience of artists and broadcasters and for access to the necessary communication cables, represents an ideal condition but one seldom found.

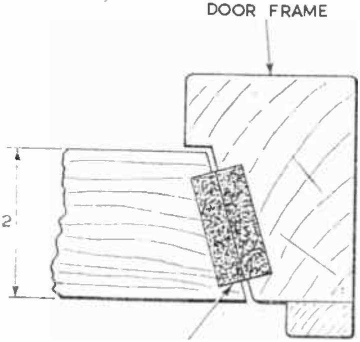
The first approach to the consideration of sound insulation is in the planning of a studio centre as a whole. It is usual to arrange as far as is practicable that groups of studios are ringed by corridors, beyond which a shell of rooms used as offices, stores, libraries, etc., provide a relatively effective sound barrier, shielding the studios from external noise.[1,2]

Rooms on floor levels, both above and below studios, should also be reserved for quiet activities, so providing a sound barrier to floors two storeys away which may also be used for studios.

Electrical generating plant, ventilating plant and maintenance workshops should also be planned to be remote from studios.

Sound Leakage

Given basically adequate sound insulation, care must be taken not to reduce this by sound leakage into the studio, which can be both air-



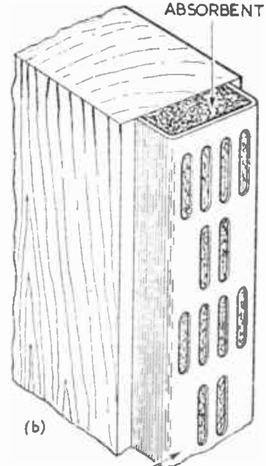


FIG. 1.—MEANS OF OBTAINING ACOUSTICALLY SEALED DOORWAYS.

borne and structure-borne. For quietness, the ventilation system should displace a large volume of air at a slow speed. The ventilating ducts must be acoustically baffled and lined with sound absorbents,[3] whilst the joints between sections of ducts must be acoustically isolated to prevent noise from the ventilating plant being transmitted to the studio. Similarly, cable ducts should be interrupted, and the unused space in the ducts should be stoppered with a plug of absorbent material wherever they pass through studio walls.

Studio doors should be of solid construction and be acoustically sealed when closed. An automatic door closer is also necessary to prevent the door being slammed or left ajar. Fig. 1 (*a*) shows one type

TABLE 1.—VARIATION IN SOUND INSULATION OF DOUBLE ¼ IN. PLATE-GLASS WINDOWS WITH VARIATION IN SPACING

(*Due to P. E. Sabine.*)

Spacing	*Sound Reduction in Decibels*
Sashes in contact . .	33·2
1½ in. separation . . .	38·6
4½ in. separation . . .	40·1
7½ in. separation . . .	44·2
9½ in. separation . . .	46·3
13½ in. separation . .	48·2
16 in. separation . . .	48·8

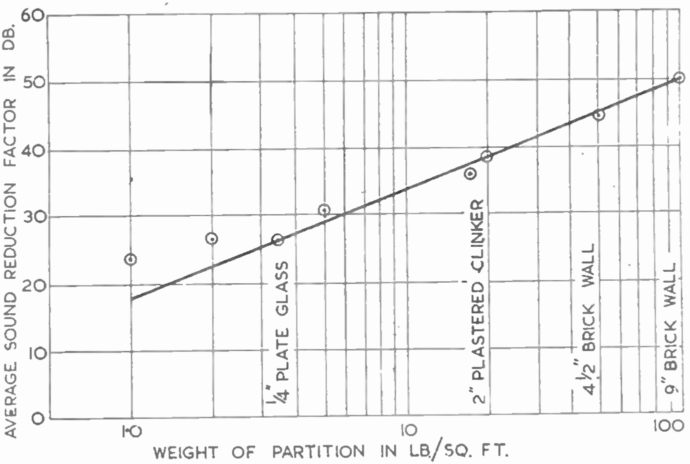


FIG. 2.—AVERAGE TRANSMISSION LOSSES OF BUILDING MATERIALS.

of door seal: another, more recent, type which had its origin in Scandinavia consists of perforated slotted metal over acoustic absorbent, and is shown in Fig. 1 (*b*).

Windows in studio walls are usually constructed of two or three sheets of plate glass spaced apart, and Table 1, due to P. E. Sabine, shows the improvement in insulation due to different air spacings between double plate-glass windows. More recently G. H. Aston of the National Physical Laboratory has published more comprehensive information regarding the sound-insulation properties of glass windows, and shows not only the variation in sound reduction with variation in spacing between double windows, but also the sound reduction variation with frequency and the effect of different frames and methods of mounting.[4]

The transmission loss in decibels due to a homogeneous partition is given by:

$$\text{Transmission loss } (TL) = 10 \log_{10} \frac{I_1}{I_2} \text{ db}$$

where I_1 is the intensity of sound incident to the partition, and I_2 is the intensity of sound transmitted by the partition.

Fig. 2 shows typical figures for the average transmission loss of a few building materials, plotted against the weight of the material in lb./sq. ft.

The transmission loss does not take into account the acoustic properties of the rooms on either side of the partition, and the total sound reduction is given by:

$$\text{Sound reduction} = \text{Transmission loss in db} + 10 \log_{10} \frac{aS}{A}$$

where aS is the total absorption in the receiving room and A is the area of the partition.

As a studio wall is seldom a simple homogeneous partition, published figures of sound insulation for such partitions should be accepted only as a rough guide. To prevent sound from a loudspeaker in a listening room being acoustically fed back to its associated microphone in a studio, thus causing a "howl round", a minimum sound reduction of some 30 db is necessary at low frequencies. When this sound insulation is obtained at low frequencies, it will generally be found that sound insulation rises with increase of frequency to the satisfactory figure of some 50 db at 1,000 c/s.

Transmission Quality

The essential considerations in the design of studios are the shape, the volume, the average reverberation time and the reverberation time/frequency characteristic. The problem of studio design differs from the design of auditoria in that the shape of a studio need not be unduly influenced by considerations of seating an audience. Although studios accommodating an audience are quite common, particularly for variety shows, the acoustical conditions required for the broadcast are of primary concern.

Shape and Size

The shape of the studios can well follow traditional design in being rectangular both in plan and elevation. Experiments have been made with studios having non-parallel, non-plumb walls, and with stepped and sloping ceilings, but there appears to be no strong evidence yet to suggest that their acoustics are superior to the studios of simpler form, which are easier and cheaper to construct. The proportions of height : width : length are important in obtaining even distribution of the characteristic resonance frequencies of a studio, and although not critical, a ratio of 2 : 3 : 5 provides a useful basis for design.

Little has been published regarding the optimum volume of studios for a given number of performers, but a popular misconception that a floor area adequate to accommodate the performers would suffice was dispelled by H. L. Kirke and A. B. Howe,[5] and their recommendations are shown in Fig. 3. For music studios the area covered by the performers should not be more than a third of the total floor area.

Reverberation

In the absence of specific faults it can be said that to a first approximation the average reverberation time and reverberation-time/frequency-characteristics, more than any other factor, determine the quality of sound transmission of a studio.

There appears to be a good deal of agreement upon the optimum value of reverberation times for studios, and a review of a number of halls and studios shows that their reverberation times fall within well-defined limits, which are indicated by the shaded areas in Fig. 4.

The higher limit would be selected when "fullness" of orchestral tone was a main consideration : the lower limit would be preferable for speech, songs, chamber music and dance music. As, however, many studios have to be used for diverse programmes, the choice of reverbera-

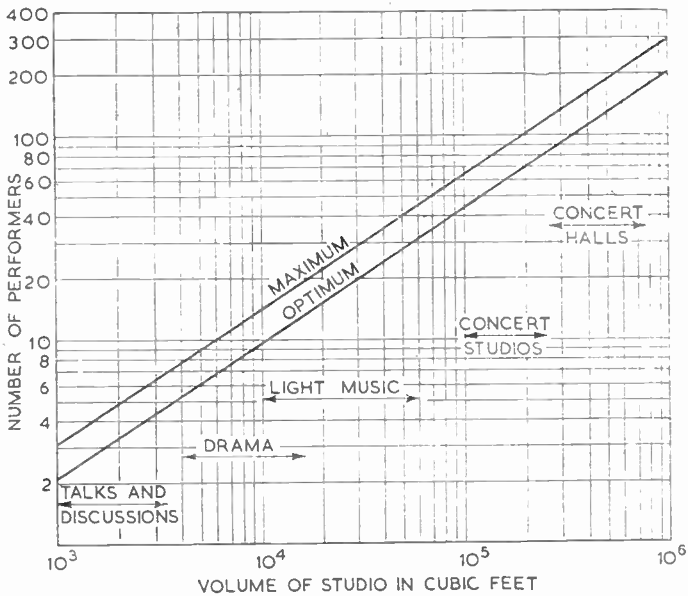


Fig. 3.—Suitable Floor Areas for Various Numbers of Performers.

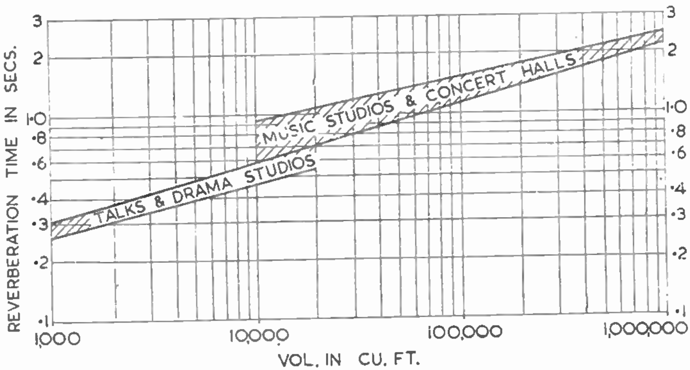


Fig. 4.—Reverberation Times.

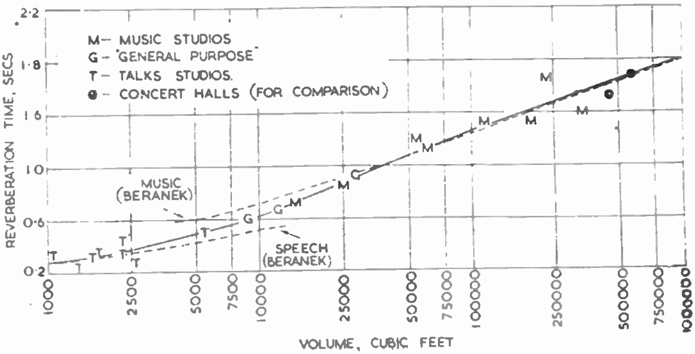


Fig. 5.—Reverberation Time of B.B.C. Studios.

tion time is a matter for compromise. The particular compromise adopted by the British Broadcasting Corporation[6] is shown in Fig. 5.

It will be noted from Fig. 4 that for studios with a volume of about 15,000 cu. ft. a large tolerance for reverberation times is indicated. This arises because, at about this volume, the requirements of drama studios and small music studios overlap. For drama studios it must be possible to produce different conditions of reverberation to agree with the different settings of a play. Some organizations tackle this problem by equipping the studios either with roller blinds, or with hinged or sliding wall panels of different absorption characteristics. These methods do not entirely satisfy the needs of drama production, as the variation in reverberation time is not very great, and the characteristics can seldom be modified during transmission.

A better solution to this problem is to use a " live-end, dead-end " studio, arranged by drawing—or partly drawing—heavy curtains across the centre of the studio, one-half of which is built with little acoustic absorption, and the other with considerably more. This acoustic treatment may be augmented by the use of large portable acoustic screens, one side of which is highly absorbent and the other side highly reflective. By these means, and by the employment of directional microphones, it is possible to segregate the microphones into zones having widely different acoustic characteristics. Portable acoustic screens, however, obstruct the view into the studio, and also tend to curtail an artist's movement. A novel solution to this problem has been put forward in a description of a new small drama studio[7] in which a tall folding screen, with different acoustic treatment on opposite sides and with curtains above, can be drawn across the studio. The line of partition bisects the control cubicle window so that a clear view is obtained into both the " live side " and " dead side " of the studio.

Reverberation Time versus Frequency

Reverberation time, considered so far in general terms, has referred to an average figure over the middle of the audio range or to the figure at 500 c/s. Unless this figure can be maintained level over a reasonably wide frequency band, its value as a guide to optimum reverberation time is meaningless. Indeed, a decade ago it was not uncommon to find studios or halls with reverberation times which might rise in the bass region to twice the value at 500 c/s, and which fell by a like-ratio in the treble, resulting in " boomy ", muffled reproduction.

One of the most heartening improvements in studio acoustics has been the manner in which reverberation over the major part of the audio range has been brought under control, and most new studios coming into commission have flat reverberation time/frequency curves. The exceptions are the larger studios and halls, where the air attenuation causes a fall in reverberation time above about 5,000 c/s, a natural phenomenon entirely acceptable. It is, however, noted that Scandinavian countries prefer the reverberation time to rise slightly below 100 c/s, and also, for orchestral studios, prefer a rise from 250 to 3,000 c/s with the studio empty. This rise probably results in a level curve when the orchestra is in place.

Determination of Reverberation Time

Prof. W. C. Sabine of Harvard University, the pioneer in this work, deduced an equation for reverberation time: [8]

$$\text{Reverberation time } (t) = \frac{0{\cdot}05V}{aS}$$

where t = time taken for sound in a room to decay by 60 db;
V = volume of the room in cubic feet;
a = average coefficient of absorption of S;
S = total area in sq. ft. of all surfaces of the room;
$aS = a_1S_1 + a_2S_2 + a_3S_3 + \ldots$;
S_1 = area of absorbent with coefficient of absorption a_1;
S_2 = area of absorbent with coefficient of absorption a_2.

Later, C. F. Eyring derived an equation [9] more applicable to large rooms or studios with large amounts of absorption. This gives:

$$\text{Reverberation time } (t) = \frac{0{\cdot}05V}{S[-\log_e(1-a)]}$$

This reduces to Sabine's formula when a is less than 0·2.

At high frequencies, carpets, heavy fabric curtains and furnishing materials provide useful absorption, which, however, increases in effectiveness with increase in frequency. Tables of absorption coefficients of these and of special absorbents and of common building materials appear frequently,[10, 11, 12, 13] and will not be repeated here, as interest is in more recent developments. Glass wool in various thicknesses between 1 and 4 in. is a most effective absorbent; it is vermin proof and does not absorb moisture. Covered with open-weave plastic material or faced with wooden slats of a particular cross-section, referred to as "Copenhagen treatment", it is extremely effective at high frequencies, whilst used in conjunction with hard perforated panels it becomes very effective in the medium-frequency range from 250 to 2,000 c/s.[7, 14]

In the low-frequency range, below 250 c/s, recourse must be made to the use of Helmhlotz resonators or to membrane or panel absorbers.

Helmholtz resonators have been used frequently on the Continent,[15] but the first use reported in this country was for the rebuilt Studio No. 1, Swansea, for the B.B.C.[14]

The particular arrangement used in this studio consisted of groups of resonators arranged in rows projecting from the surface of the wall, so providing diffusion as well as absorption. This arrangement, referred to as in "line array" has characteristics midway between those of isolated resonators and plane arrays.[16] By this means an absorption of 0·7 at resonance, i.e., absorption per resonator compared to a perfect absorber, has been obtained at a frequency as low as 90 c/s.

Membrane absorbers are probably the most attractive for use at low frequencies, as they are relatively simple to construct. The theoretical analyses of their behaviour is not so simple, but C. L. S. Gilford has presented a most comprehensive analysis,[17] confirmed by subsequent measurements, and a report of nearly twenty studios acoustically treated successfully by this means. The principal concert studio of the B.B.C. has been retreated acoustically with membrane absorbers and reported as most satisfactory in an article by T. Somerville and H. R. Humphreys.[18] This article also shows photographs of the pulsed-gliding-tone method of investigating irregularities in sound decay, a technique first reported in 1951.[19]

Diffusion

As experience has been gained in the choice and control of reverberation time and of reverberation time/frequency characteristics, so has the effect of the proper diffusion of sound been thrown into prominence. For some time opinion, influenced by investigations into this aspect of acoustics carried out in America, has favoured hemi-cylindrical diffusers. Recent work, in this country [20, 21] and in Denmark,[22] would seem to establish the superiority of rectangular diffusers beyond dispute. Furthermore, traditional architectural style may more easily encompass the use of rectangular diffusers in future building, with resulting improvement in æsthetic appeal.

Specific Defects

Even when the major problems have been solved, it might occur that a distinct "coloration" appears in the output from a studio, and it should be noted that heating radiators, lamp fittings and other studio fixtures may be resonating. Ventilating ducts may behave as Helmholtz resonators, and selectively either augment or attenuate the sound output. Inadequate attention to the reverberation characteristics of studio listening rooms, or inadequate sound insulation of echo rooms, may mar an otherwise satisfactory studio suite.

Finally, the sound output from even a good studio may be impaired if the selection and placing of microphones and artists are not meticulously carried out. Recent articles [23, 24] reviewing this technique have led to an appreciation and better understanding of this art.

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R. D. P.

BROADCASTING STUDIO EQUIPMENT

Although in the past many different arrangements of studios have been used in the preparation, presentation or recording of a programme, it is almost universally agreed that the most flexible arrangement is to provide an associated control cubicle immediately adjacent to each studio, with a large observation window built in the partition wall between the cubicle and studio. In the case of large studios of height spanning two floors, a cubicle on the upper floor level gives such an improved view into the studio as to compensate for the less easy access between cubicle and studio.

The studio control desk is situated immediately in front of the observation window, and carries all the controls necessary for controlling a programme.

All cubicles should be equipped with the means to enable the following operations to be performed :

(1) The amplification and mixing of the output from microphones, gramophones and incoming line contributions to the programme.

(2) The control of the permitted dynamic range of a transmission within certain defined limits as indicated on a particular meter.

(3) The aural monitoring of the acoustical balance and artistic and technical quality of the programme.

(4) " Talk-back " from the producer, via a microphone in the cubicle, and a loudspeaker in the studio to the artists in the studio.

(5) The operation of lights to signal the commencement and end of a transmission.

(6) The operation of light signals to " cue " or signify to an artist, groups of artists, or effects or gramophone operators the point to " come in " on programme.

(7) The aural identification of the correctness, and visual check of the level, of an incoming contribution previous to its being faded into the programme (pre-fade listening).

(8) Telephonic communication with any point on the engineering telephone exchange.

(9) The sending of " tone " to " line-up " the programme chain.

In drama and in variety studios the following additional facilities may be needed :

(10) Reverberation effects.

(11) Frequency distortion effects.

(12) A feed from any microphone or combination of microphones to the studio public-address equipment.

(13) Direct telephonic communication to the points of origin of any incoming contributions to the programme.

Programme Volume

Before an amplifying and mixing system can be designed it is necessary to know the maximum range of sound intensity which may occur. This is usually quoted as 80 phons, and is attributed to the output of a full symphony orchestra. In practice, owing to the ambient noise level in a concert hall or concert studio, it is much lower than 80 phons. The output of the B.B.C. Symphony Orchestra of 120 musicians, playing Elgar's Symphony No. 2, has been measured as 100 phons for the loudest fortissimo chord and 50 phons for pianissimo playing by the stringed instruments.[1]

A high-quality ribbon microphone with an output impedance of 600 ohms will produce at these extremes of volume a minimum level of -100 db and a peak level of -50 db (with reference to 1 mW into 600 ohms). The characteristics of different types of microphone are discussed in Section 31 " Microphones ", and it will be assumed here that by the use of local amplifiers, transformers and attenuators as necessary, their outputs may be adjusted to make them interchangeable in the studio.

The next information required is the peak level at which the amplified programme must leave the studio. This is specified by the British Broadcasting Corporation as $+8$ db peak, and the lowest peak permitted is -14 db, but provided the lower figure is indicated at least once every 20 seconds for orchestral works, the level is not adjusted. This means in practice that the total range is some 30 db, which is adequate for a large number of symphonic compositions. Nevertheless, it is necessary at times to compress the programme from a range of 50 to 22 db, and considerable artistic ingenuity is employed in accomplishing this (see " Broadcasting Studios ").

Amplification and Gain Control

The minimum basic amplification required may now be assessed. To raise the minimum peak-level input to the minimum permitted peak-level output requires a gain equal to the difference between -14 and -100 db, i.e., 86 db. Similarly, the minimum gain employed is the difference between the maximum peak-level input and the maximum peak-level output, i.e., the difference between -50 and $+8$ db $= 58$ db.

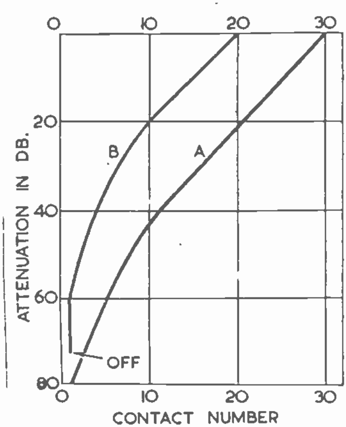


Fig. 6.—Main Gain Control.

The difference between the two gains of 86 and 56 db is 28 db and this represents the attenuation range over which the programme is to be regulated by a main gain control. This is usually a stud-type attenuator. In the interests of smooth control over the control range, and some 20 db below the step-to-step attenuation should not be greater than 2 db. After this the attenuation between steps may gradually be increased, since the ear is less sensitive to change of level as the actual level falls. The law of a typical main gain control is shown in Fig. 6.

The best position for the main gain control in the amplifier chain

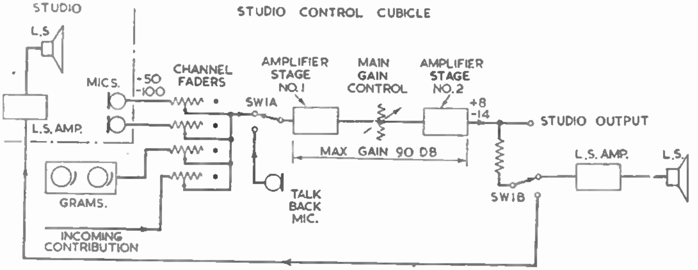


FIG. 7.—BASIC AMPLIFIER CHAIN.

Notes: 1. For one channel only faded into circuit, the figures +8/−14 denote level of μ and minimum peaks relative to 1 mW/600 ohms. 2. Switches SW1A and SW1B ganged and separated by an earthed metal screen.

may now be decided. It should not be placed previous to the amplifier chain, as the signal-to-noise ratio would be seriously decreased, nor should it be placed after the amplifier chain, as this would entail the use of unnecessarily high output power. The main control, therefore, is placed almost invariably between two stages of amplification.

The same arguments apply to the choice of the best position for the channel faders which are necessary for fading each source smoothly into and out of the programme. If the faders are placed in their rightful place, immediately preceding the main gain control, a stage of amplification is necessary for each channel. This step may be avoided, provided that certain precautions are observed, by the use of a series fader for each channel previous to amplification. As each additional source is faded into circuit the signal level and signal-to-noise ratio fall, due to the additional shunt impedance across the circuit, and the level must be restored by simultaneously advancing the main gain control. The decrease in signal-to-noise ratio can usually be neglected, due to the fortuitious circumstance that the types of programme which require a number of sources in circuit at one time generally have an initially high signal-to-noise ratio, whilst the type of programme generating the low levels mentioned earlier need seldom involve the use of a number of channels at one time. The attenuation curve of a typical channel fader is given in Fig. 6.

Amplifier Chain

The basic amplifier chain may now be shown as Fig. 7. The amplifier stages may consist of individual valves in cascade with a simple potentiometer between them. If the programme meter is also contained in the same unit as the amplifier, the arrangement represents an economical and compact assembly, and the ability to change the whole unit *en bloc* in case a fault develops commends it for use where simplicity and ease of maintenance are of first importance.

Design Refinements

For the more sophisticated needs of a large broadcasting organization, more complicated designs are necessary to meet the varied requirements already set out, and to satisfy the ever-increasing finesse needed in the presentation of variety, drama and music. A large number of channels becomes necessary, and a group control is needed to enable a number of sources to be faded in and out of programme by a single control, whilst the channel faders which set the degree of mixing need not be disturbed. It must be possible to by-pass the group control with at least one independent channel, so that a source may be faded up whilst the group is faded down, so preserving continuity of programme. A further refinement is to use two group controls, to which any combination of channels may be switched as desired. The channel and group mixers can be conveniently of the constant-impedance type, a configuration also suitable for the main gain control, for it can no longer be a simple potentiometer between two valve stages. The increased gain necessary to overcome the increased losses means that each amplifier stage of Fig. 7 now becomes a complete amplifier, and the link between them is of relatively low impedance.

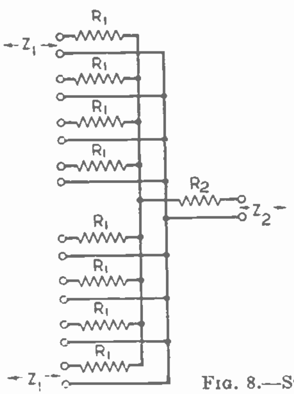
Mixing

The actual mixer losses depend on the circuits employed, which are as various as the number of organizations using them. The star mixer to be described has been found simple and effective, particularly for a large number of sources and where a constant impedance at the junction of all units is required throughout the chain

Referring to Fig. 8: If N is the number of sources of impedance Z_1 which are to be mixed to a common impedance Z_2, and if both Z_1 and Z_2 are to "see" matched impedance, then if M represents $\frac{Z_1}{Z_2}$ the values of the resistances are:

$$R_1 = \frac{M^2 + MN(N-2)}{N^2 - M} \times Z_2$$

$$R_2 = \frac{N^2 - M(2N-1)}{N_2 - M} \times Z_2$$



In the most usual case when $Z_1 = Z_2 = Z$ (and $\therefore M = 1$)

$$R = R_1 = R_2 = \frac{N-1}{N+1} \times Z$$

The loss introduced by these mixers is given by $20 \log_{10}(1/N)$, which, it will be noticed, is independent of the impedance of Z_1 and Z_2.

For the sake of completeness, it should be mentioned that if the number of sources N does not exceed a critical number Nc, given by $Nc = M + \sqrt{M^2 - M}$, then R_1 must be calculated for $N = Nc$. R_2 should be

Fig. 8.—Star Mixer.

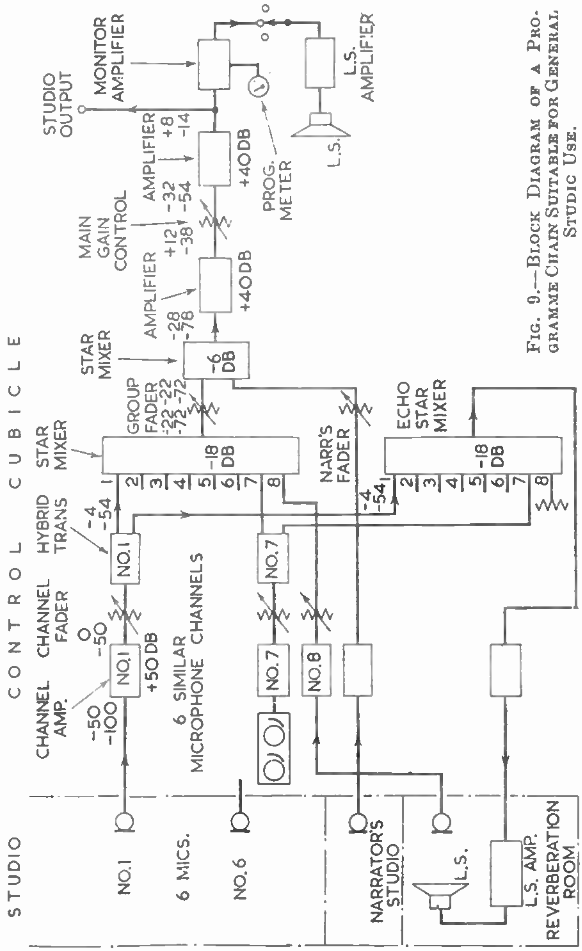


Fig. 9.—Block Diagram of a Programme Chain Suitable for General Studio Use.

zero, and the output channel Z_2 terminated with a resistance equal to $\dfrac{Z_1 + R_1}{Nc + N}$ in parallel with Z_2 and the mixing loss $= 20 \log_{10} \dfrac{2(Nc - M)}{Nc^2 - M}$.

It is more practical in initial design to ensure that if possible the number of sources is at least equal to the first whole number above Nc. The additional loss will most probably be negligible, whilst the additional source connection made available may prove useful, and until the need to use it arises it may be terminated in its matched resistance.

Considering two mixers, one suitable for eight sources of 600 ohms mixed to a 600-ohm load, and one suitable for joining two 600-ohm sources to a 600-ohm load.

The first gives $R = \frac{8 - 1}{8 + 1} \times 600 = 466$ ohms.

The second gives $R = 200$ ohms.

The relevant losses $= 20 \log_{10} \frac{1}{8} = 18$ db

and $20 \log_{10} \frac{1}{2} = 6$ db respectively.

A basic diagram (Fig. 9) can now be drawn of a programme chain suitable for general studio use, and the gains required of the amplifiers may be determined.

It was previously ascertained that the minimum gain required to raise the minimum peak-input level to the minimum peak-output level of −14 db was 86 db. To this must be added the losses of 4 db, 18 db and 6 db due to the insertion of the hybrid divider and the two star mixers in the chain. If, say, 16 db additional attenuation is allowed as gain in reserve on the main gain control, the total gain needed is $86 + 4 + 18 + 6 + 16 = 130$ db. Three amplifiers are shown in cascade in Fig. 9, but this is a matter of convenience. If the gain is equally divided between them we get the odd figure of 43 db gain for each. In practice, an amplifier would be used with its gain switchable to 40 or 50 db, and pre-set to suit a particular studio. The levels shown in Fig. 9 indicate that the penultimate amplifier would have to handle a peak output of +12 db without distortion, whilst the gain control still held 16 db of gain in reserve. If necessary the gain of the channel amplifier could therefore be set 10 db lower than that shown and the gain control advanced. However, if further losses were anticipated, as would be the case if the total number of channels were increased, or if special effect units, or equalizers, were inserted in the chain, the gain of the channel amplifier would be restored to 50 db.

Artificial Reverberation

It is not often that arrangements are made for reverberation to be applied to more than one or two channels at one time, although details of a most comprehensive arrangement have been published,[1] which is shown simplified in Fig. 9. The output of each channel fader is divided by a hybrid transformer, which is essential to prevent howl-back from the reverberation-room microphone to its own loudspeaker. Hybrid coils introduce loss in this direction of transmission only. Alternatively, buffer amplifiers must be used, one output of which is connected to the main chain-star mixer, and the other to an echo-star mixer. The output of the latter is routed via amplifiers to the loudspeaker in the reverberation chamber, whence a microphone picks up the reverberant sound, and the resulting electrical output becomes a source and appears as Channel 8 in the diagram. Ganged faders associated with the two outputs from the hybrid transformer may be pre-set to vary the ratio of direct to reverberant sound.

Ideally, the reverberation chamber should be able to be varied in size to be able to give different aural illusions, for the ear is not only sensitive to reverberation time but also to the delay time of early reflections. The development of electronic means of artificial reverberation control may eventually prove superior to small reverberation rooms.

Monitoring

The cubicle loudspeaker used for aural monitoring should be able to be switched away from its own studio output to monitor the programme to which the studio is to make its contribution. It should be able to be attenuated instantly by a fixed amount whilst a telephone is in use. The studio loudspeaker should monitor the programme on all occasions when the studio is in service and all the studio microphones are faded out, e.g., during a period when only gramophones or outside sources, being part of the studio programme, are faded up. In this way the artists in the studio may join the programme more smoothly.

To permit proper control of the dynamic range of a transmission, a programme meter is required. There are two main types in general use. In this country and on the Continent a peak programme meter is generally used, which conforms reasonably closely to the U.I.R. C.C.I.F. Specification of 1935, whilst in America a programme volume meter is preferred.[3] The former consists of a rectifier and logarithmic amplifier driving a milliammeter.

The British specification including the meter calls for the pointer to reach 80 per cent full-scale-deflection in 4 milliseconds, and to return to rest in 3 seconds. The charging-time constant of the rectifier-circuit necessary to satisfy this specification is 2·5 milliseconds and the discharging time constant 1 second.

A minor difference from British practice is noted in some Continental countries, where a programme meter is called upon to have a rise time of 10 milliseconds for 80 per cent full-scale deflection and a return time of 2 seconds.

The scale of the 2½-in.-diameter instrument used in this country is shown in Fig. 10. The pointer, the figures and calibration marks are in white against a black background, and make the meter easy to read. The scale is approximately logarithmic, the space between the figures represent 4 db. except between the figures 1 and 2, where the difference is 6 db. The meter is calibrated by causing it to read steady tone of zero decibels level (with respect to 1 mW across 600 ohms) when the meter should indicate 4. This represents "line-up" level, at which point the transmitter modulation is 40 per cent. Peaks of programme indicating 6 on the meter and equivalent to +8 db peak level

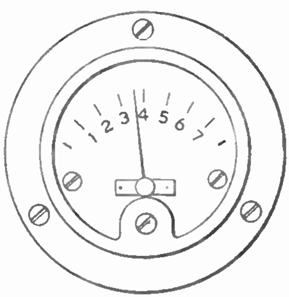


FIG. 10.—PROGRAMME METER.

therefore indicate 100 per cent modulation. Other forms of meter and scale are used, including a light beam moving across a linear translucent scale.

A rehearsal talk-back circuit may be added to the arrangements shown in Fig. 9, by enabling the cubicle microphone to be switched to the input of the second amplifier in the chain in place of the studio feed, at the same time disconnecting the cubicle loudspeaker. For transmission, however, a separate cubicle microphone amplifier is necessary whose output feeds only studio headphones.

It is possible that some programmes, e.g., when presenting two dance bands alternately, need a greater number of sources than can be accommodated on the channels available. In such a case a change-over relay may be fitted to the input of the requisite number of channels, and all these relays energized via a single key, upon the operation of which all the microphones of one band are substituted for those of another. It should be remembered that the cue lights of each band should be transferred at the same time.

Cue lights should be fitted as near as is conveniently possible to the artists: white lamps may temporarily dazzle anyone reading a script, and green lamps of from 6 to 15 watts have been found preferable. The lamp filament should not have a high thermal lag, or a quick " flick " on the lamp may pass unnoticed. If one switching contact on the cue key is arranged to switch on a small green cue lamp on the control desk a fraction later than a second contact on the key switches the studio cue, it will assure the operator that the cue has been properly transmitted.

The present trend in the fashioning of control-cubicle equipment is towards a self-contained desk which accommodates the necessary amplifiers, power-supply units, relays and switches. With miniature components available this arrangement becomes most attractive, particularly from the installation point of view. For large studios, however, there is still a place for the earlier arrangement of equipment, in which only the necessary controls were mounted on the desk, most of the associated equipment being contained in an adjacent cabinet. This arrangement generally provides more room at the desk, and greater accessibility to equipment in both the desk and the cabinet; whilst induced hum, microphonic noise and acoustic noise from relays and mains transformers are generally less troublesome.

Acknowledgment

The author wishes to express his thanks to the Chief Engineer of the British Broadcasting Corporation for permission to publish this review, and to his colleagues who have provided much helpful information and advice.

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R. D. P.

TELEVISION STUDIO PLANNING AND LAYOUT

Small studio installations are most conveniently and economically housed in a single-storey building. Large installations comprising several studios with full facilities for large-scale presentations may require a multi-storey construction. In all planning, provision for future expansion should be kept in mind.

Fig. 11 is a typical accommodation plan, which brings out the main features of a large studio with associated facilities. Accommodation for artists, technical personnel, office staff and visitors is grouped into separate areas, so that their paths do not cross unnecessarily in the course of their varied activities. In commercial television the transmitter is commonly incorporated in the studio building, with the transmitter hall, power-distribution equipment, valve store and a light workshop grouped together in one section. In these circumstances the building is sited at the highest accessible level above the general terrain, near the centre of the region to be served, and if a high radiator is to be erected, the area of the site must be large enough to embrace the stay anchorages. The mobile outside broadcast unit is garaged near the studio, so that camera and control equipment can be transferred quickly to augment the main studio equipment for full-scale productions.

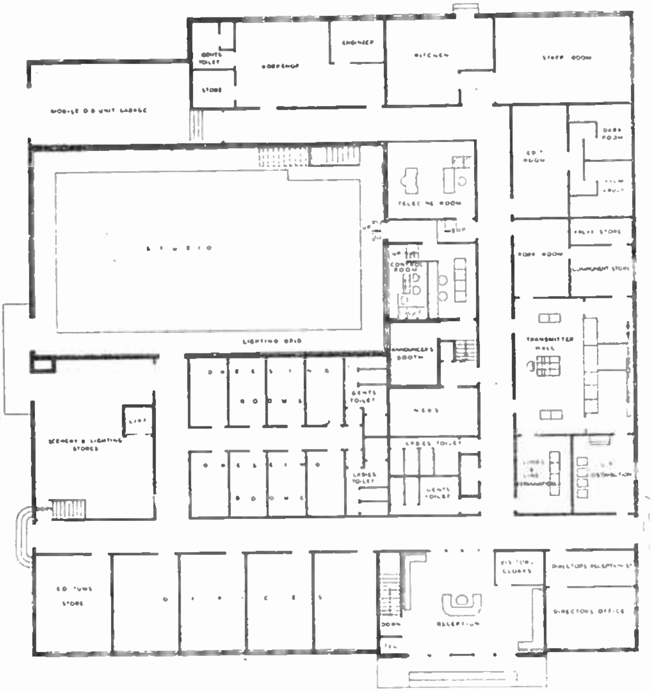
The particular arrangement shown is planned to allow for future extension by transferring the dressing-rooms to a basement and re-siting the toilet facilities, thus leaving the dressing-room space available for a second studio and the toilet space for a second control room and master control room.

Floor Areas

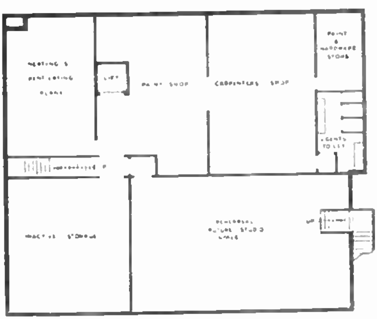
Studio floor areas vary with the equipment to be installed, the scale of programme production to be catered for and in the case of large studios with the amount of space allotted for an auditorium. The area must be sufficient to permit free movement of cameras and microphones around the sets. Typical floor dimensions for the largest general-purpose studios are 120 ft. × 70 ft., for medium-size studios 70 ft. × 45 ft., and for small studios 50 ft. × 25 ft. Heights of 50 ft. or more may be necessary to provide for certain acrobatic acts. In the arrangement shown, the main studio has a floor area 65 ft. × 40 ft., sufficient to accommodate three or four sets and camera channels, allowing for continuous production.

A gallery about 4 ft. wide erected 10–12 ft. above floor level around the studio walls, and accessible from the control room, is an extremely useful feature for giving access to the catwalks for the lighting grids and for taking downward camera shots. Background lighting units can be clamped to the hand-rails, the tops of sets anchored to the gallery and curtains hung in any desired position to serve as a background for announcements and interviews.

The studio floors must be solid, level and surfaced with smooth, resilient material to ensure even, quiet running of camera dollies and microphone stands and subdue the noise of movements of personnel. These requirements are met by a 6–12-in. layer of concrete, covered by thick, hard-wearing linoleum or hard rubber sheet. Property stores and artists' dressing-rooms should be situated for convenience near by. Studios and property rooms should preferably connect at ground level,



GROUND FLOOR PLAN



BASEMENT PLAN

FIG. 11.—TYPICAL ACCOMMODATION PLAN FOR A TELEVISION STUDIO CENTRE.

and must have doors large enough for easy transfer of " props ", with access to outside for large articles such as cars and animals.

Studio and Control-room Facilities

The function of a control room is to direct and monitor all operations in the studios and establish control between the studio and the transmitter. A typical facilities diagram is shown in Fig. 12. A separate control room adjoins each studio, equipped with sound-proof windows and stepped up to a level 3 or 4 ft. above the studio floor level, to afford an unobstructed view of all studio operations.

In commercial studios it is customary to group the camera control operators, vision mixer, audio operator, supervising engineer and producer with his secretary together in the control room. During rehearsals the scene designer and lighting engineer may also be present, to check the settings on the monitoring tubes as they would be seen by viewers. The producer sits on a dais behind the camera-control units looking into the studio, and uses the monitoring tubes on the control units for previewing each picture channel. This arrangement gives him full facilities for calling his vision mixer to cut the various cameras at will, and economizes in both personnel and equipment.

This scheme, however, has limitations from an operational point of view, and the present trend is towards separation of the individual groups. The camera-control-unit operators are accommodated in a room adjoining the studio and at studio floor level, the control room being situated immediately above and divided by means of a motor-operated window into a video and audio section. The producer sits in the video section facing a row of high-grade picture monitors, which repeat the control-unit monitors, and sideways to the control-room window. A " talk-back " network enables him to maintain continuous contact with the camera crew, microphone operators and technicians, while the audio controller in the audio section is able to check his programme level free from the distraction of the producer's commentary. During camera rehearsals the dividing window between the video and audio sections is lowered to allow the groups to co-operate as a team.

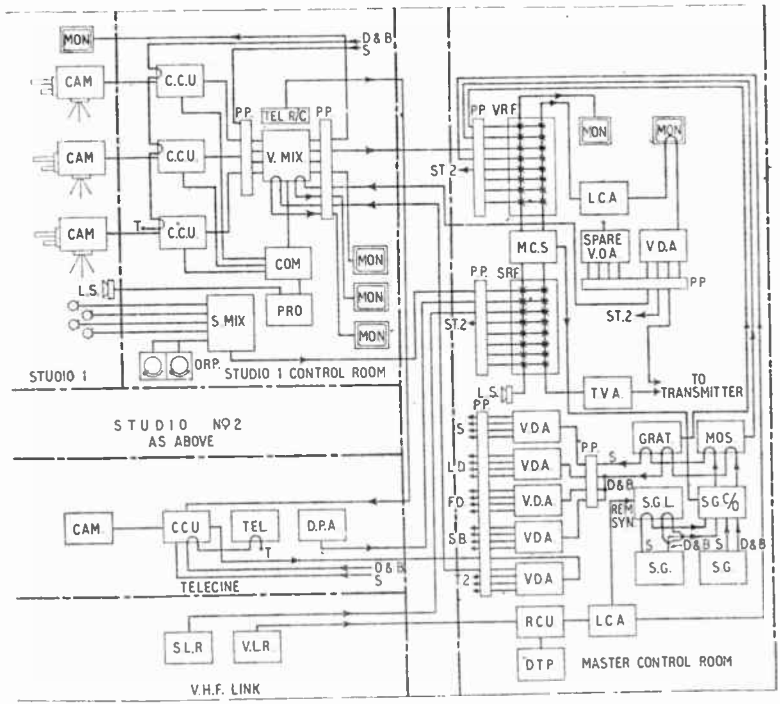
Where there are two or more studios, a master control room becomes essential, to co-ordinate all internal programmes and outside broadcasts received by co-axial cable or U.H.F. link from distant sources. This is achieved by tie-lines and video-frequency and audio-frequency patching panels. Economies in personnel and cables are made possible by placing the teleciné room adjacent to the control room, with the editing room and film storage near by.

The equipment for a projection room includes teleciné equipment, or alternatively a 16-mm. or 35-mm. film projector and camera, film cameras, and a slide projector or monoscope tube for " stills " and captions.

Air-conditioning plant or a forced-air ventilating system must be installed to remove the heat generated by the lighting equipment and compensate for seasonal variations of temperature in studios and control rooms. The plant should be designed to maintain the temperature at a level between 70° and 80° F. for reasonable comfort. Economies in heating and heating charges can often be made by ducting the warmed air to other parts of the building during the presentation and rehearsal of programmes.

Fig. 12.—Typical Television Studio Control Room Facilities.

CAM.	Camera
C.C.U.	Camera Control Unit
COM.	Communication Unit
D & B	Drives and Blanking
D.P.A.	Dual Pec Amplifier
DRP.	Disc Reproducer
D.T.P.	Dual Telephone Panel
F.D.	Field Drive
GRAT.	Grating Generator
L.D.	Line Drive
LS.	Loudspeaker
M.C.S.	Master Control Switching
MON.	Monitor
MOS.	Monoscope Camera
P.P.	Patch Panel
PRO.	Production Console
R.C.U.	Receiver Control Unit
S	Synchronizing
S.B.	System Blanking
S.G.	Synchronizing Generator
S.G.C/O	Synchronizing Generator Change-over Unit
S.G.L.	Synchronizing Generator Locking Unit
S.L.R.	Sound Link Receiver
S.MIX.	Sound Mixer
S.R.F.	Sound Relay Frame
T	Termination
TEL.	Telecine Equipment
TEL.R/C	Telecine Remote Control
T.V.A.	Trap Valve Amplifier
V.D.A.	Vision Distribution Amplifier
V.L.R.	Vision Link Receiver
V.MIX.	Vision Mixer
V.R.F.	Vision Relay Frame



Studio Acoustics

Studio walls and ceilings are either built of or treated with acoustic insulating materials, mainly to exclude outside noise. The problem of eliminating undesirable reflections in television studios is simpler than in sound-broadcasting studios, which are relatively devoid of furnishings. Reflections are largely broken up or damped by the wings and " props ". Noise may be admitted through windows, doors and other openings, door clearances, by conduction through the solid materials of walls and by structural resonances. The maximum noise level that can be tolerated in studios is 10–15 db above the threshold of hearing, and in recording rooms 8–12 db.

Precautions taken to reduce noise include the use of double doors with a sound lock. Control rooms are acoustically insulated from studios, the programme being monitored on loudspeakers and heard under the same conditions as the listener. Air noise introduced by ventilating ducts is reduced by admitting the air through ducts of generous cross-section at low velocity, and by fitting noise filters or lining the ducts with absorbing material.

In small commercial studios designed for demonstrations and announcements, reflection is controlled by the liberal use of sound-absorbing materials on the walls and ceilings, such as draperies, plaster, acoustic panels and slag or glass wool sandwiched between perforated metal sheets, wire netting or asbestos board. In general, the acoustic absorption of all materials increases with frequency, and is improved by increasing the thickness and leaving an air space between the material and the wall or ceiling.

Draperies hung a few inches from the wall are effective absorbers, and are economical and adaptable. Plaster is moderately efficient, but does not withstand rough treatment. Acoustic panels have excellent absorbent properties at all but the lowest frequencies, but being porous, they trap dust and discolour. Painting prevents discoloration, but greatly reduces the absorption efficiency. Covered mineral wools are most effective, and the coverings can be painted without serious loss of efficiency.

W. E. P.

TELEVISION STUDIO LIGHTING

The methods used in television studio lighting are founded largely on the technique evolved for ciné film production and the theatrical stage. In television studios action is continuous, and there can be no re-taking. The equipment must satisfy both technical and artistic requirements, be readily adaptable for rapid changes of the set and silent in operation. It must be sufficiently flexible for the fourfold purpose of general illumination of the set, the subject, the background and the creation of special effects. The lighting intensity must be adjusted to suit the sensitivity of the camera tube in use and the tones of the settings. Picture quality and dramatic atmosphere are profoundly influenced by the ratio of highlights to shadows.

Lighting units fall into three classes : floodlights, banked floodlights and spotlights. Incandescent lamps, rated from 100 to 5,000 watts, find wide application to both general and spot lighting. They are light in weight, silent and more compact and adaptable than gas discharge tubes, but are deficient in radiation at the violet end of the spectrum,

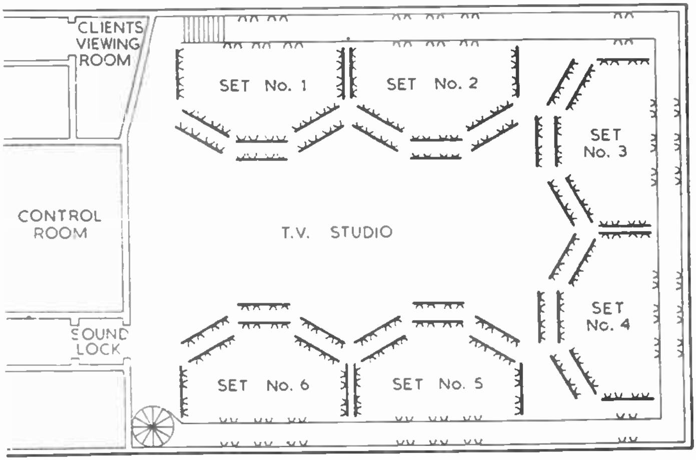


FIG. 13.—TYPICAL STUDIO LIGHTING SCHEME FOR SIX SETS.

and radiate an appreciable amount of heat, which must be extracted by the ventilating system. Gas discharge lamps generate little heat, and are richer in ultra-violet radiation than daylight, but their maintenance costs are higher. Compact source discharge lamps are sometimes used for spot lighting. Carbon arc lamps are a powerful and efficient source of pure white light, but are cumbersome. They are occasionally used for illuminating outside events and special effects.

Banked floodlights for wide-angle illumination of the set are mounted on movable overhead frames or grids, ranged along the walls, suspended from the roof or attached to the tops of the sets, about 15 ft. above floor level, as shown in Fig. 13. These lamps may be supplemented by low-level floodlights. High- and low-level sources are carefully blended to eliminate facial shadows and shadows cast by microphone booms and cameras, and to prevent the shadows of actors obscuring each other while in movement. The intensity varies up to 120 ft.-candles, according to whether it is required to simulate interior, exterior, day or night light.

Subject lighting is concentrated directionally and projected at an angle to the viewing line, to contrast the subject with the background and give an illusion of depth. More than one source may be used to emphasize particular features and colour values. This form of lighting is provided by spot lights or "modelling lights", equipped with Fresnel lenses, reflectors and shutters, designed for beam divergencies from 10° upwards. The intensity is somewhat higher than for general lighting. For close-up portraiture, general and background lighting are supplemented by individual lights to enhance the effect of make-up.

All lighting is grouped into separate circuits, controlled by switch-fuse ways and dimmers on a distribution panel, situated at a convenient control point. To avoid interference, cables are run in separate channels

TABLE 2.—STUDIO EQUIPMENT PARAMETERS

	B.B.C. *(British)* *405-line*	*R.M.A.* *(U.S.A.)* *525-line*	*C.C.I.R.* *(International)* *625-line*
Vision-signal Distribution:			
Impedance of co-axial cables	75 ohms	75 ohms	75 ohms
Polarity: white positive	Yes	Yes	Yes
Non-composite signal level, peak-to-peak voltage	0·7	1·5 *	0·75
Composite signal level, peak-to-peak voltage	1	2 *	1
Distribution of Drives, Blanking and Synchronizing:			
Impedance of co-axial cables	75 ohms	75 ohms	ohms
Polarity: negative	Yes	Yes	Yes
Level, voltage peak-to-peak	2	4	2
Drives:			
Line-drive width	8 μs	6·5 μs	5 μs
Leading edge of line drive relative to leading edge of line synchronization	−4 μs	−1·6 μs	−1·6 μs
Field (frame †) drive width	400	500	500
Leading edge of field (frame †) drive coincident with leading edge of field (frame †) blanking	Yes	Yes	Yes
Line Waveform:			
Line period (H)	98·8 μs	63·5 μs	64 μs
Line blanking period	17(0·17 H) μs	11·5 (0·18 H) μs	11·5 (0·18 H) μs
Front porch	1(0·01 H) μs	1·6 (0·025 H) μs	1·6 (0·025 H) μs
Line-synchronizing pulse width	10(0·1 H) μs	5 (0·08 H) μs	5·75 (0·09 H) μs
Field (Frame †) Waveform:			
Field (frame †) period (U)	20,000 μs	16,667 μs	20,000 μs
Field (frame †) blanking period	1,450 (0·07 U) μs	1,100 (0·065 U) μs	1,600 (0·08 U) μs
Leading edge of field (frame †) blanking coincident with leading edge of first field (frame †) synchronizing pulse	Yes	No	No
Leading edge of field (frame †) blanking coincident with leading edge of first equalizing pulse	—	Yes	Yes
Field (frame †) synchronizing pulse width	40 (0·4 H) μs	27 (0·43 H) μs	26 (0·41 H) μs
Equalizing pulse width	—	2·5 (0·04 H) μs	2·9 (0·045 H) μs
Number of field (frame †) synchronizing pulses	8	6	6
Number of equalizing pulses before and after field (frame †) synchronizing	0	6	6
Picture/Synchronizing Ratio:			
Blanking level, reference peak white	30%	25%	25%

May be reduced in practice. † British terminology.

from video-frequency and audio-frequency cables. Overhead lights are fed from connectors fixed at suitable points and connected to the distribution panel by cables run in metal cable troughing in the walls or along the studio ceiling. Low-level lamps and spot lights are fed from sockets conveniently disposed around the skirting.

W. E. P.

TELEVISION STUDIO EQUIPMENT

The normal complement of television apparatus in a studio is three or four cameras on the studio floor, together with one film transmitter for film inserts in the studio programme. The cameras are variously mounted, depending on the nature of the programme, e.g., camera cranes, dollies or trucks and pedestals. All cameras are energized all the time, and the producer in the control room has presented to him simultaneously pictures from all cameras so that he can arrange his choice of angle and view before using any particular camera for transmission. Under the control of a vision-mixing operator are facilities for cutting, fading and dissolving from camera picture to camera picture.

By means of a microphone, the producer can give instructions to cameramen, sound operators, dolly pushers, etc., who all wear headphones. Connected to this talk-back network is a small radio transmitter, so that the studio manager who is in charge of all the artists, scene shifters, etc., on the studio floor can also hear the producer with the aid of a small pocket radio receiver. During rehearsals, the producer can also talk to the artists by means of a loudspeaker in the studio.

In all modern television cameras the cameraman is provided with an electronic viewfinder which presents him with the picture from his own camera so that he can focus and adjust his angle of view conveniently.

A turret, carrying three or four lenses of different focal lengths, makes the selection of the angle of view convenient and rapid.

Each camera is connected by a multi-wire flexible cable to a rack of equipment in the television apparatus room adjacent to the studio control room. This rack contains all the necessary power supplies, pulse forming and amplifying circuits, and mixing stages for adding to the picture signal emanating from the camera the appropriate suppression and synchronizing signals. The output signal from this rack, or camera channel, is a complete television video signal of 1 volt amplitude (0·3 volt synchronizing signal, 0·7 volt picture signal).

In the camera channel are the necessary variables for adjusting picture gain, black level, etc., for the proper adjustment of the picture for transmission. The operator has a picture monitor and a waveform monitor to display clearly the picture signal waveform to assist in these adjustments.

The precise layout and disposition of equipment varies from studio to studio, but Fig. 14 shows a typical example.

Film Transmitters.

Two types of film transmitter are used: (*a*) a continuous-motion projector in conjunction with a television studio camera, and (*b*) a flying-spot film scanner. The former is used exclusively for films which form part of studio programmes, whilst the latter is used for newsreels, feature films, etc.

(*a*) CONTINUOUS-MOTION PROJECTOR. The principle of the continuous-motion projector is indicated in Fig. 15. The film is illuminated by a conventional lamp arrangement and an image of the film frame, F_1, is formed by the mirror, M_1, and the lens, L, on the sensitive surface of a television camera. As the film moves down, the mirror, M_1,

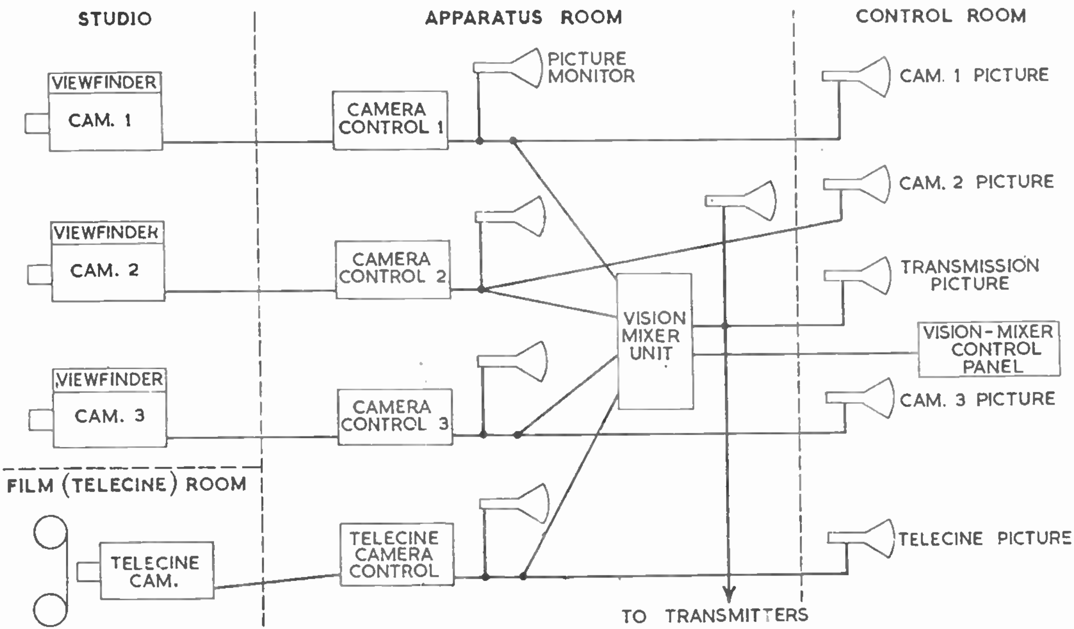


FIG. 14.—TYPICAL LAYOUT OF EQUIPMENT IN A TELEVISION STUDIO.

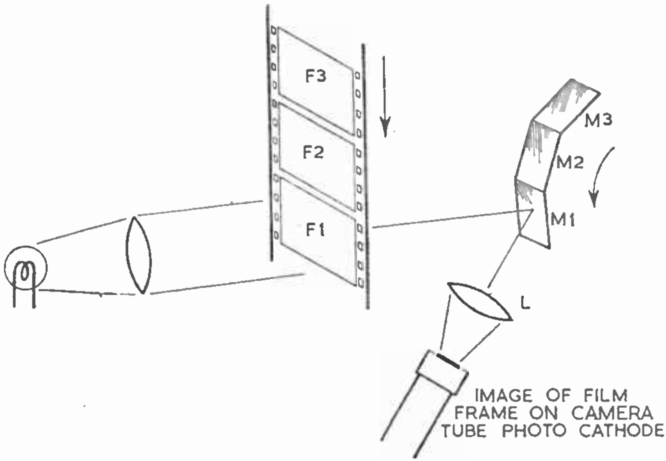


FIG. 15.—THE PRINCIPLE OF THE CONTINUOUS-MOTION PROJECTOR.

moves in sympathy to keep the image on the camera stationary. As the film frame, F_1, passes from the illuminated source, the next frame, F_2, becomes illuminated, and by means of mirror, M_2, an image is formed in the camera exactly superimposed on the image of F_1. Finally, as F_1 passes completely from the illuminated area, F_2 is fully illuminated. And so on, by mechanical optical means, the machine dissolves from film frame to film frame, all the time presenting a stationary image of each successive frame on the sensitive surface of the camera.

The television camera converts the optical image into a television signal in the same way as it would a studio scene.

(*b*) FLYING-SPOT FILM SCANNER (SEE FIG. 16). A raster is formed on a high-intensity cathode-ray tube, A. Two identical images of the raster are formed by the split optical system, B, on the film, D, one image being shuttered off by the shutter, C. Behind the film the photo-electrical cell, E, collects the light from the raster which passes through the film.

As the light spot traverses the screen, and its image traverses the film, an electrical signal will be generated in the photoelectric cell corresponding to the picture density in the film, i.e., corresponding to the picture brightness in the original scene.

The film is moved continuously and smoothly. For the odd lines of the television picture the top image, I_1, is used. When these are complete, the shutter obscures the image, I_1, and releases the image, I_2, and the even lines are then scanned, giving the complete interlaced picture.

Outside Broadcasts

In essence, the equipment used for television outside broadcasts is similar to that used in the studio, but it has to be more compact and

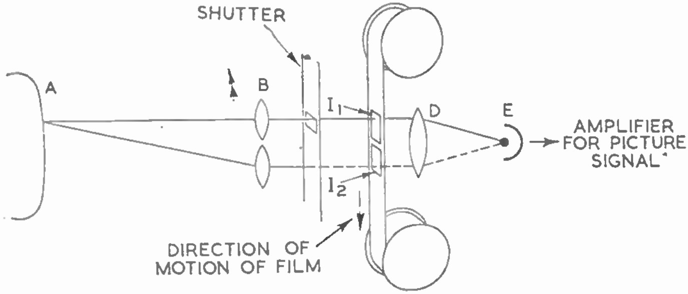


FIG. 16.—THE FLYING-SPOT FILM SCANNER.

more flexible, both with regard to facilities and capability of working over a wide range of conditions of light and weather.

The normal complement is three cameras. The operating staff, producer, etc., with the apparatus, work in a mobile control room. At many outside broadcasts, e.g., from theatres, the equipment is removed from the van and set up in a room in a building.

The cameras themselves are located up to a maximum of about 1,000 ft. from the control room in positions of vantage. The producer has a simultaneous display of the pictures from all three cameras, as in the television studio control room.

By contrast with a studio production, where the producer can position and move his cameras at will, generally speaking on outside broadcasts the camera positions are dictated by circumstances which bear little relation to artistic effect, e.g., cameras must often be located a long way from the centre of action and cannot be moved. This leads to the use of a great variety of long-focus lenses in order to obtain close-ups, say, of the batsman at the wicket. This in turn means very often a great deal more perspective distortion than is desirable.

One interesting lens used in outside broadcasts is the Zoom lens. The focal length of this lens can be varied by as much as 5 to 1 without change of focus, so that the angle of view can be changed while the camera is " on the air ". This is often used to good effect on sporting events to come into close-up on incidents of interest and then, so to speak, to retire discreetly when the interest widens.

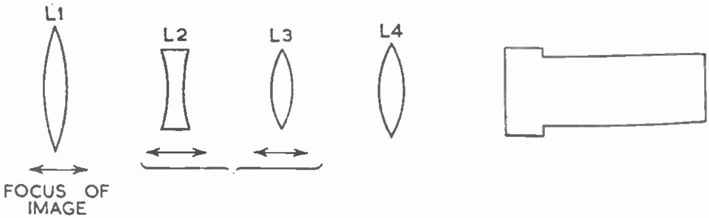


FIG. 17.—THE ZOOM LENS.

The lens itself comprises four elements, which can move relative to one another in a complex relationship as shown in Fig. 17.

The front lens, L_1, is moved to focus the image. L_2 and L_3 are moved to alter the overall focal length, and L_4 forms the image of the scene on the photo-cathode of the pick-up tube.

Cameras

The eye of the television camera is the pick-up tube. In British television several types of pick-up tube are in use. It is true to say that the completely satisfactory pick-up tube has not yet been devised, and therefore a choice has to be made with programme and artistic requirements on the one hand and technical considerations on the other.

The factors which must be considered in choosing a television camera tube for a particular function are: good resolution, sensitivity (i.e., the lighting required for good picture), freedom from unwanted distortion and contrast characteristic. The only factor in the camera design which is affected appreciably by the television receiver is the contrast characteristic, so a brief outline of this problem will show its importance.

It is well known photographically that for the satisfactory reproduction of a scene the relationship between the reproduced brightness of a particular part of the scene and the brightness of the original point in the scene should be (very) approximately linear. Some deviation is permissible, and the amount is largely a matter of opinion (see Fig. 18).

In a television system, assuming that all the amplifiers, radio links, etc., are distortionless (an assumption which is substantially true for the purposes of this discussion), then the two controlling elements in the relationship between original scene brightness and reproduced scene brightness are the television camera and the cathode-ray picture tube in the receiver.

In the receiver cathode-ray tube the relationship between voltage applied to the grid and the brightness of the screen is not linear, but is approximately a power curve of the form

$$B = KV^x$$

where B = brightness of screen;
K = constant depending on the particular tube;
V = applied grid volts;
x = coefficient (see Fig. 18).

For most television receiver tubes the value of x lies in the region of 2·5.

If therefore the overall characteristic is to be approximately linear, the television camera must have a characteristic which is the inverse of the receiver characteristic, i.e.:

$$v = k\frac{1}{b^x}$$

where v = output voltage;
k = constant depending on particular camera;
b = original scene brightness;
x = coefficient of power loss of television receiver tube (see Fig. 18).

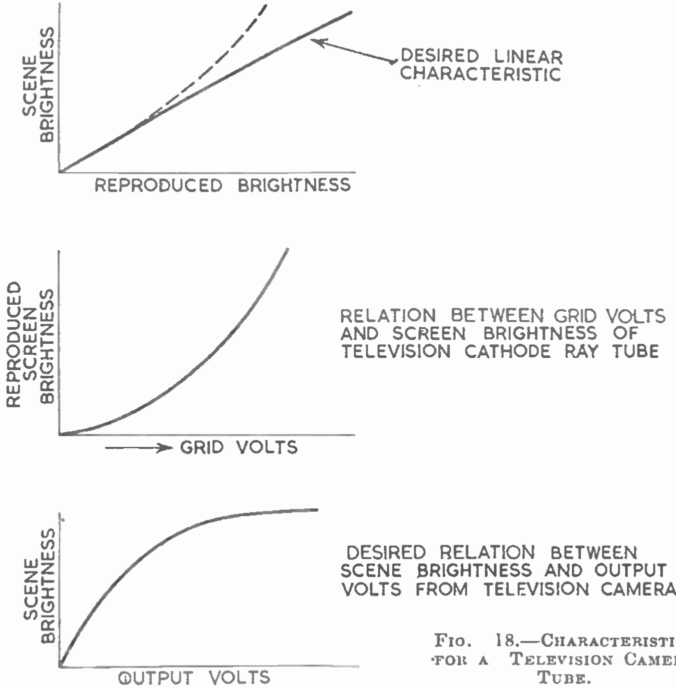


Fig. 18.—Characteristics for a Television Camera Tube.

For studio productions, the more important factors are good contrast characteristic, good definition and freedom from unwanted distortions. Factors not so important are sensitivity and ability to work over a large range of lighting conditions. In studios, the cameras in use are Emitrons, C.P.S. Emitrons (C.P.S. = Cathode Potential Stabilized) and Photicons, and P.E.S. Photicon (P.E.S. = Photo Electron Stabilized).

The most modern studios in Britain employ 4½ in. image orthicon cameras. The B.B.C. White City Television Centre is based on these cameras for all except Presentation Studios.

On outside broadcasts the cameras must be able to work with extremely bright light, e.g., summer sunshine, and very little light, e.g., the interior of a cavern. Therefore, extreme flexibility and adaptability is the dominant factor, and such matters as contrast characteristic and contrast range are secondary. On outside broadcasts, the camera tube is universally the Image Orthicon.

Image Orthicon Tube

A brief description of the mechanism of the Image Orthicon Camera Tube is given below, and a sectional schematic layout is shown in Fig. 19.

TABLE 3.—SUMMARY OF CAMERA TUBES

Camera Tube	*Use*	*Contrast Characteristic*	*Contrast Range*	*Sensitivity*	*Remarks*
Emitron	Studio	Good	Good	Low	Obsolescent. Recognized by white flare often at bottom of picture, and poor depth of focus.
C.P.S. Emitron	Studio	Good	Fair	Medium	Recognized by smooth rendering of dark tones and absence of "flare". Occasionally unstable in bright parts of picture.
Photicon	Studio	Good	Good	Medium	Recognized by white flare often at bottom of picture.
P.E.S. Photicon	Studio	Good	Good	Medium	White flare negligible.
Image Orthicon	Outside Broadcasts	Fair	Fair	Extremely Sensitive	Recognized by typical black halo round bright parts of picture.

The lens focuses the optical picture on the photo-cathode. Electrons are emitted proportional to the incident light. and are drawn towards the target by the voltage applied to the accelerator grid, their paths being kept parallel by the magnetic and the accelerating fields.

The electrons pass through the copper-mesh screen, which has 750 squares to the inch, in front of the glass target, which they strike, releasing secondary electrons, which are collected by the copper screen. This screen is held at a fixed potential of about 2 volts, with respect to the cathode of the gun. Since the mesh and the glass are in close proximity (between 1 and 2 thou. inches), the surface of the glass is stabilized against change of potential for all values of incident light.

The emission of secondary electrons leaves a pattern of positive charges on the front side of the glass target corresponding to the distribution of light over the photo-cathode. These charges leak away in two ways :

(*a*) by leakage across the surface of the glass; and
(*b*) by conduction through the glass to the rear side.

The thickness and quality of the glass is so adjusted that the transfer of charge by the second mode predominates (the temperature of the glass is an important factor in this).

By this transfer of charge, a positive electron image of the scene appears on the rear side of the target.

This rear side of the target is scanned by a beam which strikes it orthogonally. This is achieved by the use of a combination of the axial magnetic field produced by the focusing coil and the transverse magnetic fields of the scanning coils.

Because of the presence of the Decelerator Electrode, fixed at a small potential with respect to cathode, the scanning beam strikes the target at a very low velocity, so that no secondary electrons are emitted. The beam gives up electrons depending on the charge on the glass. The remainder of the beam returns by substantially the same path to the No. 2 grid of the electron gun. (There is a hole in this to permit the ejection of electrons from the gun.) This grid forms the No. 1 dynode of a five-stage electron multiplier. The output from the electron multiplier (of the order of 100 mV) is amplified by conventional means.

The scanning beam is focused by the electrostatic field of grid No. 5.

The beam is fired axially down the magnetic field along the tube by adjustment of the current and position of the alignment coil.

The secondary electrons from dynode No. 1 of the multiplier are focused on to dynode No. 2 by the electrostatic field from grid No. 3.

The quality of the picture is affected by the conductivity of the glass target. If the conductivity is too low, smearing will occur; if it is too high, resolution will be lost. It is necessary, therefore, to keep the

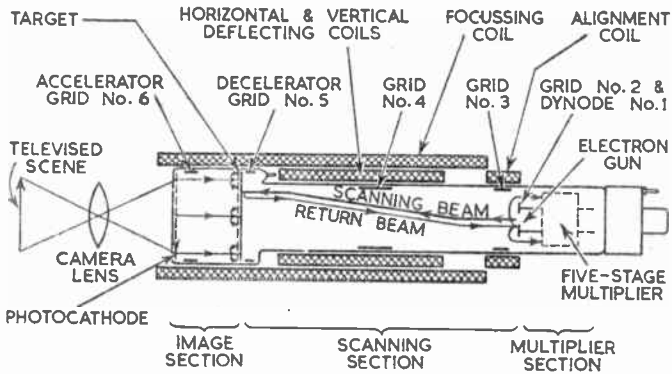


FIG. 19.—THE IMAGE ORTHICON CAMERA TUBE.

temperature within fairly close limits 35–60° C. (This is achieved by the use of a heater and blower, depending on external conditions.)

Hysteresis in the glass target gives the tube a memory of a few seconds, and this is also minimized by keeping the temperature within the prescribed limits.

The image orthicon has two regions of operation :

(*a*) the low-light region, where the signal output is proportional to the light on the photocathode;

(*b*) the high-light region, where the signal output is substantially independent of the average light on the photocathode.

It is usual to operate the tube in the high-light condition, though pictures can be obtained in the low-light condition at a sacrifice of signal-to-noise ratio, smearing and resolution.

The mechanism of the tube in the high-light region is approximately as follows :

Suppose a picture comprising only black and white squares (similar to a chequerboard) is presented to the camera, the brightness of the white parts being, say, x ft. lamberts. Then nx photoelectrons per unit area will pass from the photocathode to the target. These in turn will

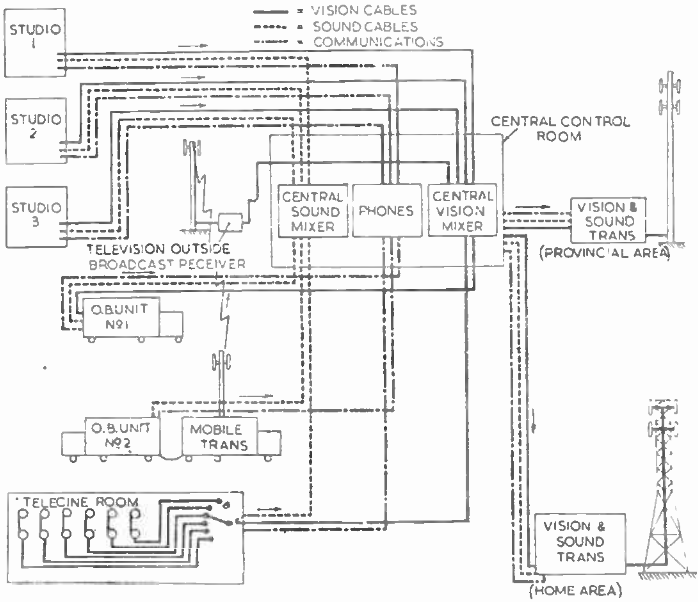


FIG. 20.—THIS FIGURE SHOWS THE RELATIONSHIPS AND GENERAL CONNECTIONS BETWEEN THE CENTRAL CONTROL ROOM CONTRIBUTING AND DISTRIBUTING NETWORKS OF A TELEVISION SYSTEM.

release $m(nx)$ secondary electrons. Of these, some will pass to the mesh, and some will fall back on the target close to their point of release (they cannot travel far because of the close proximity of mesh and target). Those which fall back on the target will give the effect of a local shading signal.

If the illumination of the scene is now increased to, say, $10x$ ft. lamberts, $10m(nx)$ secondary electrons will be released at the target. Because the number released is now larger, the target will tend to rise farther in potential and so draw back more secondaries. That is to say, proportionally more secondary electrons do *not* pass to the mesh, and the charge on the target due to changes in the incident illumination of the photocathode is substantially unaltered. All that happens is a change in the local shading signal.

This local shading signal under some circumstances is very visible; for example, a bright light in the picture is often surrounded by a region of black, although the remainder of the picture should be, say, grey.

The overall effect of this is to give the tube a kind of A.V.C., or compression characteristic, which can be extremely useful. For example, detail in dark regions of an otherwise light picture is very clear. (A dark doorway in a brightly lit house front is a good example.)

On the other hand, it gives the tube a peculiar contrast characteristic, which is at times rather inartistic.

In spite of this defect, its enormous sensitivity and its flexibility give it the versatility required for outside broadcast use.

Improved versions of this tube are now available in which these defects are less objectionable due to the inclusion of an additional mesh electrode which increases the decelerating field for the scanning beam in front of the target.

T. W.

TELEVISION OUTSIDE-BROADCAST UNITS

These units have several functions: they provide a link between outside programme sources, such as sporting events, theatres and open-air displays, and the television centre; they house the equipment necessary to form a mobile control room; they convey the cameras and accessories to the scene of the event. This type of vehicle is built on a 5–7-ton chassis. As picture monitors have to be viewed in partial darkness, the windows are fitted with shutters or blinds and the vehicle is usually fully air-conditioned. A sliding hatch and ladder behind the driver's cab give access to a cat-walk on the roof, which is used for mounting the micro-wave radiators linking with the main studios, and can also be used as a camera position. Cable drums for the camera and power-supply connections are stowed in a well at the rear.

The basic equipment for a maximum of three camera channels, programme and communication links with the main studios is shown schematically in Fig. 21. Each camera channel has a separate control unit, which contains the control and amplifier circuits for the camera and a picture and waveform monitor, under the supervision of a camera-control operator. Communication facilities between the operator and the camera-man are provided by headphones and a microphone. A synchronizing generator furnishes the "synch", blanking, line and field pulses for all the camera-control units.

Synchronized video signals are fed from the camera control units to the vision mixer, situated in the producer's position, which enables him to mix, cut or fade selected inputs and send the picture to line or radio link. An independent picture monitor is also fitted in the producer's position for cueing and general monitoring purposes.

The corresponding sound channels include three microphones, a combined sound amplifier and mixer, a control unit, a monitoring loudspeaker and a power-supply unit.

Functionally, the area can be divided into three parts: the control, production and supply sections, as shown in Fig. 22. The production position, occupied by the producer, is placed behind the control position, which is occupied by the camera-control operator, but at a slightly higher level, to enable the producer to overlook the control operator's monitors. Both positions may be arranged to face either the front or rear of the vehicle. Behind the control position are the racks which accommodate the synchronizing generators, regulated power-supply units for camera controls and a main-voltage control unit. Cameras are stored either under the control units or in the racks.

The "Roving Eye"

The "Roving Eye" is a specialized version of the outside-broadcast television unit, equipped for televising outside events while in motion.

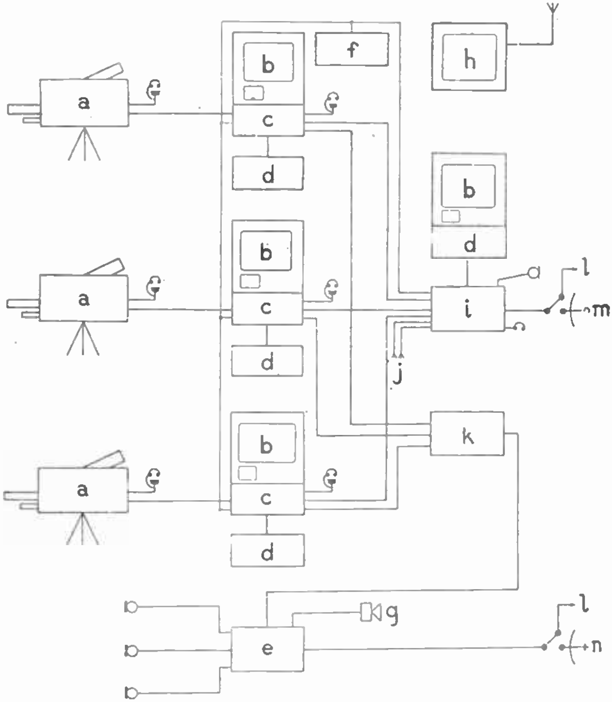


FIG. 21.—FACILITIES DIAGRAM FOR MOBILE TELEVISION CONTROL ROOM.

(a) Cameras; (b) picture and waveform monitors; (c) camera control units; (d) power supply units; (e) amplifier and mixer; (f) sync. generator; (g) loudspeaker; (h) radio check receiver; (i) vision mixer; (j) remote inputs; (k) communications unit; (l) line; (m) vision link; (n) sound link.

The fore part of the vehicle behind the driver's cab is occupied by a camera mounting, which permits the camera to be elevated through a sliding hatch in the roof and panned in any direction. In front of the camera is a position for the commentator, who is equipped with a picture monitor.

Programmes are relayed to the television centre by separate low-power V.H.F. vision and sound transmitters feeding a demountable Yagi aerial on the roof, which is capable of being rotated and orientated towards the television centre. An independent communication type V.H.F. transmitter–receiver with a whip aerial provides for inter-communication with the main station. Camera control, synchronizing, monitoring and power-supply equipment is a simplified edition of that carried in a fully equipped outside-broadcast unit.

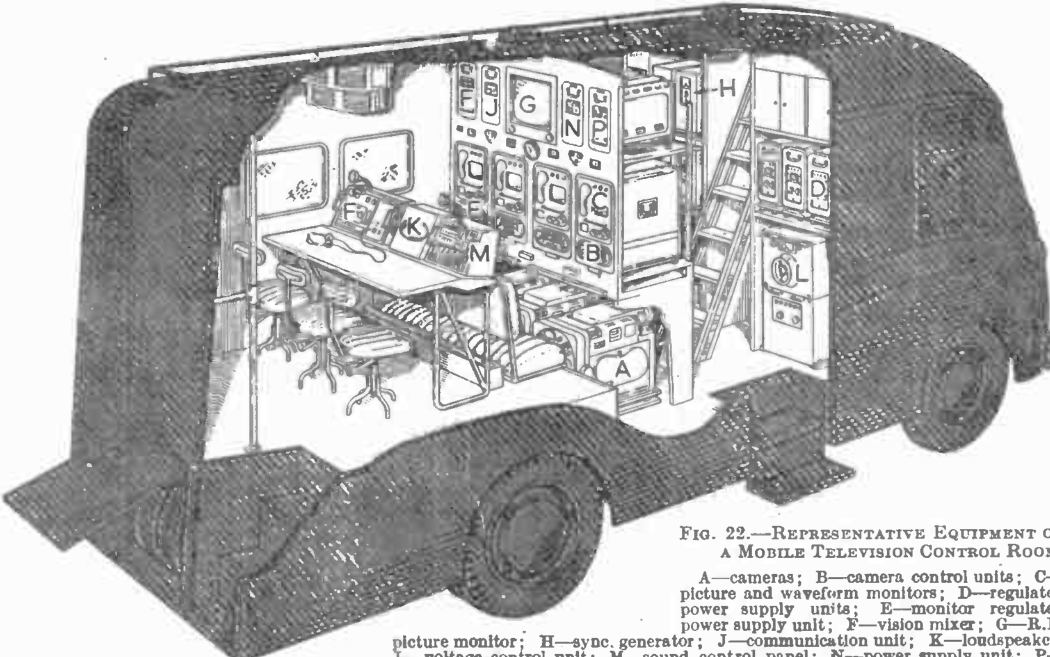


FIG. 22.—REPRESENTATIVE EQUIPMENT OF A MOBILE TELEVISION CONTROL ROOM.

A—cameras; B—camera control units; C—picture and waveform monitors; D—regulated power supply units; E—monitor regulated power supply unit; F—vision mixer; G—R.F. picture monitor; H—sync. generator; J—communication unit; K—loudspeaker; L—voltage control unit; M—sound control panel; N—power supply unit; P—sound amplifier unit. *(Courtesy Marconi's Wireless Telegraph Co. Ltd.)*

CAMERA TUBES

Basic Principle

An early device used for picture pick-up was the Farnsworth Image Dissector. An optical image was focused on to a photo-emissive surface, and the photocurrent from each "picture point" was measured once per frame. The principal limitation of the system was that only the energy originating from a picture point during the examination of that point was available for transmission.

In the modern camera tube the train of signals can still only relate at any one instant to one picture point, but the energy transmitted is that stored electronically for a large fraction of the time between scans. The gain in sensitivity depends on the ratio of "storage time" to "examination time"; in a high-definition system the examination time becomes very small and the gain becomes very large. For example, a tube operating on British Standards (405 lines and 4/3 aspect ratio) may be regarded as forming a picture with 405 × (4/3 × 405) or approximately 200,000 picture points. Hence the examination time is 1/200,000 of frame time. So that if the storage time is frame time, there is a theoretical gain in sensitivity of 200,000.

The modern camera tube, then, consists essentially of a photo-sensitive surface which can meter the light content of a picture, a storage surface on which the picture content is stored as a charge image, and a means for scanning the charged surface with a beam of electrons to neutralize the charge image or restore the surface to some equilibrium condition.

Camera tubes fall conveniently into two groups according as they are scanned with a low- or high-velocity beam. In the first group are the C.P.S. (Cathode Potential Stabilized) Emitron, the Image Orthicon and the Photo-conductive Tube (Vidicon and Staticon). In the second group are the Emitron (now obsolescent) and the modern equivalents of the Super-Emitron (the Midget Super-Emitron, the Photicon, the Eriscope and the Image Iconoscope).

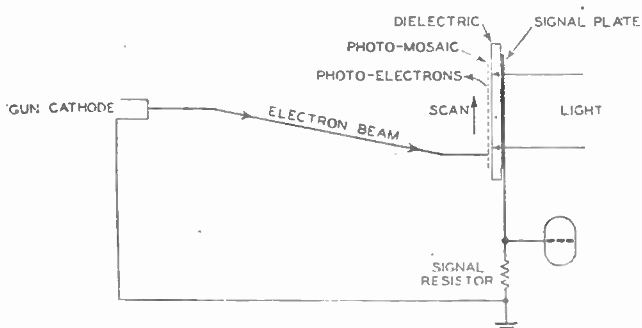


FIG. 23.—THE BASIC PRINCIPLE OF THE C.P.S. EMITRON.

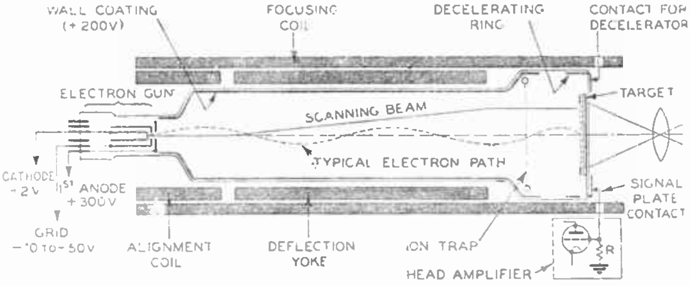


FIG. 24.—THE C.P.S. EMITRON.

The C.P.S. Emitron

In the C.P.S. Emitron the light-sensitive surface and the charge-storage surface are combined in a photo-mosaic (Fig. 23) consisting of an array of photo-emissive elements on the surface of a dielectric, each element being insulated from neighbouring elements. The mosaic surface is scanned with a low-velocity beam; without illumination this surface will stabilize at such a potential that in effect no more electrons can land. This potential is substantially that of the gun cathode. When a picture is imaged on to the mosaic surface, photo-electrons are liberated and withdrawn to the wall anode, thereby building up a positive charge image, the charge on any element being proportional to the light falling on that element between scans. The charge pattern is stored by virtue of the mosaic insulation and also by the capacitance of each element to a common transparent metallic film, the signal plate, formed on the other side of the dielectric sheet. Each element on being scanned is restored towards cathode potential; the charges thereby induced in the signal plate leak through the signal resistor R, and produce a voltage fluctuation on the grid of the first valve of the amplifier. Since there is complete collection of the photo-electrons over the whole frame period, this tube operates with full storage.

The Image Orthicon

A description of the Image Orthicon camera will be found on p. 4-31.

The Midget Super Emitron

The Super Emitron is a tube similar to the high-velocity-scanned Emitron, but having an added image section, again to separate the functions of photo-emission and charge storage. Not having a double-sided target, this tube employs oblique scanning (Fig. 25) by an electro-magnetically focused gun. The photo-electrons are focused on to the storage target by a combination of electrostatic and magnetic fields. The accelerating potential is 500 volts, so that there is secondary emission multiplication at the storage target. Since the storage target, however, is stabilized near the potential of the wall anode, not all these secondaries are collected, but a large proportion fall back on to the storage surface. As these electrons have considerable emission velocity

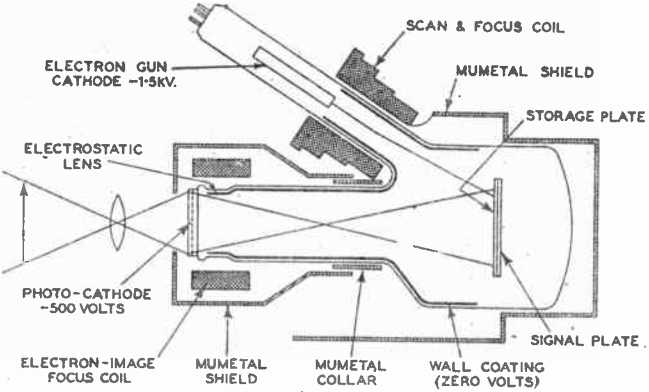


FIG. 25.—THE MIDGET SUPER EMITRON.

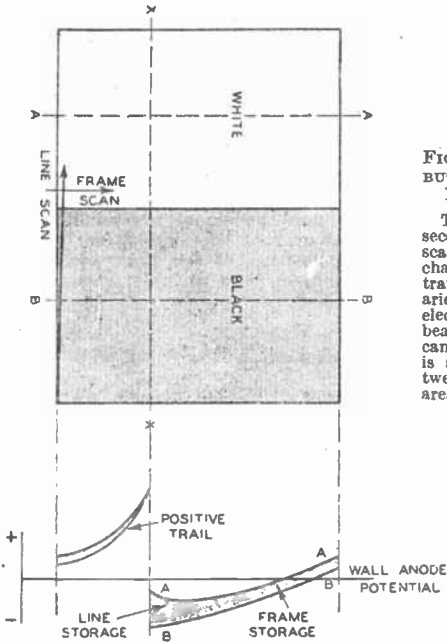


FIG. 26.—POTENTIAL DISTRIBUTION ON THE TARGET OF THE MIDGET SUPER EMITRON.

The scanning beam liberates secondaries, which are widely scattered so that a trail of positive charge is left by the beam. This trail is able to collect the secondaries liberated by the photo-electrons from a given area as the beam approaches that area. It can be seen that the stored signal is a small differential effect between a light area A, A, and a dark area B, B.

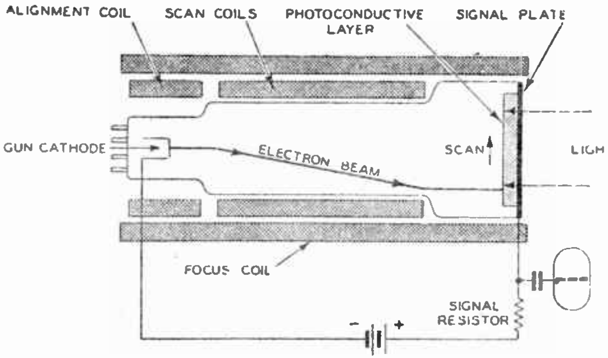


FIG. 27.—THE BASIC PRINCIPLE OF THE PHOTO-CONDUCTIVE TUBE.

they are able to spread, so that only a few secondaries fall back on their place of origin. Thus an area corresponding to a highlight gradually acquires positive charge compared with its surroundings, and this process is continuous during the whole frame interval. In addition to the frame storage effect there is a storage effect which operates for a few line periods only. The scanning beam liberates secondaries which are also widely scattered so that a trail of positive charge is left by the beam (Fig. 26), and this positive trail is able to collect the secondaries liberated by the photo-electrons from a given area as the beam approaches that area. With both these storage effects the storage efficiency of the Midget Super Emitron is still only 6 per cent of that which would be obtained if all the photo secondary electrons could be collected.

The Photo-conductive Tube

The last type of tube currently used for picture pick-up, although of limited application, is the photo-conductive tube. The basic features of this type of tube are shown in Fig. 27. The optical image is projected through a transparent metal signal plate on to a layer of photoconductive material deposited directly on to the signal plate. The scanned surface is again stabilized near cathode potential and the signal plate is held a few volts positive with respect to the cathode. Under illumination, the resistance of the layer is reduced according to the intensity of the illumination, and positive charge leaks through to the scanned side, building up a charge image during the whole frame interval. The charge image is scanned off by a low-velocity beam, and the signal is taken from the signal plate. As the charge image is nearly proportional to the light received between scans, this tube also operates with full storage.

Performance Characteristics

The practical performance of all tubes falls far short of ideal, and a review of some of the following performance characteristics, most of which can only be judged subjectively, will inevitably be largely an account of failings:

Sensitivity and signal/noise ratio	Spurious signal
Definition	Exposure time and lag
Contrast rendering	Geometry, life
Stability	Colour response

A summary of the practical performance of camera tubes will be found in Table 3, but more detailed information is given below.

Sensitivity and Signal/Noise Ratio

There is no absolute method of assessing sensitivity. Table 4—compiled by D. C. Birkinshaw—compares the sensitivities of current tubes in terms of the illumination required on a scene to give a good-quality picture with an equal, arbitrary depth of field. The depth of field is, of course, set by the optical system, the *f* number relating to the size of the optical image required by the sensitive surface of the tube.

TABLE 4.—CAMERA-TUBE SENSITIVITIES

Tube	*Lens Aperture*	*Incident Light (foot candles)*
Emitron	*f* 17	8,000
Midget Super Emitron	*f* 2·3	125
Photicon	*f* 4·2	700
C.P.S. Emitron	*f* 6·3	120
Image Orthicon	*f* 4·6	30

In practice, the incident illumination in studios does not exceed 300 f.c. (foot-candles), and where necessary focal depth is sacrificed. In " outside broadcasts " the illumination may vary from about 2,000 f.c. to below 10 f.c. Indeed, modern sensitive cameras can produce intelligible pictures, although of poor quality, at illuminations below 1 f.c. (The illumination in an artificially lighted domestic room is usually within the range of 3–10 f.c.)

One limit to sensitivity in camera tubes is set by the noise generated, either in the tube itself or in the first stage of amplification. Noise appears in a picture as white flecks which tend to mask picture detail. The length of the flecks depends on the frequency of the disturbances. Low-frequency noise produces long flecks, while high-frequency noise appears as small " snowstorm " flecks. Noise arises from the following causes: random fluctuations in the photo-current about a mean value, known as shot noise, a similar fluctuation in the beam current, a thermal agitation noise in the signal resistor and, lastly, a shot effect in the first valve of the amplifier. The sources of noise in the tube and the resistor noise have energies distributed over the whole of the range of frequencies to be transmitted, while the amplifier noise, by suitable amplifier design, tends to be concentrated in the high-frequency end of the spectrum. In estimating the signal/noise ratio required to give a noise-free picture, the more disturbing effect of flat or " white " noise as against peaked or " coloured " noise, has to be considered.

The Image Orthicon derives its sensitivity from the multiplication in the image section, followed by the noise-free amplification in the electron multiplier. This raises the output current to a level where the contribution of noise by the amplifier is negligible. The resultant signal/noise ratio then is set by the signal/noise ratio before amplification in the multiplier, since both signal and noise are amplified together. It is unfortunate that the only solution yet achieved to the problem of making a practical double-sided target necessarily produces one of low capacitance; hence only a small charge can be stored and the shot noise is relatively high. Moreover, in cathode-potential-stabilized tubes it is not possible to produce more than about 30 per cent modulation of the return beam, which therefore has a large idle component contributing noise. Finally, this noise has the disturbing flat characteristic. The anomalous result is that, although the Image Orthicon is a very sensitive tube, it has a poor optimum signal/noise ratio.

The C.P.S. Emitron, with its much higher storage capacity, is limited by the noise generated in the first stage of the amplifier. Since, also, this noise is peaked, this tube can give a sensibly noise-free picture. Both these tubes will continue to give information, however, when the light is poor and the signal/noise ratio is correspondingly low.

The Midget Super Emitron gives a fairly low-noise picture under good lighting conditions, although in this tube the beam can contribute noise. At low-light levels, however, this type of tube becomes limited by shading.

Definition

It is generally assumed that a camera tube operating on 405 lines with a 4/3 aspect ratio should be capable of resolving $405 \times 4/3 = 530$ vertical black and white lines with 100 per cent modulation to give a balanced picture. This is equivalent to saying that the response should be flat to 2·7 Mc/s. In practice, it is doubtful whether any tube can achieve 50 per cent modulation of 530 lines. Definition in a tube may be limited by the following factors: beam focus and pulling of the beam by highly-charged areas (low-velocity tubes), target leakage, cross talk between scanning and image sections (image-section tubes), and the mosaic structure in the C.P.S. Emitron. High-frequency losses can be corrected by boosting the top response of the amplifier, although such "aperture correction" will reduce the signal/noise ratio.

Contrast Rendering

The receiving tube has a gamma of about 2·2 (gamma = log signal/log light). To reduce the overall gamma of the system to 1·5, which is still high, the signal applied to the tube must have a gamma not exceeding $1{\cdot}5/2{\cdot}2 = 0{\cdot}68$. The C.P.S. Emitron has unity gamma over its normal working range, and must be corrected by a non-linear amplifier. A prerequisite for gamma correction is a true black level as a zero reference, as can be provided during frame return time when the beam is cut off. As long as the light range of the scene is not too great, the C.P.S. Emiton will give a picture with fine tone gradation.

No other tube has a fixed gamma law. In the Image Orthicon the tonal gradation depends on the mesh potential and the light content of the picture. In low-velocity tubes the voltage excursion of the scanned surface must always be controlled to within fairly close limits. If it is too high the beam will be defocused when scanning the more positive areas; if, on the other hand, the potential is too low, then the discharge efficiency of the beam falls, resulting in poor beam modulation and incomplete discharge of the storage surface. The target capacitance of the C.P.S. Emitron is chosen to give the right voltage swing for the charge available from the photo-emission in frame time. The target of the Image Orthicon, however, has limited capacitance, and the voltage excursion is controlled by the mesh potential. When a "highlight" reaches mesh potential, the secondaries from the highlight will no longer be collected by the mesh, and will redistribute themselves on the target, some falling on the darker areas surrounding the highlight, producing a dark "halo" to the highlight (Fig. 28). Thus a highlight is able to preserve its contrast even if the surroundings are so bright that their corresponding charge images are also subject to limiting. The tube will operate under adverse lighting conditions by virtue of this effect, but it does so only at the expense of losing fine tone gradation.

The contrast rendering of the Midget Super Emitron is generally good, but will depend on scene content to a small extent, since in this tube too there is a redistribution of the secondary electrons, which reduces the brightness of a highlight together with the surrounding area, but the effect is less localized than in the Image Orthicon. In practice, this tube can be said to have a gamma varying from 1·0 at low light to 0·3 at highlight levels, so that again this tube is adaptable to adverse lighting conditions. A gamma-correction circuit cannot be used, as there is no fixed black level.

Stability

This is the ability of a camera tube to produce a picture when overloaded with light. As explained above, the Image Orthicon and the Midget Super Emitron are able to give a picture when overloaded, at the expense of picture quality. The C.P.S. Emitron, however, works in a metastable condition. Should a highlight appear sufficiently bright to charge the mosaic within frame time to such a potential that on the average each beam electron will liberate one secondary electron, then it is clear that the beam will not be able to discharge the mosaic. On subsequent scans the mosaic will be charged by loss of secondaries very quickly, the mosaic will stabilize at wall anode potential, and the picture is lost. This tube is very much more stable than previous tubes of a similar type, owing to the structure of its mosaic, which is made by evaporating the base layer of the sensitive surface through the interstices of a metal mesh. The strips of dielectric surface between the elements exert a marked biasing effect on the photo-emission. This has no effect within the normal working range, but a very small highlight will be biased off very quickly no matter how bright it is. Due to this biasing effect, the tube will stand an overload on one-eighth of the area of about twenty times the light to give peak white signal.

It is possible to render the C.P.S. Emitron completely stable by providing a mesh electrode close to the scanned surface; the mesh is held near the critical potential mentioned above, and no element can acquire a potential above the mesh potential. So that even under extreme lighting conditions the mosaic is held to a potential at which the beam can still discharge efficiently.

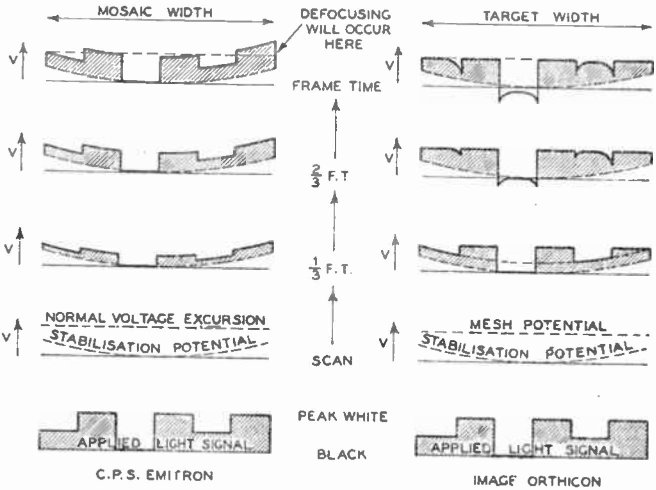


FIG. 28.—BUILD-UP OF THE CHARGE IMAGE.

Spurious Signals

Any non-uniformity in the sensitivity of a photo-surface, or any variation in a secondary emission surface, will produce "background". This type of defect occurs in all tubes. The C.P.S. Emitron has no other source of spurious signal.

The Image Orthicon has the following additional possible sources of background. If the beam electrons have some lateral energy, the amount of which varies according to the amplitude of scan, then there will be a variation in the potential at which the target is stabilized, and hence a variation in the voltage excursion. This will produce variations in signal and gamma (Fig. 28). (A similar variation of target potential may occur in the C.P.S. Emitron, but the voltage swing is unchanged, and only slight defocusing will result.) Secondly, there will be background due to the return beam scanning a small area of the gun anode: this defect, which includes a white spot due to the limiting aperture, is minimized by slightly defocusing the beam. There may be further shading introduced by non-uniform multiplication of the return beam. Finally, there is the possibility of "ghost" images being formed which are the result of the secondary electrons from a saturated "highlight" arriving back on the target in partial focus.

High-velocity tubes have a serious form of shading due to the non-uniform generation and redistribution of the secondaries released by the scanning beam. Increasing the beam current in high-velocity tubes increases the signal, but shading also increases, and at a faster rate. This effect rather than noise sets the illumination lower limit for high-velocity tubes. The modern small versions of the original Super Emitron have reduced shading signals, and in one development of the Photicon

(P.E.S. Photicon) they are still further reduced by providing an auxiliary source of electrons to balance the redistribution of secondaries. Residual shading can be corrected with varying success by injecting signals with sawtooth and parabolic waveforms at both line and frame frequency, but these may require adjustment as the scene content changes.

Exposure Time and Lag

The exposure time depends on the storage mechanism. In the C.P.S. Emitron the exposure time is frame time. In the Midget Super Emitron the exposure time is still frame time, although the storage efficiency is only 6 per cent, as this tube stores inefficiently for the whole frame time. The exposure time in the Image Orthicon will depend on how long it takes for the highlights to start saturating: darker parts of the picture may have an exposure time which is the same as frame time. Frame time exposure ($\frac{1}{25}$ second) will produce blurring on fast-moving objects. This should not be confused with the lag resulting from the incomplete discharge of the storage target, an effect which becomes noticeable when the light is not sufficient to produce the optimum voltage excursion.

This brings us to the fundamentals of the photo-conductive tube. The layer, which is in effect its own storage dielectric (the dark-resistance being high), is very thin, since photo-conductive materials used hitherto produce highly absorbing layers, and with most materials the tube will operate only if the whole thickness of the layer is illuminated. Owing to the thin layer, the capacitance between the storage surface and the signal plate is correspondingly large, so that the tube stores a high charge at a low potential. Consequently the discharge efficiency of the beam is poor, resulting in long lag on moving pictures. The effect is minimized by scanning the smallest area allowed by the beam focus, thus reducing the capacitance and stored charge and allowing the beam to be used more efficiently. In addition to the discharge lag there may be hysteresis of the light-sensitive effect, which also produces lag. In tubes employing materials not subject to the light-absorption limitation, so that thicker layers can be used, hysteresis appears to nullify the expected improvement in lag.

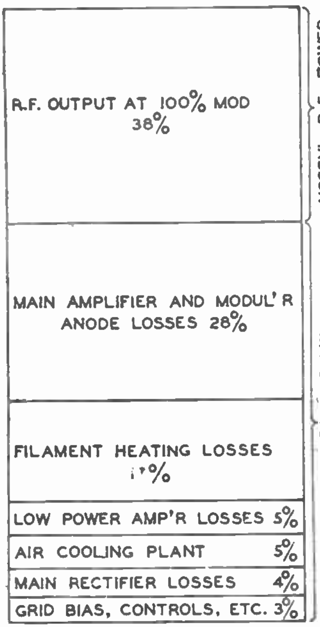
5. TRANSMITTER POWER PLANT

5. TRANSMITTER POWER PLANT

TRANSMITTER POWER REQUIREMENTS

Primary Power Supply

Low-power transmitters for fixed stations, for which the power input does not exceed 5 kW, usually derive energy from a single-phase service. This source may be one phase and neutral of a three-phase public supply or a small single-phase engine-alternator. Transmitters above this rating are invariably designed for operation from a three-phase supply system, since a single-phase load greater than this may seriously unbalance the phase loading, particularly where a small engine alternator is the source of supply.



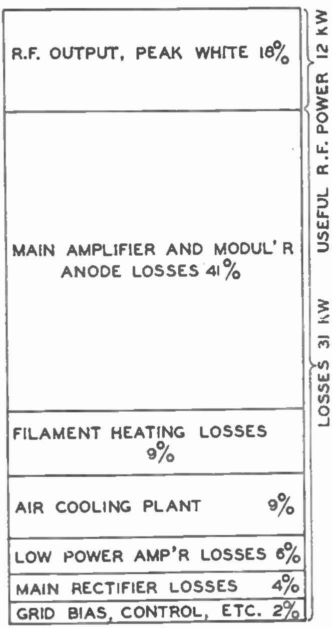


Fig. 1.—(*left*) Efficiency and Losses 150-kW H.F. Broadcasting Transmitter Class B Modulation.

Fig. 2.—(*right*) Efficiency and Losses 12-kW V.H.F. Television Transmitter.

The chief advantages of a three-phase supply are that the weight of copper in supply and distribution cables is only 75 per cent of that required for a single-phase supply; three-phase or six-phase rectification for anode power is more efficient than single-phase rectification, and smoothing filters for the higher rectification frequency can be made much smaller and cheaper; three-phase squirrel-cage motors for cooling plant and other auxiliaries are cheaper and require less maintenance.

Light mobile transmitters are designed to operate from 6-volt or 12-volt vehicle batteries or from 12-volt or 24-volt D.C. aircraft supply. A small rotary transformer, rotary converter, a vibrator power pack or transistor invertor converts the D.C. to A.C. Filaments are heated directly from the battery or D.C. source. Transceivers and "walkie-talkie" equipments usually rely entirely on dry batteries. Military pack sets commonly obtain their supplies from hand-driven or pedal-driven generators or small motor generators.

Transmitter Efficiency

The division of load in a transmitter is illustrated by the typical power-utilization diagrams of Fig. 1 for a high-power broadcast transmitter and Fig. 2 for a television transmitter. In general, the efficiency of conversion to radio frequency power increases as the output power is increased and as the frequency is reduced, and seldom exceeds 60 per cent. The following table shows the average conversion efficiencies to be expected from transmitters of various types and ratings. These figures are typical, and some variation will be found with transmitters of different design.

TABLE 1.—OVERALL EFFICIENCY OF TRANSMITTERS

Type of Transmitter	*R.F. Output Power*	*Frequency Range*	*Overall Efficiency:* $\frac{\text{R.F. power output}}{\text{A.C. power input}}$ (%)
M.F. amplitude-modulated broadcasting	0·25–5·0 kW carrier	525–1605 kc/s	22–34
M.F. amplitude-modulated broadcasting	5–100 kW carrier	525–1605 kc/s	34–60
H.F. amplitude-modulated broadcasting	0·25–5·0 kW carrier	2·5–26·0 Mc/s	20–30
H.F. amplitude-modulated broadcasting	5–100 kW carrier	2·5–26·0 Mc/s	30–43
V.H.F. amplitude-modulated broadcasting	2–20 kW carrier	42–88 Mc/s	26–33
V.H.F. frequency-modulated broadcasting	2–20 kW	85–100 Mc/s	34–48
V.H.F. television	0·5–5·0 kW peak	42–88 Mc/s	16–28
	5–50 kW peak	42–88 Mc/s	28–33

TABLE 1 (*contd.*)

Type of Transmitter	*R.F. Output Power*	*Frequency Range*	*Overall Efficiency:* $\frac{\text{R.F power output}}{\text{A.C. power input}}$ (%)
H.F. telegraph, C.W	0·1–2·5 kW key down	2·0–27·5 Mc/s	17–33
H.F. telegraph, C.W.	2·5–20 kW key down	2·0–27·5 Mc/s	33–40
H.F. double-sideband telephony	2·5–20 kW carrier	2·0–27·5 Mc/s	29–36
V.H.F. telegraph/telephone	10–100 W	100–156 Mc/s	8–12

Total kVA and Power Factor

The determination of the total kVA and overall power factor of a station must take into account the different reactive effects of the various parts of the equipment. The total kVA is calculated by adding separately the power components in kW and the reactive or wattless components in reactive kVA taken by each piece of apparatus, and adding the results vectorially, as shown in Fig. 3.

Power component in kW	$P = \text{kVA} \times \cos\phi$
Reactive component in kVAr	$= P \tan\phi$
Total kVA	$= \sqrt{(\text{total kW})^2 + (\text{total reactive kVA})^2}$
Overall power factor	$= \dfrac{\text{total kW}}{\text{total kVA}}$

It is convenient to tabulate the kVA and power factors for each piece of equipment, and obtain values of tan ϕ for corresponding values of cos ϕ from trigonometrical tables, as in the example Table 2.

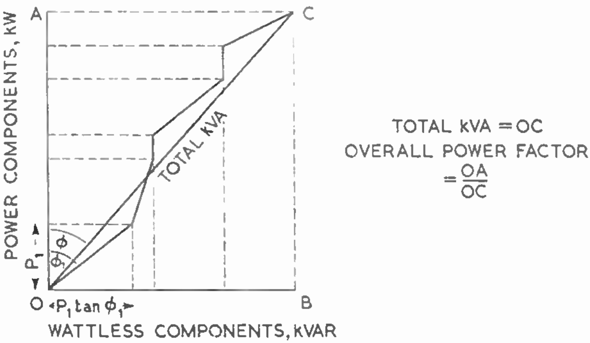


FIG. 3.—TOTAL kVA AND OVERALL POWER FACTOR OF TRANSMITTING EQUIPMENT.

TABLE 2.—kVA AND POWER FACTORS

Equipment	*kVA*	*P.F. (cos φ)*	*kW (P)*	*tan φ*	*kVAr (P tan φ)*
Main H.T. rectifier anodes	25·0	0·95	23·9	0·33	8·25
Main H.T. rectifier filament heating	0·5	1·0	0·5	0	0
Auxiliary H.T. rectifier anodes	0·5	0·87	0·44	0·57	0·29
Grid-bias rectifier anodes	0·75	0·85	0·64	0·62	0
Auxiliary and grid-bias rectifier filament heating	0·2	1·0	0·2	0	0
Amplifier filament heating	3·0	1·0	3·0	0	0
Air cooling fan	1·5	0·85	1·27	0·62	0·93
Controls and auxiliaries	0·3	0·83	0·25	0·67	0·2
Totals	—	—	30·2	—	10·14

$$\text{Total kVA} = \sqrt{30{\cdot}2^2 + 10{\cdot}14^2} = \underline{31{\cdot}9 \text{ kVA}}$$

$$\text{Overall P.F.} = \frac{30{\cdot}2}{31{\cdot}9} = \underline{0{\cdot}947 \text{ or } 94{\cdot}7\%}$$

The load power factor of modern high- and medium-power transmitting equipment is usually of the order of 95 per cent. Consequently, the use of power-factor correcting condensers is seldom justified by the saving in power costs.

Power Distribution Diagrams

One of the first steps in planning the equipment of a high-power transmitting installation is to construct a power-distribution diagram of the kind outlined in Fig. 4. These diagrams are built up from a knowledge of the voltage and current ratings and anode dissipation of the valves, obtained from characteristic curves. From this starting point the kVA or kW ratings of rectifiers, transformers and other conversion plant in each branch are estimated. Table 3 is a useful guide to the average efficiencies and power factors of the types of equipment in common use. Finally, the total kVA input is determined by the method described in the preceding section. The complete diagram furnishes the total power, overall

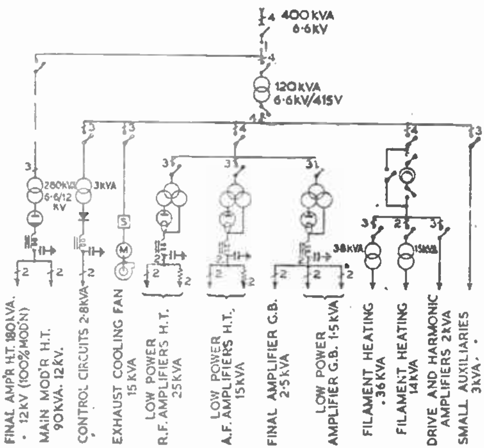


FIG. 4.—POWER DISTRIBUTION DIAGRAM FOR 150-kW M.F. BROADCAST TRANSMITTER.

efficiency and power factor of the installation, and data for the preparation of detailed specifications of conversion plant, cooling equipment, switchgear and cables.

TABLE 3.—AVERAGE EFFICIENCIES AND POWER FACTORS OF CONVERSION AND COOLING PLANT

Equipment	*Efficiency (%)*	*Power Factor*
Three-phase or single-phase alternator		
10–50 kVA	86–90	—
50–100 kVA	90–92·5	—
100–500 kVA	92·5–94·5	—
500–1000 kVA	94·5–95	—
D.C. generator, compounded		
1–5 kW	75–82	—
5–10 kW	82–87	—
10–50 kW	87–91	—
50–100 kW	91–93	—
100–500 kW	93–94	—
Three-phase induction motor/D.C. generator		
1–5 kW	62–70 *	0·75–0·80 *
5–10 kW	70–77	0·80–0·83
10–50 kW	77–82	0·83–0·86
50–100 kW	82–85	0·86–0·87
Three-phase squirrel-cage induction motor		
1–5 B.H.P.	75–85 *	0·81–0·86 *
5–10 B.H.P.	85–88	0·86–0·88
10–50 B.H.P.	88–91	0·88–0·91
50–100 B.H.P.	91–92	0·91–0·92
Three-phase synchronous motor		
50–100 B.H.P.	91–94	1·0–0·9 †
100–500 B.H.P.	94–96	1·0–0·9
D.C. shunt-wound motor		
Fractional B.H.P.	65–75	—
1–5 B.H.P.	76–83	—
5–10 B.H.P.	83–86	—
10–50 B.H.P.	86–90	—
50–100 B.H.P.	90–93	—
Rotary transformer		
50–250 W	50–60	—
Three-phase or single-phase transformer		
Fractional kVA	80–93	—
1–5 kVA	94–96·5	—
5–10 kVA	96·5–97·2	—
10–50 kVA	97·2–98·2	—
50–500 kVA	98–98·4	—
Cold-cathode steel-bulb mercury-arc rectifier with auxiliaries		
200–1000 kW, 2–12 kV	93–95	0·96

TABLE 3 (*contd.*)

Equipment	*Efficiency* (%)	*Power Factor*
Cold-cathode glass-bulb mercury-arc rectifier with auxiliaries		
25 kW, 100 V	78	0·92
50 kW, 250 V	87	0·92
200 kW, 600 V	93	0·93
Hot-cathode mercury-vapour rectifier with auxiliaries		
50–200 kW, 2–12 kV	94–95	0·95
Metal rectifier with transformer		
25–250 W, single-phase full wave .	50–65	0·7–0·85
250–2500 W three-phase full wave .	65–83	0·8–0·9
Vibrator converter	50–60	—
Motor-driven centrifugal fan		
0·05–0·1 air H.P.	7–12 ‡	0·75–0·78
0·1–0·25 air H.P.	12–20	0·78–0·8
0·25–0·5 air H.P.	20–35	0·8–0·81
0·5–1·0 air H.P.	35–37	0·81–0·82
1·0–2·0 air H.P.	57–65	0·82–0·84

* Efficiency and power factor improve with increase in synchronous speed.

† Power factor varies from lag to lead with increase in excitation.

‡ Overall efficiency, air H.P. divided by input power to motor.

POWER TRANSFORMATION

Economics of Transformation

The practice of duplicating transformers ensures continuity of supply if a defect occurs in one of them. A saving in first cost can be made by rating each transformer for two-thirds of the station load with all transmitters operating at full power. The iron and copper losses of two similar transformers sharing a load in parallel are less at full load than the losses of a single transformer of equivalent rating. British Standard Specification B.S. 171 : 1936 stipulates that a transformer shall be capable of handling 50 per cent overload for 5 minutes following normal full loading, or for a longer period after light loading. Correspondingly longer periods are permissible with a lighter overload. Thus, if one transformer should fail, the other can take over the full load for a short period or maintain continuous service at two-thirds of the normal full load.

Transformers will operate satisfactorily in parallel on the same supply if the secondary voltages are the same and in phase at all loads. This means that they must be alike in (*a*) voltage ratio, (*b*) polarity, (*c*) phase rotation, (*d*) impedance and (*e*) resistance/reactance ratio. If the impedances differ the load will divide between the transformers in the

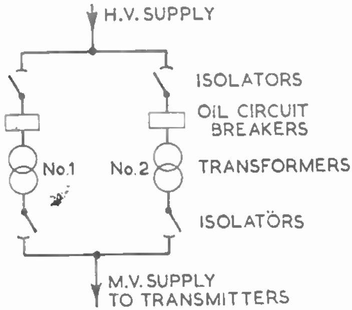


FIG. 5.—DUPLICATION OF SUPPLY TRANSFORMERS.

inverse ratio of their impedances, the terminal voltages will not be the same and circulating currents will flow. Means must, of course, be provided for isolating either transformer in the event of a fault or for separate operation. Fig. 5 illustrates the basic requirements.

Transformer losses and heating increase with the loading. The load that can be carried is, therefore, limited by the temperature rise above the surrounding air. As there is a thermal lag, the temperature rise corresponds to the average load, so that transformers for transmitters in which the load varies under modulation or keying may be rated below the peak demand. For example, a supply transformer for a Class B modulated broadcast transmitter, rated for a continuous load equal to that at average modulation, which is about 30 per cent of the peak, will comfortably deal with modulation peaks. If the transformer complies with the British Standard Specification, it will also be capable of dealing with an overload of 20–30 per cent for 30 minutes, and it will not, therefore, overheat excessively on a sustained full-modulation test of the transmitter for that period of time.

With high-power broadcast transmitters a saving in transforming equipment is made possible by taking power for the main rectifier transformer, which forms the bulk of the load, direct from three-phase, high-voltage lines, and transforming down to a medium voltage only for grid bias, filament heating and auxiliary supplies, as shown schematically in Fig. 6.

Uses of Auto-transformers

Auto-transformers can be used economically to step up or step down the supply voltage where it differs from that for which the transmitter is designed. Since the primary turns are included in the secondary winding, the volume of copper in the windings is less than in an equivalent double-wound transformer. For a given transformation ratio k, the saving in copper compared with a double-wound transformer is

$$1/k \times 100 \text{ per cent}$$

The saving is appreciable with low transformation ratios: for example, with a 2 : 1 ratio it is 50 per cent; with a 10 : 1 ratio it is only 10 per cent.

Three-phase auto-transformers are usually Y-connected, as in Fig. 7. This permits the insulation to be reduced at the neutral point when it is joined to the neutral of a four-wire supply. The neutral is usually brought into the station to obtain the phase-to-neutral voltage, (which is $1/\sqrt{3}$ of the line voltage) for auxiliary apparatus and control circuits

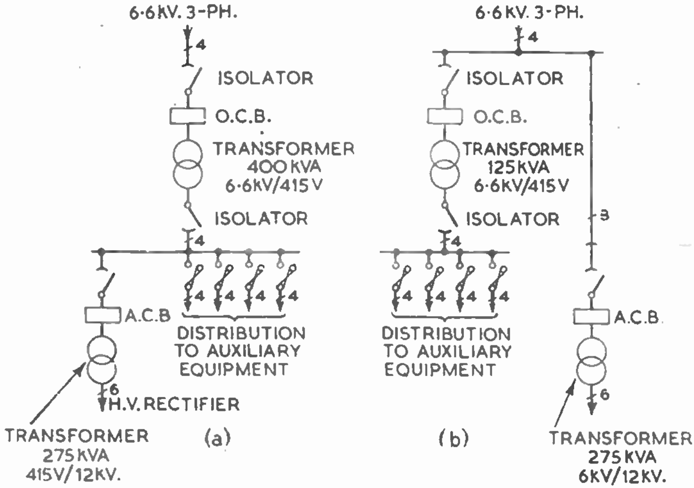


FIG. 6.—TWO TRANSFORMATION SCHEMES FOR HIGH-POWER TRANSMITTING EQUIPMENT.

(*a*) Medium-voltage distribution. (*b*) High-voltage supply to main rectifier, medium-voltage distribution to auxiliaries.

Where a four-wire supply is available, the neutral point of the transformer must be connected to the neutral of the supply, for the following reasons :

(1) It permits the use of cables and apparatus insulated only for the line-to-neutral voltage.

(2) If the neutral were left floating, out-of-balance conditions or the occurrence of a fault between a line and neutral might lead to excessive rise of voltage on the neutral conductor (see Fig. 8). Under such conditions the insulation of connected apparatus would have to be good for the line-to-line voltage.

(3) A floating neutral would require to be switched in the same way as the line conductors to ensure that no voltage is present on connected appliances when the three-phase switch is open.

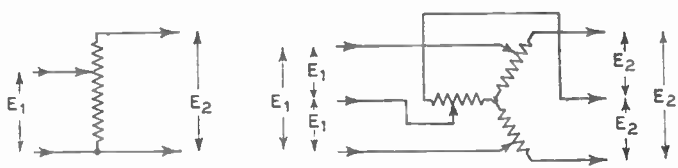


FIG. 7.—ONE-PHASE AND THREE-PHASE AUTO-TRANSFORMERS.

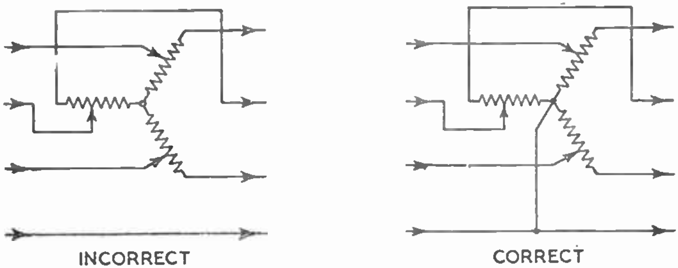


FIG. 8.—THREE-PHASE AUTO-TRANSFORMER—THREE-PHASE, FOUR-WIRE SUPPLY.

An auto-transformer should not be used under the following conditions :

(1) On a three-phase, three-wire supply where it is required to earth the neutral (see Fig. 9). Supply authorities usually prohibit earthing at any point in the distribution system other than at the source, as the introduction of a second earth connection would unbalance the earth-leakage protection circuits.

(2) For high-ratio step-down of voltage, when there is a danger of high voltages appearing in the low-voltage circuit under fault conditions.

Transformer Rooms

Precautions must be taken against fire risks in the housing of oil-filled supply transformers, modulation transformers and smoothing-chokes. Such equipment must be accommodated separately from other transmitting equipment. If the transformer chamber is built into the main building, entry must only be possible from outside, and the door must be normally kept locked, the key being in the possession of an authorized person. To prevent the possibility of burning oil escaping, there must be no direct connection between the chamber and other

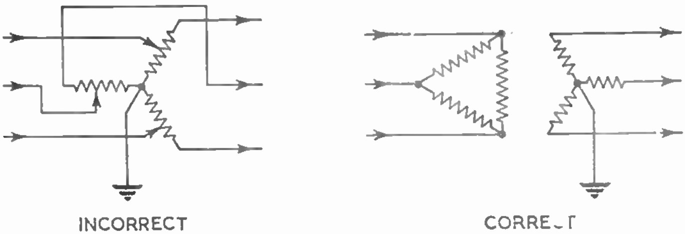


FIG. 9.—THREE-PHASE AUTO-TRANSFORMER—THREE-WIRE SUPPLY, NEUTRAL REQUIRED.

rooms, nor any interconnecting pipes conveying air, gas or water. Cable connections must be sealed where they pass through the wall.

The chamber must be constructed of non-combustible materials and ventilated to the outer air independently of the rest of the building. Ample access must be provided for safe access to plant and for removing faulty units without disturbing other equipment. The accessories should include approved fire extinguishers or an automatic spraying system.

Oil-drainage ducts should slope towards a sump in the chamber, which must in turn discharge outside the building. The sump should preferably not connect to the public drainage system, because of the risk of water " backing up " the drain and flooding the chamber during heavy rain.

VOLTAGE AND FREQUENCY VARIATIONS OF SUPPLY

Effects of Voltage Variation

Voltage variation at the supply terminals results from the inherent regulation of the supply network associated with :

(*a*) changing loads on the supply lines, independent of the transmitter load, or

(*b*) the fluctuating load of the transmitter itself.

Variation caused by (*a*) is slow and occasional, or may be characterized by a sudden rise or fall of voltage. This can be corrected by an automatic induction regulator having a throughput rating sufficient to cover those parts of the equipment affected by the variation. Induction regulators are capable of reducing variations up to ± 10 per cent to within ± 1 per cent, but they are slow acting and require from 60 to 120 seconds to correct the full range of variation.

Variations allied with the operation of the transmitter result chiefly from the load on the main rectifier fluctuating with signalling or modulation, the auxiliary load remaining substantially constant. Simple telegraph transmitters are designed for on/off keying, which in effect switches on and off the anode current of the final amplifier. The load swings rapidly at the signalling speed between a maximum on " mark " and a minimum on " space " at periodic intervals of time ranging from ½ to 6 cycles of a 50-cycle supply, depending on the signalling speed.

The load of amplitude-modulated telephone, broadcast and television transmitters fluctuates irregularly, according to the degree and frequency of modulation. At all modulation frequencies above the rectified A.C. ripple frequency the smoothing filter of the main rectifier is effective in smoothing out the peaks, so that the terminal voltage assumes a steady mean value corresponding to the average load. Heavy peaks of modulation at the lowest frequencies, however, may be reflected back to the terminals. The load of frequency modulated transmitters, on the other hand, is constant, since the radio frequency and not the amplitude varies under modulation.

The automatic induction regulator is too slow in action to correct for fluctuations of this kind. Two alternative methods are available :

(1) Load equalization.

(2) Quick-acting flux-regulating transformers or thermionic voltage stabilizers.

Load equalization is applicable only to communication transmitters keyed by the on/off method. This method makes use of an absorption load incorporated in the transmitter, consisting of a D.C. amplifier controlled by keying, which completes a circuit across the H.T. supply on "space". This circuit contains a resistance capable of dissipating an amount of power equal to the difference between the "mark" and "space" loads.

Quick-acting flux-regulating transformers are connected in the supply leads to apparatus which requires stabilizing, in the same way as a transformer, or may replace a transformer to serve the dual purpose of voltage regulation and transformation. These regulators are capable of correcting voltage variations within a fraction of one cycle of the supply frequency, and of maintaining the voltage constant to within ± 1 per cent or better with variations of ± 10 per cent or more.

Alternatively, thermionic voltage stabilizers may be fitted in the output circuits of small rectifiers to control the D.C. voltage. They are practically instantaneous in action, and will control the voltage to within $\pm 0{\cdot}1$ per cent, but they have the disadvantage of a relatively high inherent voltage drop.

Voltage variations at the input of transformers and static rectifiers are repeated at the output. A.C. motor-generators and A.C. motor-driven equipment, such as valve-cooling fans and water-circulating pumps are, for all practical purposes, insensitive to voltage variations. The speed of induction motors and synchronous motors is determined by the frequency, and remains nearly constant over a wide voltage range. The R.F. stability of crystal oscillators is very little affected by changes in the anode and filament voltages of the associated valves, but the frequency of valve oscillators can vary widely with voltage changes.

In order to meet certain performance specifications, it is often necessary to stabilize the whole A.C. input to broadcasting and television transmitters by installing an induction voltage regulator, but more usually it is sufficient to stabilize only the auxiliary equipment indicated in Table 4.

TABLE 4.—AUXILIARY EQUIPMENT REQUIRING STABILIZATION

Equipment	*Type of Voltage Regulator*
Filament transformers . .	Induction regulator or flux regulating transformer
Filament rectifiers . . .	Induction regulator or flux regulating transformer
Grid bias rectifiers . . .	Thermionic stabilizer
D.C. control motors for timed operations	Thermionic stabilizer

Effects of Frequency Variation

All inductive apparatus is, to some extent, affected by frequency variation. Assuming the applied voltage to remain constant, the current in an inductive circuit varies inversely as the frequency and in a capacitive circuit directly as the frequency. Induction regulators and moving-coil regulators are responsive to frequency change, which degrades the voltage constancy. The speed and output voltage of A.C. motor-generators, rotary transformers and converters varies directly as the frequency. Rectifiers are substantially unaffected by normal variations of frequency. The R.F. constancy of crystal oscillators and valve oscillators is not impaired, provided the valve supplies are voltage stabilized.

Voltage and Frequency Tolerances

Under normal conditions voltage variations at the terminals of a public supply system are not allowed to exceed ±6 per cent of the declared voltage. The maximum allowable frequency variation on time-controlled supply systems is ±1 per cent. Performance specifications for transmitters vary so widely for different services, that voltage and frequency tolerances cannot be closely defined to meet every case. The tolerances given in Table 5 are useful as an approximate guide.

TABLE 5.—VOLTAGE AND FREQUENCY TOLERANCES

Apparatus	*Maximum Permissible Voltage Variation* (%)	*Maximum Permissible Frequency Variation* (%)
Main H.T. and grid-bias rectifiers	± 6	±1
H.T., D.C. motor generators	± 6	±1
Filament heating transformers	± 2½	±1
Filament heating rectifiers	± 2½	±1
Crystal oscillators (frequency constancy 1 part in 10^5)	± 6	±1
Compensated valve oscillators (frequency constancy 4 parts in 10^5)	± 2	±1
Motor-driven air fans, water pumps and controls	± 6	±1
Automatic voltage regulators	±10	±1

VOLTAGE REGULATORS

Induction Voltage Regulators

The induction regulator is a form of transformer, in which one winding can be moved with respect to the other to produce a variable secondary voltage. The primary or movable winding is wound on a rotor and connected across the line. The secondary is wound on a stator and connected in series with one line.

Referring to Fig. 10, current in the primary coil P produces a flux which links with the secondary S in a varying degree as the rotor is turned, and induces a voltage in it, which either adds to or subtracts

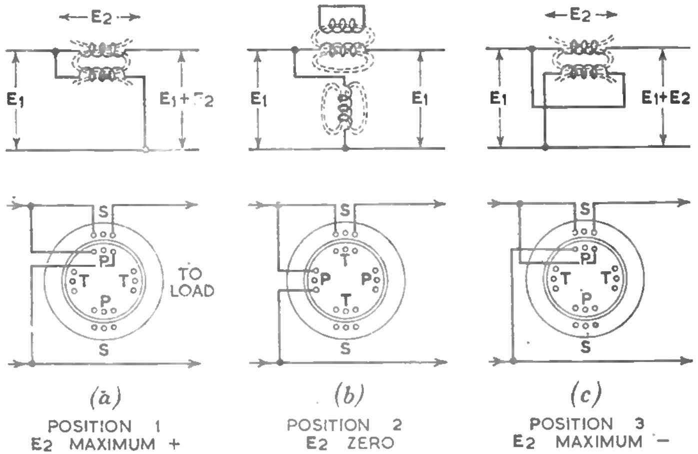


FIG. 10.—SCHEMATIC DIAGRAM OF SINGLE-PHASE INDUCTION REGULATOR.

from the line voltage. In position 1, Fig. 10 (*a*), most of the flux links with the secondary, and the induced voltage E_2 is a maximum. When the rotor is turned through 90° to position 2, Fig. 10 (*b*), none of the flux links with the secondary, and the induced voltage is zero. Rotation through a further 90° to position 3, Fig. 10 (*c*), produces another voltage maximum, but of reversed phase. Thus, rotation through 180° varies the line voltage from a maximum value $E_1 + E_2$ to a minimum $E_1 - E_2$. The correct degree of rotation compensates for the rise or fall of line voltage.

In the mid position the leakage flux and the reactance drop in the secondary would, in the ordinary way, be excessive. To overcome this, a short-circuited tertiary winding T is wound on the rotor. The heavy current produced in the tertiary winding by linkage with the secondary largely annuls the flux producing it and reduces the reactance drop.

When designed for automatic regulation, a voltage-discriminating relay, fitted with two pairs of contacts, controls the turning of the rotor shaft. Any rise or fall of line voltage causes one pair of contacts to close and starts a motor coupled by a worm drive to the rotor shaft, turning it either in a forward or reverse direction. Rotation continues until a position is reached which restores the line voltage to its normal value. At this point the contacts open and the motor is stopped.

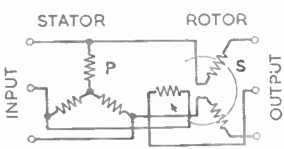


FIG. 11.—CONNECTIONS OF THREE-PHASE INDUCTION REGULATOR.

The principle of the three-phase induction regulator is basically the same as for the single-phase regulator. Fig. 11 shows the simplified connections. The stator is wound with three windings connected in star formation across the lines, and the rotor with three windings

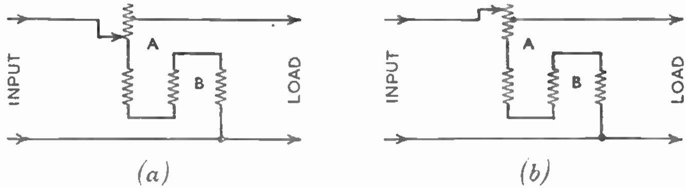


FIG. 12.—SINGLE-PHASE AUTO-TRANSFORMER VOLTAGE REGULATOR.

connected separately in series with the lines. Rotation through 180° covers the full range from maximum buck to maximum boost.

Auto-transformer Voltage Regulators

Figs. 12 and 13 show two adaptations of the auto-transformer principle to automatic voltage regulation. In Fig. 12, A is the variable portion of the winding which covers the range of voltage correction required. This winding is tapped at a large number of points along its length to give very small voltage increments. The tappings are brought out to a group of contacts, to which connection is made by a sliding brush, actuated by a reversing capacitor motor. B is the fixed portion of the winding, which is multi-layered on the core in the conventional manner. Fig. 12 (*a*) and (*b*) show respectively the adjustments for maximum and minimum transformation ratios to give maximum boost and buck.

Fig. 13 is a modified arrangement incorporating a small booster transformer, which is suitable for higher throughput powers. The voltage applied to the boosting primary P is varied by two brushes moving in opposite directions over contacts which select tappings on the main winding. With this arrangement the current in the brushes is determined by the turns ratio of the booster transformer, and is only a fraction of the load current. The voltage induced in the secondary winding is added to or subtracted from the line voltage according to the positions of the brushes, as shown in Fig. 13 (*a*) and (*b*), and confines the output voltage within the limits required.

The control consists of a moving-coil voltmeter movement fitted with two pairs of contacts and fed by D.C. obtained from a metal rectifier connected across the A.C. output terminals. An important advantage

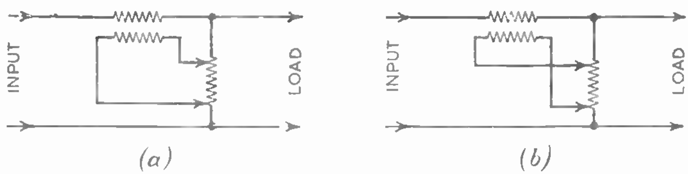


FIG. 13.—SINGLE-PHASE AUTO-TRANSFORMER VOLTAGE REGULATOR WITH BOOSTER TRANSFORMER.

of this method is that the performance is unaffected by frequency variation. If the voltage rises or falls beyond specified limits, the appropriate contacts close a contactor circuit, which actuates the control motor. Limit switches are fitted to prevent over-running of the brush mechanism. To eliminate contact chatter and hunting of the regulator, compensating resistances are incorporated in the relay circuit to hold the contacts closed until restoration of the voltage is complete.

Thermionic Voltage Regulators

Thermionic voltage regulators, when incorporated in the D.C. output circuits of rectifiers, correct for both variations of load and A.C. supply voltage. There are many circuit variations, but the basic feature common to all is a variable impedance regulating valve connected in series with the line. A typical circuit is shown in Fig. 14.

V_1 is a low impedance regulating triode, capable of passing full-load current with a low voltage drop. V_2 is a control pentode fed from the line through a suitable voltage-reducing resistance R_1. When V_2 passes current, the voltage drop in the neon lamp N in series with the cathode lead, biases the cathode positively with respect to the negative line. A tapping on a voltage divider R_2 connected across the output terminals biases the control grid to a value less positive than the cathode, and therefore negatively with respect to it. A second tapping on the divider provides a suitable screen grid voltage.

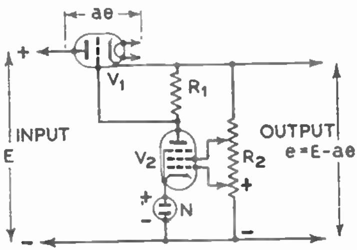


FIG. 14.—D.C. THERMIONIC VOLTAGE STABILIZER.

The principle of operation is as follows: A rise in line voltage makes the control grid of V_2 more positive and causes the valve to pass a larger current. The increased voltage drop in R_1 reduces the anode voltage and biases the grid of the regulating valve V_1 more negatively. This action is augmented by the fact that the resistance of a neon lamp falls with rise of voltage. As a consequence the current increases to a greater extent than it would with an ohmic resistance of equal value. Provided R_1 and R_2 are suitably adjusted, the voltage change across valve V_1 largely compensates for the rise of line voltage. If the line voltage falls, the reverse action occurs.

In practice 100 per cent regulation is never attained in any voltage regulator. If e is the residual voltage variation applied to the control circuit and a is the voltage gain of the circuit, the correction voltage opposed to the line voltage variation E will be ae, so that:

$$\begin{aligned} e &= E - ae, \\ &= E/(1 + a) \\ &\simeq E/a \end{aligned}$$

since a is always much greater than unity.

Hence variations in the rectified output voltage, together with any residual ripple voltage after smoothing are reduced to $1/a$ of the original values. Gain factors of the order of 500 are commonly obtainable with well-designed circuits.

ENGINE GENERATORS

Mains Supply versus Generating Plant

Power for a transmitting station is most economically obtained from a public supply system. The chief arguments in favour of purchasing power as compared with generating at the station are :

(1) It can be generated more efficiently and cheaply in bulk.
(2) There are little or no capital and renewal costs or depreciation charges.
(3) It eliminates the cost of shift and maintenance staff for generating plant.
(4) Failure of supply is less probable and usually of shorter duration.
(5) Voltage and frequency stability are better.
(6) Additional power for future expansion of services is more quickly and cheaply made available.

There are, however, many sites where conditions render generating plant essential. Engine generators are installed :

(1) At remote sites where it is either uneconomical or impracticable to extend power lines from a public supply system.
(2) As a stand-by to the mains supply.
(3) Where the public supply is unreliable, e.g., a small hydro-electric source where the water supply may fail in the dry season.

Diesel alternators are the most suitable for all but very low power requirements. They have a high thermal efficiency, are quick-starting and compact. Although Diesel alternators are made to give outputs as low as 1 kVA, for powers below 3 kVA petrol-engine alternators are a serious rival.

Anti-vibration Precautions

The performance of thermionic apparatus may suffer from the vibration produced by engine alternators. Cyclic impulsing of the engine is conducted through a solid foundation and transmitted some distance if the sub-soil is of a rocky nature. This trouble is sometimes accentuated by resonances occurring in parts of the building structure. Vibration must be reduced to a minimum, particularly when receivers are installed in the same building, by isolating the foundation from the surrounding soil.

Fig. 15 shows the basic method of isolating a foundation. After excavating the soil, a concrete raft is laid, and the sides of the excavation are concreted to prevent water seeping through from the sub-soil. On this raft a 3-in. layer of resilient material, such as "Coresil" or "Mascolite" is laid. The main foundation block is then formed in

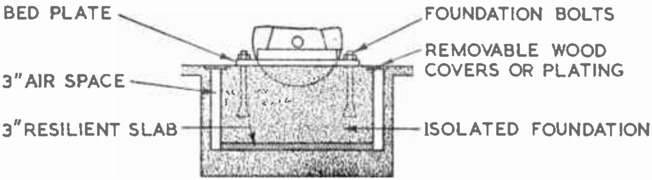


FIG. 15.—METHOD OF ISOLATING FOUNDATION BLOCK.

shuttering on the isolating material, leaving a 3-in. air gap all round, between the block and the sides of the excavation. At floor level the air gap is closed by laying loose wood covers or steel plating.

An alternative method, which is suitable for loads up to about 12 tons and eliminates the necessity of preparing a special foundation, is to erect the machine on anti-vibration mountings. These mountings are made of steel and a special grade of rubber, welded by a rubber-to-metal process. They are bolted to the bedplate and the concrete foundation and spaced at suitable intervals, so as to share the load evenly and deflect equally.

Yet another form of mounting depends for its resiliency on a stiff helical spring, damped for compression and elongation by pads of resilient material. The spring and pads are housed in a circular mount equipped with fixing bolts and a tension-adjusting nut.

Stabilization of Engine Alternators

The voltage constancy of an engine alternator is governed by :

(1) The voltage regulation of the alternator.
(2) The speed/load variation of the engine.
(3) The cyclic irregularity of the engine.

Good voltage regulation demands special attention when the source of supply is an engine alternator. The percentage load variation on a small generator is likely to be heavier than is normally encountered on a public supply system. Voltage drop due to regulation of the alternator and the fall of voltage with speed both increase with the load, and their effects are additive.

Voltage and frequency vary directly as the speed of the alternator. The extent to which the engine speed varies with changes of load depends on the response time of the engine governor. The governor will correct the speed with sustained load changes, but its response is too slow to follow rapid changes due to signalling. Inertia of the machine, reinforced by the mass of the flywheel, tends to curb sudden changes, but cannot correct for a sustained change. Slow-acting automatic voltage regulators only partially compensate the voltage for rapid variations of load. Both the regulator and the governor take up some intermediate position corresponding to the average value of a swinging load. Thermionic voltage regulators exert effective control with both rapid and sustained changes, but since they have no control over the speed, transitory changes of frequency will always occur.

Automatic voltage regulators for stabilizing the voltage of alternators are designed to control the field excitation of the alternator or its exciter. They are classified according to the principles of operation as :

(a) Rheostatic.
(b) Vibrating contact.
(c) Thermionic.

Rheostatic Voltage Regulators

One form of this type is shown in Fig. 16. The excitation control rheostat is wound with resistance wire, from which tappings are brought out at intervals along the winding to a group of contacts assembled closely together. The control circuit consists of a transformer T, and a bridge-connected rectifier B, fed from one phase of the alternator output, which energizes an electromagnet E. When the alternator voltage rises or falls, the armature of the electromagnet causes a contact arm A to move in either direction from its normal position across the contact studs and vary the amount of resistance in the exciter field circuit. Regulation of the excitation of the alternator restores the output voltage to its nominal value, and the regulating arm then moves back automatically to its normal position. The relatively short travel of the arm ensures rapid correction of the voltage variation.

Transformer T_2 is introduced to damp the system and prevent hunting when the excitation voltage changes too rapidly. Its primary is connected across the output from the exciter. Under steady conditions no e.m.f. appears in the secondary, but when a sharp variation occurs, a damping pulse of e.m.f. is induced in the secondary in opposition to the original surge.

A variation of the rheostat type is the carbon pile regulator, which has a quicker response than the moving-contact type. A bank of thin carbon discs, assembled on an insulated frame, takes the place of a wire-wound element, but the control circuit is very similar. The disc assembly has the property of varying its resistance when subjected to pressure by the movement of a lever attached to the electromagnet.

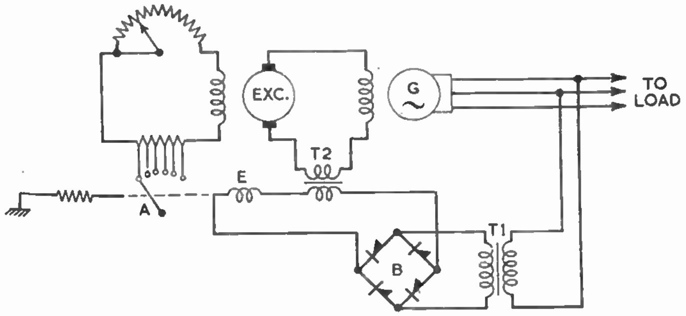


Fig. 16.—Automatic Rheostatic Voltage Regulator for Control of Alternator.

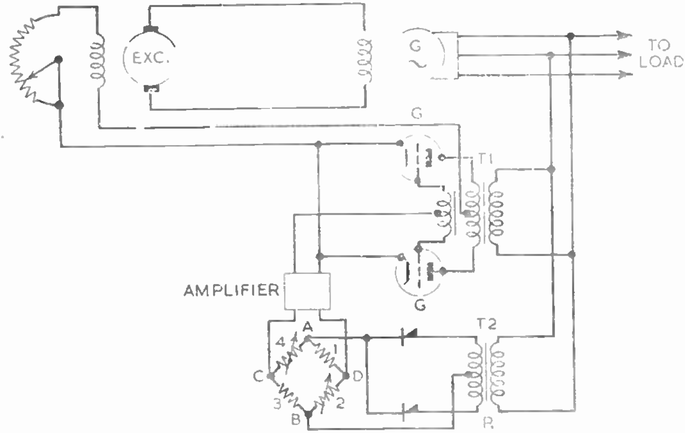


FIG. 17.—THERMIONIC VOLTAGE REGULATOR FOR CONTROL OF ALTERNATOR.

In this type the total resistance is determined mainly by the number of contact surfaces and the pressure exerted between them rather than by the resistivity of the material.

Vibrating Contact Regulators

This type has largely given place to rheostatic and thermionic control. It depends for its action on the rapid opening and closing of a short-circuit path across a fixed regulating resistance in series with the exciter field by a vibrating relay. The rate at which the short-circuit is made and broken and the duration of the short-circuit is controlled by the deviation of the alternator voltage, and determines the average excitation current of the alternator.

Thermionic Regulators

These regulators are highly sensitive and quick-acting, as they have no moving parts. Fig. 17 illustrates the principle of the circuit. The exciter field current is derived from two gas-filled rectifying valves G connected in a full-wave rectifying circuit and a transformer T fed from one phase of the line. A second rectifier R and transformer T2, also fed from the line, supply the D.C. control voltage. The rectified voltage is applied across diagonally opposite points A and B of a resistance bridge. Elements 1 and 3 in opposite arms are fixed resistors with a low-temperature coefficient of resistance. The complementary elements 2 and 4 are special metal-filament lamps having a high positive temperature coefficient; that is, their resistance increases with increase of current. All four elements are adjusted to have equal resistance values when the supply voltage is normal. In this condition the bridge is balanced and the voltage across C and D is zero. When the line voltage rises or falls, the change of current in the bridge circuit upsets the balance, and a voltage exists across C and D, the polarity depending on whether the voltage rises or falls. The out-of-balance voltage is amplified and applied as a negative bias to the grids of the gas-filled rectifiers when the line voltage is high, and as a positive bias when the line voltage is low. The corresponding reduction or increase in the excitation current restores the line voltage approximately to its correct value.

Cyclic Irregularity

The cyclic rise and fall in speed associated with the non-uniform torque of a reciprocating engine is reflected in the electrical performance of the alternator. The extent to which the speed deviates from the nominal uniform value is defined by the *coefficient of cyclic irregularity*. This is the ratio of the maximum change in angular speed ($\omega_{max.} - \omega_{min.}$) during one engine cycle to the mean speed ω. Its effect on the alternator is defined by the *angular deviation*. This is the displacement in electrical degrees of the rotor in either direction from the position it would occupy if the rotation were uniform. Although the use of a heavy flywheel largely curbs the disturbance, there is always some residual fluctuation of voltage and frequency. Since both voltage and frequency vary directly as the speed, and the speed fluctuates alternately above and below the mean value, it follows that, if k is the coefficient of cyclic irregularity, the percentage voltage or frequency change from the mean value is equal to $100k/2$. In terms of the angular deviation, it can be shown that

$$v' = f' = 1{\cdot}75\ \theta_d f_e/f_o \text{ per cent}$$

where k = coefficient of cyclic irregularity;
θ_d = angular deviation in electrical degrees;
f_e = frequency of cyclic impulsing
(= no. of expansion strokes per rev. × revs./sec.);
f_o = mean frequency of alternator.

It is noteworthy that, for a given supply frequency, the cyclic irregularity is reduced as the speed is increased. Since the number of pairs of alternator poles are reduced at high speeds, there is also a corresponding reduction in the angular deviation.

Motors and motor-generators are little affected by cyclic irregularities of the supply. The inertia of the machine tends to maintain uniform rotation and smooths out the impulses. Transformers will pass on the disturbance, but its effect on filament-heating transformers can be neglected, because of the thermal lag of the filaments. Rectifiers will also pass the disturbance, but the smoothing filters will suppress fluctuations at all frequencies except those lower than the ripple frequency of the rectifier. This can sometimes cause undesirable modulation of signals via the grid-bias rectifiers, but the modulation frequency is usually sub-sonic and passes unnoticed.

Voltage fluctuations may be reduced by the use of quick-acting thermionic voltage regulators, and if these are fitted, a maximum angular deviation of 2½° may be permitted. With slow-acting regulators, a tolerance of 2½° is satisfactory for a power supply to a C.W. telegraph transmitter, but a tolerance of ½° is to be recommended for a high-quality broadcast or television transmitter.

AUTOMATIC GENERATING PLANT FOR UNATTENDED STATIONS

V.H.F. repeater and other unattended stations must be safeguarded against service interruptions by failure of the power supply. To ensure continuity of supply, an automatic stand-by Diesel-alternator or petrol-driven alternator is installed when power is normally obtained from the mains. At sites where supply mains are not available, duplicate generating plants are installed, one set acting as a reserve to the other. Since the site is usually isolated, the plant must be capable of operating unattended for periods of four weeks or more.

Small automatic plants are designed to start automatically from a 6-volt or 12-volt battery, which is kept charged by a rectifier while the engine is running. A remote starting switch at the terminal station closes a contactor, which connects the battery to a starting motor. Provided the engine is stationary, the motor engages with the engine and disengages as soon as the engine fires. Under normal conditions, the engine, once started, continues to run until the starting switch is opened.

Automatic voltage regulation is essential. Self-regulating alternators, designed to operate without the use of an automatic voltage regulator, are particularly suitable for unattended working. In these machines an automatic excitation winding built into the generator ensures a constant output voltage within ±2½ per cent on all loads. Once the hand-operated regulator is set for a given output voltage, no further adjustment is needed.

The controls and safeguards necessary for fully automatic and reliable working are :

(1) Switching on and off the mains supply or starting and stopping the generating plant from the terminal station.

(2) Local control of generating plant for testing, either on or off load.

(3) Automatic disconnection of the load, starting of the reserve engine and delayed switching of the load to the reserve generator under any of the following faulty conditions :

- (*a*) Failure of mains supply or first engine.
- (*b*) Excessive deviation of voltage or frequency.
- (*c*) Abnormal rise of temperature of cooling water.
- (*d*) Abnormal fall of pressure of lubricating oil.

(4) Automatic switching of load back to the mains supply and stopping the engine about 1 minute after the mains supply has been restored to normal.

(5) Automatic voltage regulation.

(6) Automatic replenishing of fuel oil in the service tank when oil falls to a pre-determined level.

(7) Automatic control of ventilation in the engine room while the engine is running.

(8) Automatic fire-extinguishing equipment.

Local and remote indication must be provided by red and green indicating lamps and/or audible warning of the following :

(1) Mains or generator supply normal or defective.
(2) Diesel-alternator on or off load.
(3) Failure of engine to start.
(4) Failure of oil or water.
(5) Correct setting of switches for remote control.

As a continuous check on efficient running, recording meters are fitted on the station switchboard to register the number of hours on load and the kilowatt-hours generated.

SWITCHGEAR

Types of Switchboard

The great variety of purposes for which A.C. power is required at transmitting stations makes a low-voltage A.C. distribution switchboard a virtual necessity. Suitable types of switchboard for power supply and distribution are :

(*a*) ironclad unit type ;
(*b*) flat front, unit assembly type ;
(*c*) flat front panel type.

Ironclad boards are built up of standard ironclad switches or switch-fuse boxes, busbar chambers and distribution fuse units mounted on an angle or channel steel frame, designed for floor or wall mounting. Additional units are easily fitted for future extension of the services. These boards are relatively low in cost. Although they lack the appearance and finish usually associated with radio equipment, this need not be regarded as a drawback if they are installed separately. In the unit assembly type the units are manufactured to standard dimensions, the larger sizes being multiples of the smaller ones, and standard busbar boxes the same size as the largest switch unit. This allows any combination to be arranged in a symmetrical assembly by the addition of blanking panels, if necessary, as shown in Fig. 18. Units are withdrawable from the front for servicing and quick replacement of fuses. This type of board is compact, neat in appearance and is easily extended by adding standard units.

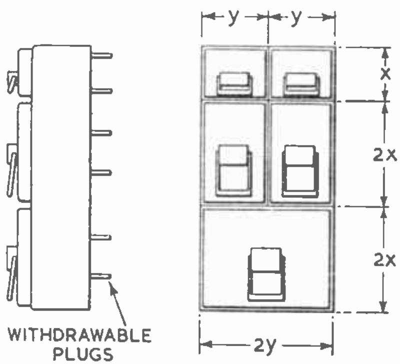


Fig. 18.—Arrangement of Units in Unit Assembly Switchboard.

Panel-type boards are suitable for distribution, generator supply and control, and a variety of other purposes where miscellaneous equipment has to be mounted. They consist of sheet steel or Sindanys panels bolted to a framework and fitted with expanded steel or sheet steel end screens. Provision is made for rear access by doors at either end, which are normally kept locked.

Switchboard Equipment

The equipment for low-voltage, three-phase distribution in a multiple transmitting station is relatively simple. Individual circuit-breakers are unnecessary for currents which do not exceed 600 amperes. Modern high-rupturing-capacity fuses have many of the merits of circuit-breakers, and are much less costly. They have close discrimination, are capable of clearing the heaviest fault currents likely to be encountered, do not deteriorate, and can be relied on for consistent performance. The inverse time/current characteristic of fuses is a valuable feature in preventing unnecessary interruptions due to occasional flash-over.

Circuit-breakers have the advantage that they can be re-set instantly and adjusted for closer overload setting than H.R.C. fuses. Overload and no-volt coils are fitted as standard practice. Overload coils are usually shunted by time-lag fuses to prevent the breaker tripping on momentary overloads, such as flash-backs in the main rectifier. On a three-phase system with unearthed neutral two overload coils give sufficient protection, but three are necessary for protection against earth faults if the neutral is earthed. If the distribution system is an extensive one, earth-leakage protection may also be included to provide closer discrimination on earth faults and ensure that the breaker will open before the leakage current attains a dangerously high value.

4-in.- or 6-in.-scale meters are normally fitted, but $2\frac{1}{2}$-in. scales are often more suitable for small wall panels, such as battery charging boards. Long-scale instruments can be read more accurately than ordinary sector scales, but the instrumental error is the same for both types. The movements of ammeters which indicate a fluctuating load should be well damped. Ammeters and wattmeters should have full-scale readings 30 per cent higher than the normal full load reading to allow for overload indication. Full-scale readings of voltmeters should be such that the normal reading appears at the centre of the scale.

The following is the minimum equipment recommended for various purposes :

(1) *Three-phase Low-voltage Distribution* (*Exclusive of Supply Authorities Metering*).

- (*a*) 3 ammeters or 1 ammeter with ammeter switch.
- (*b*) 1 voltmeter with switch.
- (*c*) 1 frequency meter.
- (*d*) 1 power-factor meter (optional).
- (*e*) Switch-fuse ways for each transmitter or main distributor.

If the load on each phase is balanced, one ammeter is sufficient: if unbalanced, three are necessary. A frequency meter is not required when the frequency is time-controlled. Fig. 19 illustrates a typical simple distribution system.

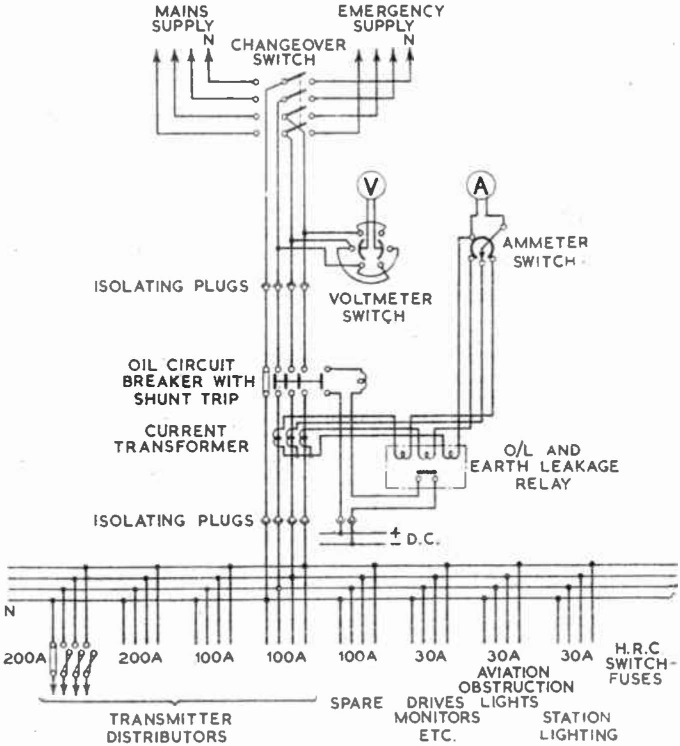


FIG. 19.—TYPICAL LOW-VOLTAGE DISTRIBUTION SWITCHBOARD FOR TRANSMITTERS.

(2) *Single Engine Alternator-exciter.*

As (1), but with the following additions :

(*f*) 1 excitation ammeter.
(*g*) 1 integrating kWH meter.
(*h*) 1 oil circuit-breaker with O/L and N/V releases.
(*j*) 1 set isolating links, if breaker is not the isolating type.
(*k*) 1 alternator or exciter field regulator.
(*l*) 1 automatic voltage regulator.

If the load is balanced, a single-phase kWH meter is fitted, and the readings are multiplied by three; if unbalanced, a polyphase meter is necessary. This instrument enables the units generated to be logged against the fuel consumption. From these records the average fuel consumption per unit generated, the efficiency and the generating costs can be computed.

(3) *Two or More Engine Alternator-exciters Run in Parallel*

(*a*) 1 or 3 ammeters.
(*b*) 1 voltmeter with switch.
(*c*) 1 frequency meter.
(*d*) 1 synchroscope.
(*e*) 1 excitation ammeter.
(*f*) 1 integrating kWH meter.
(*g*) 1 power factor meter.
(*h*) 1 oil circuit-breaker with O/L and N/V releases.
(*j*) 1 set isolating links.
(*k*) 1 alternator or exciter field regulator.
(*l*) 1 automatic voltage regulator.

Items (*e*) to (*l*) are for each machine.

The synchroscope is essential for paralleling, to indicate when the incoming machine has been run up to the same speed as the machines on load, and its e.m.f. is exactly in phase with the busbar voltage. When these conditions are fulfilled, the incoming machine can be switched on to the busbars without disturbance to the busbar voltage or risk of damage to plant. The power-factor meters and kWH meters are necessary to avoid disturbance to the busbar voltage when de-paralleling. To carry out this operation smoothly the outgoing machine should neither: (*a*) be supplying wattless current to the circuit, or take a leading wattless current from the other machines; nor (*b*) be delivering energy to the load or be motored by the other machines. The first condition is satisfied by adjusting the excitation of the outgoing machine to bring the power factor to unity: the second by reducing the motive power of the machine until it is just rotating at the speed of the other machines without generating, and observing when the disc of the kWH meter just ceases to rotate. The outgoing machine can then be taken off the bars without disturbing the busbar voltage.

VALVE COOLING EQUIPMENT

Power Losses in Transmitters

The electrical losses dissipated as heat in a transmitter, which have to be taken into account in determining the most suitable type of cooling system, are made up predominantly of the valve anode losses. Filament heating may amount to as much as 70 per cent of the anode losses. Grid bias is derived from the driving power supplied by the stage preceding the stage considered, and is relatively very small. A certain amount of heat is also generated by R.F. resistance in the coils and dielectric materials of the tuned circuits. For the purpose of estimating the cooling requirements, the total heat losses are equal to the difference between the average total power input and the average R.F. power output.

It is convenient to consider the methods of cooling in common use under five headings:

(*a*) Natural air cooling.
(*b*) Assisted natural air cooling.
(*c*) Forced air cooling.
(*d*) Water cooling.
(*e*) Vapour cooling.

Natural and Assisted Air Cooling

Natural cooling by air convection is satisfactory for low-power transmitters where the total losses do not exceed 500 watts. To assist convection, the back, sides and top of the cubicle or container are liberally louvred or perforated. When the heat dissipation exceeds this value, but does not exceed about 3 kW, it is desirable to augment the natural circulation by fitting an extractor fan in the top of the cubicle. Propeller-type fans can be used where the head losses do not exceed 2 in. of water column. The total amount of heat liberated in a small room housing several low-power transmitters may exceed the limit for comfort. If the ambient air temperature is high, external wall or roof-mounted ventilating fans should also be fitted.

It requires an air displacement of approximately 3,100 cu. ft./min. per kW dissipated for a temperature rise of 1° F. Hence, if P is the total power loss in kW and the temperature rise of the air is not to exceed T° F., the air displacement in cu. ft./min. of an extractor or ventilating fan is given by

$$Q = 3{,}100\ P/T$$

Forced Air Cooling

Air cooling is less efficient than water cooling where large amounts of heat are involved, but it has the following important advantages:

(*a*) There are no R.F. and D.C. losses, as there are in water insulating columns.
(*b*) The cost is lower and the circulatory system simpler.
(*c*) Leakage and obstruction are less likely to occur.
(*d*) Failure due to freezing is eliminated.

The rate of flow is directly proportional to the losses, but varies widely with the type of valve, depending on the surface area of the cooling fins and the emissivity of the anode. For a given power P in kW released as heat, the theoretical flow in cu. ft/min. can be calculated from

$$Q = 5.92\ PT_0/T_1$$

where T_0 is the absolute temperature of the incoming air (= °C. + 273) and T_1 is the temperature rise of air in °C. Normally, the temperature of the outgoing air is not allowed to exceed 180° C. The actual flow recommended by the valve manufacturer may be anything from 50 to 150 cu. ft./min. per kW dissipated.

Intake and exhaust ducting usually have a large cross-sectional area to reduce air velocity and noise and keep the head losses as low as possible. This permits low-pressure, single-stage centrifugal fans to be employed. Single-stage fans are made for a wide range of duty, varying from 10 to 15,000 cu. ft/min. at total pressures up to 15 in. w.g. The brake horse-power absorbed by a fan can be calculated from:

$$\text{B.H.P.} = Qh_t/6350\eta_t$$

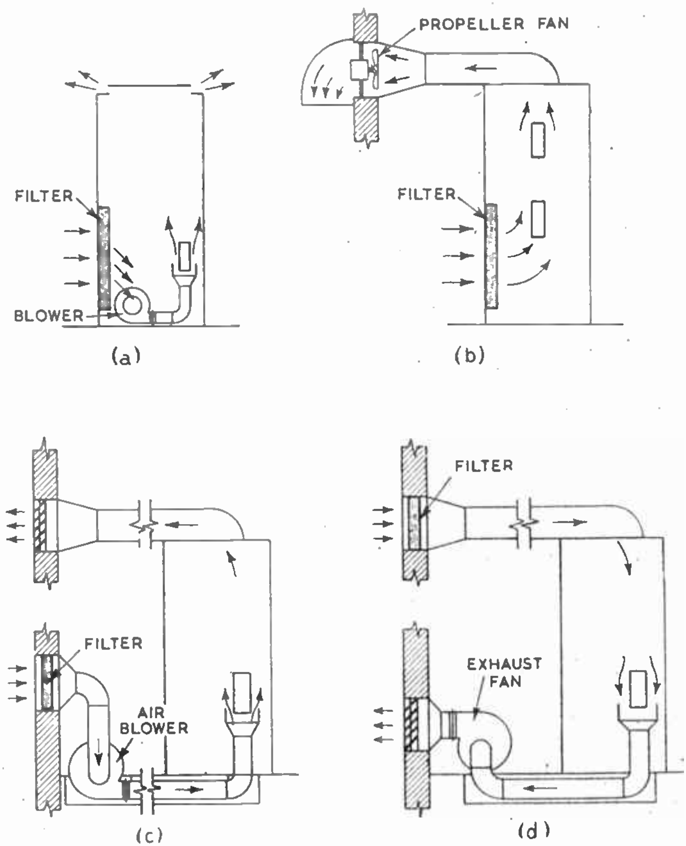


FIG. 20.—FORCED-AIR COOLING SYSTEMS.

where Q is the volume of flow in cu. ft/min., h_t the fan total pressure in in. w.g. and η_t the fan total efficiency. The total efficiency of a fan varies from about 20 per cent for volumes of 100 cu. ft./min. to 75 per cent for 7,000 cu. ft/min.

Fig. 20 (*a*) to (*d*) show schematically four common variations of the air-circulating system. In (*a*) cool air is drawn from the transmitter room through glass wool or fabric filters built into the back panel of the transmitter by a centrifugal fan mounted inside the cabinet. Air discharged from the fan and ducted into funnels which accommodate the valves, and finally released through the top of the transmitter. In (*b*) an external extractor fan replaces the blowing fan and exhausts the heated air outside. Both methods are practicable with small volumes of air, when the intake velocity is less than, say 6 ft./sec., and does not create excessive draught in the room.

High-power transmitters are cooled by the methods outlined in Fig. 20 (*c*) and (*d*). With either method the cabinets or cubicles must be of airtight construction. In Fig. 20 (*c*) cool air is drawn from outside the building, filtered and conducted through a floor duct into valve funnels at the bases of the cubicles, and finally discharged through an upper duct to atmosphere. In (*d*) the air is filtered, drawn through an overhead duct and extracted through a floor duct, being finally exhausted outside the building at low level. This method has the advantage that the pressure inside the cubicle is below atmospheric pressure, so that any inward leakage tends to assist cooling, whereas in (*c*) the pressure is above atmospheric pressure, and outward leakage tends to degrade the cooling efficiency and overheat the room. In practice, the overhead duct may be a vertical shaft terminated by a cowled roof intake or a louvred cupola.

Air Ducts

Galvanized sheet steel air ducts are often installed where noise is not important. Straight lengths of circular or rectangular cross-section with the necessary bends and transformation sections are fabricated to suit the particular requirements of the installation. Circular ducting of a given gauge will withstand a greater collapsing stress than rectangular ducts of equivalent cross-section. Rectangular ducts, on the other hand, are better adapted for connection to cubicles and installing in confined spaces. Ducts constructed of plywood, masonry or reinforced building board are better noise attenuators than metal ducts.

Head losses in ducts are produced by air friction with the surfaces, eddying at bends and at sudden changes of cross-section. Fig. 21 shows graphically the head loss in straight, smooth galvanized steel ducts of circular cross-section against volume of flow. The head loss in ducts of rectangular section can be estimated by considering the duct to have an equivalent diameter. For ducts of the size used with transmitters the equivalent diameter is approximately equal to the smaller side multiplied by a factor determined by the ratio of the sides, and given in Table 6.

TABLE 6.—EQUIVALENT DIAMETER OF DUCTS OF RECTANGULAR SECTION

Ratio of sides .	1/1	1·5/1	2/1	3/1	4/1
Diametrical factor .	1·1	1·33	1·52	1·83	2·0

Bends are treated as equivalent lengths of straight duct. Fig. 22 shows the equivalent length in terms of the ratio of inner radius of bend to diameter of duct. To reduce turbulence, the inside radius should be at least equal to the diameter of the duct.

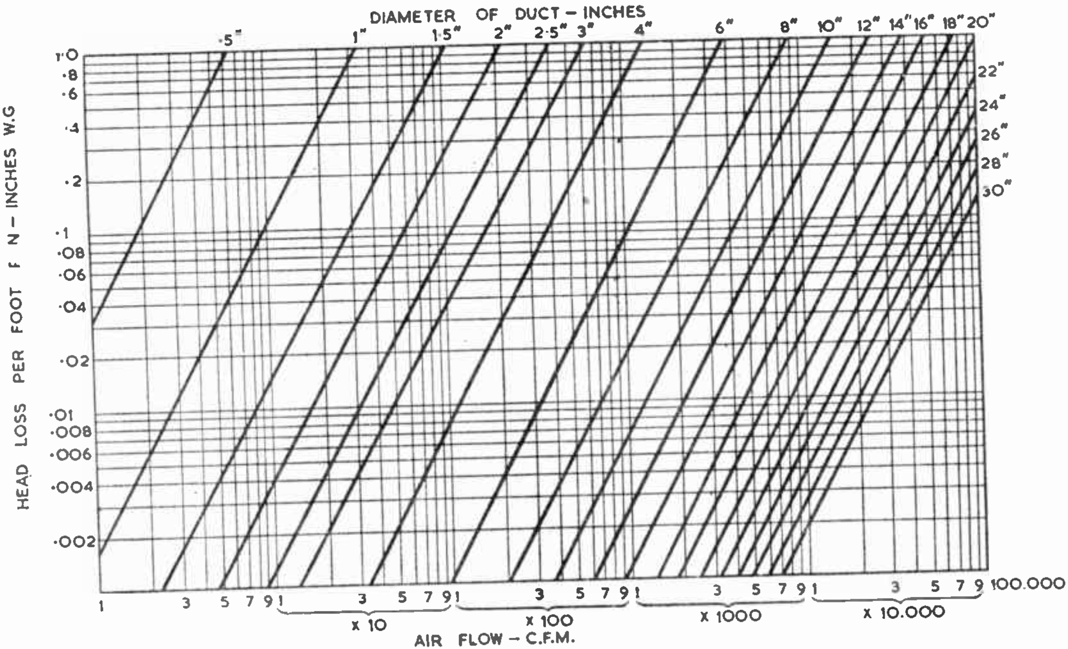


FIG. 21.—HEAD LOSS IN AIR DUCTS OF CIRCULAR CROSS-SECTION.

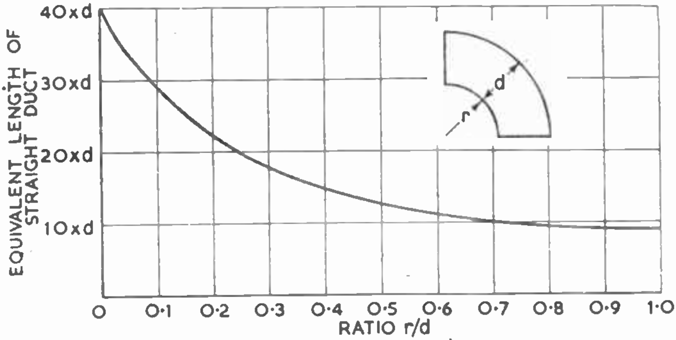


Fig. 22.—Equivalent Lengths of Radiused Bends in Air Ducts.

Noise in Air-cooling Systems

The problem of noise abatement has a special interest for radio installations, where the noise level must be of a very low order. A rough criterion for a transmitting installation is that noise generated by an air-cooling system should not exceed the average level of background noise in the transmitting room, but in some cases monitoring facilities may necessitate an even lower level.

Noise in an air-cooling system originates in several ways. The chief components of system noise are mechanical and air noise produced in the fan and motor, noise inherent to the movement and pulsations of the air stream, vibration and drumming of ductwork caused by the presence of sharp obstructions, resonances, eddies and vibration transmitted from the fan or its mounting.

Low-speed fans are less noisy than high-speed fans, but for the same duty the fan and motor are larger and more costly. To install a low-speed fan is, however, not necessarily the best way of coping with the problem. High-speed fans produce most of their noise at high frequencies, and it is well known that absorbent treatments are more efficient in attenuating noise of high frequency. Small fans are often built into the transmitter framework, but large fans are invariably installed in an adjoining room. Transmission of noise via the fan casing and foundation is eliminated by fitting a flexible joint between the discharge opening and the delivery duct, and mounting the fan and motor on rubber feet or a resilient pad.

Silent-running fans of similar construction to those designed for organ blowing are eminently suitable for mounting in transmitter cubicles. The casings of these fans are lined with sound-absorbent material, and special attention is given to the balance of the motors and impellers and the design of the bearings to ensure smooth torque and freedom from vibration.

Much improvement can be made by fabricating ducts of rigid sound absorbent material, such as Celotex, wood or hardboards sandwiched with slag wool, or by lining metal ducts with thin rubber sheet or other damping material. It is important that the inside surfaces should be smooth, since rough surfaces increase the pressure drop.

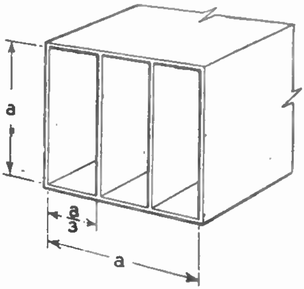


FIG. 23.—METHOD OF REDUCING NOISE BY DIVISION OF AIR DUCT.

Noise attenuation varies directly as the length of duct and as the ratio of perimeter to cross-sectional area. A useful device for increasing the ratio without reducing the cross-sectional area is to sectionalize the duct by inserting thin partitions or "splitters," as shown in Fig. 23. The attenuation in decibels achieved in this way can be expressed by

$$a = klp/A$$

where k is a constant depending on the absorption coefficient of the material of the duct, l is the length of duct, p the length of the perimeter and A the cross-sectional area. For example, a square duct of sides a will have a perimeter/area ratio of $4a/A$. The introduction of two splitters extending the length of the duct will increase the ratio to $8a/A$ and double the attenuation.

Water-cooling Systems

The basic elements of a valve water-cooling system are a motor-driven centrifugal pump, an air blast or exchange water cooler, valve jackets with helical or spiral water-insulating columns to insulate the anode from earth, and electrical cut-outs, either thermally or flow operated, to safeguard the valves against failure of the circulation. A typical cooling system is shown in Fig. 24.

The pump draws cold water from a closed storage tank through a foot valve and strainer, and distributes it through a pipe system and the inlet insulating columns to the valve jackets. Tangential entries in the jackets give the water a swirling motion round the anodes. After traversing the jackets, the heated water is discharged tangentially through the outlet insulating columns. It is then cooled by passage through the cooler and returned to the tank, in readiness for recirculation.

In some systems the cooler is inserted in the circuit on the delivery side of the pump. The pressure drop round the circuit is analogous to the voltage drop in an electric circuit. Pressure is highest at the pump delivery, and falls progressively round the circuit, being finally reduced to zero as the water is returned to the tank. This system therefore relieves the pressure on the valve jackets, gaskets and insulating columns, but it suffers from the disadvantage that the storage tank and the pump run hot.

The capacity of the tank should be sufficient for at least 10 minutes circulation at the normal rate of flow, with reserve capacity for draining the system. It is advantageous to place the tank at a low level to permit the system to be completely drained into it, and the pump at the same level to avoid the necessity of priming.

Valve-cooling water must be free from solids in solution or suspension. Mains water and water that has been softened precipitate hard lime and other deposits on the hot anodes, which is difficult to remove. These

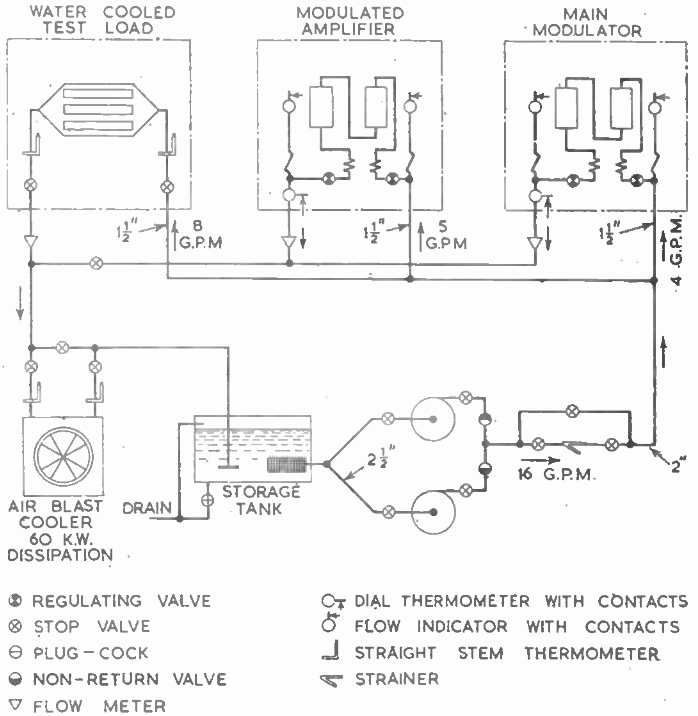


FIG. 24.—SCHEMATIC ARRANGEMENT OF VALVE WATER-COOLING SYSTEM.

deposits have low thermal conductivity, which causes the anodes to overheat and favours the formation of local hot spots. An overheated anode expels occluded gas and impairs the vacuum of the valve. For these reasons distilled water or filtered rain-water are used.

If a large surface area of water is exposed to the atmosphere or excessive turbulence or spraying is allowed to occur, air is dissolved. When the water is heated in a cooling system, air in solution is liberated and exerts a corrosive action on ferrous components, causing the red and black oxides of iron to be deposited on the anodes and connecting pipes. Metallic oxides in suspension increase the leakage current, and the black oxide is particularly difficult to remove. To prevent aeration, a closed circulating system is used, and the water, after circulating, is returned to the storage tank with a minimum of turbulence.

Electrolytic action with water which is usually slightly acid is another trouble to be avoided. The presence of zinc, zinc alloys or galvanized parts in contact with the water in the system cause insoluble zinc to be electrolytically deposited on the anodes. Pumps are fitted with gun-

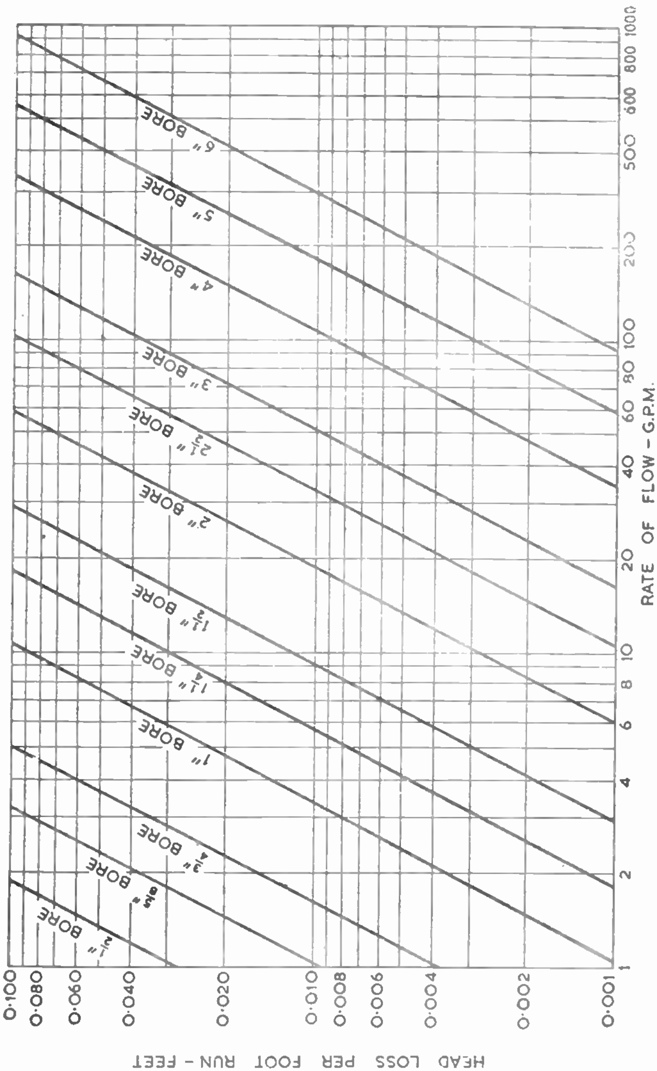


Fig. 25.—Head Loss in Copper Water Pipes ½–6 in. Bore.

metal impellers, and only the outside surfaces of tanks are galvanized. Pipes are usually of copper and pipe fittings of brass.

The rate of flow of water is proportional to the total power dissipated in the valve. If P is the total dissipation of kW and T the permissible temperature rise in °C., the rate of flow in gal./min. is given by

$$Q = 3{\cdot}2P/T$$

The temperature of the water in the jacket must not approach boiling point as the formation of steam bubbles causes unequal heating of the anodes. Normally the temperature of the outlet water is not allowed to exceed 65° C., with a maximum temperature rise across the jackets of 15° C. Heat is transferred more efficiently if the velocity is not too low, and in practice the mean velocity is never less than 2 ft./sec.

The pressure or equivalent head in feet to be provided by the pump is the sum of the velocity head and the static head. Velocity head is the dynamic head due to movement of the water column, which is absorbed in overcoming frictional losses and eddies in the pipes, bends, valves, strainers and sudden enlargements and contractions in the system. Static head is the lift above some datum level, for example the pump level, and includes the suction lift.

Frictional-head losses vary directly as the length of pipe and as the (volume of flow)^2. Fig. 25 shows the head loss in feet in circular smooth metal pipes when run full bore. It should be noted that in course of time the bore becomes encrusted with deposits which increase the head losses. To allow for this a suitable percentage should be added to the figures obtained from the graphs. Pipe fittings are considered as equivalent straight lengths of pipe, which are obtained by multiplying the head loss per foot run for straight pipes by the factors given in Table 7. Head losses in the cooler and valve jackets are usually specified by the maker.

TABLE 7.—FRICTIONAL HEAD LOSSES IN COPPER WATER-PIPE FITTINGS

Radiused bend	10	× Head loss per foot of straight pipes
Right angle elbow	30	
Gate valve, full open	15	
Non-return valve	25	
Suction entry	30	
Strainer	35	

For a given flow, the diameters of pipes must be sufficient to ensure that the head losses in pipes and fittings do not exceed, say, 10 per cent of the total system losses. Excessive head losses entail the provision of pumps of increased delivery head and horse-power.

Vapour Cooling

The principle of vapour cooling was investigated as long ago as 1934, and was the subject of a joint patent (British Patent No. 432,891, 1935) by P. E. Privett and Marconi's Wireless Telegraph Co. Ltd. Although theoretical expectations were fully realized in these early experiments, further development work was temporarily suspended in favour of

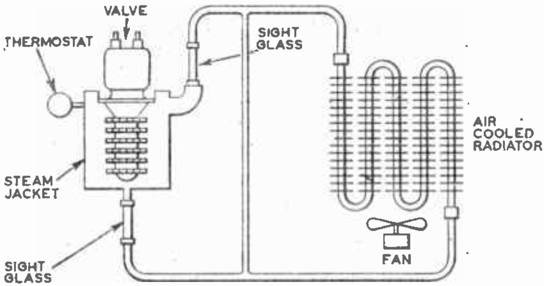


FIG. 26.—VAPOUR-COOLED VALVE SYSTEM.

forced-air cooling, which was then coming into vogue. It is only during the last decade that the system has been successfully developed and applied by the *Compagnie Française Thomson-Houston* in France, and more recently in Britain and the U.S.A.

It is a matter of common knowledge that when a liquid is vaporized heat is absorbed. The heat extracted does not raise the vapour temperature, but is used in changing from the liquid to the vapour state, and remains latent until it is given up again during condensation. In a vapour-cooled valve water is allowed to circulate round the anode at a rate just to cause the water to boil. The steam produced is allowed to pass off freely into a condenser, from which the condensed water is recirculated. Thus heat is continuously absorbed and the anode is maintained at a constant temperature.

The latent heat of steam at 100° C. is 538 calories. By definition this is the quantity of heat required to convert 1 gm. of water into steam. Theoretically, if the condensed water were returned to the valves at 100° C., it would require less than ½ gallon of water per hour to absorb the heat equivalent of each kilowatt dissipated, or less than 3 per cent of the quantity required for water cooling. In practice, since extra water is required for safety, and a drop of temperature occurs during condensation and passage through pipework, the rate of flow is somewhat higher.

Each pound of steam liberated carries the maximum quantity of heat possible, whereas with water cooling inefficient heat transfer necessitates a large excess of water, which reduces the already low temperature difference between the water and the cooling air. With high-power transmitters, the greater temperature difference of the evaporative system makes it possible to use it effectively for the additional duty of space heating the station building.

The anodes of vapour-cooled valves are specially constructed to prevent the formation of insulating air films and bubbles, which would otherwise produce hot spots and quickly cause serious damage to the anodes. For this purpose the surface of the anode is broken up by a series of radial projections, which present the appearance of co-axial sprocket wheels.

The elements of a complete system are shown in Fig. 26. The valve is enclosed in a steam jacket, and water is circulated upwards and round the anode. Steam is delivered from the upper end to an air-cooled radiator, and the condensate is recirculated to the valve jacket. Vaporization generates the pressure necessary to stimulate circulation, so that separate water pumps are unnecessary. Since the pressure developed is proportional to the rate of vaporization, which in turn depends on the rate at which heat is being dissipated by the anode, the rate of flow is self-adjusting. For low powers up to about 5 kW the radiator is cooled by air convection, and the whole apparatus forms a self-contained unit. For higher powers cooling is usually fan-assisted, and the cooling equipment is installed outside or on the roof of the station building.

Advantages of Vapour Cooling

Several advantages over conventional methods of cooling are claimed for the vapour cooling system:

(1) The process is more efficient. Compared with water cooling, the rate of flow is much lower. This reduces water costs and makes it possible to use smaller bore pipework.

(2) The higher resistance of the narrow water column and the better insulating properties of steam make it possible to employ a shorter length of insulating pipe.

(3) The system is a closed one, and water losses are negligible. Since the water is distilled in the cooling process, tap water need be added only occasionally to make good the losses which occur after the initial filling.

(4) For equal duties, the system occupies a much smaller space than either an air- or water-cooling system.

(5) It is safer than air or water cooling. Water flow and heat exchange are automatically controlled by the amount of power dissipated. Except for thermostatic cut-outs embodied in the valve jackets, flow safeguards are unnecessary.

(6) Silent operation. As there are no large centrifugal fans, the problem of noise abatement does not arise.

(7) There is no cleaning or renewal of clogged air filters, and the relatively small radiator needs less attention than the large air blast coolers required for water cooling.

ANODE POWER SUPPLY

Anode Power Conversion

The chief means of converting low-voltage A.C. to high-voltage D.C. for the supply of anode power are the high-voltage rectifier and the motor-generator. The development of high-voltage hot-cathode and cold-cathode mercury-arc bulbs has led to a general preference for static rectifiers. They are silent in operation, highly efficient, low in cost compared with rotary machines and the absence of moving parts simplifies maintenance.

Anode power demands vary from less than 1 ampere at a few hundred volts for low-power transmitters to as much as 20 amperes at 15 kV for high-power work. These diverse requirements are met by the use of three main types of rectifier :

(1) Metal rectifiers for low-power transmitters.
(2) Hot-cathode mercury-vapour rectifiers for low and medium power.
(3) Cold-cathode mercury-arc rectifiers for high power.

A detailed treatment of the principles and circuits of rectifiers may be found in Section 25.

In radio technique the ripple voltage present in the rectified output must be reduced to a negligibly low value by the addition of smoothing filters, in order to prevent modulation of signals. Polyphase rectification greatly reduces the residual ripple voltage and makes it possible to economize in filter components. In practice three-phase, full-wave rectifiers are employed as far as possible. The degree of ripple voltage permissible in the D.C. output from a filter depends on the class of transmission. Maximum peak values of ripple voltage, expressed as a percentage of the D.C. voltage, which are commonly used as a basis of design are :

Telegraphy	0·5 per cent
Commercial telephony	0·1 per cent
Broadcasting and television	0·05 per cent

It is usual with low- and medium-power transmitters for all the R.F. and A.F. amplifier stages to be fed from a common rectifier. The full voltage is applied to the final stage, and voltage-reducing resistances, incorporated in the transmitter, are connected in series with the individual anode supply leads to drop the voltage to the working values. There are, however, certain limitations to the use of series resistance. Unless the current taken by the earlier stages is small compared with the total current, the power loss and heat dissipated in the resistance may be excessive.

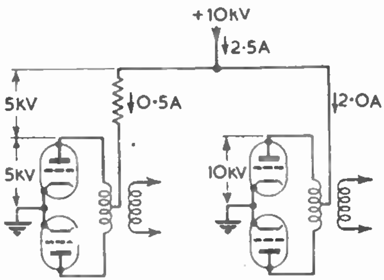


FIG. 27.—POWER AMPLIFIERS FED BY A COMMON RECTIFIER.

Consider, for example, a final amplifier taking 2 amperes at 10 kV and a penultimate stage taking 0·5 ampere at 5 kV, as shown in Fig. 27. The loss in a series resistance in the penultimate stage will be 5 × 0·5 = 2·5 kW or 10 per cent of the total D.C. power. If this stage operates as a Class C or Class B amplifier,

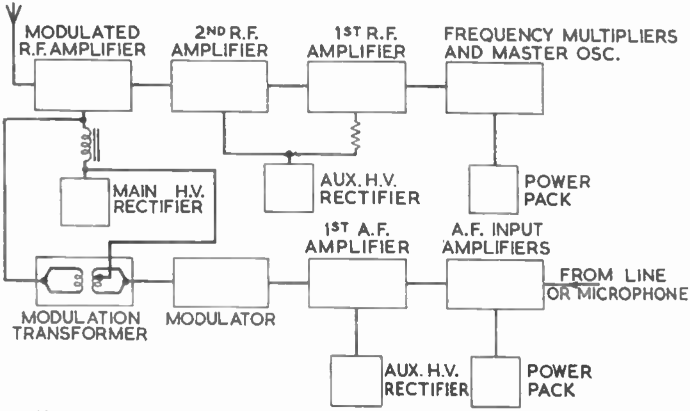


FIG. 28.—HIGH-POWER BROADCAST TRANSMITTER FED BY SEPARATE HIGH-VOLTAGE RECTIFIERS.

the anode voltage regulation may be seriously degraded. Assuming the current to vary from 0·5 ampere to zero, the anode voltage will swing from 5 to 10 kV. In telegraph transmitters keyed by the on/off method, this usually has no objectionable effects, but in low-level modulated transmitters it will produce amplitude distortion. The remedy in such cases is to feed the modulated amplifiers from a separate rectifier. In many high-power transmitters the main rectifier serves only the final power amplifier and the low-power stages are fed by one or more auxiliary rectifiers, as in Fig. 28.

Regulation of Anode Power

The facility of being able to regulate the anode voltage is useful for testing at low voltage, making initial adjustments and reducing the R.F. output of a transmitter. Induction regulators and moving-coil regulators provide the smoothest method of regulating the primary voltage. These appliances may be either hand-operated or motor-operated by push buttons from a control desk.

The principle of operation of these regulators is the same as for voltage regulators, but the range of control and the internal kVA is greater. Voltage regulators usually have a range of ±10 per cent. A power regulator capable of controlling from full to quarter power would have a range of full to half voltage or 50 per cent. Induction and moving-coil regulators are rated by their internal kVA, which is calculated as follows:

Single-phase regulators (secondary current × max. boost voltage)/1,000.

Three-phase regulators 3(secondary current × max. boost voltage)/1,000.

M

A regulator designed to buck and boost about the mean voltage is rated for only half the internal kVA of one that bucks over the whole range. A fixed ratio auto-transformer preceding or combined with the regulator can be employed to step down the supply voltage by half the overall voltage range to reduce the size of the regulator. Alternatively, the regulator may be designed to buck and boost equally about the input voltage. This method eliminates the auto-transformer, but requires that the primary winding of the rectifier transformer must be wound or tapped for the maximum boost voltage.

Step-by-step methods of control include auto-transformers with tapping switches and tapping switches on the primary winding of the rectifier transformer. Either off-load or on-load selection may be used, but for testing and occasional reduction of power, on-load adjustments are unnecessary.

GRID-BIAS SUPPLIES

Methods of Biasing

There are three possible ways of providing biasing voltages for the control grids:

(*a*) Cumulative grid biasing.
(*b*) Self-biasing.
(*c*) Bias rectifier.

The cumulative grid method of biasing (Fig. 29) is confined to unmodulated amplifiers, as it introduces distortion. Its action is too complex to admit of mathematical treatment here. In simple terms, if the charging and discharging effect of the capacitor C and the interaction of the anode circuit is neglected; the grid-cathode circuit behaves as a diode in the absence of a fixed biasing voltage. The D.C. component of grid current is confined by the capacitor C to the path through the leak resistance R, and produces a voltage drop across it. Adopting the electronic convention that current flows from negative to positive, the direction of flow is such as to bias the grid negatively with respect to the cathode, by an amount determined by the product of current and resistance at any instant.

The self-biasing method (Fig. 30) is restricted to very low-power transmitters. This method makes use of the well-known principle

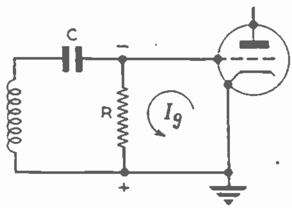


Fig. 29.—Leaky Grid Method of Biasing.

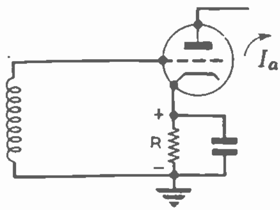
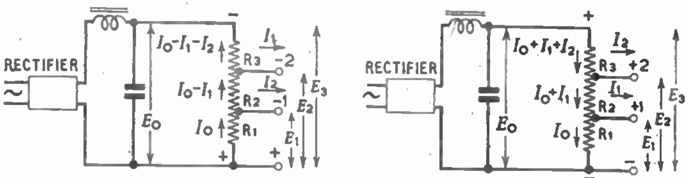


Fig. 30.—Cathode Resistance Biasing.



(a) GRID NEGATIVE BIAS (b) GRID POSITIVE BIAS

FIG. 31.—GRID-BIAS RECTIFIER WITH VOLTAGE DIVIDER.

adopted in receiver design, in which the biasing potential is obtained from the voltage drop produced by the D.C. component of anode current in a resistance connected in series with the cathode lead.

Voltage Dividers

The objection to applying the self-biasing principle to medium- and high-power transmitters is the power loss and heat generated by the use of resistors. It is more usual to bias the power-amplifier stages by means of a common rectifier and tap the required voltages from a voltage divider. This method is satisfactory provided the grid current in the final stage does not fluctuate excessively, as for example in Class C amplifiers. Heavy fluctuations in current drawn from one tapping can be a source of interstage coupling, and in such cases it is sometimes necessary to obtain the bias voltage for the final stage from a separate rectifier.

A voltage divider serves the dual purpose of a distributor and, to a limited extent, of a voltage regulator. The greater the ratio of the bleeder current in the divider to the total current tapped, the better is the regulation. Fig. 31 (a) and (b) show the directions of current flow for grid negative bias and grid positive bias respectively. It should be noted, however, that in Class A amplifiers the grid is never driven positive, and consequently no grid current flows. The resistance of each section is determined from the sectional voltage drops as follows:

Let E_0 = full-load output voltage from rectifier;
I_0 = full-load output current from rectifier;
$I_1\ I_2\ I_3$ = indicated grid currents tapped from divider;
$E_1\ E_2\ E_3$ = tapped voltages on load;
$R_1\ R_2\ R_3$ = sectional resistances.

Then

Resistor	*For Negative Bias*	*For Positive Bias or H.T. Supply*
R_1	$= E_1/I_0$	$= E_1/I_0$
R_2	$= (E_2 - E_1)/(I_0 - I_1)$	$= (E_2 - E_1)/(I_0 + I_1)$
R_3	$= (E_3 - E_2)/(I_0 - I_1 - I_2)$	$= (E_3 - E_2)/(I_0 + I_1 + I_2)$

Glow-gap Dividers

When voltage stability of an exceptionally high order is necessary, wire-wound voltage dividers are replaced by special gas discharge tubes or glow-gap dividers, as shown in Fig. 32. These tubes have the additional merit that they compensate for both variations of load current and mains supply voltage, and remove to a great extent any residual ripple voltage. A constancy closer than 0·1 per cent is readily obtainable with variations of supply voltage up to 10 per cent.

The tubes are filled with a mixture of neon and other gases at low pressure and contain a number of cup-shaped electrodes. When a voltage exceeding a certain critical value is applied, a discharge takes place. A gaseous discharge, unlike a metallic conductor, does not obey Ohm's law. The voltage drop across the discharge varies imperceptibly with change of current. Thus, a substantially constant voltage is maintained at the electrodes, regardless of the current tapped off. A resistance R_s is connected in series with one supply lead to limit the current and reduce the tube voltage when the tube strikes, to a value just sufficient to maintain the discharge and absorb the voltage variations. The total resistance in series, including the internal resistance of the rectifier and the resistance of the smoothing choke, must be such that the series voltage drop at the working current is at least half the tube drop. To maintain the ignition, additional resistances R_1, R_2, R_3 are joined between each inner electrode and one of the outer leads to maintain the ignition.

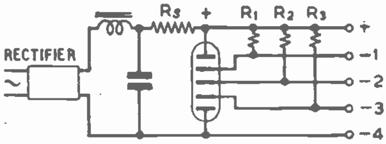


Fig. 32.—Grid-bias Rectifier with Glow-gap Divider.

The overall efficiency of rectifiers fitted with gas discharge dividers is obviously low, because of the high resistance losses, but this is usually unimportant, since the losses are small compared with other transmitter losses.

FILAMENT HEATING

Filament Heating Requirements

Large transmitter valves have thick tungsten filaments, which are heated by low-voltage A.C. or D.C. The thickness, which determines the heating current and voltage, is a compromise between some minimum value which will give an adequate expectation of life and some maximum, above which the amount of heating power becomes uneconomical. Current and voltage requirements vary from about 1 ampere at 6 volts for the smallest values to 430 amperes at 35 volts for the largest water-cooled types.

A.C. heating by means of step-down transformers is more efficient than D.C., which involves conversion losses, but it introduces a certain amount of modulation of the emission current. This is due partly to deflection of the electron stream by the alternating magnetic field set up round the filament; partly to thermal variation at twice the A.C.

frequency; and partly to voltage variation between grid and filament and between anode and filament at the A.C. frequency. The last of these causes is easily eliminated by connecting the earthy end of the grid and anode circuits to a centre tap on the transformer heater winding or the centre point of a feed-equalizing resistor, as in Fig. 33. With indirectly heated valves the connection is made to the cathode, which has no direct connection with the heater. Modulation of the emission, in general, can be appreciably reduced by the use of negative feed-back circuits.

The resistance of cold tungsten filaments is about 7 per cent of the resistance when heated. Consequently, they must be switched on gradually, in order to limit the heavy initial current and prevent damage or distortion of the filaments. Usually, the transitory starting current is not allowed to exceed 1½ times the normal heating current. The time allowed for a filament to reach its full working temperature is directly proportional to its diameter, and for a filament 1 mm. diameter is about 30 seconds.

The life expectation of a valve is adversely affected by excessive filament voltage, and valve-life guarantees are given subject to close tolerances of the voltage. A continuous ovorvoltage of 5 per cent may halve the life of a valve. Transmitters under constant supervision require close manual adjustment and regular attention to avoid any sustained rise of voltage. In cases where the supply voltage is unstable or the transmitter is remotely controlled, an automatic voltage regulator should be installed to stabilize the main supply voltage within ±1 per cent.

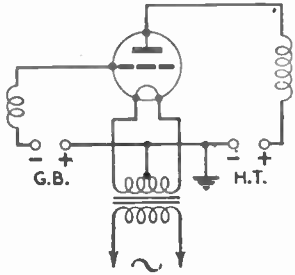


FIG. 33.—CENTRE TAP FEED RETURN FOR ELIMINATING GRID-CATHODE MODULATION BY A.C.

Filament Switching

The principle methods of switching A.C. heated filaments are:

(a) Stepped or variable resistance in series with the transformer primary circuit (Fig. 34 (a)).

(b) Star-delta switching (Fig. 34 (b)).

(c) Tapping switch on the transformer primary winding (Fig. 34 (c)).

(d) Transformer with high leakage reactance.

Resistance in series is switched either manually or automatically by means of delayed-action contactors. Star-delta switching is a two-step method confined to three-phase heating or two-phase heating with a Scott-connected transformer. Step 1 connects the windings in star formation and reduces the starting voltage to $1/\sqrt{3}$ of the working value, but this can be further reduced to any desired value by adding resistance in series with the windings. The transformer tapping switch is essentially a manual method. On-load switching is quite practicable if close tappings giving very small voltage increments are used to prevent overheating of the winding while the switch arm bridges the studs.

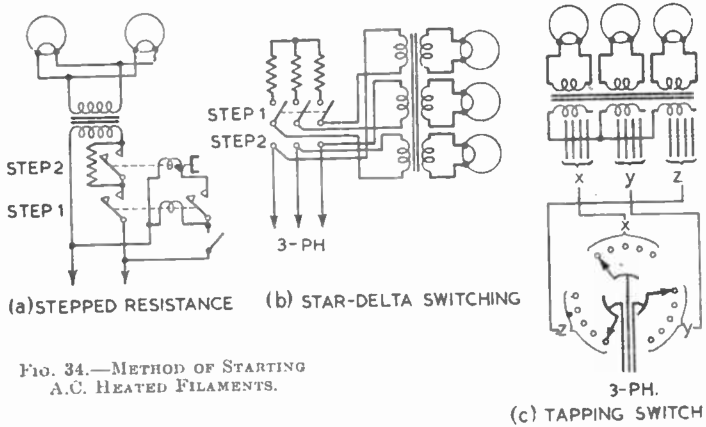


Fig. 34.—Method of Starting A.C. Heated Filaments.

High leakage reactance transformers, designed to have poor voltage regulation, are automatic in action. On first switching on, the heavy current rush due to the low load resistance of the cold filaments produces a large voltage drop in the windings and limits the applied voltage to a safe value. As the filament resistance increases with temperature, the transformer voltage rises to a steady working value. It is important to note that if a single transformer were used for two or more valves, the sudden reduction of the load following the failure of one valve would cause the filament voltage to rise excessively on the others. For this reason, a separate transformer is necessary for each valve.

D.C. Filament Heating

D.C. heating insures complete immunity from undesirable modulation of the emission. Conversion from A.C. to low-voltage D.C. is most efficiently carried out by means of metal rectifiers with step-down transformers or with D.C. motor-generators. For heating purposes the residual ripple after rectification is sufficiently small to make it possible to dispense with smoothing filters.

Similar starting methods are employed with rectifiers as with A.C. heating, namely, stepped series resistance in the primary circuit, star-delta switching of the transformer primary windings or a primary tapping switch. With motor-generators the voltage is raised slowly by means of the field regulator of the generator.

The emission from a D.C.-heated filament is greatest at the negative end. In order to equalize the emission over a period of time and improve the life of the filament, provision is made for reversing the polarity of the supply at regular intervals. When the filaments are heated by a rectifier, this is accomplished by fitting a commutating switch or change-over links in the low-voltage output. When a motor-generator is employed it is usually more convenient to reverse the field excitation.

W. E. P.

6. BROADCASTING TRANSMITTERS

6. BROADCASTING TRANSMITTERS

This first section describes the chief features of amplitude-modulated broadcasting transmitters for the medium-frequency band (300–3,000 kc/s) and the high-frequency band (3–30 Mc/s); this account also applies to transmitters operating in the frequency range 150–300 kc/s.

Information on frequency-modulated broadcasting transmitters for the very-high-frequency band is given in the second part of this section.

Before discussing the design of the transmitters, we shall briefly review the principal features of amplitude modulation.

Amplitude Modulation

In this system of modulation the amplitude of a radio-frequency carrier is varied by the modulation signal, and is at all times directly proportional to the instantaneous value of the modulating signal. Thus the carrier amplitude swings above and below its unmodulated value, as shown in Fig. 1. In this diagram the amplitude of the unmodulated carrier is represented by A and the peak variation in carrier amplitude due to modulation by a. The ratio of a/A is known as the modulation depth, and it is usually expressed as a percentage. The maximum value of the modulation depth is 100 per cent, which occurs when $a = A$; in these conditions the carrier amplitude swings between zero and $2A$. Any attempt to increase the modulation depth beyond this value results in distortion of the modulation envelope, the carrier being cut, i.e., reduced to zero amplitude for a period coinciding with the negative half-cycles of the modulating waveform signal.

The expression for an amplitude-modulated wave is:

$$(1 + K \sin pt) \sin \omega t$$

where ω is $2\pi \times$ carrier frequency, p is $2\pi \times$ modulating frequency, and K is the modulation depth.

On expansion this expression gives

$$\sin \omega t + \frac{K}{2} [\cos (\omega - p)t - \cos (\omega + p)t]$$

The first of these terms represents the carrier wave, which is of constant amplitude even during modulation. The second and third terms represent signals known respectively as the upper and lower sidebands, and are present only during modulation. The sideband amplitudes are equal and directly proportional to K, the modulation depth. These sidebands are symmetrically disposed in frequency about the carrier value, and their displacement from it is equal to the modulating frequency. For 100 per cent modulation ($K = 1$), the amplitude of the sidebands is equal to one-half that of the carrier. Each sideband thus has one-quarter the power of the carrier, and both sidebands together have one-half the power of the carrier. Thus to modulate a carrier 100 per cent with a sinusoidal signal, the power which must be supplied by the modulating device must equal one-half that of the carrier.

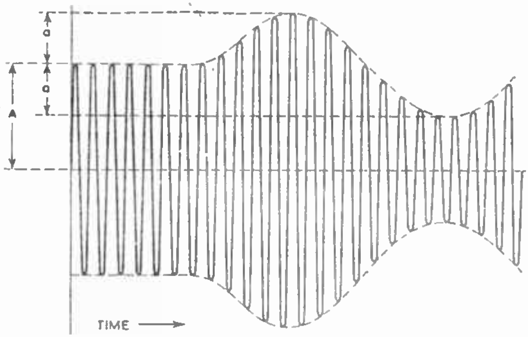


FIG. 1.—A CARRIER WAVE AMPLITUDE-MODULATED BY A SINUSOIDAL SIGNAL.

Performance of Broadcasting Transmitters

One of the principal requirements of a broadcasting transmitter is that the modulation waveform should be a faithful copy of the audio-frequency signal supplied to the transmitter; accordingly, the design must be such that attenuation and harmonic distortion are kept at a low level. Efforts are made to keep the frequency response level within ± 1 db over the modulation frequency range of 50 c/s to 10 kc/s and the total harmonic distortion to below 1 per cent for modulation depths up to 80 per cent. By careful design it is possible to keep within these tolerances even for a transmitter radiating 100 kW. It is very difficult, however, to keep harmonic distortion low for modulation depths exceeding approximately 80 per cent, and the harmonic content generally increases sharply as the modulation depth approaches 100 per cent.

The attainment of such a performance is made considerably more difficult by the necessity for high efficiency. Modern transmitters radiate such large powers and operate for such a large proportion of each day that the utmost economy in operation is essential. The overall efficiencies obtained are surprisingly high, being of the order of 35 per cent; this should be compared with the efficiency of a receiver which rarely exceeds 10 per cent.

Very great carrier-frequency stability is also required; transmitters are allocated a particular frequency (known as a channel) in the waveband on which they operate, and deviations from this frequency must not exceed certain values. For medium-frequency transmitters, the deviations must not exceed 20 c/s, and for high-frequency transmitters must not exceed 0·003 per cent of the carrier frequency.

These are possibly the most noteworthy features of modern amplitude-modulated transmitters, and the next few pages show how such a performance is obtained.

CLASSIFICATION OF AMPLIFIER TYPES

The economy so essential in high-power transmitters is largely obtained by very careful choice of operating conditions for the high-power stages. Thus, wherever possible, radio-frequency stages are operated in Class C, giving up to 80 per cent efficiency, and high-power audio-frequency amplifiers and modulated radio-frequency amplifiers are operated in Class B, giving an average efficiency of approximately 35 per cent. Class A operation is not used to any great extent because of its very low efficiency, although in some circuits such as series modulation it is essential. Before the transmitter circuits are considered in detail the chief properties of these three classes of valve operation will be briefly summarized.

Class A Operation

This is the type of amplification employed in most stages of audio-frequency amplifiers, and it is also used in some stages of transmitters. The valve is biased to near the mid-point of the linear portion of the

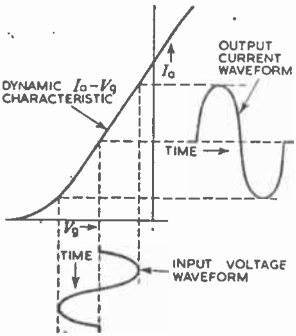


FIG. 2.—ILLUSTRATING CLASS A OPERATION.

I_a–V_g characteristic as shown in Fig. 2. Such amplifiers are used in transmitter circuits for two reasons :

(1) They introduce less distortion than Class B and Class C amplifiers, and are used in early radio-frequency and audio-frequency stages where harmonic distortion must be kept low, and where efficiency is of no great importance.

(2) The mean anode current of a Class A amplifier does not vary with amplitude variations of the input signal. There are certain stages of a transmitter, such as the modulator of a series-modulation circuit, where constant-current operation is essential and Class A amplifiers are used.

The objection to the use of Class A amplifiers elsewhere in transmitters lies in their poor efficiency, which seldom exceeds 25 per cent for a single valve. This means that the power delivered to the external load is only 25 per cent of that drawn from the H.T. source, the remaining 75 per cent being dissipated as heat in the valve. The power wasted as heat is thus three times that delivered to the load. Such an inefficient form of amplification would be very uneconomic and very inconvenient to use in high-power stages, uneconomic because the H.T.

source must supply four times the power required in the output load, and inconvenient because of the large physical size the valve must have to dissipate safely the power generated inside it. Artificial cooling is essential for high-power valves, and some of the methods adopted are described later. The more efficient an amplifier is made, the less is the power wasted as heat, and the smaller the valve can be. It is an interesting point that the power taken from the H.T. source does not change when the input signal is applied to a Class A stage, but the power lost as heat in the valve decreases by the amount supplied to the external load. In brief, the valve runs cooler when it is supplying a load; a Class B stage, on the other hand, runs hotter when it is working.

The theoretical efficiency of a perfect valve as a Class A amplifier can be calculated in the following way. If the H.T. supply is V volts and the anode current I amperes, the power taken from the H.T. source is VI watts. If the valve is perfect, the anode current can be

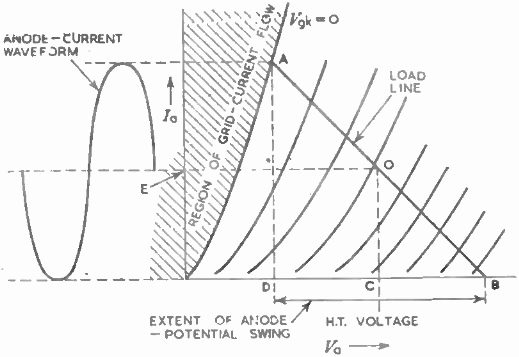


FIG. 3.—LIMITATION OF ANODE-POTENTIAL SWING BY GRID-CURRENT FLOW IN A CLASS A AMPLIFIER.

swung, without causing distortion, between 0 and 2 × I amperes, and if the load value is correctly chosen the anode potential will swing at the same time from 2 × V to 0 volts. Thus the peak value of the output current is I amperes, and that of the output voltage is V volts, giving the output power as ½ × IV watts, 50 per cent of the power taken from the H.T. supply.

In practice, such high values of efficiency are never attained and, as suggested above, 25 per cent is a good efficiency for a single Class A amplifier; it is, however, possible to obtain higher efficiencies from push-pull Class A stages.

The reason for the great disagreement between practical and theoretical efficiencies can be seen from Fig. 3, which shows a set of I_a–V_a characteristics for a triode and a load line A–B for a transformer-coupled load. The quiescent point is at O, the ordinate through which intersects the V_a axis at C, corresponding to the H.T. voltage. As shown, there is no very great difficulty in swinging the anode current between zero and twice the quiescent value E, but the lowest instan-

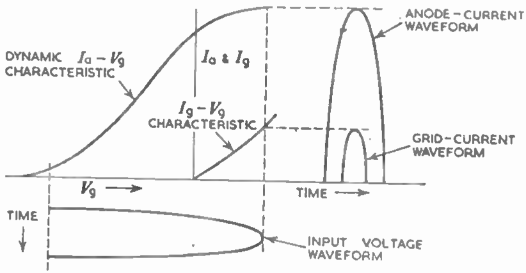


FIG. 4.—ILLUSTRATING CLASS B OPERATION.

taneous anode potential reached is D, which represents an appreciabl positive value. The anode potential cannot be swung to zero withou making the grid appreciably positive with respect to the filament, an such operation introduces considerable distortion. If the grid is kep negative with respect to the filament throughout the cycle of the inpu signal, the peak value of the anode-potential swing is limited to approxi mately one-half of the H.T. supply, and the efficiency is only one-hal of the theoretical maximum.

Class B Operation

The grid bias in a Class B amplifier is approximately equal to the cut-off bias (Fig. 4), and the input signal is great enough to cause considerable grid current. Thus the valve takes zero or very little anode current in the absence of an input signal, and the mean current rises when the signal is applied. The power taken from the H.T. supply—and the power dissipated as heat in the valve—both increase when the input signal is applied, this behaviour contrasting with that of a Class A amplifier.

The dynamic I_a–V_g characteristic of a single valve has, however, two regions of curvature which in practice cause distortion. The " bottom bend " near the point of anode-current cut-off is one such region, and the distortion caused by it is so serious that Class B amplifiers seldom, if ever, use a single valve. The effects of this curvature are to a large extent eliminated by using two valves in push–pull. The push–pull connection cancels even harmonics and results in a less-distorted output waveform. This is illustrated in Fig. 5, which shows two I_a–V_g curves for the individual valves of a push–pull pair. The effective characteristic for the stage can be obtained by adding the ordinates of the two curves, and is illustrated by the dotted line. This " composite characteristic " is clearly a better approximation to the ideal than either of the individual curves.

The second region of characteristic curvature, the " top bend ", occurs for positive grid voltages and is caused by grid current. This, in flowing through the resistance of the input-signal source, generates a voltage which opposes the input signal in polarity and reduces the effective input to the valve. Thus, as the grid is driven increasingly positive, the anode current falls more and more short of the expected

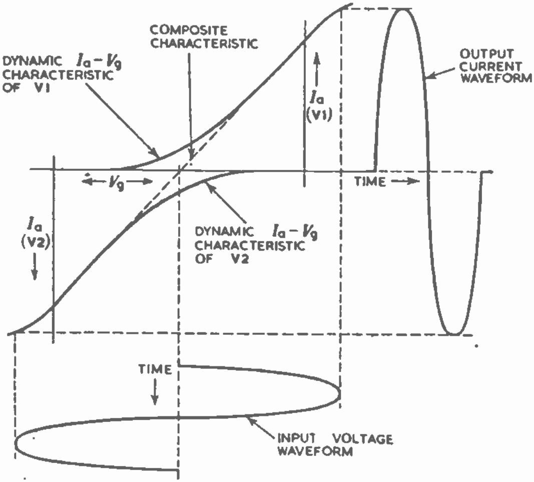


FIG. 5.—ILLUSTRATING OPERATION OF A CLASS B PUSH–PULL AMPLIFIER.

value and the non-linearity becomes more acute. In order to obtain high efficiency the anode potential must fall instantaneously almost to zero on positive peaks of the input signal. As shown in Fig. 3, this can only be achieved by driving the grid considerably positive with respect to the filament. Thus considerable grid current and the consequent "top bend" characteristic curvature are inevitable. This source of distortion is minimized in practice by arranging for the grid-signal source to be of low resistance so that the potential difference developed by grid current, in flowing through the source, is small compared with the signal amplitude. In general, the signal source resistance is made approximately $\frac{1}{5}$ the instantaneous input resistance of the Class B amplifier for a peak-amplitude input signal.

The fact that a Class B amplifier takes grid current implies, of course, that the previous amplifying stage, usually known as a driver stage, must supply appreciable power.

Push–pull Class B stages of this type with low external grid-circuit resistance can be used successfully for amplification of audio-frequency or modulated radio-frequency signals. When used as a modulator stage in a high-power transmitter the small harmonic distortion introduced is further reduced by negative feedback.

The theoretical maximum efficiency of a Class B amplifier can be calculated in the following way. Suppose the two push–pull valves have linear I_a–V_g characteristics and are biased precisely to cut-off, the H.T. supply being V volts. If the anode currents swing from zero to a peak value of I amperes and the anode potentials swing from V volts to zero, the output power is $\frac{1}{2} \times IV$ watts. This is the total power obtained from the two valves, each valve contributing one-half of each output sine wave. The mean anode current of each valve is I/π amperes, for the whole stage is $2 \times I/\pi$ amperes, and the power delivered by the H.T. supply is $2 \times IV/\pi$ watts. Thus the efficiency is given by $\frac{IV}{2} \times \frac{\pi}{2 \times IV} = \pi/4$ or 78·54 per cent. In deriving this figure it is assumed that the anode potential of each valve swings to zero volts. If this occurred in practice, the anode potential would momentarily be less than the grid potential, which is at its positive peak when the anode potential is at its negative peak. The input voltage is generally adjusted to give an anode potential swing equal to, say, 80 per cent of the H.T. voltage, this percentage being known as the " voltage utilization factor ". This reduction in anode swing causes an equal reduction in efficiency, which, for a " voltage utilization factor " of 80 per cent is thus $0{\cdot}8 \times 78{\cdot}54 = 63$ per cent. This is the efficiency for sinusoidal input signals of constant amplitude. For such signals, however, Class C amplification can be used to give higher efficiencies, and Class B operation is generally used for audio-frequency and modulated radio-frequency signals for which the amplitude is constantly varying, depending on the nature of the programme.

If the valve characteristics were straight, the efficiency would be independent of the input signal, the mean anode current being always a constant fraction of the peak value. Because of the "bottom bend", however, it is difficult to assess the precise value of grid potential at which the anode current is zero, and it is usual to operate the valves with a small standing anode current. For small inputs, therefore, the operating conditions approximate to those of a Class A amplifier and the efficiency is low. For larger inputs, the efficiency rises to a maximum of, say, 63 per cent as calculated above. In practice, the signal amplitude varies, and the efficiency varies with it. The average efficiency achieved over any given period depends on the fraction of the period which is occupied by large amplitude signals, and is thus again dependent on the nature of the programme. A typical value for the average efficiency of a Class B amplifier is 35 per cent.

The mean anode current varies with the magnitude of the input signal, but if the mean anode voltage falls appreciably as the anode current rises distortion will result. To keep the H.T. voltage constant, in spite of variations in anode current, the internal impedance of the H.T. source must be very low. One way of achieving this is by use of a mercury-arc rectifier with smoothing chokes and H.T. transformer secondaries of low resistance.

Class C Operation

The grid bias in a Class C stage is two or three times the cut-off bias (Fig. 6), and the input signal is large enough to give considerable grid current. The anode voltage utilization factor is approximately 80 per cent. Anode current flows only during a fraction of each cycle of radio-frequency input, and the anode current waveform is not similar to that of the input signal, but consists of a series of pulses occurring at the rate of one per cycle of input signal. The anode load consists of a parallel L–C circuit resonant at the frequency of the input signal. The impedance of such an anode load is high at the resonance frequency but low at other frequencies. Thus the valve gives good amplification at the frequency of the input signal but little amplifica-

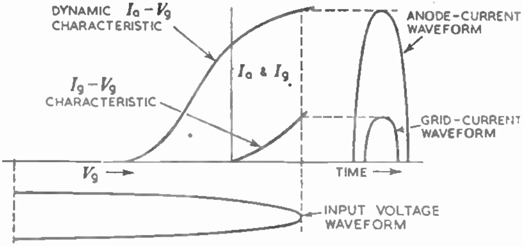


FIG. 6.—ILLUSTRATING CLASS C OPERATION.

tion at harmonic frequencies, and the waveform generated across the L–C circuit is a good approximation to a sine wave, in spite of the pulsating form of the anode current. An amplifier of this type can have a very high efficiency, approaching 85 per cent for a constant-amplitude, sinusoidal input signal, and such amplifiers are used in all high-power stages of a transmitter, such as modulated amplifiers and previous amplifiers, where the input is of constant amplitude. Class C stages may be single-sided or push–pull, but high-power modulated amplifiers are usually push–pull. The use of push–pull increases efficiency slightly, and has the useful property of giving very little output at even harmonics of the carrier frequency. Moreover, it is easier to cool two medium-sized valves than one large one.

Class C amplifiers cannot be used for amplification of modulated radio-frequency signals. If such signals are applied to such a stage, the modulation envelope of the radio-frequency output generated across the anode L–C circuit is distorted; in fact, there are gaps in the output (i.e., periods of zero signal) corresponding to periods of deep modulation of the input signal. The reason is that a Class C stage cannot respond to signals whose amplitude is insufficient to overcome the static bias and, when the input amplitude falls instantaneously to below this value (as it does during deep modulation) the valve momentarily loses its input.

We shall now consider the applications of these types of amplifier in the various stages of broadcasting transmitters, beginning with the generation of the carrier.

CARRIER SOURCES

Medium-frequency transmitters are generally used to radiate programmes within an area immediately surrounding the transmitter, and are usually required to operate on a particular carrier frequency for very long periods, possibly of several years. High-frequency transmitters, on the other hand, are usually employed to radiate programmes to distant areas, possibly on the opposite side of the earth, and, in order to give reliable reception throughout a period of many hours, it is necessary to make changes in carrier frequency possibly as frequently as several time per day. To meet these different requirements, two types of carrier source are required. One type, used with medium-frequency transmitters, gives exceedingly high stability but is not specifically designed to facilitate rapid changes in frequency. The other type is designed for use with high-frequency transmitters, and can quickly be reset to any frequency required; the frequency stability, though high, need not be of the exacting standard necessary in medium-frequency transmitters.

1. Carrier Sources for Medium-frequency Transmitters

Medium-frequency carrier sources contain a master oscillator, the frequency of which is controlled by a crystal. Certain crystals, notably those of quartz, tourmaline and Rochelle salt, exhibit piezo-electric properties; this is to say they undergo mechanical deformation if an electric potential is applied between opposite faces of a thin slab of the material. Conversely, they develop e.m.f.s when mechanically deformed, and, at the frequency of mechanical resonance, the electrical impedance measured between electrodes in contact with opposite faces, is similar to that of an L–C circuit with a Q-value which may be 100 times that obtainable from conventional inductors and capacitors. A high Q-value is an essential requirement in the frequency-determining element of an oscillator of high stability, and such crystals are clearly well suited to the needs of master oscillators.

The natural resonance of a crystal is inversely proportional to its thickness, and thus crystals can be ground to give resonance at any desired frequency within a certain range. The thickness of a crystal varies in general with its temperature, and the resonance frequency is thus frequency-dependent. To achieve the highest stability the crystal is usually mounted within a thermostatically controlled oven. The temperature of the oven is controlled within a narrow range, and the oven is usually fitted with an alarm device giving audible or visible indication if the temperature deviates by more than a predetermined amount from the normal value.

A simplified schematic diagram of a carrier source is given in Fig. 7. The master oscillator, which may contain more than one valve, is followed by a buffer stage which is included to isolate the oscillator from the following equipment to make the oscillator frequency independent of any variations in the value of the load connected to the carrier source. The carrier source also includes an A.G.C. system which is included for a number of reasons. In addition to stabilizing the output level of the carrier source, it keeps the oscillation amplitude at a value low enough to compel the oscillator valve(s) to operate in Class A conditions. This gives better frequency stability than if the valve is allowed to develop a large amplitude which would necessitate grid-current flow. It also ensures that the waveform generated has a very small harmonic content and is a good approximation to a sine wave.

Crystals can be ground to give resonance at any frequency in the medium-frequency range, and in most medium-frequency transmitters

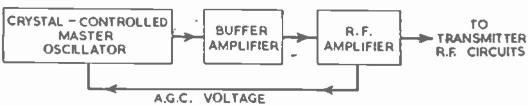


Fig. 7.—Block Schematic Diagram of a Typical Carrier Source for a Medium-frequency Transmitter.

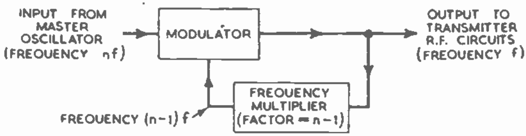


FIG. 8.—BLOCK SCHEMATIC DIAGRAM FOR ONE TYPE OF FREQUENCY DIVIDER.

the master oscillator can be connected directly to the radio-frequency circuits of the transmitter. There is no need, and indeed it would be practically impossible, to grind crystals to resonate precisely at a desired frequency. The frequency of the crystal-controlled oscillator can be adjusted within a very small range about the resonance value of the crystal by means of a small variable capacitor connected in series with the crystal in the master-oscillator circuit.

More detailed information on quartz oscillators will be found in Section 44.

Although transmitters operating at frequencies within the range 150–300 kc/s could also be controlled by a crystal-controlled master oscillator operating at carrier frequency, it is sometimes preferred to use a master oscillator at a higher frequency, in the region of 1 Mc/s, the carrier frequency being derived from this by means of a frequency divider.

Frequency Dividers

There are many types of frequency divider, and the block schematic diagram of one type of frequency divider used in transmitter equipment is shown in Fig. 8. The input to this divider is obtained from a master oscillator and has a frequency of nf, the divider output, which is used to feed the transmitter radio-frequency circuits, having a frequency of f. The factor by which the frequency is divided is thus n. The divider includes a modulator, one input to which is the frequency nf. The second input has a frequency of $(n - 1)f$ and is obtained from a frequency multiplier (giving a multiplication factor of $n - 1$), which is fed from the divider output. (Details of frequency multipliers are given later.) The modulator produces two outputs; one at a frequency equal to the difference between the frequencies of the two inputs. The other has a frequency equal to the sum of those of the two inputs. The difference frequency is given by $nf - (n - 1)f = f$. The sum frequency is given by $(n - 1)f + nf = (2n - 1)f$, which is far removed from f and can be rejected by use of an output filter tuned to f.

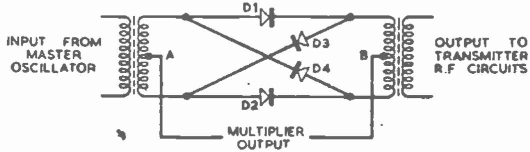


FIG. 9.—RING MODULATOR WHICH CAN BE USED IN THE DIVIDER CIRCUIT OF FIG. 8.

The modulator may be a ring type comprising four rectifiers as shown in Fig. 9. This is a balanced modulator, and e.m.f.s. applied between points A and B, the centre taps of the windings, do not appear in the output of the modulator. Nevertheless, when such an e.m.f. makes point A positive with respect to point B, rectifiers D1 and D2 conduct, allowing the input signals to reach the output. When point A is made negative with respect to B, rectifiers D3 and D4 conduct, again allowing the input signal to reach the output, but this time in reversed phase. When, as in a divider, the e.m.f. applied to A and B is itself sinusoidal, the input signal is reversed at the frequency of the multiplier output. This is an example of "commutator" modulation, and two of the products are signals at sum and difference frequencies as mentioned above.

The short-term stability of carrier sources employing a thermostatically controlled crystal is of the order of 1 part in 10^7, implying that the variations of a carrier at 1 Mc/s is $\pm 0{\cdot}1$ c/s, very much smaller than the 20 c/s mentioned above.

2. Carrier Sources for High-frequency Transmitters

If a high-frequency transmitter is required to operate on a single-carrier frequency only, the master oscillator can be controlled by a crystal as already described. However, for broadcasting purposes, it is not generally considered practical to use a crystal with a resonance frequency much higher than approximately 3 Mc/s, mainly because of the mechanical fragility of the crystal, which would need to be very thin to operate at higher frequencies. This difficulty is avoided by use of a master oscillator which operates at a frequency much lower than that of the transmitter. The crystal is ground to a submultiple of the carrier frequency, and the master oscillator is used to drive frequency multipliers, the output of which is fed to the transmitter radio-frequency circuits. For example, a transmitter to operate on 15·26 Mc/s could employ a crystal-controlled master oscillator at 954 kc/s, these frequencies having a ratio of 16 : 1.

If, however, a high-frequency transmitter is used to maintain a 24-hour service to a distant point, the carrier frequency must be changed at intervals. In general, each change in frequency necessitates a change of crystal and a change of multiplier ratio. At a site containing a number of high-frequency transmitters, each maintaining a service to a different part of the world, the total number of crystals required is very large. The need for crystals can be avoided, however, by use of high-Q L–C circuits and, provided precautions are taken, the stability of a carrier source employing an L–C circuit is adequate. L–C circuits are more flexible than crystals in that their resonance frequency can readily be changed over a wide range.

For satisfactory results using an L–C circuit as the frequency-stabilizing element, great attention must be paid to the construction of the inductor and capacitor, and both must be situated within a thermostatically controlled oven. To enable the variable capacitor to be adjusted to a given frequency a slow-motion drive is essential, and it must be free of backlash to permit a given frequency to be reset with accuracy. In general, the carrier frequency is derived from the output of the oscillator L–C by means of frequency multipliers.

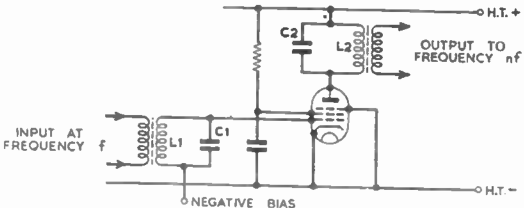


FIG. 10.—BASIC CIRCUIT FOR ONE TYPE OF FREQUENCY MULTIPLIER.

Frequency Multipliers

The simplest type of frequency multiplier consists of a Class C amplifying stage such as that shown in Fig. 10. The input has a frequency f, and the output nf, where n is an integer which in practice may have any value up to 5. The input-signal amplitude is considerably larger than the grid base of the amplifier, and the I_a–V_g characteristic of the amplifier is non-linear. The anode current of the valve consists of a succession of pulses which contain in addition to a component at the frequency of the input signal other components (harmonics) at frequencies which are multiples of that of the input signal. The required component is selected by tuning the output circuit L_2C_2 to the frequency of that component. The grid base of the amplifier can be made very small (to give efficient generation of harmonics) by using a pentode valve with a very low value of screen–grid potential.

In general, the amplitude of the harmonics decreases as the order increases, and there may be insufficient amplitude at, say, the tenth harmonic, to give adequate output. Moreover, the selectivity necessary to select the tenth harmonic but reject the ninth and eleventh may necessitate an impractically-high Q-value in L_2C_2. Thus the multiplication factor is generally limited to 5 or less, and higher factors are achieved by use of two or more multiplying stages in cascade. Thus a multiplication by 12 is obtained by use of multipliers with factors of 3 and 4.

TRANSMITTER RADIO-FREQUENCY STAGES

The power output of the carrier source is usually of the order of a few watts, and this is raised by subsequent radio frequency amplifiers in the transmitter proper to the value required in the aerial. The valves used in the carrier source, multipliers and dividers are generally small receiving types and, because of their low anode dissipation, no precautions are taken to secure economy in operation. In the succeeding radio-frequency stages, however, the valves are progressively larger as the power to be handled increases, and economy of operation becomes essential. Wherever possible, therefore, the valves are operated in Class C conditions, for which, under favourable conditions, 80 per cent of the power supplied to the anode from the H.T. source can be converted into useful power. The sequence of valves constituting the radio-frequency amplifier are power amplifiers and not voltage amplifiers as in the early stage of audio amplifiers; each valve has to supply the power necessary to drive grid current through the following stage. Early radio-frequency stages commonly employ tetrodes or pentodes, but later stages, particularly where the anode power exceeds approximately 500 watts, usually consist of triodes. To avoid instability, triode radio-frequency amplifiers are neutralized (as explained later), but their efficiency, under Class C conditions, is as high as that of pentodes, and their use avoids the difficulty of cooling the screen grid experienced with high-power pentodes.

Valve Cooling

The cooling of high-power valves raises a number of problems. Small valves dissipating up to approximately 500 watts do not, in general, require artificial cooling. If the construction of the transmitter is such that the air heated by the valve can readily escape by flowing upwards, and if cold air can flow in underneath the valve, the natural vertical flow of air over the valve envelope is sufficient to cool it adequately.

Larger valves handling power up to approximately 5 kW are artificially cooled by blowing a blast of cold air over the envelopes and, to make the cooling efficient, the air comes into direct contact with the anode which is arranged to form part of the envelope, glass parts of the envelope being joined to the metal anode by an air-tight seal. As a further aid to cooling, the anode is fitted with fins to give it a large surface area.

Valves dissipating power in excess of approximately 10 kW, such as those used in the final stages of high-power transmitters, usually require water-cooling in addition to air-blast cooling : in such valves the anode is constructed in the form of a water jacket through which cold water is circulated, the inner wall of the jacket forming part of the valve envelope. The cooling water comes into electrical contact with the anode and, since the latter may have a potential of up to 10 kV, precautions are taken to minimize radio-frequency power loss and H.T. leak through the resistance of the water path between anode and earth. Both losses can be reduced to negligible proportions by purifying the water (to increase its resistance) and by arranging that the length of the water path between anode and earth is very long. The latter is achieved by using long rubber or plastic hose-pipes to lead the water to and from the water jacket, the hoses being coiled up to conserve space. Water-cooled valves are also air-blast cooled, the air being directed against the sides of the glass envelope and in particular at the points where the filament leads pass through the glass walls.

The modern tendency in broadcasting transmitter design is to dispense with water cooling and to use air-blast cooling only. This is feasible, even for a high-power transmitter, if the high-power stages employ several relatively small valves instead of a few large ones. By this means the effective cooling surface is considerably enlarged and the air-blast cooling made more efficient. For medium-frequency transmitters each large valve can be simulated by a number of small valves in parallel. Such an arrangement leads to large input and output capacitances, however, and in a high-frequency transmitter it is difficult to secure high enough anode loads for satisfactory operation. In such transmitters, therefore, the tendency is to use push–pull circuits or inverted amplifiers, both of which lead to lower input and output capacitances.

Two circuit diagrams of transmitter radio-frequency stages are given

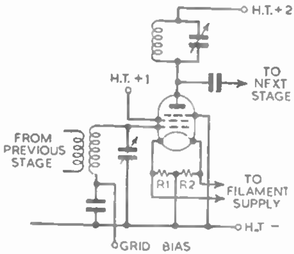


FIG. 11.—CIRCUIT DIAGRAM OF TYPICAL LOW-POWER TRANSMITTER RADIO-FREQUENCY STAGE USING A PENTODE.

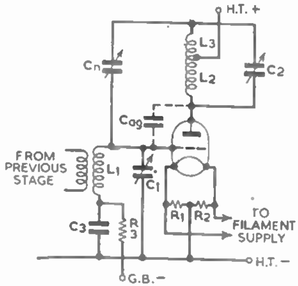


FIG. 12.—SIMPLIFIED CIRCUIT DIAGRAM OF TYPICAL HIGH-POWER TRANSMITTER CURRENT-FREQUENCY STAGE USING A NEUTRALISED TRIODE.

in Figs. 11 and 12. Fig. 11 shows a low-power stage employing a tetrode and, except for the directly-heated filament, is very similar to the type of radio-frequency amplifier used in receivers. The two resistors R_1 and R_2 are equal and have the effect of providing an artificial centre-tap of the filament for the H.T. negative connection. The resistors commonly have a value equal to five times the filament resistance (when hot). This circuit technique is extensively used in transmitters, and such resistors are indicated in most of the circuit diagrams of this section.

Neutralizing

Fig. 12 shows the circuit of a high-power stage using a triode. The anode–grid capacitance of triodes varies from approximately 2 pF in a small receiving type valve to 50 pF for a large transmitting valve; any value in this range is sufficient to cause instability at medium and high frequencies when the anode and grid circuits are tuned to approximately the same frequency. This tendency is eliminated in transmitter radio-frequency amplifiers by arranging the circuit in a balanced form in which the feedback from anode to grid via the anode–grid capacitance is counterbalanced (neutralized) by feedback to the grid of a voltage in anti-phase to the anode potential. The anti-phase voltage is obtained, as shown in Fig. 12, from the *L–C* circuit, which is connected to the anode and is effectively earthed at the point where the H.T. is introduced. The amplitude of the anti-phase voltage fed back to the grid is critical, and is adjustable by the capacitor C_n. The correct setting for C_n is found by experiment in the following manner. The H.T. is removed from the stage to be neutralized, the radio-frequency input to the valve from the previous stage being still present, and a lamp or some other radio-frequency-indicating device is clipped across some of the conductors of the coil L_2. When neutralizing is not perfect, some radio-frequency power is transferred to L_2–C_2 from L_1–C_1 via the anode–grid capacitance or the neutralizing capacitor, and the lamp or meter registers the presence of power. The neutralizing capacitor is adjusted to give minimum radio-frequency power in the anode circuit.

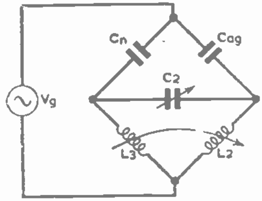


FIG. 13.—CIRCUIT OF FIG. 12 REDRAWN AS A BRIDGE CIRCUIT.

The circuit of Fig. 12 can be redrawn as a bridge circuit as shown in Fig. 13. When C_n is properly adjusted, signals applied to the grid (V_g) do not appear across C_2, i.e., there is no transfer of power from grid to anode circuits. If the coupling between L_2 and L_3 were perfect, the value of C_n for balance would be independent of frequency. In practice, however, the coupling is not perfect, and the neutralizing capacitor must be reset after any change in carrier frequency.

The value of neutralizing capacitance necessary to prevent instability depends on the position of the H.T. tapping point on the anode tuning inductor, increasing as the tapping point is moved away from the anode. In a radio-frequency stage using a relatively small triode, the position of the tapping point is usually so chosen that the neutralizing capacitance is at least several times the anode–grid capacitance of the valve. In this way a neutralizing capacitor of a reasonably large capacitance is required. In larger valves, however, where the anode–grid capacitance is of the order of 50 pF, the neutralizing capacitance can be equal to it. This requires the H.T. tapping point to be at the centre of the inductor, a requirement also necessitated by the use of push–pull radio-frequency amplification.

The use of neutralizing has a subsidiary benefit in high-frequency transmitters, namely that it reduces the effective input capacitance of the triode.

Grid Bias

The grid bias for the radio-frequency amplifier in Fig. 12 is derived from two sources. Part is obtained from a D.C. source, usually a generator, and the remainder is obtained from grid-current flow. The valve behaves to some extent as a grid-leak detector, and on the peaks of large positive-going signals takes grid current which, in flowing through C_3, charges it, the resulting potential between the plates biasing the grid negatively. During negative half-cycles the capacitor C_3 discharges into the resistor R_3, but C_3 is charged again on subsequent positive half-cycles. Provided the time constant R_3C_3* is large compared with the period of the radio-frequency input, the voltage across C_3 does not vary very greatly during the cycle, and the capacitor can be regarded as a source of relatively constant negative bias for the valve. The voltage of the bias so obtained is dependent on the amplitude of the

* The value of the time constant R_3C_3 is more critical in an amplifier of modulated radio-frequency. Here it must be small compared with the period of the highest audio-frequency in addition to being large compared with the smallest radio-frequency, the condition which is also required of the components in a diode detector.

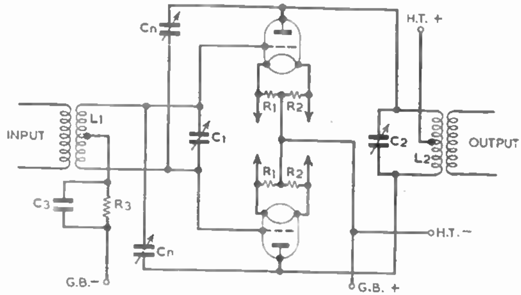


FIG. 14.—SIMPLIFIED CIRCUIT DIAGRAM OF TYPICAL HIGH-POWER TRANSMITTER RADIO-FREQUENCY STAGE USING TWO NEUTRALIZED TRIODES IN PUSH–PULL.

radio-frequency input to the valve, and if the drive falls the negative bias also falls. In this way the amplitude of the radio-frequency output from the valve is rendered more independent of variations in input than if fixed bias only were used.

Some fixed negative bias is, however, essential in addition to the bias derived from the grid current. If all the bias were obtained from grid current, and if the radio-frequency input to the valve failed, the valve would be without bias and an abnormally high anode current would flow. Such a large current is likely to impair the emission of the valve, and this possibility is avoided in high-power stages by arranging to have fixed bias approximately equal to the cut-off value for the valve to be present all the time.

Push–Pull Radio-frequency Amplifiers

Transmitter radio-frequency amplifiers are frequently of push–pull type, and a typical circuit diagram of a high power push–pull radio-frequency amplifier is given in Fig. 14. Push–pull is used for a number of reasons, some of which have already been mentioned. It gives a slight increase in efficiency and cancels even harmonics of the carrier frequency. The suppression of carrier harmonics generated in the radio-frequency amplifiers of a transmitter is very important, and there are international obligations to keep the harmonic content below a certain value. Some suppression is achieved by filters at the output stage, but the use of push–pull materially helps in this respect. It is essential to operate Class B amplifiers of modulated radio-frequency in push–pull to minimize envelope distortion. Push–pull is adopted in high-frequency transmitters, in preference to parallel operation, to reduce the capacitance thrown across tuning inductors, a low capacitance being essential to give high-anode-load values.

Inverted Radio-frequency Amplifiers

Because of the large physical size of high-power valves, the inter-electrode capacitances are quite high. A typical valve has $C_{ag} = 65$ pF, $C_{af} = 25$ pF and $C_{gf} = 45$ pF. Thus the effective output capacitance,

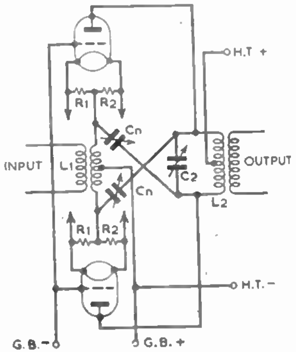


FIG. 15.—SIMPLIFIED CIRCUIT DIAGRAM OF TRANSMITTER RADIO-FREQUENCY STAGE USING TWO NEUTRALIZED INVERTED AMPLIFIERS IN PUSH–PULL.

even of a push–pull stage, incorporating two such valves is very considerable, amounting to more than 200 pF. Such a high capacitance introduces design difficulties in high-frequency transmitters, because it is almost impossible to present the valve with an anode load high enough to permit the stage to operate at high efficiency. This difficulty can be materially reduced by the use of an inverted amplifier.

Fig. 15 gives the essential features of a push–pull inverted amplifier. The control grids of the valves are earthed, and the input signal to each valve is applied between its filament and earth. The total capacitance due to inter-electrode and neutralizing capacitances across the anode-tuned circuit is less than half that obtained with a conventional push–pull circuit such as that shown in Fig. 14. In addition, the control grid acts to some extent as an electrostatic screen between input and output circuits, and capacitive coupling between the two is less than in conventional earthed-filament amplifiers. Nevertheless, it is still necessary to employ some neutralizing to give stability, and the capacitors C_n are included in Fig. 15 for this purpose. At high frequencies the lead used to connect the control grids to earth often has appreciable inductive reactance, and a capacitor is inserted in series with the lead in order to neutralize this reactance. This capacitor breaks the D.C. continuity of the grid–earth path, and a high-impedance choke is connected in parallel with the capacitor to restore the continuity.

An interesting feature of the circuit is that some power is transferred through the inverted amplifier from the driver stage to the output load of the inverted amplifier, and in practice this can amount to one-fifth of the power supplied by the inverted amplifier itself. This means, if an inverted amplifier is used as modulated amplifier, that it is impossible to achieve more than a certain depth of modulation, the (unmodulated) output of the driver stage always being present in the transmitter output. This is avoided by modulating the driver stage also.

LOW- AND HIGH-POWER MODULATION

So far in this section, methods of amplifying radio-frequency signals up to powers of approximately 100 kW have been described, but, in a transmitter, at some point in the radio-frequency chain the audio-frequency modulating signal must be impressed upon the carrier wave. The amplifying stage which finally introduced the modulating signal into the radio-frequency amplifier is described as the *modulator*, and the radio-frequency stage which is modulated is known as the *modulated amplifier*. Modulation can be carried out in the final radio-frequency stage or at an earlier stage. If it is introduced at an early point in the chain, the audio-frequency power needed for 100 per cent modulation

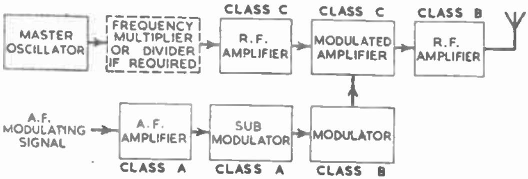


Fig. 16.—Block Schematic Diagram of a Low-power Modulated Transmitter.

is low, being in fact, equal to one-half of the radio-frequency power at the anode of the modulated valve. Such a system is termed *low-power modulation*, and a block schematic diagram of a complete transmitter employing low-power modulation is given in Fig. 16. The radio-frequency stages following the modulated amplifier are required to amplify modulated radio-frequency signals with negligible distortion, and cannot operate in Class C. Linear Class B amplifiers are necessary and, as previously mentioned, their efficiency is only approximately 35 per cent. This disadvantage largely offsets the advantage of the low audio-frequency power requirements.

When modulation is carried out in the final stage, the audio-frequency power required is considerable, being 50 kW to modulate a radio-frequency amplifier delivering 100 kW. Such a system is described as *high-power* modulation, and a block schematic diagram of a transmitter using high-power modulation is given in Fig. 17. It is difficult to generate such audio-frequency power economically, at the same time keeping harmonic distortion low. On the other hand, all the radio-frequency stages in a high-power modulated transmitter, including the modulated amplifier, can operate in Class C, and hence at high efficiency. Thus the disadvantages of the high-power modulator tend to be offset by the economy of the radio-frequency stages. There is therefore very little to choose between low-power and high-power modulation, and both systems have been used in high-power transmitters. The modern tendency seems, however, to be swinging in favour of high-power modulation.

Modulation Systems

Amplitude modulation is usually achieved by arranging for the gain of a radio-frequency amplifier to be proportional to the instantaneous value of the audio-frequency modulating signal. In most of the circuits

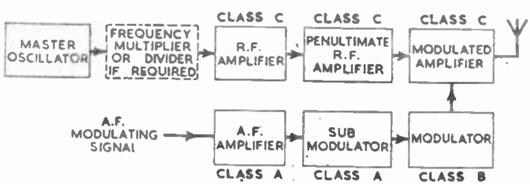


Fig. 17.—Block Schematic Diagram of a High-power Modulated Transmitter.

used for this purpose the radio-frequency signal is applied to one electrode of a valve and the audio-frequency signal to another; the anode current, being dependent on the voltage on both electrodes, contains modulated radio frequency. An obvious method is to apply the radio-frequency signal to the control grid of a pentode and the audio-frequency signal to the suppressor grid, a circuit with the advantage of requiring negligible audio-frequency power. Although such modulating circuits are used, the gain of the valve is not linearly related to the suppressor-grid voltage, and the distortion introduced precludes the use of such an arrangement in a broadcasting transmitter.

The most linear method of amplitude modulation is to apply the audio-frequency signal to the anode of the modulated amplifier, the radio-frequency signal being applied to the grid. This system, known as anode modulation, is extensively used in broadcasting transmitters in spite of the need for considerable audio-frequency power, the anode circuit of the radio-frequency amplifier representing quite a low impedance. The good linearity of such a system of modulation is at first

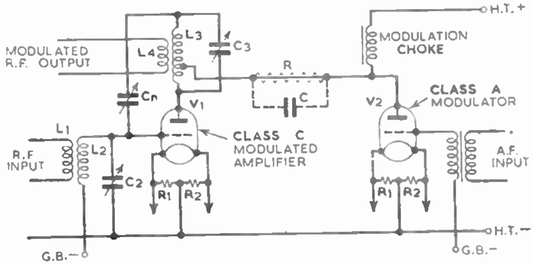


FIG. 18.—BASIC FORM OF HEISING OR CONSTANT-CURRENT MODULATION CIRCUIT.

sight somewhat surprising, because the static relationship between the anode voltage and anode current of a triode is not, in general, a very good approximation to a straight line. In a modulated amplifier, however, the conditions are dynamic, the grid being biased considerably beyond cut off, but having a large-amplitude radio-frequency signal impressed upon it which gives pulses of anode current. Provided the valve is thus operated under Class C conditions, the relationship between mean anode current and anode potential is very nearly linear and the radio-frequency output of the valve is directly proportional to the anode potential. By swinging the anode potential at an audio frequency, a linearly-modulated radio-frequency output can be obtained.

Heising or Constant-current Modulation

Fig. 18 gives the circuit of one method of obtaining anode modulation. V_1 is a Class C modulated amplifier and V_2 a Class A modulator, the two anodes being bonded and connected to the H.T. line by an audio-frequency choke, which for the purpose of explanation will be assumed of infinite impedance. The load impedance at V_2 anode is, in fact, the dynamic anode A.C. resistance of V_1 and, since the ratio of anode current to anode voltage for a Class C amplifier is linear, this is also equal to the D.C. resistance of V_1 (i.e., the ratio of H.T. voltage to mean anode current in the absence of an audio-frequency modulation signal). Because of this load the anode potential of V_2 (and hence V_1) varies when an audio-frequency signal is applied to V_2 grid and the radio-frequency output from L_4 is amplitude modulated. For positive voltages at V_2 grid, the anode current of V_2 increases, causing its anode voltage to fall. This causes an equal fall in V_1 anode potential and a consequent fall in V_1 anode current. Under ideal operating conditions the fall in V_1 anode current just counterbalances the increase in V_2 anode current, and the H.T. current for the two valves is steady during modulation; hence the name constant-current modulation.

There are a number of disadvantages to the simple circuit of Fig. 18. Firstly, it is impossible to achieve deep modulation without incurring serious distortion. To obtain 100 per cent modulation, the H.T. voltage at the anode of V_1 must be swung between zero and twice the H.T. supply voltage. It is quite impossible to obtain such a large amplitude signal from V_2 whilst still operating it under Class A condi-

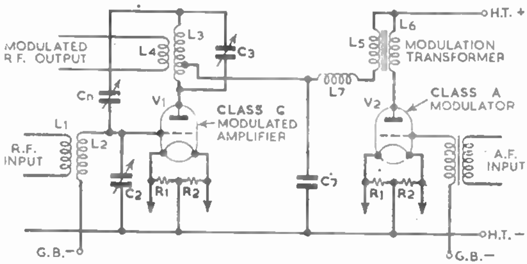


FIG. 19.—ANODE-MODULATION CIRCUIT INCLUDING A MODULATOR TRANSFORMER.

tions; in general, for the maximum possible grid swing of a Class A amplifier the peak anode swing is approximately equal to one-half the H.T. supply voltage. In the circuit of Fig. 18 this also causes the H.T. supply of V_1 to swing an equal extent, giving 50 per cent modulation. Any attempt to obtain greater than 50 per cent modulation by increasing the audio-frequency input to the modulator results in grid current in V_2 and severe distortion of the modulation envelope.

This difficulty can be overcome by including an RC combination in series with the H.T. feed to V_1 anode, as shown in dotted lines in Fig. 18. The voltage drop across R due to V_1 anode current ensures that the quiescent anode voltage of V_1 is less than that of V_2. By making the quiescent voltage at V_1 anode equal to approximately one-half that at V_2 anode, the modulator can achieve 100 per cent modulation, even though the potential swing at its anode is only half that of its H.T. supply. The resistor R is shunted by a capacitor C, the reactance of which at audio frequencies is low compared with the anode D.C. resistance of V_1. This condition ensures that the audio-frequency swing developed at V_1 anode is approximately equal to that at V_2 anode, and is not attenuated by R.

An alternative circuit permitting 100 per cent modulation with the

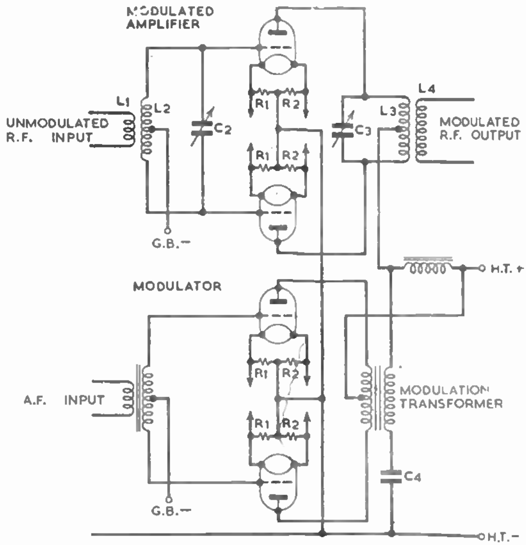


FIG. 20.—CLASS B MODULATOR WITH PARALLEL-FED MODULATION TRANSFORMER.

same H.T. supply to both modulator and modulated amplifier is shown in Fig. 19. In this the two valves are coupled by means of a modulation transformer and, if L_5 is greater than L_6 the audio-frequency voltage swing at the anode of V_1 is greater than that at the anode of V_2, thus making full modulation possible. For example, the transformer might require a 2 : 1 turns ratio to achieve full modulation. In such a circuit the effective load at V_2 anode is one-quarter the anode resistance of V_1, and the valves should be chosen so that this is at or near the optimum value for V_1. The incremental inductances of L_5 and L_6 should be such that their reactances at the lowest modulation frequency (say 30 c/s) is at least several times the resistance to which the winding is connected. This may necessitate a bulky component, particularly if the transformer core is polarized by the steady components of the anode currents of the two valves. This polarization can, however, be minimized by so connecting the transformer that the effects of the two steady components are in opposition. The components L_7C_7 are included to prevent radio-frequency currents generated in the modulated amplifier from entering the modulator.

Class B Modulator

In high-power transmitters it is customary to use a Class B modulator, and Fig. 20 gives a simplified circuit illustrating such a stage and also a push–pull modulated amplifier. Neutralizing is omitted for the sake

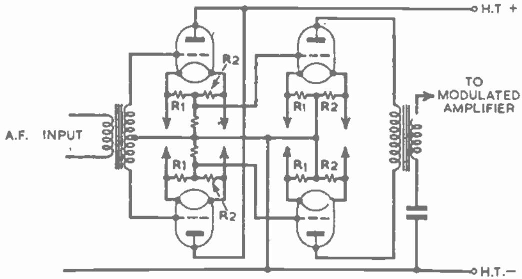


FIG. 21.—CATHODE FOLLOWERS USED AS DRIVERS OF A CLASS B MODULATOR. FOR SIMPLICITY SOURCES OF GRID BIAS ARE IGNORED.

of simplicity. This circuit may be regarded as typical of those used in the final stage of a high-power modulated transmitter. The use of push–pull in the modulator avoids steady polarization of the modulation transformer from the primary winding, because the steady components of the modulator anode currents flow in it in opposite directions. By using choke-capacitance coupling between the secondary winding and the modulated amplifier, the D.C. components of the anode currents of the modulated amplifier are removed from the secondary winding, avoiding polarization from this cause. Thus the size of the modulation transformer can be kept reasonably small.

An interesting point about this circuit is the voltage rating of the capacitor C_4. The potential across this capacitor is made up of the steady component from the H.T. supply and the varying a.f. component from the modulation transformer secondary. The latter is alternately aiding and opposing the former and, under 100 per cent modulation conditions, can equal twice the H.T. supply voltage, and the capacitor rating must exceed this (possibly 20 kV) by an adequate safety margin.

As explained earlier, the source of audio-frequency signal feeding the modulator grids must be of low resistance to minimize distortion on large-amplitude signals, and one way of achieving a low-resistance source is indicated in Fig. 20. The grids are fed from a transformer with a centre-tapped secondary winding of low resistance. An alternative circuit is shown in Fig. 21, in which each of the modulator valves is fed from a cathode follower via a low-value cathode resistor, which is sometimes as small as 20 ohms. This circuit is probably preferable to the use of a transformer if the Class B stage and the driver stage are included within a negative feedback loop. Wherever possible it is desirable to eliminate iron-cored components such as chokes and transformers from those sections of an amplifier over which it is desired to apply negative feedback. Such components introduce phase shift at the extremes of the passband and, when feedback is applied, instability sometimes occurs as a result of these phase-shifts. RC networks also introduce phase shifts, of course, but in general the performance of an RC circuit is more amenable to calculation than that of iron-cored components. On the other hand, negative feedback is sometimes used in order to reduce the harmonic distortion introduced by, say, a modulation trans-

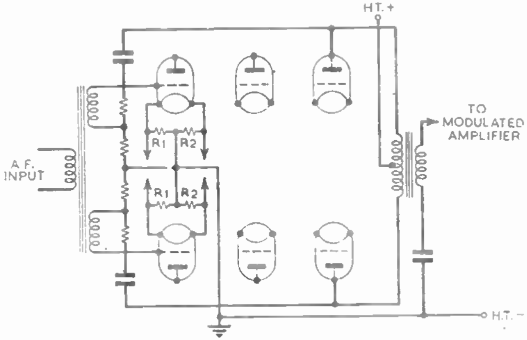


FIG. 22.—ONE METHOD OF APPLYING NEGATIVE FEEDBACK TO A MODULATOR.

former; in such a circuit the constants of the feedback network must be chosen to avoid any possible instability, and this may well be a very difficult task.

Sometimes negative feedback is applied to a section of the modulator by means of potential dividers connected between the anodes of the output stage and the grids of an earlier stage, as shown in the simplified circuit diagram of Fig. 22. In such a circuit the modulation transformer is not wholly included within the feedback loop, and the feedback circuit does nothing towards reducing any harmonic distortion introduced by this component. Moreover, feedback does not correct any deficiencies in frequency response (" top loss ") due to the leakage inductance of the modulation transformer; it does, however, reduce any attenuation distortion (" bass loss ") caused by inadequate primary inductance.

A more ambitious feedback circuit sometimes employed is that illustrated in Fig. 23. A detector is coupled to the output of the transmitter and rectifies some of the modulated radio-frequency output. The resulting audio-frequency signal from the detector is returned to an early stage of the modulator as a negative feedback voltage. Such feedback tends to correct harmonic and attenuation distortion occurring in the modulator output stage or in the modulated amplifier; in other words, it tends to linearize the modulation process. Such a feedback system needs very careful design; the detector circuit must be free from distortion, and the constants of the feedback loop must be very carefully planned to avoid instability.

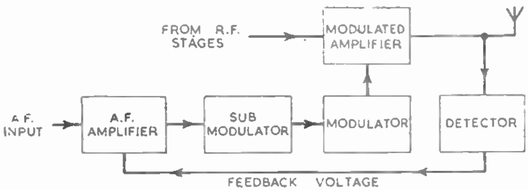


FIG. 23.—METHOD OF APPLYING OVERALL FEEDBACK TO A TRANSMITTER.

Series or Constant-voltage Modulation

The necessity for a modulation choke or transformer can be avoided by the use of a circuit such as that shown in Fig. 24, in which the modulator and modulated amplifier are connected in series across the H.T. supply. The modulated amplifier is usually a Class C stage, and modulation is achieved here, as in Heising modulation, by varying its anode H.T. supply voltage. The anode potential is swung above and below its quiescent (unmodulated) value, and the average current of the modulated amplifier is steady during modulation. In a series modulation circuit the anode current of the modulated amplifier is equal to that of the modulator, which must therefore also have constant anode current during modulation. The modulator must therefore operate under Class A conditions. Because of the low efficiency of such amplifiers, series modulation is not generally used in high-power modulated transmitters. It is more generally used as the penultimate stage of a low-power modulated transmitter.

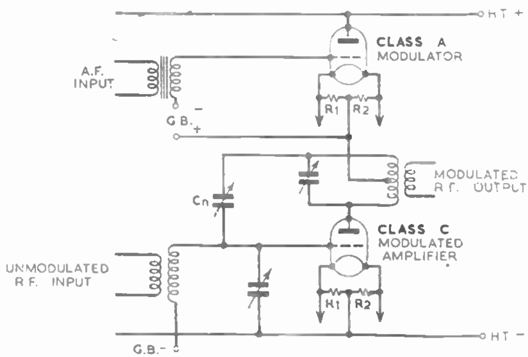


Fig. 24.—Basic Features of Series Modulation Circuit.

The modulator anode load is the anode A.C. resistance of the modulated amplifier, and the two valves are so chosen that the load is near the optimum value for the modulator. There are two possible forms for this circuit; the modulated amplifier can be "below" the modulator as in Fig. 24, or it can be "above" it. Both circuits operate in the following manner. The relationship between anode potential and anode current for a Class C stage, such as the modulated amplifier, is approximately linear, and it behaves as a linear anode load for the modulator. When an audio-frequency signal is applied to the grid of the modulator, an amplified audio-frequency signal is generated across the modulated amplifier, thereby modulating the radio-frequency output of this stage. The sum of the voltages across the modulated amplifier and the modulator is constant during modulation, being at all times equal to the H.T. voltage. For this reason this circuit is sometimes known as constant-voltage modulation, in contrast with the Heising system, which is a constant-current device.

For reasons which have already been explained, the peak audio-frequency output developed at the anode of a Class A amplifier is necessarily less than that of its H.T. supply. Moreover, to achieve 100 per cent modulation, the H.T. voltage for the modulator must be swung between zero and twice its quiescent (unmodulated) value. Thus, under steady carrier conditions, the H.T. voltage for the modulator must exceed that for the modulated amplifier, and it is commonly double that of the modulated amplifier. For example, if the total H.T. supply is 18 kV, the quiescent voltage across the modulator could be 12 kV and that across the modulated amplifier 6 kV. On a peak positive input to the modulator the voltage across this valve falls to 6 kV and that across the modulated amplifier rises to 12 kV; on a peak negative input the voltage across the modulator rises to 18 kV and that across the modulated amplifier falls to zero. On a negative input to

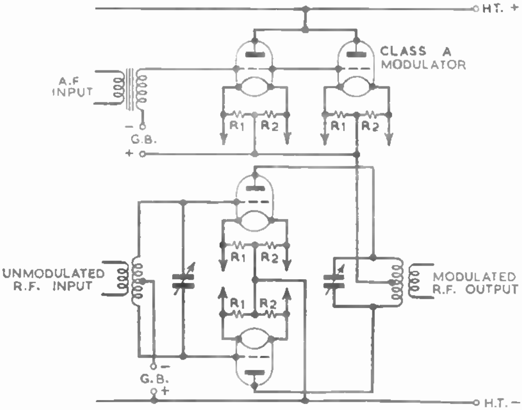


FIG. 25.—SERIES MODULATION CIRCUIT INCLUDING TWO VALVES IN PARALLEL AS A MODULATOR AND A PUSH–PULL MODULATED AMPLIFIER. NEUTRALIZING IS OMITTED FOR SIMPLICITY.

the modulator, the anode current of both valves falls instantaneously to zero. Some characteristic curvature is inevitable when the anode current of the Class A modulator falls to zero, and thus some harmonic distortion is inevitable on 100 per cent modulation. This distortion can, however, be considerably reduced by a negative-feedback circuit described later.

Another difficulty of this circuit is associated with the fact that the filament of the modulator or modulated amplifier must be at high potential with respect to earth. This necessitates insulation of the filament generator and the associated leads and meters. It is generally preferred to have the modulated amplifier at the lower potential, because this avoids the insulation difficulties which arise when the anode-tuned circuits are at the full H.T. potential (and this potential is generally very high, possibly 20 kV, for a series modulation circuit). In practice, the modulator may consist of two or more valves in parallel and the modulated amplifier of two valves in push–pull. Fig. 25 gives a simplified circuit diagram of a series-modulator including a total of four valves.

In Figs. 24 and 25 the secondary winding of the modulator input transformer is returned, via the grid-bias supply, to the modulator filament. This circuit has the merit that only a comparatively small modulating voltage is required, the modulator behaving as an amplifier. In an alternative version of the circuit, shown in Fig. 26, the modulating signal is applied between modulator grid and H.T. negative. This modification brings about an entirely different set of operating conditions which lead to a much better performance, particularly in respect of linearity of modulation. The modulator now behaves as a cathode follower, the modulated amplifier acting as a resistive cathode load. Because of the negative feedback, the gain of the modulator is now smaller and is slightly less than unity. The audio-frequency signal required at the modulator grid for 100 per cent modulation is now very

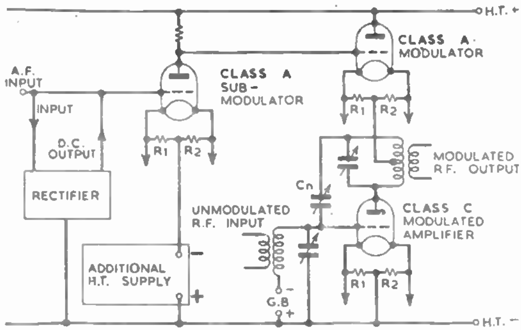


FIG. 26.—SERIES MODULATION CIRCUIT INCLUDING CATHODE-FOLLOWER MODULATOR AND PROVISION FOR FLOATING-CARRIER OPERATION.

much greater than for the circuit of Figs. 24 and 25. One way of applying this signal is indicated in Fig. 26, in which the modulator grid is connected directly to the anode of the sub-modulator, the D.C. conditions in the circuit being so chosen that the modulator grid bias is correct. To achieve 100 per cent modulation the mean anode potential of the modulated amplifier must be reduced instantaneously to zero. This requires the grid potential of the modulator, and hence the anode potential of the sub-modulator, to be reduced instantaneously to less than zero. This can only be achieved by returning the filament of the sub-modulator to a point which is considerably negative with respect to H.T. negative; as shown in Fig. 26, a subsidiary H.T. supply is used for this purpose.

Floating-carrier Operation

A modulation circuit of the type shown in Fig. 26 can very simply be modified to give floating-carrier operation. This is a mode of operation sometimes used to effect some economy in high-power-transmitter operation. During amplitude-modulation the power supplied by the modulator goes into sidebands and the modulated amplifier maintains the carrier wave at constant amplitude. During periods of shallow modulation the carrier amplitude is unnecessarily large, and economy in operation can be secured if the carrier amplitude is varied so that at any instant it is always just sufficient to accommodate the amplitude of the modulating signal. In effect, this means that the transmitter is maintained at 100 per cent modulation, and the carrier amplitude is varied in accordance with the average amplitude of the modulating signal. As already pointed out, distortion is a maximum for deep modulation, and for this reason and also for a reason connected with receiver operation, this floating-carrier technique is not carried to its logical conclusion. In practice, arrangements may be made for the carrier amplitude to vary between a half and its full amplitude, and it can be achieved using the circuit of Fig. 26 by rectifying the modulating signal and feeding the resultant undirectional voltage to the grid of the submodulator in addition to the modulating signal itself. The rectifier may be of the biased-diode type, delivering a positive-going output to the submodulator grid to increase the carrier amplitude for audio-frequency input signals exceeding a certain value. The forward time-constant must be short to enable the carrier amplitude to rise rapidly to cope with steep transients, but the decay-time constant can be long, and may be as much as 1 second.

A disadvantage of floating-carrier operation is that its use affects the performance of receivers fitted with automatic gain control. In such receivers the radio-frequency gain is dependent on the amplitude of the received carrier and, if this is varied at the transmitter, the receiver gain will fluctuate, giving rise to undesirable variations in volume of the received programme. To minimize such effects the variations in carrier amplitude are limited as described.

Adjustment of Depth of Modulation in Transmitters

When a carrier is 100 per cent modulated by a sinusoidal signal the power radiated increases by 50 per cent. Thus the aerial current increases to $\sqrt{(1{\cdot}5)} = 1{\cdot}225$ of its unmodulated value. This increase of 22·5 per cent is often used as a means of determining when full modulation is reached, and by use of a gain control calibrated in decibels in the modulating-amplifier chain, it is possible to set up any desired modulation depth with reasonable accuracy. For example, suppose it is desired to set up a broadcast transmitter so that, with a certain level of audio-frequency input, the modulation depth is 40 per cent. This input, whatever its value, is 8 db below that which gives 100 per cent modulation. Thus the desired modulation depth is obtained by increasing the gain of the audio-frequency amplifier feeding the modulator until the aerial current meter indicates an increase of 22·5 per cent. The audio-frequency gain is now reduced by precisely 8 db and the required modulation depth is obtained. This method is more accurate than attempting to read the increase in aerial current due to 40 per cent modulation. This increase is very small, being only 4 per cent, and is very difficult to read with accuracy.

Adjustment of Radiated Power of a Transmitter

It is sometimes necessary to reduce the radiated power of a transmitter, perhaps because it is causing interference at some distant point or because it is operating with a temporary aerial which will not withstand the full power. The adjustment can be carried out in the following way. Suppose it is desired to reduce the power to one-half. The H.T. supply to the modulated amplifier is reduced to $1/\sqrt{2}$, i.e., 0·71 of its normal value. Since the anode current is proportional to the anode potential for a Class C stage, the anode current also falls to $1/\sqrt{2}$ of its normal value, and the total power taken by the stage from the H.T. supply falls to one-half of normal. It is assumed that the efficiency of the modulated amplifier remains constant, and thus its power output falls to one-half. The power output of the modulator must also be reduced to one-half, otherwise distortion due to over-modulation (carrier cutting) will result, and the necessary reduction can be obtained by reducing the gain of the audio-frequency amplifier feeding the modulator by 6 db. The adjustments necessary to reduce the radiated power to other fractions of the normal output can be determined from the above example.

S. W. A.

FREQUFNCY-MODULATED TRANSMITTERS

A frequency-modulated transmitter is one in which the amplitude of the radiated signal remains constant and in which intelligence is transmitted by modulating the frequency about its mean value. Degree of modulation from 0 to 100 per cent modulation is determined by the extent to which the frequency is moved from tho mean value, and is called the deviation.

Thus for a sinusoidal modulating signal the transmitted signal is

$$e = E_0 \sin (wt + m \sin pt)$$

where E_0 is the peak amplitude of the signal;

$w = 2\pi f_0$, where f_0 is carrier frequency;

$p = 2\pi f_m$, where f_m is modulating frequency;

m = modulation index

$$= \frac{\text{Deviation of frequency from } f_0}{f_m} = \frac{\Delta f_0}{f_m}.$$

The spread of side-bands is not exclusively related to the modulation frequencies as in amplitude modulation, but is also dependent to a great degree on the modulation index, and usually extends to many times the extent of the modulating frequency.

Thus, in order to convey the same modulation information, the bandwidth of a frequency-modulated transmitter must be many times greater than that of the corresponding amplitude-modulated transmitter.

With a modulation index of less than 0·5, the frequency deviation is less than half the modulating frequency, and the band-width required to accommodate the essential part of the side-band spectrum is the same as that for amplitude modulation.

When the modulation index is greater than 0·5 there are significant frequency components extending on each side of the mean carrier

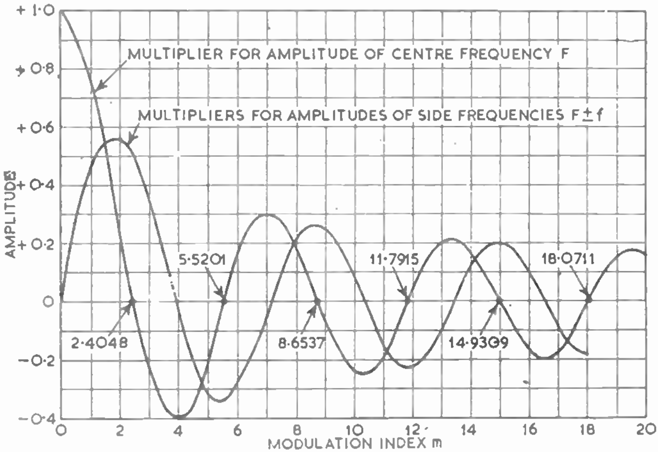


FIG. 27.—VARIATION OF CARRIER AND SIDEBAND AMPLITUDES IN FREQUENCY-MODULATED SYSTEM.

frequency between $(m + 1)$ times the modulation frequency and $1{\cdot}5\,(m + 1)$ times the modulation frequency.

For broadcast transmitters it is usual to employ a modulation index of at least 5, which necessitates a radio-frequency band-width of approximately twelve to sixteen times the modulation frequency.

A further difference exists; whereas, in amplitude modulation, the amplitude of the individual side-bands can never exceed 0·5 of the carrier amplitude, in frequency modulation the side-band amplitudes can exceed that of the carrier, and depend again on the modulation index. Conditions can be such that the carrier amplitude is zero. This occurs when the modulation index is 2·4048, 5·5201, etc.

The distribution of side-band amplitude and carrier amplitude for various modulation indices is illustrated in Fig. 27.

By the nature of the frequency-modulated signals, in which the modulation index varies inversely with the modulation frequency, the relative amplitude of noise interference varies directly as the frequency. This means that the maximum noise energy on frequency modulation occurs at the highest frequency, where the modulation is normally small.

Profit can therefore be made by employing increasing amplitude of signal at higher frequencies, or pre-emphasis at the transmitter and corresponding de-emphasis at the receiver.

In order to effect frequency modulation, it is necessary to modulate the source of frequency, i.e., at low power level. A frequency-modulated transmitter consists, therefore, of a modulated signal source followed by a chain of wide-band radio-frequency amplifiers.

Before proceeding to a discussion of the problems involved and to detailed circuit arrangements, it is necessary to have a specification of overall performance of a system. Performance specifications vary in detail, depending upon the country of origin and the type of service envisaged. As a guide, the performance specification of the British frequency-modulated broadcast service is given below.

Performance Specification for Frequency-modulated Transmitters

Frequency Stability.—The stability of the centre frequency without modulation shall be within ± 20 parts in 10^6. When modulated with a maximum deviation of 100 kc/s the mean frequency shall be constant within ± 5 parts in 10^6.

Deviation.—± 75 kc/s for normal operation. ± 100 kc/s optional.

Input Signal Level.—1 mW + 12 db in 600 ohms.

Pre-Emphasis.—Selected time constants from 0 to 50 microseconds. The normal working value to be 50 microseconds.

Audio-frequency Distortion.—Maximum distortion at deviations specified:

30–60 c/s	1 per cent for 25 kc/s deviation 1·5 per cent for 75 kc/s deviation 3 per cent for 100 kc/s deviation
60 c/s–10 kc/s	0·5 per cent for 25 kc/s deviation 1·0 per cent for 75 kc/s deviation 2·0 per cent for 100 kc/s deviation

Frequency Response without Pre-emphasis.—Level $\pm 0{\cdot}5$ db with respect to 400 c/s from 30 c/s to 10 kc/s at 75 kc/s deviation.

Frequency Response with Pre-emphasis (and Corresponding De-emphasis in Monitor).—

Level	*Max Tolerances*
$\pm$ 0·5 db	60 c/s to 3 kc/s
$\pm$ 1 db	30 c/s to 10 kc/s
$\mp$ 3 db	12 kc/s
$-$10 db	15 kc/s

all at 75 kc/s deviation.

Noise.—Frequency modulation noise level shall be better than -60 db relative to output at 400 c/s at 75 kc/s deviation. Amplitude modulation noise level for all frequencies between 30 c/s and 10 kc/s shall be better than -50 db relative to the unmodulated carrier.

Amplitude Modulation.—Not to exceed 2 per cent for 75 kc/s deviation.

Amplitude Stability.—The amplitude of the frequency-modulated signals shall be constant within ± 2 per cent with respect to the unmodulated carrier for any frequencies between 30 c/s and 10 kc/s.

Radio-frequency Harmonics.—Power at harmonic and sub-harmonic frequencies shall be less than 200 mW.

Methods of Modulation

A variety of methods of effecting frequency modulation have been used, and date from the early work of the late E. A. Armstrong.[1] The two basic methods are "direct frequency modulation" and "pre-corrected phase modulation ". In order to appreciate the two basic systems, the relation between frequency and phase modulation must be clearly understood.

Relation between Frequency and Phase Modulation

Suppose we have an L–C circuit oscillating at a frequency f_0 and we vary the capacitance across the circuit : there will be a corresponding variation of frequency. If we denote the change of frequency by the symbol Δf and assume a sinusoidal modulation frequency $f_m = \frac{p}{2\pi}$, causing the corresponding variation of capacitance, we may express this as an equation

$$f = f_0 + \Delta f \sin pt \quad . \quad . \quad . \quad . \quad (1)$$

The modulating frequency f_m may have any value, but as long as the maximum variation of oscillatory circuit capacitance, caused by the modulation, is the same, the maximum frequency deviation Δf will be constant. This is frequency modulation.

Now let us consider a stable frequency oscillating tuned circuit followed by an isolating amplifier and a driven tuned circuit. If now we vary the capacitance across the driven circuit, there will be no change of frequency, as this is determined by the unchanged oscillatory circuit. However, there will be a change of phase of the current in the driven circuit.

As before, we may assume the change to take place sinusoidally and express the condition as one of constant frequency f_0 with phase varying as, say, $\Delta\phi \sin pt$. This is phase modulation.

It is convenient to show the equivalence of phase and frequency modulation by means of vector diagrams.

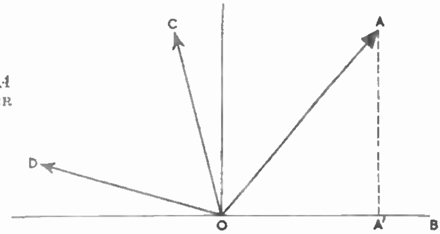
Fig. 28 shows a vector OA corresponding to the carrier frequency f_0. Conventionally this vector is assumed to be rotating counter-clockwise at a constant angular velocity $w = 2\pi f_0$ radians, and its instantaneous amplitude OA' is given by its projection on to the datum reference OB.

For a frequency f_0, the vector OA obviously makes f_0 revolutions a second. For constant frequency applications, all vectors such as OA, OC, OD keep the same angular relation, and we may conceive the three vectors as remaining fixed and the original datum OB rotating clockwise at the constant frequency.

In Fig. 29 let the fixed reference vector OA represent a carrier frequency f_0, and let the vector OB represent a varying frequency. The vector OB will thus rotate relative to the fixed vector OA at a rate determined by the difference in frequency between f_0 (vector OA) and its own frequency, and in a direction dependent upon whether its frequency is greater or less than f_0.

Thus, if the frequency of the vector OB is $f_0 + 1$ c/s the vector OB will rotate counter-clockwise at 1 revolution/second, while if it is $f_0 - 1$ c/s, it will rotate clockwise at 1 revolution/second. It is also true that the vector OB equally represents a carrier of frequency f_0 with phase varying according to the speed of rotation of OB. Thus, for the example used above, where OB makes 1 revolution/second, it may represent a frequency f_0 with phase varying at the rate of 2π radians/second.

FIG. 28.—THE VECTOR OA CORRESPONDS TO THE CARRIER FREQUENCY, f_0.



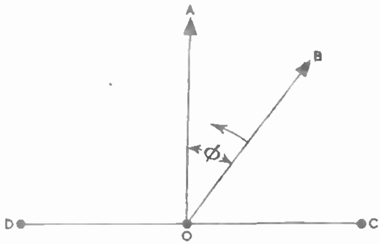


FIG. 29.—VECTOR OA REPRESENTS A CARRIER FREQUENCY, f_0, AND VECTOR OB A VARYING FREQUENCY.

Thus, if the frequency difference between OB and OA is Δf, the equivalent phase change for constant frequency f_0 will be $2\pi\Delta f$ radians/second or $\frac{2\pi\Delta f}{f_0}$ radians/cycle.

Now let us consider the special case of square-wave frequency modulation illustrated in Fig. 30 (a).

This corresponds in Fig. 29 to the vector OB rotating at constant angular velocity from OC to OD in a counter-clockwise direction and then reversing immediately and travelling clockwise back from OD to OC at the same constant angular velocity corresponding to 1 revolution/second. This corresponds to a maximum phase change of $\pm\frac{\pi}{2}$ relative to OA, which, for constant angular velocity of the vector, is linear. Thus the phase change or phase modulation corresponding to the square-wave frequency modulation of Fig. 30 (a) is a triangular phase modulation as at Fig. 30 (b).

If we now decrease the periodicity of the square wave to $\frac{1}{2}$ but maintain the ± 1 c/s deviation, it follows that the maximum phase shift is

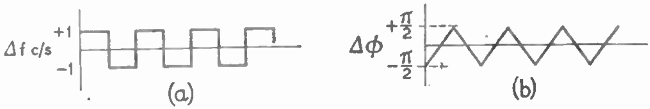


FIG. 30.—(a) SQUARE-WAVE FREQUENCY MODULATION; (b) TRIANGULAR PHASE MODULATION.

halved, i.e., the maximum phase deviation is $\pm \frac{\Delta f}{f_m} \cdot \frac{\pi}{2}$ radians, where Δf is the maximum frequency deviation.

If the frequency deviation of OB varies sinusoidally, the maximum deviation being 1 c/s, the average angular velocity of OB will be $2/\pi$ of the maximum velocity.

Hence the maximum phase shift will be $\frac{2}{\pi} \times \pm \frac{\pi}{2} = 1$ radian/second.

As is clear from the vector relations established, if the change of OB from OC to OD is sinusoidal the maximum frequency deviation will occur when OB is passing the central position OA, i.e., when the phase deviation is a minimum. Likewise, the maximum phase deviation occurs at OC and OD when the frequency deviation is a minimum.

Thus, taking the vector OA as the time datum,

Frequency deviation $= \Delta f \cos pt$ c/s . . . . . . (2)

Equivalent phase deviation $= \frac{\Delta f}{f_m} \sin pt$ radians/second . . (3)

$$(p = 2\pi f_m)$$

Including the carrier, it is clear that the expressions for frequency and phase modulation are:

$$f = A \sin\left(2\pi f_0 t + \frac{\Delta f}{f_m} \sin pt\right) \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

and

$$f = A \sin\left(2\pi f_0 t + \Delta\phi \sin pt\right) \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

This factor $\frac{\Delta f}{f_m}$ (or $\Delta\phi$) is the modulation index. Whereas for frequency modulation, the modulation index $\frac{\Delta f}{f_m}$ is inversely proportional to modulation frequency, in phase modulation, as defined, the modulation index is constant. To make phase modulation equivalent to frequency modulation, therefore, we must introduce a factor inversely proportional to frequency, as implied in equations (3) and (4).

Summarizing, we find:

For Frequency Modulation (F.M.), Δf is constant

$$m \text{ is } \frac{\Delta f}{f_m}$$

For Equivalent Phase Modulation, we start with

$$\Delta\phi \text{ is constant} \qquad \therefore \quad \Delta f \propto f_m$$

so we include a circuit to introduce a factor $\frac{1}{f_m}$

Hence $$\Delta f \propto f_m \times \frac{1}{f_m} = \text{constant}$$

as for F.M.

For Phase Modulation (P.M.),

$$\Delta\phi \text{ is constant} \qquad \therefore \quad \Delta f \propto f_m$$

$$\therefore \quad m = \frac{\Delta f}{f_m} = \text{constant}$$

To change phase modulation into equivalent frequency modulation we must make the modulating signal inversely proportional to f_m.

The pre-correction of the signal is usually accomplished in a network in the path of the modulating signal.

Pre-corrected Phase Modulation

One method to obtain phase modulation is to employ a reactance modulator valve system across an L–C circuit that is excited at constant frequency. To preserve linearity of modulation, phase shift must be limited to say $\pm\frac{1}{2}$ radian. For frequency deviation of ± 75 kc/s at the radiated carrier frequency for a modulating signal of say 30 c/s, we require $75{,}000/30 = 2{,}500$ radians of phase shift, i.e., a multiplication of 5,000 times is required. This is normally too great to be practicable in a single chain of multipliers, since, starting with a crystal frequency of say 100 kc/s, we should finish with a carrier frequency of 500 Mc/s.

It is usual in such a case, therefore, to multiply in two stages, say from 100 kc/s to 6·4 Mc/s and from 2 Mc/s to 96 Mc/s. A 4·4-Mc/s heterodyne frequency change operates on the 6·4-Mc/s signal to produce the required 2-Mc/s signal.

Another method of producing phase modulation is to add the side-bands of an amplitude-modulated carrier to a carrier in quadrature. This process is easily demonstrated vectorially using the same artifice as previously.

Thus in Fig. 31 (a) we have first a carrier frequency vector OV_1 and two side-band vectors OV_2 and OV_3.

The vector OV_2 is, say, the upper-side-band signal, and therefore rotates counter-clockwise relative to OV_1 with the modulation frequency. Likewise, the lower-side-band vector OV_3 rotates clockwise relative to OV_1. Vector addition shows that the counter rotation of the two side-band vectors in conjunction with the carrier vector OV_1 produces an amplitude-modulated signal without phase shift.

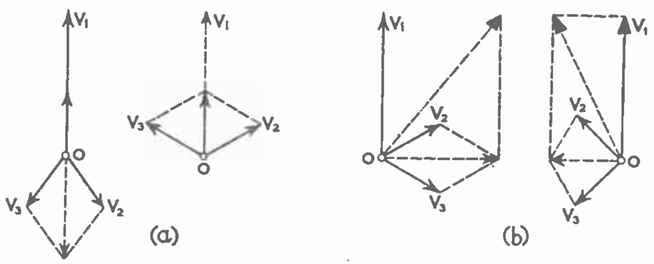


FIG. 31.—(a) VECTORIAL REPRESENTATION OF AMPLITUDE MODULATION; (b) VECTORIAL REPRESENTATION OF FREQUENCY MODULATION.

Now suppose the two side-band vectors are added to the carrier vector in quadrature, i.e., the side-bands become directly additive on the time axis instead of as previously on the amplitude axis. This is shown in Fig. 31 (*b*).

It is clear that the counter-rotating side-band vectors now produce phase modulation of the carrier and no amplitude modulation. This, then, provides a means of producing phase modulation. An amplitude-modulated signal is separated into the carrier signal component and the sideband signals component, and these components are then phase shifted 90°, and the two are then added again, transforming the modulation process thereby from that of Fig. 31 (*a*) to that of Fig. 31 (*b*).

This method was used in the original Armstrong arrangement, and is still in current use.

More recently the Serrosoid method of modulation has gained some popularity; this is described fully later.

Direct Frequency Modulation

Two broad methods are in current use.

One uses the reactance control of an L–C oscillator and the other the direct control of a crystal oscillator.

The former requires much the greater care in the stabilization of the centre or mean frequency, but is nevertheless quite satisfactory in practice, and is described in detail later. With direct frequency modulation it is possible to complete the frequency multiplication in one set of multipliers.

Circuit Arrangements : Pre-corrected Phase Modulation

The Armstrong System

This system was introduced by the late E. A. Armstrong, and paved the way for the engineering of frequency-modulated broadcast services.

The schematic diagrams of Fig. 32 illustrate an arrangement for a radiated frequency of 96 Mc/s based on the Armstrong system.

A crystal oscillator operating at a frequency of 100 kc/s drives a buffer amplifier, which th n provides outputs for two channels—one the balanced modulator and the other to an oscillator frequency amplifier. The output of the balanced modulator which contains the modulation products only and the output from the carrier-frequency amplifier, one output of which is shifted 90° with respect to the other, are combined

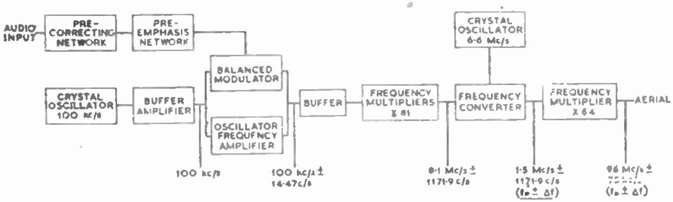


FIG. 32.—THE ARMSTRONG TYPE OF FREQUENCY MODULATION SYSTEM.

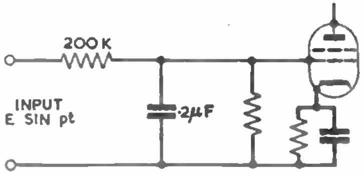


FIG. 33.—PRE-CORRECTING NETWORK.

Class A amplifier output $\frac{AE}{p}$ sin pt.

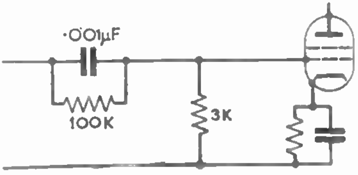


FIG. 34.—PRE-EMPHASIS NETWORK FOR TIME CONSTANT 100 MICROSECONDS.

Class A amplifier.

and fed into a second buffer amplifier. Frequency multiplication by 81 brings tho carrier frequency to 8·1 Mc/s. This is then heterodyned with a crystal oscillator and the lower difference frequency, retaining the $\times 81$ deviation, is selected and multiplied further by a factor of 64 to achieve a frequency deviation of ± 75 kc/s at 96 Mc/s for a modulation phase shift of less than $\pm \frac{1}{2}$ radian at 30 c/s.

Pre-correcting Networks.—As already stated, for phase-modulation systems it is necessary to include in the audio-frequency input circuit a network to convert the signal from the form $E \cos pt$ to the form $\frac{E}{f_m} \cos pt$.

The form of pre-correction used in the original Armstrong system is shown in Fig. 33.

Pre-emphasis Network.—This takes the form of a simple R–C network giving a rising amplitude output for increasing frequency. One example is given in Fig. 34. Equivalent R–L networks are sometimes used.

The Balanced Modulator.—A basic form of balanced modulator is shown in Fig. 35.

In this arrangement the audio-frequency signals, $e \cos pt$, operate on the valves in push–pull, while the carrier-frequency signals operate on the valves in push–push or parallel mode. The outputs of the two valves are combined in the output transformer to produce signals of the form

$$A\{\sin (w + p)t + \sin(w - p)t\}$$

which may be written in the form

$$B \cos pt \sin wt$$

which is the modulation product containing only the side-band information, i.e., with no carrier components present.

The Frequency Multiplication Stages.—These are of conventional design, and may be high-efficiency Class C stages of adequate band-width.

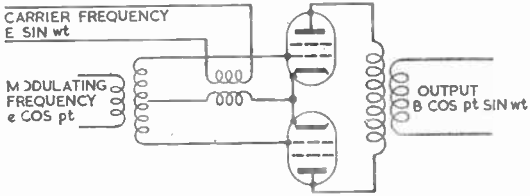


Fig. 35.—Balanced Modulator.

The Serrosoid System [2]

This system overcomes many of the disadvantages of the Armstrong phase-modulated system and provides, if necessary, an initial peak phase shift of $\pm 150°$. Normally, the system is based on $\pm 90°$ or $1\frac{1}{2}$ radians for 100 per cent modulation at the lowest frequency. For $1\frac{1}{2}$ radians and 50 c/s the peak deviation is ± 75 c/s, so, for a ± 75-kc/s deviation of the radiated signal, a frequency multiplication of 1,000 is required.

The nearest simple whole number with simple multiplication factors is 972. Thus, for example, we may take a 100-kc/s crystal frequency multiplied to a radiated frequency of 97·2 Mc/s.

Fig. 36 shows the complete Serrosoid system.

In general terms a saw-tooth waveform is derived from a stable crystal-controlled source. This is then limited at about 0·5 amplitude, and a pulse is derived at the onset of limiting. Bias modulation of the limiting action then causes the pulse to vary in time in sympathy with the instantaneous value of the modulating signal. The derived fre-

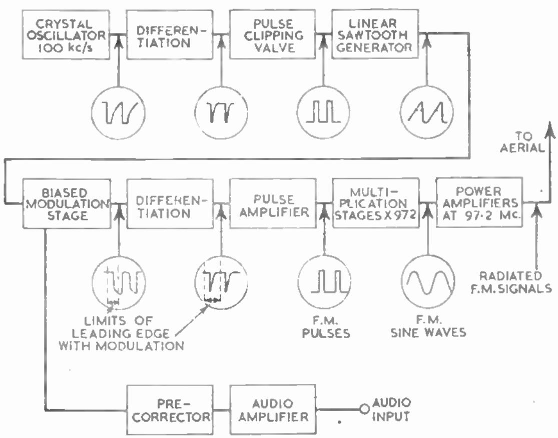


Fig. 36.—The Serrosoid System of Pre-corrected Phase Derived Frequency Modulation.

quency-modulated pulses are then fed into a conventional chain of frequency multipliers, and the frequency-modulated signals so obtained at the radiated carrier mean frequency are power amplified and radiated. Mean frequency stability is thus crystal controlled and, due to the large initial phase deviation possible, a single chain of multipliers is sufficient.

Performance of this relatively modern form of frequency modulation, first operated commercially in 1948, is good, the claims made being summarized below :

Mean Frequency Stability.—$\pm 0{\cdot}0002$ per cent.

Frequency Modulation noise in modulator 50–15,000 *c/s Measured with* 75 *Microseconds De-emphasis.* 80 db below 100 per cent modulation.

Distortion from the Non-linearity of Phase Shift for $\pm 135°$ *Peak Phase Deviation.*—0·25 per cent frequency modulation distortion.

Direct Frequency Modulation

Centre Frequency Stabilized Systems

With this form of frequency modulation the modulating signals are made to control the frequency of an *L–C* oscillator, the mean frequency of which is stabilized in relation to a quartz-crystal-controlled oscillator or by the difference between a quartz-crystal oscillator and an *L–C* frequency discriminator.

The frequency control is usually achieved by a reactance tube acting directly on the frequency-determining circuit and operated by the modulating signals.

This method is capable of producing large deviation, but suffers from the necessity to introduce complicated reactance control in order to stabilize the mean frequency.

The frequency of the directly controlled oscillator is usually of the order of a few Mc/s, and operates at a sufficiently low power level to enable the reactance tube modulation to be reasonably small and economical. The frequency is then multiplied successively in a chain of harmonic amplifiers, and finally in straight Class C amplifiers to the required power level.

A typical system is illustrated in Fig. 37.

In this system a sample of the fully modulated frequency-modulated signal is taken, and the mean frequency is compared with that of the fixed frequency derived from a crystal oscillator. When the two frequencies are identical the discriminator is designed to give zero output. As soon as a difference in frequency occurs the discriminator

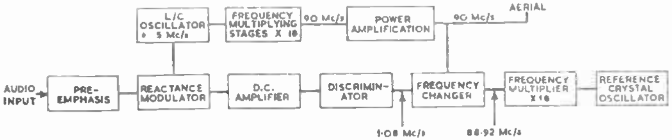


FIG. 37.—DIRECT FREQUENCY MODULATED TRANSMITTER WITH REACTANCE VALVE CONTROL OF CENTRE FREQUENCY.

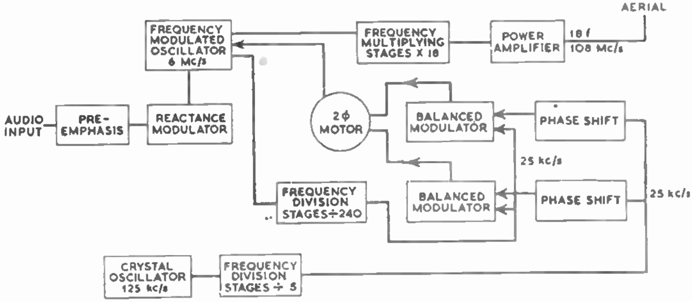


Fig. 38.—Frequency Modulated Transmitter with Motor Control of Centre Frequency.

produces a D.C. bias of correct sense to reduce the difference to a negligible amount.

There is, however, always a tolerance on the mean frequency determined by the stability of the quartz-crystal oscillator, which is easily made high enough to be disregarded, and also by the stability of the D.C. amplifier system.

More accurate stabilization of the L–C oscillator is possible by the method shown in Fig. 38, in which the frequencies of both the L–C oscillator and the crystal reference are divided down to a few thousand c/s.[3] One of these two low-frequency signals is then applied to a balanced modulator direct, and the other to a second balanced modulator through a 90°-phase-shift network.

The outputs of the two balanced modulators thus produce a frequency, at the output, that is proportional to any difference between the two input signals. The outputs are also in quadrature, with a combined

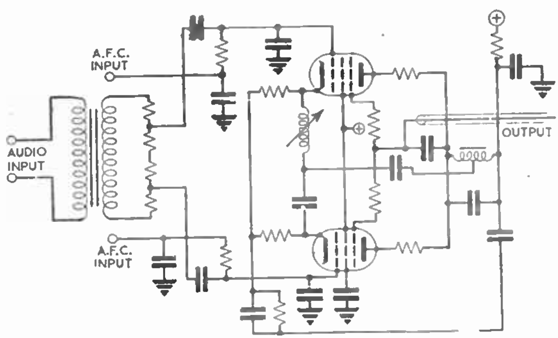
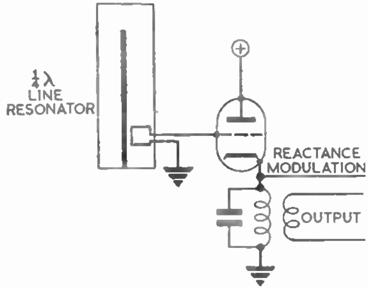


Fig. 39.—Modern Version of Reactance-controlled Oscillator Capable of Wide Deviation with Low Distortion.

FIG. 40.—BASIC FORM OF "LINE" TYPE FREQUENCY MODULATION OSCILLATOR.



vector rotation depending on whether the L–C oscillator frequency is above or below the reference frequency.

This two-phase output, suitably amplified, is utilized to drive a two-phase motor which controls the L–C oscillator tuning control directly.

Accuracy of frequency control with this system is claimed to be limited only by the heat cycle of the crystal oven. Distortion is of the order of ½ per cent and noise level in the output is 74 db below 100 per cent modulation.

A particularly interesting version of the reactance-controlled oscillator has recently been developed by W. L. Wright.

A simplified diagram is shown in Fig. 39.

In this arrangement the modulated oscillator consists of a pair of pentodes operating in a balanced oscillating circuit, with feedback taken from the anode-tuned circuit to the cathodes of the two valves through phase-shifting networks, which in one case gives 45° lead and in the other 45° lag when the mutual conductances of the two valves are equal.

Modulating signals are applied to the two grids in phase opposition so that the slopes of the valves are changed differentially, one increasing while the other is decreasing. This causes the net phase shift to change, producing frequency modulation of the output. The application of modulating signals thus produces direct frequency modulation. The inductive phase-shifting network is made variable as a balancing control, and balance is achieved by adjusting for minimum amplitude modulation. Application of A.F.C. voltage to the modulator grids is employed.

Centre-frequency Stability.—Less than ± 20 in 10^6 for ambient temperature 15–70° C.

Change of Centre Frequency with Modulation.—± 5 in 10^6.

Maximum Deviation Possible over Band 88–108 Mc/s.—± 300 kc/s.

Frequency Response.—$\pm 0{\cdot}5$ db 30–15,000 c/s without pre-emphasis.

Amplitude Modulation.—Less than 1 per cent.

Noise.—-70 db relative to 75 kc/s deviation.

Harmonic Distortion.—Less than ½ per cent for 30–15,000 c/s at 100 kc/s deviation.

Inherently Stable Systems

Probably the most simple of the inherently stable systems is provided by the simple line-type oscillator with reactance modulation.[4]

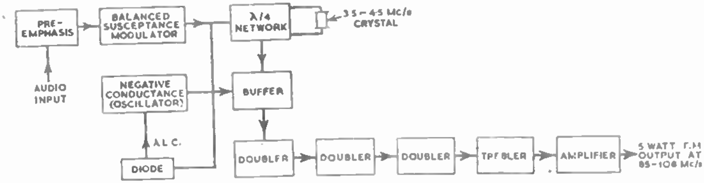


Fig. 41.—Block Schematic Diagram of "F.M.Q." Drive.

This is illustrated in Fig. 40, and will produce ± 75 kc/s deviation with low distortion directly at 88–108 Mc/s, with a stability dependent on the Q and therefore the physical dimensions of the "line". As far as is known, this has not been developed to commercial standards.

A more recent disclosure known as F.M.Q. operates by modulating a quartz-crystal oscillator directly.[5]

The elements of this system are illustrated in Fig. 41.

In one successful design a deviation of ± 1 part in 1,000 has been achieved for a susceptance change of $\pm 0{\cdot}008$ mhos by using automatic level control on the oscillator, which maintains the output at 1 volt r.m.s.

Carrier-frequency stability is determined predominantly by the modulator. The achievement of a centre-frequency stability, such that it never deviates by more than 2 kc/s, calls for a high degree of stability of the modulation valve characteristics over their life. This problem is reduced considerably by the use of a balanced modulator with reasonably matched valves having a long linear range of g_m/E_g. The arrangement in Fig. 42 has been found quite satisfactory and gives linear modulation.

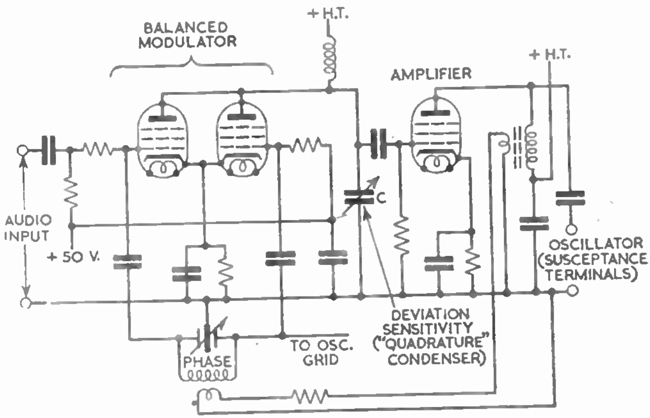


Fig. 42.—Three-valve "Susceptance Modulator" Circuit.

In practice, an ordinary quartz crystal does not behave as a simple resonator but has spurious modes of resonance which may occur within the large spectrum used for signals. The F.M.Q. system uses a special type of quartz crystal in which significant spurious resonances have been eliminated.

Overall, this system shows considerable economy in valves, and results in great operational reliability.

Performance is adequate for high-fidelity sound broadcast, and is summarized below.

Distortion (Maximum) at 75 kc/s Deviation.—Less than 1 per cent for 60–15,000 c/s.

Amplitude Modulation Noise.—

Zero deviation : −60 db.
200 kc/s deviation 400 c/s : −53 db.

Frequency Modulation Noise.—Relative to 75 kc/s deviation : −60 db.

Stability of Carrier.—Frequency change for deviation from 0 to 100 kc/s : 2 in 10^6.

Change over 48 hours : 12 in 10^6.

Frequency response without pre-emphasis : 3 db at 15 kc/s; 10 db at 20 kc/s.

References

[1] E. A. Armstrong, "A Method of Reducing Disturbances in Radio Signalling by a System of Frequency Modulation", *Proc. I.R.E.*, 1936, 24, p. 689.

[2] J. R. Day, "Serrosoid F.M. Modulation", *Electronics*, October 1948.

[3] N. J. Oman, "A New Exciter Unit for Frequency Modulated Transmitters", *R.C.A. Review*, Vol. VII, No. 1, March 1946.

[4] V. J. Cooper, British Patent 610,951.

[5] W. S. Mortley, "'F.M.Q.' Circuit Giving Linear Frequency Modulation of a Quartz Crystal Oscillator", *Wireless World*, October 1951.

A. Hund, *Frequency Modulation*, McGraw-Hill Book Co., 1942. (This book contains an extensive list of further references.)

V. J. C.

FREQUENCY ALLOCATIONS FOR BROADCASTING

Table 1.—V.H.F./U.H.F. Broadcast and Television Bands

Band I	41–68 Mc/s.	7·3–4·4 m.
Band II	87·5–100 Mc/s.	3·4–3 m.
Band III	174–216 Mc/s.	1·7–1·4 m.
Band IV	470–585 Mc/s.	64–51 cm.
Band V	610–960 Mc/s.	49–31 cm.

TABLE 2.—ATLANTIC CITY CONFERENCE ALLOCATIONS FOR BROADCASTING

Band	*Frequency*	*Wavelength*	*Area* *
Long Wave-band .	150–285 kc/s.	2000–1053 m.	Region 1.
Medium Wave-band	525–1605 kc/s.	571–187 m.	Region 1.
Medium Wave-band	535–1605 kc/s.	560–187 m.	Regions 2 and 3.
120 m. band . .	2·300–2·495 Mc/s.	130·3–120 m.	Region 1. Regions 2 and 3. 2·300–2·495 Mc/s only.
90 m. band . .	3·200–3·400 Mc/s.	93·69–88·18 m.	All Regions.
75 m. band . .	3·900–4·00 Mc/s.	76·9–75·0 m.	Region 3. Region 1, 3·95–4·0 Mc/s only. Region 2 excluded.
60 m. band . .	4·75–5·06 Mc/s. (except 4·995–5·005 Mc/s.)	63·10–59·25 m. (except 60·05–59·95 m.)	All Regions.
49 m. band . .	5·950–6·200 Mc/s.	50·39–48·36 m.	All Regions.
41 m. band . .	7·100–7·300 Mc/s.	42·23–41·07 m.	Regions 1 and 3 only.
31 m. band . .	9·500–9·775 Mc/s.	31·56–30·70 m.	All Regions.
25 m. band . .	11·700–11·975 Mc/s.	25·63–25·0 m.	All Regions.
19 m. band . .	15·10–15·45 Mc/s.	19·86–19·40 m.	All Regions.
16 m. band . .	17·70–17·90 Mc/s.	16·94–16·75 m.	All Regions.
13 m. band . .	21·45–21·75 Mc/s.	14·00–13·79 m.	All Regions.
11 m. band . .	25·60–26·10 Mc/s.	11·71–11·49 m.	All Regions.

* See Section 44, pages 20–1.

7. COMMUNICATION TRANSMITTERS

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Although basically the same in principle, modern radio transmitters differ greatly in detail from those of a generation ago. Apart from improvements in size factor and durability under extreme climatic conditions, the changes are mostly related to the vastly increased number of applications and the consequent congestion of all frequency bands, with the resulting development of measures to avoid mutual interference between services. The increasing complexity of equipment has necessitated special attention towards convenience of operation and maintenance.

The advent of transistors has made possible a spectacular reduction in size and power consumption of small portable transmitters, and despite their present-day limitations in frequency response and ambient temperature range, they make useful economies possible in the auxiliary circuits of larger equipments.

The International Radio Regulations resulting from the Atlantic City conference of 1947 set down limits of carrier-frequency tolerance for all classes of user, and additional limits are set, for example, by the British Post Office for equipment under their jurisdiction; these make mandatory for some classes the standards for sideband spread, harmonic radiation and the like, which are generally accepted as good practice in modern equipment.

Communications transmitters, although varying in size from many kilowatts down to units driven from dry batteries, have, nevertheless, many common features which are at variance with the technique of broadcast-transmitter design. Sound quality on telephony is, of course, far less important, a frequency response covering the main speech range of 250–3,000 c/s being adequate for most purposes. The frequency response is often deliberately distorted, in fact, the higher speech frequencies being accentuated in the interests of maximum clarity. Harmonic distortion of 10 per cent or even more at 90 per cent modulation is relatively unimportant, and for equipment working off normal mains-supply frequency a hum level of 30 db below maximum modulation is adequate. A satisfactory service may in fact be achieved with an aerial power sufficient only to give 10 db signal-to-noise ratio for 30 per cent modulation at the associated receiver.

Microphones are generally of types designed for close working to cater for conditions of high background noise, and carbon microphones of the telephone type are still in frequent use. Frequency modulation is becoming popular, particularly for the smaller portable equipment operating on the V.H.F. bands.

Aerials vary from a simple receiving-type wire or whip at one extreme to complex directional arrays at the other, but are rarely such a dominant feature of the installation as with broadcast equipment.

Telegraphy, whether by continuous-wave, modulated-continuous-wave or frequency-shift methods, is in frequent use, and sets a series of problems unmatched by any corresponding ones in the broadcast sphere.

BASIC TRANSMITTER DESIGN

The special requirements of various forms of communications service are best considered in relation to the basic sections of a primitive transmitter. These consist fundamentally of a frequency generator, power amplifier and modulator, together with arrangements for coupling the oscillator to the power amplifier and the power amplifier to the aerial, and means of interrupting the carrier wave for keying purposes.

Oscillator Stage

Single-valve oscillator circuits are shown in Figs. 1–4. Fig. 1 depicts a tuned-grid circuit, in which resistor R1 serves to bias back the valve to the operating part of its characteristic at the onset of oscillation, by virtue of the flow of grid current; this will result in a Class C condition of operation, in common with other oscillators not limited by automatic-gain-control. This circuit suffers from the defect that variations in valve input capacitance, which are most marked during the valve warming-up period, have a direct influence on the frequency of oscillation, particularly at the high-frequency end of the band.

Matters are improved by reversing the positions of tuned and tickler windings, since valve-anode capacitance, in general, suffers far less variation. Further improvement is obtained by stabilizing H.T. voltage, e.g., with a gas discharge tube. The tuned circuit itself is rendered less dependent upon temperature variations by the use of low-expansion materials. The coil may be wound with stressed or fired-on silver windings on a ceramic former, or with silver-plated invar wire, whilst a variable capacitor of substantially zero temperature coefficient (5–10 parts per million per ° C., compared with 60–80 parts per ° C for normal types) can be achieved by the judicious use of nickel in the construction.

Recently capacitors with compensating sections controlled by bimetal elements have appeared on the market, enabling a zero or a slightly negative temperature coefficient to be achieved, and facilitating compensation for residual variations in other components. The alternative form of compensation, by means of ceramic capacitors of high negative temperature coefficient (approximately 800 parts per million per ° C.)

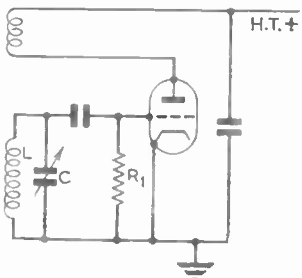

FIG. 1.—SIMPLE OSCILLATOR.

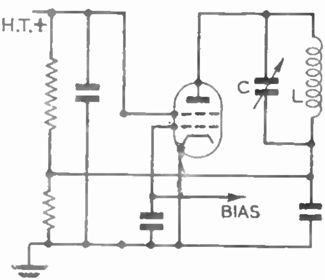

FIG. 2.—DYNATRON OSCILLATOR.

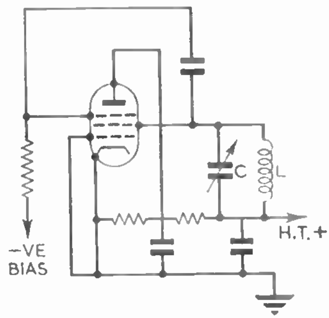

FIG. 3.—TRANSITRON OSCILLATOR.

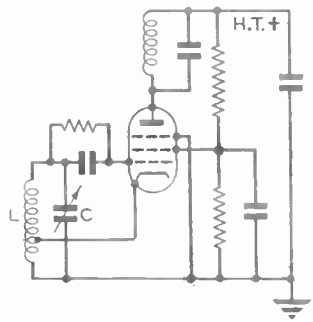

FIG. 4.—ELECTRON-COUPLED OSCILLATOR.

depends upon the selection of components free from scintillation, which produces the effect of small, discontinuous jumps in frequency.

Full compensation over the entire frequency band is, of course, not obtainable where tuning is effected by capacitance variation, but matters are considerably improved by the use of a compensating capacitor in series with the variable condenser, as well as one in parallel. High-value ceramic condensers of negative temperature coefficient can now readily be obtained with the necessary " flutter free " performance.

A further source of trouble in damp climates lies in variation of capacitance with condensation of moisture on the capacitor vanes. It is now possible to obtain shaft bushings which permit rotation of the spindle while preserving the hermetic seal.

Figs. 2 and 3 show circuits utilizing single coil windings, with consequent simplification of coil structure. The dynatron circuit of Fig. 2, requiring a two-terminal inductance, functions by virtue of the negative resistance present in the anode characteristic of a tetrode, which is produced as a result of secondary emission from the anode caused by the high screen/anode voltage ratio. This appears effectively in shunt with the anode-tuned circuit of higher " dynamic resistance " $Q\omega L$. The circuit suffers frequently from short valve life and the need to select individual valves; it has declined in popularity since the development of alternative negative resistance circuits, such as the transitron, which depends upon the effect of a negatively-biased suppressor upon the screen impedance of a pentode valve (Fig. 3).

The so-called electron-coupled oscillator of Fig. 4 requires a tap on the coil, but, with simple switching of one or more quartz crystals, it can be transformed easily into an effective crystal oscillator circuit.* Frequency stability of the order of $\pm 0{\cdot}05$ per cent may be obtained, with careful design, in an L–C oscillator using this circuit.

Still less drift is given by two-valve circuits, such as the Franklin oscillator (Fig. 5), or the keyed twin-valve oscillator of Fig. 6, but modern statutory limits of frequency tolerance, coupled with arduous climatic requirements, have relegated the simple variable frequency oscillator to a subsidiary role, for stand-by and emergency use.

* Patent No. 555,750.

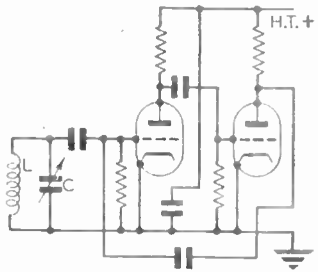

FIG. 5.—FRANKLIN OSCILLATOR.

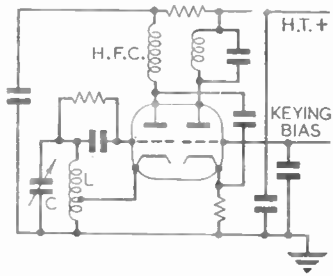

FIG. 6.—TWIN-VALVE KEYED E.C.O.

Transistor Oscillators

Transistor oscillators while relatively free from the initial warming-up drift of valve circuits, are, at the present stage of development, inherently less stable with variations of ambient temperature. With light coupling to the tuned circuit, however, stabilities of the order of 0·1 per cent are at present obtainable. Fig. 7 shows a transistor *L–C* oscillator of the Clapp type, with frequency determined by inductance L tuned by the capacitance of C1, C2, and C3 in series. C2 and C3 are made as large as possible to reduce the effect on frequency of variations in transistor impedance. The transistor should, of course, be of the high-frequency type, with an α-cut-off frequency several times the highest frequency required.

An arrangement suitable for crystals is shown in Fig. 8. The circuit L2–C2 is tuned to give maximum output, which, provided that there is tight coupling between L1 and L2, will be at the crystal series resonant frequency; otherwise the frequency of oscillation will be somewhat displaced, the resultant crystal reactance then tuning out the leakage inductance. If the crystal is one intended for parallel resonance, it will in addition be necessary to wire in series with it a capacitance equal to the specified load capacitance in order to obtain the correct frequency. With crystals whose series resistance is too high for oscillation, the addition of a second transistor with "grounded collector" connection,

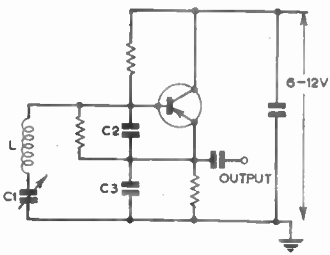

FIG. 7.—TRANSISTOR *L–C* OSCILLATOR.

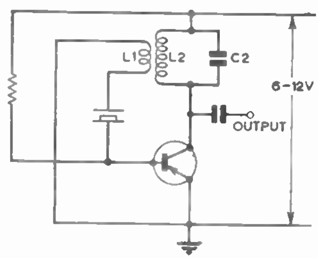

FIG. 8.—TRANSISTOR CRYSTAL OSCILLATOR.

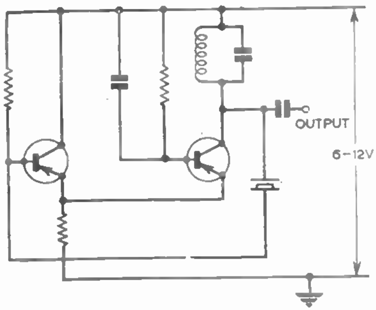

FIG. 9—TWO TRANSISTOR CRYSTAL OSCILLATOR.

to give a high input impedance, will overcome this difficulty. The circuit is best arranged as in Fig. 9, with common emitter coupling, in order to dispense with the coupling winding.

High-stability Circuits

Present-day stability requirements can be met using variable frequency sources of the additive type, which function by mixing the output of a relatively low-frequency variable-frequency oscillator (V.F.O.) with one or another of the harmonics of a quartz-crystal oscillator; using the heterodyne voltage to give an overall frequency accuracy, after the necessary filter circuits, of many times the accuracy of a corresponding V.F.O. on the same frequency.

In a further elaboration of this circuit the resultant frequency is compared with the output of a separate oscillator. which is automatically corrected by means of a motor-driven compensating capacitor driven from the difference frequency. This method has the advantage that no spurious signals originating from inadequately rejected harmonics of the crystal oscillator can reach the aerial; but problems of "hunting" in the output frequency arise unless a virtually inertialess tuning motor is used.

Both these methods of frequency synthesis are, of necessity, far more elaborate and costly than a single variable circuit, but can be made to give a frequency stability of better than $\pm 0{\cdot}001$ per cent.

Crystal Oscillators

With modern low-temperature coefficient cuts of crystal and no special precautions for stabilizing supplies, oscillators can readily be made to give stabilities of $\pm 0{\cdot}005$ per cent over a wide temperature range. For frequencies between 2 and 15 Mc/s, it is possible to obtain a precision A–T cut with limits of $\pm 0{\cdot}0015$ per cent over the temperature range $-20°$ to $+70°$ C. A stability approaching this figure is also obtainable from A–T overtone crystals, which operate at a frequenc close to the third or fifth harmonic of the fundamental frequency. For the frequency band 90–250 kc/s, the G–T cut is available for applications where the relatively large size and increased fragility do not count as a disadvantage, giving a stability of about 0·0006 per cent over the range 0–60° C.

Crystal Ovens

For operation on other frequency bands, as well as for the more exacting requirements of single-sideband and V.H.F. transmitters, a frequency stability is required better than can be obtained over a wide temperature range with present-day crystals. These requirements are met by maintaining the crystal temperature substantially constant in a thermostatically controlled oven. Such ovens range from elaborate units housing a number of crystals, possibly with inner and outer compartments each fitted with contact thermometer, heater and associated relay to reduce the temperature differential to a negligible value over the whole ambient temperature range, down to small plug-in devices housing one or two crystals and fitted with a 6- or 12-volt heater winding controlled by a bimetal thermostat.

The success of such small ovens depends upon the reliability of the thermostat; in some respects, however, design problems are easier than in larger ovens, owing to the generally lower thermal inertia and the closer thermal coupling between heater, thermostat and crystal. Efforts are being made to standardize the essential features (including base-pin connections and heater voltage) of the small plug-in ovens to facilitate interchangeability. Typical requirements for a unit to contain two style " D " crystals are:

Heater 6/12 volts, A.C. or D.C.
Crystal temperature, 75° ± 3° C.
Differential, not more than 2° C.
Warm up time, less than 10 minutes.

With a 6-watt heater winding, it is feasible to meet these requirements over an ambient temperature range of −20° to 70° C. The temperature differential gradually increases as the margin between controlled temperature and ambient temperature becomes smaller, but if control down to −20° C. ambient is unnecessary, can be lessened by a reduction of the heater voltage.

A novel alternative [1] to oven control is a specially constructed A–T cut crystal unit incorporating a thermistor in close proximity to the quartz plate. By connection of the thermistor in a suitable impedance network, the effective load capacitance is arranged to vary with temperature by an amount sufficient to cancel out the natural frequency drift of the quartz element.

Frequency Trimming

With the reduction in rate of frequency change with ambient temperature, other factors, such as the initial adjustment error of the crystal unit, become of increased significance. It may then be necessary to " pull " the crystal frequency by variation of load capacitance from the specified standard value.[2] The capacitance change needed for a given frequency change varies from one crystal to another; it depends largely on the motional inductance, which varies even between crystals of the same frequency and cut. Owing to variation.

in the dimensions of the quartz element between one manufacturer and another, crystals of the same cut and frequency may differ by some 3 to 1 in motional inductance. For frequencies between 3 and 15 Mc/s the possibility, unless the cut is stipulated in advance, of the crystals being either A–T or B–T increases the possible variation to about 5 to 1. Clearly, the necessary capacitance change will also depend upon the total value of load capacitance. The following table shows the approximate limits likely at various frequencies, in the case of a load capacitance of 30 pF.

Frequency (kc/s)	*Type of cut*	*Capacitance change for frequency variation of 10 p.p.m * (pF)*
40– 200	5° X	0·4 – 0·8
90– 250	G–T	1·3 – 1·6
100– 250	D–T	1 – 4·2
150– 500	C–T	1·5 – 5
400– 800	E–T (Overtone)	4 – 6·2
1,000–20,000	A–T	0·8 – 2·3
3,000–22,000	B–T	1·25 – 3·5
15,000–75,000	A–T (3rd Overtone)	7·7 –12·5

* p.p.m. = parts per million.

Unless adequate capacitance variation can be provided, the advice of the manufacturer of the particular crystals concerned should be sought. These notes have assumed the use of a "parallel" oscillator, i.e., with the crystal operating as an inductance tuned by the load capacitance and its own shunt capacitance in parallel. The same general considerations apply, however, to "series" oscillators, which are increasingly popular for very low frequencies and for overtone crystals. Crystals intended for parallel circuits can be used in series oscillators by wiring in series a capacitor of the standard load-capacitance value. For trimming, this capacitor is then adjusted in the appropriate direction, assisted if necessary by tuning out the "pin-to-pin" capacitance by means of shunt inductance.

Further information on crystal oscillators is given in Section 44 and reference [2].

Compensation for Ageing

Another factor of importance in close-tolerance frequency control is the phenomenon of ageing of the crystal unit; this can cause serious difficulty in the case of V.H.F. mobile and marine equipment, where regular frequency checks on precision measuring gear are not always possible. There is, as yet, no generally accepted standard for ageing beyond the proviso that twelve months after manufacture the crystal unit shall still be within the specified limits for frequency and activity. Where these limits have been narrowed with the help of frequency trimming, the drift at the end of a year may well be several times the maximum tolerable value. At present, no reliable prediction is possible of the direction of frequency drift or of the effect of temperature or drive. The phenomenon is, in fact, less a reflection upon the properties of quartz crystals than on the extreme constancy of physical properties necessary for modern precision frequency control.

Fortunately, the rate of drift nearly always becomes smaller as time passes; two readjustments of trimming, one when the crystal is six months old and a second after another twelve months, are normally found sufficient. In the absence of local frequency measuring apparatus, it is of advantage, with mobile equipment, to calibrate the trimming control in terms of load-capacitance value. New or used crystals can then be correctly set with the exact load capacitance determined at a central depot.

Buffer Stages

Drive for the power amplifier is usually obtained from the oscillator via one or more buffer amplifiers; these serve to isolate the frequency standard from changes of impedance in the grid circuit of the power amplifier consequent upon variations in its anode loading. Often the buffer stages are required to function also as frequency multipliers. Class C operation is usual, both in the interests of high efficiency and for the production of a signal current in the anode circuit rich in harmonics. Tetrode or pentode valves are most frequently used, for their high power gain as well as for stable operation when required to amplify at fundamental frequency. Occasionally, when gain and output requirements are moderate, the anode circuit may be aperiodic, to simplify tuning operations and give still better isolation of the oscillator.

For efficient frequency multiplication, when necessary, bias and drive are both considerably increased, and push–pull operation is favoured as a means of amplifying odd harmonics.

Power Amplifier

The power-amplifier stage may consist of triode, tetrode or pentode valves operated singly, in parallel or in push–pull. For the higher powers, directly heated tungsten or thoriated tungsten cathodes are in use, and care must be taken to keep heater-voltage variations within the closer limits required for these valves than for oxide-coated types. The initial heater-current surge must, in many cases, be limited to about 50 per cent in excess of the running current, by the use of current limiting devices, such as a supply transformer of high leakage reactance.

Class C operation is general, with the grid circuit biased back to approximately twice the cut-off voltage, for maximum output with high efficiency. An exception is the Class B radio-frequency amplifier, used for distortionless amplification of a previously modulated signal, with critical adjustment of standing current and amplitude of drive.

In all Class C circuits, but particularly the power amplifier, the tuned circuit fulfils the function of storing energy during the conducting pulse and releasing it during the rest of the cycle. For stable operation, the energy $1/2\ C\,.\,E^2$ stored in the capacitor at peak voltage (E) must approximate to at least twice the energy dissipated per cycle. This corresponds to an effective " Q " in the loaded tuned circuit, expressed as reactance/resistance or, approximately, volt amperes/watts ratio, of 4π. There is no objection to exceeding this value, with corresponding benefit to harmonic content, apart from the increase of coil losses with increased circulating current. For most practical purposes, the peak voltage E may be taken as four-fifths of the H.T. voltage.

In practice, with a continuously-tuned power-amplifier circuit, a variation of at least two to one in effective Q is permitted, to minimize waveband switching. This corresponds to a reactance ratio of 4 : 1 in the variable inductance or capacitor for a frequency coverage of one octave.

The valve capacitances alone exceed the above values in some circumstances, in which case efforts are best made to tune by variation of inductance. With V.H.F. equipment the working Q may unavoidably be as high as 60, and some loss of efficiency must be tolerated. For example, a popular V.H.F. double tetrode of the 40-watt class, rated at 50 watts output for an anode voltage of 500 volts, has an output capacitance of 7 pF. Assuming an output of 680 volts r.m.s. for push–pull operation, the load impedance is 9,200 ohms; at 150 Mc/s, the approximate resistance/reactance ratio will be 64. Further allowance should be made for residual capacitance in the tuning elements or anode lead-offs. In a case such as this, the "line" resonant circuits sometimes favoured form little more than a convenient means of varying the tuned-circuit inductance.

One of the directions of progress in this field lies in the development of valves with ever lower capacitance values for a given impedance or power rating.

The type of double tetrode mentioned above, consisting of separate assemblies mounted side by side in one envelope, is now falling out of use. This is because an output capacitance of little more than 2 pF is achieved, with similar characteristics, by the latest double tetrodes of "butterfly"-type construction. In these valves the electrode assemblies are spaced one on each side of a common cathode: internal neutralizing is provided, and the genuinely symmetrical construction makes it possible largely to dispense with external decoupling. All supply voltages are then fed through high frequency chokes thus avoiding the problems of spurious oscillations of the "parallel" type which so often follow in the wake of routine decoupling measures at these frequencies.

The scope of transistor power amplifiers lies initially in low radio-frequency, low power equipment. With types at present in production in the U.K., a power output of 3 watts at 500 kc/s is obtainable from a push–pull pair using a 12-volt supply, providing a power gain of some 10 dB. The situation will doubtless improve rapidly, however, with the introduction of high-frequency transistors already in production in the United States.

Coupling to Aerial

With an effective tank-circuit Q of 12 or thereabouts, the impedance ratio at resonance to that at overtone frequencies is not high enough to give sufficiently low harmonic voltage across the circuit, for radiation purposes. Modern regulations permit a maximum radiated level of 40 db below fundamental, or 200 mW (whichever is less), for individual

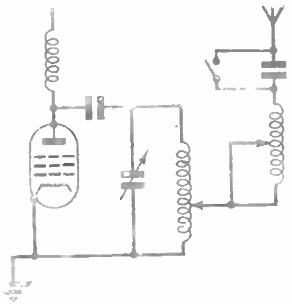

FIG. 10.—DIRECT OUTPUT COUPLING.

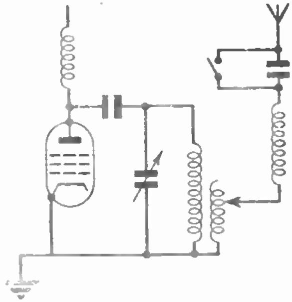

FIG. 11.—INDUCTIVE OUTPUT COUPLING.

harmonics. To comply with this, it is necessary to devise an aerial-coupling circuit which discriminates against the higher frequencies.

A selection of output circuits is shown in Figs. 10–14. Figs. 10 and 11, with closely-coupled inductive or direct coupling, give no harmonic voltage reduction in the aerial circuit other than that afforded by the aerial loading coil. In Fig. 11, however, loose coupling will afford a degree of harmonic reduction by virtue of the series leakage reactance thus introduced.

The once-popular circuit of Fig. 12 is worse, since lack of tight coupling between turns of the coil can give rise to spurious series resonances in the aerial circuit, which may actually accentuate some harmonics. To a smaller extent, this trouble can be encountered with other circuits when an excessive lead inductance separates the valve anodes and the tank capacitor.

Pi-Coupling

The well-known "pi" filter circuit of Fig. 13 gives adequate harmonic reduction for small transmitters, and dispenses with the need of a

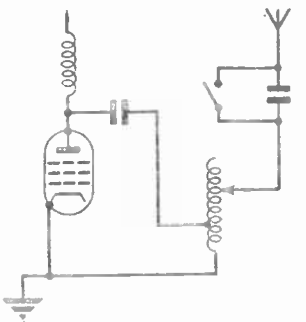

FIG. 12.—"ANODE TAP" OUTPUT.

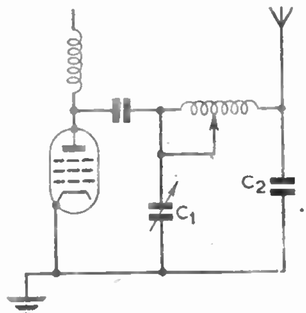

FIG. 13.—"Pi" CIRCUIT.

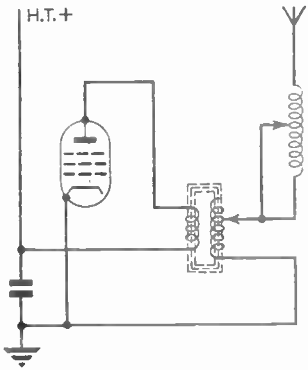

FIG. 14.—WIDE-BAND RADIO-FREQUENCY TRANSFORMER AERIAL COUPLING.

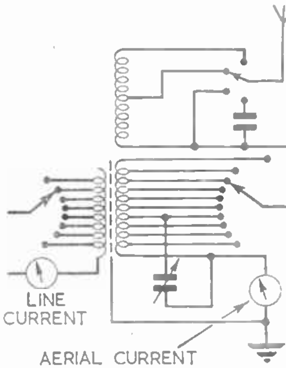

FIG. 15.—SIMPLE AERIAL COUPLING UNIT.

separate aerial tuning inductance over a comparatively wide tuning range. This circuit can be further simplified for use with short aerials by omission of the second capacitor C2, which is replaced by the aerial capacitance.

A comparatively recent development for use with aerials of high Q factor, which in practice is inevitable on the lowest-frequency bands, is the use of a wide-band radio-frequency transformer to match the valve anode to the tuned-out impedance of the aerial (Fig. 14). The only tuning control in this case is the loading coil, which, with the aerial itself, forms the tank circuit. In one remotely controlled low-frequency transmitter using this arrangement all the earlier stages are aperiodically coupled by wide-band transformers. This extension of audio-frequency technique, which very much simplifies remote frequency changing, is claimed to work satisfactorily up to 500 kc/s with an output of several kilowatts, using the best modern core materials.

Where two valves are to be used in the power amplifier stage, push–pull connection is advantageous in reducing even harmonics, provided that some form of electrostatic screening is introduced to minimize capacitative coupling to he output circuit. Push–pull operation offers a further advantage in easy neutralizing of residual grid–anode capacitance, which is necessary even with tetrodes in the larger sizes and at very high frequencies; this offsets to some extent the difficulties of providing balanced drive and a balanced output circuit. When linear amplification of a modulated signal is required, with a Class B (radio-frequency) mode of operation in the power amplifier, the use of push–pull enables very low values of harmonic radiation to be achieved.

For the larger installations a separate aerial-coupling unit is usually fitted under the aerial, and fed via a co-axial or open-wire transmission line. Such a unit, shown basically in Fig. 15, can be made to feed the aerial over a wide band of frequencies. Further discrimination against harmonics beyond that afforded by the second (aerial) tuned circuit is given in this case by the increased transmission-line losses for frequencies other than resonance; the high reflection coefficient and increased power dissipation at harmonic frequencies is, however, generally insufficient to embarrass the transmission line.

High-voltage Problems

The main problem presented by the aerial-loading circuit, whether fitted inside the transmitter or in the form of a separate unit, lies in the provision of high Q inductances and switches capable of withstanding the very high voltages at the aerial terminal and at the same time capable of dissipating a substantial part of the total output power. Thus, with a typical short aerial at a low frequency, the resistive component of the aerial/earth impedance might be 10 ohms and the capacitive component 1,000 ohms. A power of 100 watts would correspond to a current of 3·1 amperes, and the voltage at the aerial terminal would be 3,100 volts r.m.s. without modulation. To radiate this power when using a loading coil of Q factor 350, approximately 30 watts extra must be allowed for loading-coil losses. Even then a relatively low proportion is actually radiated, since the bulk of the 10 ohms represents earth losses rather than radiation resistance.

At medium frequencies rotating solenoid coils with inductance adjustment by a sliding roller contact (controlled by winding handle and counter mechanism) or fixed solenoids with rotating rollers, have become popular in radio-frequency output circuits; a wider range of adjustment, combined with considerable space saving, is possible when using these components with ceramic pot capacitors than was previously possible with fixed inductance–variable capacitance circuits. To ensure low contact losses under all conditions, heavy silver plating with a rhodium flash coat is usually considered necessary.

For lower frequencies, coils wound in sections with multi-strand litz wire (sometimes of 243 strands or more) are necessary, tapped between sections and with some form of continuous adjustment such as an oil-filled variable capacitor. Variometers, also popular for this purpose, are of much more constant "Q" over their range of adjustment than might be expected, owing to the influence of "proximity effect" upon the radio-frequency resistance of the windings.

For aerials longer than a quarter-wavelength, the tuning reactance takes the form of a high-voltage series-connected variable capacitor, or perhaps a bank of fixed ceramic capacitors in series with a variable inductance for fine tuning.

Adjustment

Transfer of power from transmitter-tank circuit to aerial-coupling unit is usually best effected by variable inductive coupling for balanced transmission lines; in the case of co-axial lines, direct connection to an adjustable tap on the tank coil is an alternative possibility. At the aerial end the line winding may be tightly coupled to the low-voltage end of the aerial coil, with efficient electrostatic screening in the case of balanced lines. Some form of turn adjustment must be provided to enable the correct matching impedance to be presented to the line under all conditions, and to facilitate this, a line voltmeter and line ammeter are sometimes fitted in the larger installations. Where a "rotating coil" type of primary winding is fitted for this turn adjustment, this must usually be carried inside the aerial coil, and precautions such as winding with copper strip taken to minimize eddy current losses.

For tuning the aerial circuit, a thermoammeter is usually provided in series with the loading-coil earth connection, when a coupling unit is in use; if an ammeter is required at the high-voltage end of the coil, thorough shielding will be necessary, including in most cases a wire gauze cover, joined to one terminal. In the smaller transmitters with direct aerial connection, a radio-frequency meter is frequently omitted, and the power-amplifier anode current loading taken as a guide to correct setting-up; alternatively, a small wideband transformer and rectifier-type meter can be made to give a sufficiently reliable indication of aerial current up to a few Mc/s.

When a dipole aerial is in use, coupling may be made direct from the power-amplifier tank circuit via a 70-ohm co-axial line, with little effect on loading or on radiation pattern compared with a balanced feed. It may also be convenient, when feeding a rhombic aerial from a small transmitter, to use co-axial output and perform the transformation to a balanced 600-ohm feeder by means of a tuned line transformer or an aperiodic wideband transformer fitted at some convenient spot.

Modulation

On telephony, amplitude modulation is predominant in all but the V.H.F. bands.

Anode modulation is most commonly used, since it combines ease of setting-up with relatively low audio distortion at high modulation levels. Where the screens of tetrode valves are supplied via a voltage-dropping resistance from the anode supply, the dropper may simply be fed from the full modulating voltage; a peak voltage of somewhat less than the anode H.T. will give 100 per cent modulation in most cases. In the case of some valves the screen dropper may be returned to the unmodulated supply without detriment. With larger valves, where a fixed-voltage screen supply is in use, a separate winding furnishing the appropriate proportion of the full audio voltage is usual; in many cases, however, it is found possible to modulate fully, with low distortion, by inserting a choke in the screen supply and modulating only the anode.

In all cases care must be taken to ensure that adequate reserves of cathode emission and of radio-frequency grid drive are available. The cathode bias resistance, if employed, must be de-coupled down to the lowest modulation frequencies. The tank circuit blocking capacitor appears in shunt across the modulator output winding and may give rise to problems at the highest audio frequencies.

In transmitters of 500 watts and above, anode modulation may be applied to the radio-frequency driver stage, and the power amplifier stage run as a Class B radio-frequency amplifier. In practice, to avoid excessive audio distortion, the power-amplifier grid bias must be set to give an appreciable standing current, in the absence of the carrier, and the degree of radio-frequency drive carefully controlled. Complications arise with power-amplifier valves which are driven positive at peaks of modulation, in which case the resultant grid current necessitates a driver stage of the requisite good regulation. In the larger equipments stability of operation becomes a major problem under Class B conditions.

Suppressor grid modulation is popular in the smallest equipments,

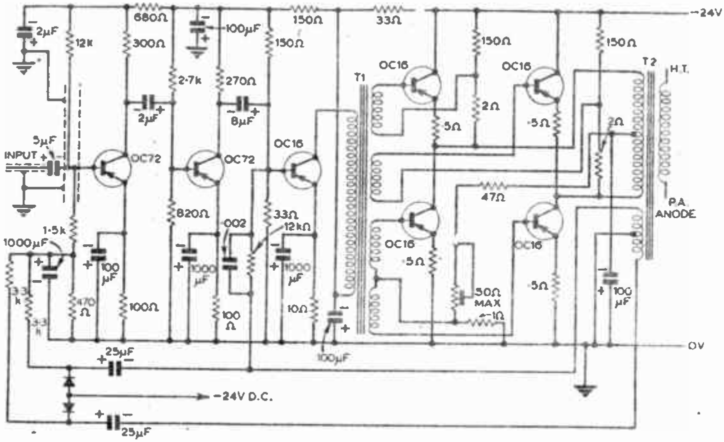

Fig. 16.—20-watt Transistor Modulator with Vogad.

specially designed valves, such as the Philips PE1/100, enable 90 per cent modulation to be obtained with relatively low distortion by application of an audio voltage to the biased-back suppressor grid. Setting up is simplified, since the valve is run in the Class C mode, without drive adjustments. Since the peak of modulation occurs at zero suppressor volts, there is no appreciable suppressor-grid current, and virtually no power is required for modulation.

Fully transistorized modulators are practicable for anode modulation of low-power transmitters, and afford valuable economies where batteries are the source of supply. Fig. 16 shows a modulator suitable for a 30-watt transmitter, using four 12-volt power transistors in a Class B bridge amplifier circuit with a 24-volt supply. A single transistor of the same pattern is used as a Class A driver, while the preceding stages provide microphone amplification with voice-operated gain adjustment (VOGAD). An A.F. output of 20 watts is obtainable with 5 per cent total harmonic distortion, for a peak current in the output transistors of 2 amps. If the H.T. for the accompanying valve-operated transmitter P.A. stage is obtained from a transistor D.C. converter, the overall power consumption can be reduced to well under half that of former equipment of the same output. Protection against accidental reversal of supply is essential in such circuits.

Keying

For most communications purposes, hand-speed keying at speeds of up to 40 words per minute is adequate, and is often used as a stand-by to telephony service under adverse conditions. Examination of the Morse characters for the typical word "Paris" will reveal that the dot length for 40 words per minute is approximately 30 milliseconds. Difficulty is frequently encountered in persuading relays and crystals to follow at this speed without serious clipping. Even for lower-speed keying, the less-skilled operators themselves tend to clip dots to about this value, and it is essential for reliable service that comparable clipping should not occur in the transmitter.

A number of refinements are essential in the basically simple technique of interrupting supplies to the oscillatory circuit before present-day requirements are met. First, with a variable-frequency oscillator, some change of frequency during the pulse is inevitable if the oscillator stage itself is keyed. This manifests itself as a warble in the note heard in the receiver, and is increasingly troublesome at higher radio frequencies. Secondly, some rounding-off of the pulse is necessary in order to avoid transients and excessive sideband spread as a result; this is sometimes achieved by the use of inductive and capacitative elements in the circuit of the keying relay contacts, and by the judicious employment of the modulation transformer inductance in transmitters with telephony facilities. Furthermore, direct interference with adjacent receivers due to sparking at key or relay contacts must be avoided by means of suitable radio-frequency suppression.

A further source of interference with local reception when buffer keying is in use lies in direct pick-up from the oscillator under "key-up" conditions, due to incomplete shielding of this stage. Screening of the necessary degree of perfection is difficult to achieve on a production basis, and wherever possible oscillator keying is favoured in the smaller equipments.

Quartz-crystal oscillators are relatively free from the "chirrup" mentioned above, but in some cases are reluctant to start with the necessary promptness; while circuit conditions have some influence, the position is somewhat obscure, since two crystals which otherwise appear identical may differ greatly in keying aptitude. When specifying crystals for this purpose, it is usual to stipulate the keying speed. A further trouble, frequently encountered in low-frequency circuits, is a tendency for oscillations to decay at too slow a rate when the key is lifted, owing to the high Q of the crystal; this does not prevent the valve from responding to keying, but may embarrass the local receiver if "listening through" facilities are in use. The tendency to "ring" may be overcome by heavy resistive damping of the crystal, preferably applied only at "key up".

The modern tendency is to avoid crystal keying despite the elaborate screening then necessary to avoid the radiation from affecting the local receiver. The lower amplitudes of drive which are demanded by modern plated crystals, particularly when improved frequency stability is required, help to reduce radiation.

With an H.T. supply that is inductively smoothed and only lightly loaded at "key up", damped oscillations are likely in the smoothing system, with a resulting adverse effect on the shape of the radiated characters.

To conserve the relay contacts, it is preferable to key at a point of relatively high impedance and low voltage, usually in a grid circuit. In addition, since the guaranteed relay life of perhaps a few million operations may be reached after quite a short period of service on telegraphy, the keying relay is, where possible, made a plug-in item.

In many small equipments, however, the permissible dissipation of the power-amplifier valve or valves would be exceeded by the mere removal of drive, on account of the use of grid-leak bias in this stage and/or the employment of a series resistor in the H.T. line for screen feed. In this case, both oscillator (or buffer) and power-amplifier stages may be keyed simultaneously, if necessary through the grids or screens of both stages. Alternatively, a resistance in the negative H.T. line may be keyed to provide a biasing-off voltage for the requisite valves. These methods are preferable to the once-popular method of opening the valve-cathode connection, where there is danger of breaking down the cathode-heater insulation before the cut-off voltage is reached.

Relays of sufficiently high rating for these purposes and capable of keying at 40 words per minute are readily available.

"Listening Through" (Break-in)

A more difficult application is the combining of "listening through" facilities with the use of a common aerial for transmission and reception. "Listening through", or "break-in" as it is sometimes known, enables the operator to receive incoming signals between the characters of his own message, and allows the distant operator to break in immediately clarification is wanted, often avoiding repetition of the entire message. Separate aerials are used for this purpose in the higher-power stations, and the difficulties are chiefly confined to avoidance of back wave and of excessive key clicks in the receiver. Where an aerial changeover relay is needed, a type must be chosen with a contact assembly of low inertia, combined with high-voltage insulation and large current-carrying capacity. This necessitates a correspondingly large coil and an efficient magnetic circuit. The relatively high inductance may then result in delayed current build-up in the coil, sufficient to slow down closure of the contacts, unless additional series resistance is included to reduce the time constant.

Typical Keying Circuit

The circuit of Fig. 17 illustrates these various points. With this arrangement a high-frequency transmitter of the 50-watt class is operated with an associated receiver, using a high-voltage, low-loss "type 3,000" aerial changeover relay X1. The relay inductance and resistance are approximately 35 henrys and 1,000 ohms respectively, giving a time constant L/R of 35 milliseconds. Series resistor R1, fed from the 500-volt H.T. supply, reduces the overall time constant to 4 milliseconds; to this is added a further 4–5 milliseconds, before contact closure, due to inertia in the moving parts. Use of a back contact on the key enables the keying "mark" to be anticipated slightly, thus reducing the delay.

The transmitter radio-frequency contacts are protected from sparking by energizing the keying relay X2 via the second set of contacts on X1, whilst, on lifting of the key, relay X2 (of the light-duty, high-speed type) interrupts the carrier before X1 drops out. The remaining contacts on X1 enable the receiver to be suitably desensitized to give side tone at "mark" by direct reception of the transmitted carrier. Change of resistor R2 enables this drop-out time to be controlled up to 120 milliseconds approximately, since the degree of damping of the coil influences the rate of collapse of the magnetic flux. This gives choice of receiver recovery time and permits "break in" at 40 words per minute between letters or between words, as desired.

Components R3, L1, C1 are fitted to smooth out the transients at

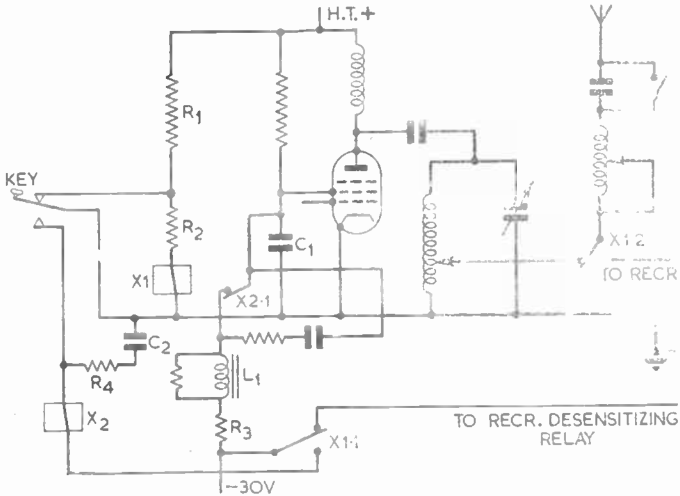

FIG. 17.—TYPICAL KEYING CIRCUIT.

beginning and end of the keyed impulse, and C2, R4 serve to suppress radio-frequency interference due to sparking at the key contacts.

Since common-frequency working is usual on A1 telegraphy, the aerial changeover contacts are situated at the base of the loading coil thus minimizing the voltage stress, and the receiver is fed at low impedance from the tuned-out aerial circuit.

Modulated C.W. (A2) is less commonly used, but is still popular in marine applications. Design problems are eased if the audio-frequency signal is interrupted at the same rate as the radio-frequency keying impulses, enabling momentary overmodulation, during carrier build-up and decay, to be avoided. A high-power audio oscillator directly modulating the power-amplifier stage is in favour nowadays, except for high-power transmitters, particularly where no telephony operation is required; where a telephony-type modulator stage is in use, measures may be needed to avoid excessive rounding off of the keyed pulse, due to the inductance of the modulation transformer.

For some A2 applications, notably V.H.F. telegraphy, the tone only is keyed, with a continuous carrier. This avoids the difficulty of fluctuations in carrier frequency, which become more severe the higher the frequency, and permits the use of full automatic-gain-control in the receiver.

Components and Finish

The need for transmitters to operate under arduous climatic conditions and to withstand rough handling without ill effect has led to radical changes in component design. With these changes has come the evolution of generally accepted performance standards, and the development of suitable humidity chambers and vibration tables for accelerated life tests, both of components and of complete equipments.

In this country the *Radio Components Standardization Committee*, arising out of the earlier *Inter-Services Component Committee*, has issued a comprehensive series of Guides, List and Specifications for various types of component for Service use, all subject to continuous review. A thorough system of type-approval testing to the specified climatic and durability requirements enables the quality of various component manufacturers' products to be certified. Later specifications on somewhat parallel lines, issued by the *Radio Industry Council* and *British Standards Institution*, recognize the need for classes of less-elaborate components suitable for more temperate conditions.

The requirements of the services are dealt with in their latest form by the Defence Specifications, published by H.M.S.O. The ultimate aim is to issue comprehensive British Standards covering also components for general applications, with guidance on their use.

Humidity Protection

For many purposes, the use of wax impregnation of small coils and capacitors gives sufficient protection, withstanding two or three cycles of heat and cold at 100 per cent relative humidity without ill effect, provided that the impregnation is thorough and flawless. Other varnishes giving similar protection are often preferred, on account of the tendency of some waxes to encourage mould growth. In some instances the impregnation is followed by treatment with an enveloping medium for added protection. Many varnishes have the effect of increasing coil losses at radio frequencies; the use of solutions of polystyrene is without ill effect, but does not always give lasting protection on account of its tendency to craze. For some components, such as transformers, not used at radio frequency, a bituminous compound is often used as the enveloper to follow vacuum impregnation.

These relatively simple methods of protection are sufficient even for tropical climates, where the apparatus is stored and worked on sites with reasonable air conditioning. The subsequent development of silvered ceramic terminals, however, has facilitated the development of fully sealed components, suitable for extreme conditions of heat and humidity. Early transformers and chokes of this class were considerably more bulky than their open counterparts, largely on account of inferior ventilation. With the development of " C " cores, built up from laminated, grain-oriented, silicon iron, however, the permissible magnetic flux density has considerably increased; as a result, a complete range of fully sealed transformers no bigger than earlier open types is now available. Such transformers are filled with oil or dry nitrogen, or sometimes evacuated before sealing. Whilst the permeability of the " C " core material is very high compared with normal silicon iron, the residual air gap between core halves, despite grinding to close tolerances, limits the effective permeability in the smaller sizes; consequently, interleaved laminations of conventional types, using nickel–iron alloys, are still employed for such items as audio input transformers, mounted and sealed in the standard " C " core case.

A substantial item in the price of a " C " core transformer is the very high cost of the terminal bushes, and difficulties are sometimes encountered in production and in the field owing to leakage of air and oil seals; consequently, a type of open " C " core construction has recently been developed, with attempts to achieve full climatic protection by processing with solventless varnishes. While such processes were initially very costly and of somewhat uncertain effectiveness, it is now possible to make a substantial saving compared with the enclosed type, for virtually equal protection. A still more recent development in transformers is the use of the resin-cast technique to give a completely potted assembly at relatively low cost.

Paper capacitors are now available in hermetically sealed cases employing special dielectric impregnants suitable for use at 100° C.

FIG. 18.—A MEDIUM-POWER TRANSMITTER, GIVING 750 WATTS ON C.W. AND 500 WATTS CARRIER ON M.C.W. OR TELEPHONY, IN THE BANDS 137–550 kc/s OR 2–25 Mc/s.
(*Redifon Ltd.*)

For the smaller values, a complete range of metal-clad tubular capacitors is made for the same temperature rating; these include compact and high-capacitance electrolytic types. Ceramic capacitors are available, wax-dipped or enclosed in a protective outer ceramic tube, while the larger transmitting-type ceramic capacitors rely for protection on their non-tracking qualities and the expulsion of surface moisture through the heat generated in normal use. The ceramic type of variable trimmer has, however, fallen largely into disuse, owing to the tendency to trap moisture by capillary action. To replace these, there is available a solid dielectric trimmer with an effective seal on the screw adjustment.

Wire-wound resistors nowadays are protected by vitreous-enamelled or lacquered coatings, depending on rating and tolerance, and in the larger sizes are fitted in clips for easy replacement. The vitreous surface is not entirely moisture-proof, partly owing to a tendency to craze, and trouble is sometimes experienced when running at very low loading, owing to electrolytic corrosion of the wire by trapped moisture. At normal rating such moisture would be driven off by heat before damage occurred. Silicone-protected wire-wound resistors, now available, are claimed to be more impervious to moisture. Carbon-composition resistors are usually of the ceramic-enclosed type in the smaller values, with carbon-film types used for high stability and higher wattage ratings; the tendency is for a very slow but continuous increase in resistivity with use.

In the main structure of equipments, corrosion, where encountered, is found to be largely electrolytic in origin; it is good practice to keep "contact potential" low by avoiding juxtaposition of unprotected metals whose potential differences in sea-water, referred to a calomel electrode, differ by more than 0·5 volts. This entails, for instance, the use of tin or chromium, rather than nickel, as a protective coating for brass in contact with cadmium plate or aluminium. Steel itself is likely to be zinc- or cadmium-plated, or else passivated before painting, whilst special alloys of aluminium are available for use under tropical conditions.

When steel is to be left unpainted, as in chassis contruction, the usual finish is cadmium plate followed by passivation, since unprotected cadmium is itself liable to rust after a short period.

L. P. L.

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SINGLE-SIDEBAND TRANSMITTERS

The requirements of the transmitting equipment for long-distance, fixed-service, radio-telephony are maximum useful output power per unit capital cost and unit running cost, minimum occupied band-width per intelligence channel in the radiated emission and high linearity so as to minimize loss of intelligibility due to distortion (and consequent cross-talk between channels in multi-channel operation).

These requirements are met by single-sideband operation. The basic type of emission is designated A.3a (S.S.B.), comprising a low-level pilot carrier, together with a single sideband containing the intelligence. A substantial increase in effective radiated power results as compared with conventional amplitude or double sideband, which is designated A.3 (D.S.B.). The two types of emission are compared in Table 1, based on equal total sideband power, as this condition gives equal received signal-to-noise ratio in appropriate receiver band-widths.

Thus an S.S.B. transmitter rated at 1 kW peak envelope power output (P.E.P.) will give the same received signal-to-noise ratio in half the band-width required for D.S.B. operation at a D.S.B. carrier power of 2kW. Or again, for equal transmitter peak envelope powers, S.S.B. will give 9 db better signal-to-noise ratio at the receiver and occupy half the bandwidth.

TABLE 1.—COMPARISON BETWEEN DOUBLE- AND SINGLE-SIDEBAND OPERATION

Characteristic	*D.S.B.*	*S.S.B.*
Sideband power, kW	1	1
D.S.B. transmitter carrier power, kW	2	—
Transmitter peak envelope power, kW	8	1
Transmitter final radio-frequency stage valve emission in arbitrary units	4 *	1
Transmitter peak radio-frequency voltage at output stage in arbitrary units	2·8	1
Transmitter final radio-frequency stage anode dissipation in kW for sideband power of 1 kW (corresponding to 100 per cent modulation)	1·5	1
Anode dissipation in kW for 0 per cent modulation	1	0·25
Approximate total power in kW taken from mains for sideband power of 1 kW	10	3
Receiver output power (arbitrary units)	2	1
Receiver band-width (arbitrary units)	2	1
Receiver noise power output (arbitrary units)	2	1
Receiver signal-to-noise power ratio (arbitrary units)	1	1
Receiver signal-to-noise power ratio if an S.S.B. receiver is used to receive one sideband of the D.S.B. transmission	0·5	—

* D.S.B. high-level modulation of Class C Power Amp.

A logical development from the basic S.S.B. emission is Independent Sideband operation designated as A.3b (I.S.B.), whereby a second, but completely independent, sideband is also associated with the same pilot carrier and radiated on the opposite side of it. Further intelligence channels are handled in the second sideband, and cross modulation, which would show itself as in cross-talk is minimized by carefully maintaining linearity of amplification throughout the transmitter.

Modern transmitters in this class consist of two main parts. These are :

(a) Drive Unit generating the required type of emission at low-power level and at a fixed frequency.

(b) Frequency Translator and linear Power Amplifier.

In the foreseeable future frequency stability will be so greatly improved, both in transmitters and receivers, that it will no longer be necessary to transmit a pilot carrier, A.F.C. being rendered superfluous.

Drive Unit

In the drive unit the S.S.B. or I.S.B. emission is generated at relatively low frequency (usually 100 kc/s or less), using balanced modulators and crystal lattice filters. The carrier frequency is balanced out in the modulator circuit and later reinserted at the desired predetermined low level to be radiated and act as a pilot for automatic-frequency-control and automatic-gain-control purposes in the receiver.

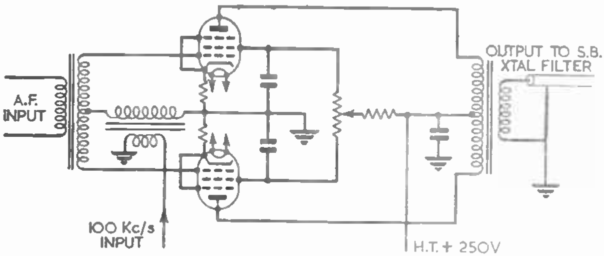

FIG. 19.—SIMPLIFIED CIRCUIT OF TYPICAL BALANCED MODULATOR.

A typical balanced-modulator circuit is shown in Fig. 19. In this circuit provision is made for adjustment of the balance after a change of valves, etc., but it is usual practice to include a narrow band stop crystal filter following the modulator completely to eliminate the carrier frequency, thus rendering the circuit independent of component drift, etc., which would upset the balance. By using a low frequency such as 100 kc/s, there is no difficulty in suppressing one sideband of the modulation process whilst allowing the other sideband to pass through with low attenuation. For I.S.B. operation two balanced modulators are provided. One balanced modulator in association with a band-pass filter pass-

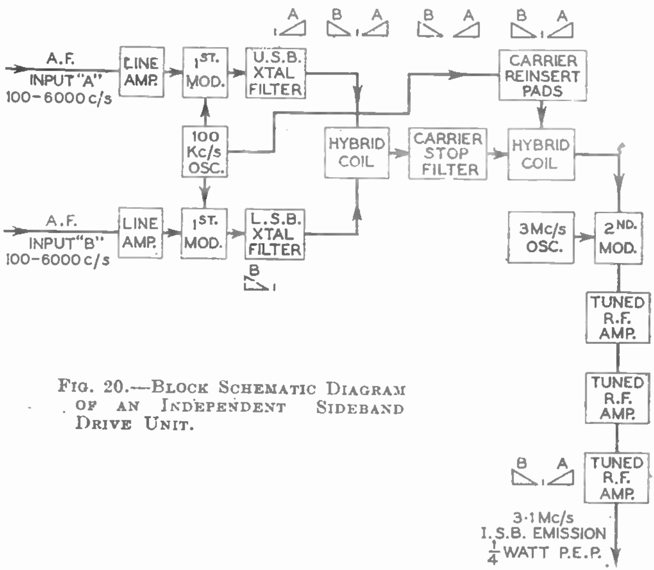

FIG. 20.—BLOCK SCHEMATIC DIAGRAM OF AN INDEPENDENT SIDEBAND DRIVE UNIT.

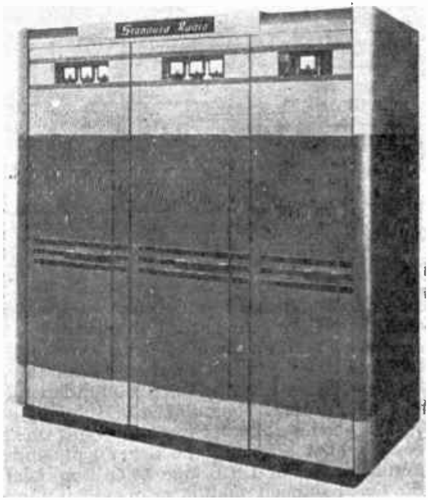

FIG. 21.—D.S.12 4 kW PEAK POWER TRANSMITTER.
(*Standard Telephones and Cables Ltd.*)

ing 100·1–106 kc/s furnishes the upper sideband (U.S.B.), and the other, in association with a band-pass filter, passing 94–99·9 kc/s furnishes the lower sideband (L.S.B.). A block schematic diagram showing the general arrangement of a typical drive unit is shown in Fig. 20. At this stage the two separate single sidebands are combined by means of a hybrid coil, and then the pilot carrier is added by means of a further hybrid coil. This results in an I.S.B. emission centred on 100 kc/s, which as a whole is then translated to an intermediate radio-frequency, usually between 1·5 and 3·5 Mc/s according to particular requirements. 3·1 Mc/s is a widely used and popular figure for this intermediate frequency. Filtering and amplification take place at the intermediate frequency (3·1 Mc/s), care being taken to maintain high linearity. The output power from the drive unit is usually between $\frac{1}{4}$ and $\frac{1}{2}$ watt.

Frequency Translator and Power Amplifier

The second main part of the equipment is the transmitter proper, which starts off with a frequency translator stage or mixer. Here again, a balanced-modulator circuit is often used for convenience. The S.S.B. or I.S.B. signal from the drive unit at 3·1 Mc/s is mixed with a frequency furnished by a suitable beating oscillator (usually crystal controlled) to give the required working or radiated frequency. Radiated frequencies below 10 Mc/s are obtained as lower sideband of this mixing process, i.e., $F_r = F_0 - 3{\cdot}1$ Mc/s: radiated frequencies above 10 Mc/s are obtained as upper sideband, i.e., $F_r = F_0 + 3{\cdot}1$ Mc/s.

This practice enables the overall range of the beating oscillator to be minimized. Its range can be further minimized by arranging to double and quadruple the basic oscillator frequency for parts of the band. A

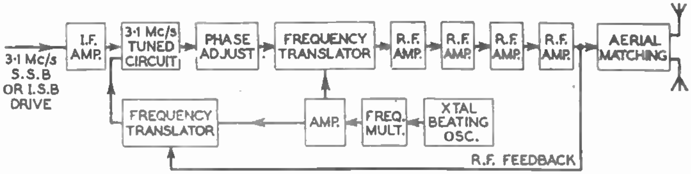

FIG. 22.—BLOCK SCHEMATIC DIAGRAM OF 4-kW TRANSMITTER USING RADIO-FREQUENCY NEGATIVE FEEDBACK.

definite lower limit to the radiated frequency range is set by the choice of the intermediate frequency, e.g., for an intermediate frequency of 3·1 Mc/s the lower limit of radiated frequency is about 4 Mc/s. To extend lower than this a lower value of intermediate frequency must be adopted, e.g., with an intermediate frequency of 2·1 Mc/s radiated frequencies down to 2·5 Mc/s can be obtained.

It is necessary to provide several tuned circuits tuned to the final radiated frequency, so as to reject all the unwanted products of the frequency translation process. The frequency changer stage itself has to be carefully designed to eliminate coincidence beats (i.e., 3·1, 6·2, 9·3 Mc/s . . .).

Following the frequency translation stage are several stages of radio-frequency amplification all at the final radiated frequency. These stages can be operated class A or class AB, and the final stage is usually worked closely approaching class B in order to get the highest possible efficiency consistent with linearity. Sometimes arrangements are made

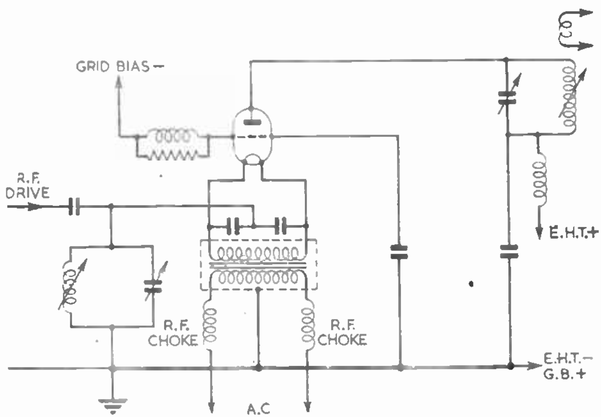

FIG. 23.—GROUNDED-GRID LINEAR RADIO-FREQUENCY POWER AMPLIFIER CIRCUIT.

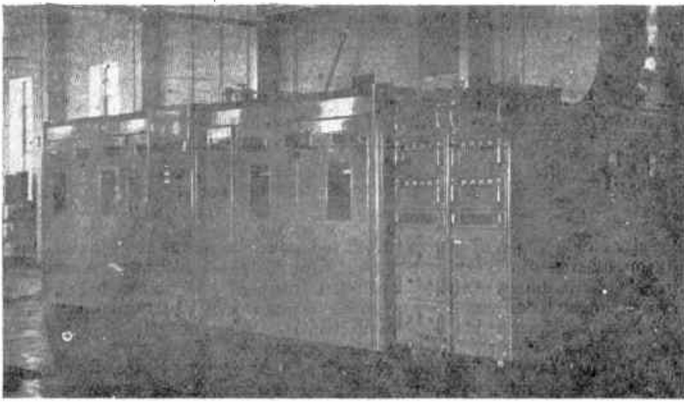

FIG. 24.—D.S.13. 40-kW PEAK POWER TRANSMITTER AND A1406/A1407 I.S.B. DRIVE AND MONITORING EQUIPMENT.
(*Standard Telephones and Cables Ltd.*)

for switching in additional grid-bias voltage to reach class C operation where the transmitter may be required to be operated on telegraphy service instead of telephony.

R.F. Negative Feedback

In order to maintain high linearity at the same time as high efficiency, radio-frequency negative feedback can be applied to transmitters of this type with considerable advantage A transmitter capable of giving 2 kW P.E.P. with an acceptably low degree of distortion on S.S.B. or I.S.B. operation will give some 4–5 kW P.E.P. at the same distortion figure with 12 db of negative feedback applied.

Fig. 21 shows an example of a modern transmitter rated at 4 kW P.E.P. in which radio-frequency negative feedback is used.

The application of negative feedback at radio-frequency is a little more complicated than at lower frequencies, and with the particular class of transmitter under discussion there is the further difficulty that the input frequency is different from the output frequency. There has thus to be a frequency change in the feedback path. The output frequency may be, for instance, anything between 4 and 28 Mc/s, whereas the input frequency to the transmitter from its drive unit is fixed at 3·1 Mc/s. In practice, the same beating oscillator which is used for the frequency translation in the forward direction is also used for the frequency change in the negative feedback path. In this way frequency errors are eliminated, but it is still necessary to adjust the phase, as with a relatively small change in circuit constants the feedback could become positive instead of negative. A typical block schematic diagram of a transmitter using negative feedback is given in Fig. 22.

Another useful arrangement is to employ a grounded-grid circuit in the final amplifier stage. This gives an inherent negative feedback effect, and, provided not too great a stage-gain is attempted, i.e., provided adequate drive is available from the penultimate stage, high linearity results. A typical circuit of a grounded-grid amplifier stage is given in Fig. 23. Fig. 24 shows an installed 40 kW peak power transmitter together with two I.S.B. drive units and monitors.

To simplify the tuning and adjustment of these transmitters, it is modern practice to gang as many stages as possible and to calibrate the common control directly in frequency. To achieve this, mechanical rigidity and accuracy of the component parts and tuning elements are essential, and electrical trimmers of the pre-set type are fitted to obtain accurate alignment over the whole of the frequency range. This technique lends itself to motor-controlled tuning; and in certain modern applications remotely controlled tuning covering up to, say, ten spot frequencies is provided. Where this is done, it is usual to multiply up the number of frequency-translator stages correspondingly, i.e., provide ten of these also, each one being selected as required by a low-loss radio-frequency contactor system.

Monitoring Equipment

With this class of radio transmitter it is important that the linearity of operation should be maintained over long periods. To facilitate this, relatively elaborate monitoring equipment is usually provided in association with the S.S.B. or I.S.B. drive unit. At this point the method of rating the power output of single-sideband transmitters should be explained.

As already indicated in the table of comparison between S.S.B. and D.S.B., the power output of an S.S.B. transmitter is normally stated as the peak envelope power (P.E.P.). This is the r.m.s. power reached at a peak of modulation, and coupled with it the permissible distortion has to be stated. The distortion is expressed in terms of the level of a particular cross-modulation product, or products, resulting from the application of two equal test tones at levels such that when they swing in phase the peak envelope power of the transmitter is just reached. This test is known as the two-tone test. In practice, the procedure for carrying out this test is as follows.

The monitoring equipment contains two audio-frequency oscillators which provide the two test tones. Widely used tone frequencies are 1,775 and 1,100 c/s. These are applied to one sideband of the I.S.B. drive unit, which in turn drives the radio transmitter working into a dummy aerial load. A portion of the output from the radio transmitter is picked off and taken to a monitoring receiver which demodulates it in stages, first to the intermediate frequency (e.g., 3·1 Mc/s), then to 100 kc/s and finally to audio-frequency. The output of the monitoring receiver will contain not only the original two test tones but also cross-modulation products such as the third-order product $2f_1-f_2$, the fifth-order product $3f_1-2f_2$, etc. By means of a narrow filter the third-order product is picked out and its level measured in comparison with the level of one of the test tones, using a volume indicator and calibrated attenuator.

In the past, a widely accepted standard of distortion measured in this way was for the third-order product to be -28 db relative to the level of test tone, but with the introduction of multi-channel operation on S.S.B. and I.S.B. it has become necessary to improve on this standard. Modern transmitters for this service require to meet a distortion figure of approximately -36 db relative to test tone.

As has been stated, the levels of the test tones are equal and such that when they add in phase the peak envelope power of the transmitter is just reached. This means that the level of each tone is −6 db relative to the peak envelope power. Therefore the distortion standard can equally well be quoted as −42 db relative to peak envelope power.

Measurement of Power Output

A word or two should be given concerning the measurement of the output power in the dummy aerial load. Perhaps the easiest method is to cause the dummy aerial load to heat either a flowing column of air or a flowing column of water. The temperature rise along the column is measured, and with a knowledge of the rate of flow the r.m.s. power can be easily calculated. The r.m.s. power figure obtained in this way is then multiplied by 2 to give the peak envelope power figure. Other methods are sometimes used, such as removing one of the test tones, measuring the r.m.s. power in the aerial load and then multiplying by 4, or again, measuring the output power as the product of current times voltage in the load, use being made of a cathode-ray oscilloscope to measure the peak voltage. The former of these alternative methods gives a rather optimistic answer, and the latter of the two methods a rather pessimistic answer. Care must be taken when discussing the performance of a single-sideband transmitter to have a clear understanding of the method which will be used for testing.

Pilot Carrier Level

Monitoring equipment can also incorporate means for measuring the level of the pilot carrier and its constancy with variation in modulation. With the aid of additional external test gear, measurements of noise in each sideband can also be made. Such noise figures are usually expressed in db below the peak envelope power or below the pilot carrier level.

Multi-Channel Operation

In the early days of the use of S.S.B. and I.S.B., the non-linearity of the available transmitters was such that there was serious danger of cross-talk when two intelligence channels were handled simultaneously. To reduce this trouble it became common practice to displace the intelligence channel in one sideband by a few thousand cycles. This practice could result in a reduction in cross-talk of between 5 and 10 db. However, such a practice was obviously very wasteful in band-width occupied by the emission. With the greatly improved linearity now available from modern transmitters it is possible to package intelligence channels adjacent to one another throughout the available band-width of the emission. Care, however, has to be taken to ensure that the peak envelope power of the transmitter is not exceeded due to the addition in phase of two or more of the intelligence channels.

Where it is a question of telephone channels only, it is not found necessary to reduce the level of each channel by the full anticipated amount, because of the diversity factor as between the various telephone conversations taking place. Thus, four simultaneous telephone channels may be handled by one I.S.B. radio transmitter, and the levels set so that each of the four telephone channels can in turn peak up nearly to the full peak envelope power rating.

Where it is a case of voice-frequency telegraphy only, i.e., the application to one sideband of a complete system of voice frequency tones, then there is no corresponding diversity factor, and the level of each tone is determined by the total number of simultaneous tones, e.g., for two telegraph channels the level of each is set at about 6db below the P.E.P. In these considerations the telegraph channels referred to are assumed to be frequency modulated (or two-tone). For four telegraph channels the level of each tone is set −12 db relative to the P.E.P., etc. Operation with mixtures of telephone and telegraph channels is also fairly common practice, and this introduces yet another factor. It is well established that the signal-to-noise ratio necessary for satisfactory operation of such a telegraph channel is some 12 db lower than that required for commercial quality telephony. Thus it is not necessary to transmit the tones of the telegraph channels in a mixed system at a very high level. A further reason for keeping the level low is to minimize the danger of cross-talk from the telegraph tones into the associated telephone channels. As an example, if four telegraph channels are handled in a mixed system associated with telephone channels, the level of each telegraph tone might be set at about −18 db relative to P.E.P.

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Smaller Single-sideband Transmitters

Single-sideband operation is also found of advantage in the smaller transportable class of transmitters, effecting valuable economies in size and power consumption for a given range, particularly if the S.S.B. generator is kept as simple as possible. One system of S.S.B. generation in a compact unit is the filter method, using crystal networks or mechanical filters; another is the phasing system in which R.F. and A.F. signals are subjected to a 90° phase shift. An entirely new system, introduced recently, is known as "The Third Method".[1,2] which can provide a relatively simple and compact generator, particularly when transistors are used in the amplifier and power-supply stages.

This method, using four balanced modulators, enables simple low-pass filters of moderate stability to be used instead of the bulky and elaborate sideband filters and carrier stop circuits found in larger commercial units. The intelligence is mixed in two separate balanced

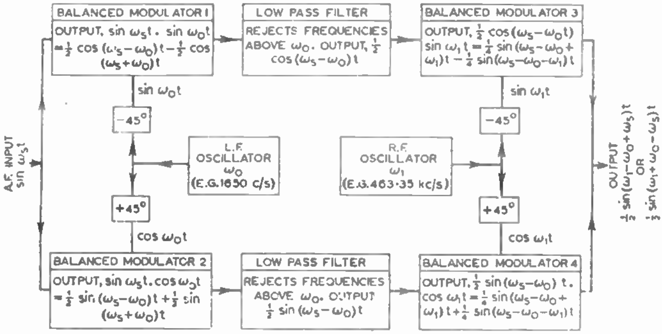

FIG. 25.—BLOCK DIAGRAM OF PORTABLE S.S.B. GENERATOR.
Reversing the output of balanced modulator 3 or 4 gives choice of upper or lower sideband.

modulators with a continuous signal of the mean audio frequency (e.g., 1,650 c/s) from an A.F. oscillator; reactances are employed to advance the phase of the oscillator signal by 45° in the first case, and retard it by 45° in the second case. In each case, the balanced modulator output, an unintelligible mixture of normal and inverted components, is passed through a low-pass filter to remove the addition products; then fed, in each case, to a further balanced modulator, where it is mixed with the output of an R.F. oscillator of the appropriate frequency. The signal from this oscillator is also advanced in phase by 45° in the first case and retarded by 45° in the other. On combining the resultant signals, half of the components cancel, the other components reinforcing each other as a single-sideband carrier-suppressed R.F. output. Reversing one-half of the connections to the combining unit will result in an output of lower, instead of upper-sideband form. The "appropriate" R.F. oscillator frequency mentioned above is not the R.F., of which the upper or lower sideband is required, but is displaced down or up from it by the mean audio-frequency. In the above case, for instance, the appropriate oscillator frequency for a 465-kc/s *lower* sideband output would be 463·35 kc/s. For this type of radio-telephone equipment this may be a most convenient frequency at which to generate the single-sideband output; particularly when it includes a spot frequency receiver with a 465 kc/s I.F. This is because the oscillator crystal for each channel in the receiver can often be used in generating the transmitter signal for that particular channel. Where, as is usual, the receiver local oscillator is 465 kc/s higher than the signal frequency, the R.F. mixer in the transmitter generator will produce an upper sideband output. The phase and frequency relationships in this system are shown in Fig. 25.

The relative simplicity of the A.F. low-pass filters can be judged from the fact that the lowest unwanted frequency, with an audio bandwidth of 300 – 3,300 c/s, is 300 + 1,650 c/s = 1,950 c/s; while the highest wanted frequency is 3,300 – 1,650 = 1,650 c/s.

The additional number of balanced modulators, which can be of the simple "ring modulator" type, is a small price to pay for this simplification, particularly since the same components can be used in the corresponding receiving circuit.

A transistorized single-sideband generator and detector of this type can be contained in a 19-in. rack-mounted unit, 3½ in. high.

MARINE TRANSMITTERS

The special requirements of ship-borne transmitters are governed by the Ministry of Transport for ships registered in this country, and are defined in the Merchant Shipping (Radio) Rules of 1952. These pay special attention to the easing of congestion in the marine wavebands and to factors influencing the safety of life at sea; dates are fixed after which all seaborne equipment must comply with these requirements. The inspecting and advisory authorities are the Post Office, who also issue specifications for voluntary radiotelephone and medium-frequency transmitters.

Generally, marine gear must work from D.C. supplies of relatively poor regulation, either 220, 110 or 24 volts, with H.T. usually derived from a rotary transformer; modern ships of higher tonnage, however, are nowadays frequently equipped with a 50-c/s A.C. supply. In all classes, extensive measures are often necessary for suppression of interference caused by electrical machinery and appliances. The use of A.C.-operated equipment with rotary converters is growing in popularity, since this avoids difficulty in the choice of transmitting valves suitable for series-heater operation and complications with bias circuits when one pole or the other of the supply may be earthed.

For specification purposes, the essential equipments are classified according to bands and function.

Long-wave Radio-telegraph Sender

Aerial power in this band (100–160 kc/s) is limited to 3 kW, and must be capable of progressive reduction in steps of 6 db or less, to 200 watts in cases where maximum output exceeds this value. At these frequencies efficient radiation is difficult to achieve, despite the low earth resistance, on account of the relatively short aerials provided. In order to reach the rated power without excessive loss into the smallest specified aerial, a high Q-loading coil, capable of dissipating considerable power and of withstanding exceptionally high voltages, is necessary. Thus, at one extreme, the aerial constants are 3·6 ohms + 300 pF; to achieve even 200 watts into this load, a current of 7·5 amperes is needed, with a resulting voltage of 37,500 volts r.m.s. at 100 kc/s, and with some 500 watts or so dissipated in the aerial tuning inductance.

With an aerial circuit of such low decrement, high tone-modulation levels and high-speed keying would be impossible, and type A1 (unmodulated) signals only with a keying speed of up to 30 bauds (approximately 40 words per minute) are called for. Sideband spread is limited to not more than 5 per cent of the radiated power appearing outside ±100 c/s from the nominal carrier frequency when transmitting dots at this keying speed. In practice, this limit will not be exceeded by an almost square pulse, provided that spurious peaks are absent.

Listening-through facilities are demanded at keying speeds up to

30 words per minute, and a separate aerial is invariably fitted for the associated receiver.

A frequency tolerance of 0·10 per cent is allowed by international regulations, and at least three predetermined channels, including the calling frequency of 143 kc/s, must be amenable to selection by adjustments taking not more than 10 seconds. For checking purposes, a dummy load with a radio-frequency indicator must be provided, as for other types of marine transmitters.

Medium-wave Radio-telegraph Sender

Four channels are required in this band (405–525 kc/s), one channel being the international distress frequency of 500 kc/s. Both A1 and A2 (M.C.W.) transmission must be possible at keying speeds of up to 30 bauds. A minimum aerial power of 50 watts is specified, with some concession towards the low-frequency end of the band. To provide this power into the specified aerial impedance under the damp-heat tests which form part of the Post Office climatic and durability requirements, an output approaching 100 watts will be needed from the power-amplifier stage. Ready reduction of power must be provided in steps of 6 db or less, down to 5 watts. This reduction may be more conveniently secured, in the smaller sizes of equipment, by some such device as reduction of the power-amplifier screen voltage rather than by adjustment of anode H.T. supply; compensation for the decreased loading on the modulator is, of course, necessary. The modulator itself is likely to be a self-excited oscillator, since no telephony service is called for. With A2 waves the permitted sideband spread for 95 per cent of the total power, at 30 bauds, is extended to $\pm 2{,}500$ c/s.

Short-wave Radio-telegraph Sender

One calling frequency and two working frequencies are necessary in each of the maritime mobile bands between 4 and 23 Mc/s. The permitted frequency tolerance of 0·02 per cent obviously points to quartz crystal control, although at the present date many ship installations do not reach this standard. Since channels in the 4, 6, 8, 12 and 16 Mc/s bands are always allotted in harmonic sequence to any vessel, considerable crystal economy is possible by arranging to use an initial oscillator frequency of about 2 Mc/s in each case.

Whilst a 50-watt transmitter can be made to satisfy the specification, outputs of up to 1 kW, or by special arrangement even more, are allowed. For the smaller units, it is frequently convenient to combine the functions of medium-wave and short-wave senders in the same transmitter, in which case radio-telephony facilities may also be provided.

Emergency Radio-telegraph Sender

This equipment, reserved solely for the transmitting of signals on the 500-kc/s international distress frequency, is operated entirely from secondary batteries. A minimum aerial power of 15 watts (M.C.W.) and full operation within 6 seconds of switching on are required, necessitating the use of directly heated valves or temporary boosting of the heater voltage * for the first few seconds.

* Patent No. 669,524.

While fully transistorized emergency transmitters have already been made, they cannot be considered an economic proposition until R.F. transistors of higher power rating are available.

The alarm signal comprises twelve dashes of accurately controlled duration; three or four successive dashes of this length and spacing are capable of alerting the auto-alarm receivers in other vessels within range. An automatic code sender must be fitted to enable this signal or an SOS call to be sent out in the absence of an operator, if necessary.

There is no objection to the use of the emergency transmitter for normal communication service on the medium waveband, provided that certain of the requirements of the medium-wave sender specification are also met.

Lifeboat Transmitters

In this case, facilities for M.C.W. transmission are necessary on both 500 and 8,634 kc/s, with full operation on whichever channel is required within 30 seconds of switching on. The requirement of a rectangular modulation waveform simplifies modulating arrangements and ensures maximum audibility in distant receivers. The local receiver is built into the same unit and shares the same aerial, using an aerial change-over relay actuated by the send/receive switch. Both fixed and mobile equipments are specified.

Output requirements for the fixed installation are for not less than 50 watts at 500 kc/s and not less than 15 watts at 8,634 kc/s, with power supplied from a 24-volt secondary battery of sufficient capacity for 4 hours of reception plus 2 hours of continuous transmission. The battery, of "non-spill" type, is rechargeable from a dynamo coupled to the lifeboat engine and from the ship's main supply, which furnishes power also to heaters fitted in the equipment case to lessen risk of freezing-up. Connections to the ship's mains must be so arranged as not to impair launching of the boat.

The minimum requirements for the aerial/earth system, which are specified in some detail, stipulate that failure of the aerial ammeter or "visual" radio frequency indicator (both of which are necessary) shall not break the aerial circuit; this points to the use of a current transformer rather than a directly connected lamp and meter.

A waterproof lamp of 3–15 watts must be provided for illumination of operating instructions and controls, whilst all controls must be suitable for operation by a person wearing thick gloves.

The portable lifeboat equipment is operated by hand-generator, with indicator lamp to show when the generator speed is within the correct range. Radio-frequency output is 3–5 watts approximately, and the equipment must float and remain serviceable after a drop of 30 ft. into water. A 30-ft. single-wire aerial is used, with an earth wire cast over-board, and a simple aerial tuning control and tuning indicator are fitted.

Automatic keying devices of similar performance to those used in the ship's emergency transmitter are required in each case.

Radio-telephony Marine Transmitters

Ship-borne transmitters with speech modulation are predominantly used for linking with telephone networks on land, in the case of larger vessels, or in self-contained radio-telephones working similar equipments in other ships or shore stations.

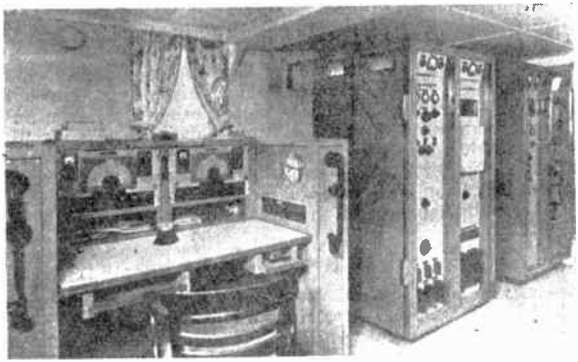

FIG. 25.—THE RADIO CABIN OF THE *Abraham Larcen* WHALING FACTORY SHIP.

(*Redifon Ltd.*)

Post Office radio-telephone specifications are issued as a guide to the requirements of "voluntarily" or "compulsorily" equipped ships, with maximum permitted power outputs of 50 and 100 watts respectively. Whilst the frequency-stability limits indicate the use of quartz-crystal control for both transmitter and receiver, the number of channels required in the receiver (and the excessive frequency errors in many old installations) render a continuously tuned receiver more practicable. In the compulsorily fitted equipment the specification insists upon a single switch for frequency selection of certain channels.

Simplex working (sometimes two-frequency) is usual, employing a common aerial with change-over relay for transmission and reception, the change over being effected by a voice-operated relay or by a switch adjacent to the microphone. Duplex operation, when employed, requires a high degree of receiver preselection, owing to the proximity of transmitting and receiving aerials.

The band employed for small ships' radio-telephones is 1·6–3·7 Mc/s, and in some circumstances facilities for telegraphy are permitted on certain channels when a qualified operator is carried. The transmitter-frequency selector switch must give special prominence to the telephony distress channel (formerly 1,650 kc/s, now changed to 2,182 kc/s, which serves also as a calling frequency). A frequent feature, although not required by the specification, is the provision of a "loud hailer" energized, when desired, from the transmitter modulator.

Owing to the extreme congestion in this band, stringent requirements are laid down for sideband spread, necessitating the use of efficient audio filters (cutting sharply above 3 kc/s), together with peak-clipping devices to prevent overmodulation.

Larger telephony transmitters are used with individual Post Office approval on ships of the highest tonnage. In these cases, use is made of the 4-, 8-, 12- and 16-Mc/s bands, on which telephony is allowed with a frequency tolerance of $\pm 0{\cdot}005$ per cent. The larger ship installations are now often equipped with single-sideband telephony transmitters.

LAND MOBILE TRANSMITTERS

Units of power rating up to 2–4 kW output are commonly set up in mobile installations for the medium-frequency and high-frequency bands, whilst up to about 100 watts output is used for V.H.F. mobile applications. Such equipment, whilst not subject to rigid specifications and standardization, such as is enforced for airborne gear, must be built to a high standard of robustness and be capable of withstanding severe climatic extremes. The units will operate to a large extent in uninhabited country under military or field-survey conditions, with a minimum of attention and maintenance over long periods. For the larger equipments separate vehicles are used for power supply and receiving purposes, with remote control of the transmitter, but with outputs of up to about 100 watts a single vehicle or trailer can form a self-contained transmitting and receiving station, with whip aerial or telescopic aerial mounted on the body; power can be supplied from a built-in petrol–electric A.C. generator or from secondary batteries charged from a D.C. generator. Much of the radio-frequency output is liable to be absorbed in the metalwork of the vehicle, unless care is taken to bond together major items, such as engine, frame and bodywork, with heavy copper braid or strip, whilst for all but the smallest whip aerials a radial counterpoise or buried earth is a necessity.

For short-distance work the V.H.F. bands are popular, giving a range somewhat greater than the optical limit, with very compact aerial systems, for a radio-frequency output of less than 50 watts. Horizontal polarization is generally favoured, except in circumstances where a circular polar diagram is necessary, as, for instance, in communication with aircraft, and a directional array can readily be brought into service by means of a telescopic mast mounted on the vehicle. For aeronautical work two vertical dipoles can conveniently be mounted on one mast without serious distortion of the polar diagram of either; for this purpose a non-conducting top section is used, with the aerials mounted one each side and separated vertically by about half a wave-length. An assembly of this type must, of course, be securely guyed for protection against wind.

On the high-frequency bands, range is interrupted by the skip distance, depending on frequency and time of day, but beyond this point effective communication is frequently maintained for several hundreds of miles with an aerial power of less than 100 watts. In this case the telescopic mast is insulated and used as the radiator, an aerial capacity of some 120 pF being obtainable with a total height, including the top whip, of 42 ft.

Whip aerials alone are frequently used on high frequency for shorter-range work, and the aerial capacity in this case may be no more than 50 pF, for example, using a 12-ft. whip. Auxiliary loading inductance may be necessary at the lower end of the band, or else the capacity may be suitably increased by the use of two or more whips spaced apart and connected in parallel. Where limited mounting space is available, a useful increase of capacity may be secured by the use of a two-whip assembly of V form, using a single-base insulator.

In the case of larger and lower frequency transmitters such ready mobility is not feasible, and a typical 64-ft. mast assembly, for use with a 2-kW transmitter, involves several hours of erection time. Two of these masts, which are split up into 8-ft. interlocking tubular sections of 3 in. diameter assembled on the ground, are erected 100 ft. or more apart by portable block and tackle and guyed at alternate sections using a pivoted base-plate and three buried guy-support plates in each case. The corresponding radial earth comprises twelve wires each 200 ft. long, carried on drums.

At the other extreme are small radio-telephone sets used by taxis, police cars, etc., operating on V.H.F. channels. Frequency modulation is popular for this application, making possible extreme compactness of transmitter and power unit. V.H.F. transmitter–receiver equipment is described in Section 8.

Meteor Burst and "Moon Bounce" Communication Systems

An interesting development in the field of communications transmission is the use of reflection from meteor trails, investigated recently by the Canadian Defence Research Board.[3, 4] The highly ionized trail left by a meteor entering the earth's upper atmosphere persists for a period amounting, in many cases, to several seconds. This trail can be used for two-way communication by the use of specially oriented transmitting and receiving aerials. Communication is intermittent, but the effect is so marked, even with meteors too small to be visible, and the number of such meteors is so vast, that practical use can be made of it with special equipment.

The intelligence to be transmitted is recorded on magnetic tape at normal speeds, stored as long as necessary and then transmitted in intermittent high-speed bursts. The mechanism is triggered off by the reception from the distant position of a monitor signal on a slightly different frequency, indicating the temporary existence of a transmission path. Use of a transmission rate of 600 words per minute, which is feasible with special high-speed low-inertia tape equipment at both ends of the link enables a mean rate of 60 words per minute to be maintained.

A carrier power of 100 watts at approximately 50 Mc/s has been successfully used over a distance of 600 miles, using five-element Yagi arrays. The system has a very high degree of privacy, and is immune from the normal ionospheric disturbances. Unlike the conventional "forward scatter" systems, for which much higher powers are used, it is unlikely to cause interference with other services.

"Moon bounce" communication has recently been achieved at 200 Mc/s between sites on the earth's surface several thousands of miles apart. The extreme signal attenuation due to the great length of transmission path was overcome by the use of highly directional aerial systems.

There are distinct advantages in the absence of fading and the constancy of signals received at various points of the earth's surface, despite the complexity of equipment and the need for the moon to be above the horizon at both transmitting and receiving sites.

GROUND-TO-AIR TRANSMISSION

Airborne communications equipment is discussed in Section 19, and the information here will be confined to ground equipment for the communication service.

The ground counterparts of medium-frequency, high-frequency and V.H.F. airborne transmitters generally follow conventional lines, being usually mounted together in standard racks on a site at some distance from the Control Tower, and where possible, remote also from the associated receivers. Carrier powers of up to 2 kW on the high-frequency band and up to 50 watts at V.H.F. are commonly used. Facilities are provided at the Control Desk for modulation of a number of channels simultaneously, and a typical remote-control unit will incorporate channel-indicator lamps to enable incoming signals from a number of receivers to be identified individually and dealt with singly or in combination.

At least one stand-by transmitter of each type is usually provided ready for instant use, particularly on the V.H.F. channels, where failure during close-range work would be a serious occurrence. For the same reason, separate single-channel remote-control units are often favoured, built into a compact assembly on the main control desk.

Amplitude-modulated telephony is predominant for short-range airport use of V.H.F., with "VOGAD" circuits incorporated in the transmitters to maintain a high average level of modulation. The ground transmitter usually employs a quartz crystal oscillating at one-eighteenth or one-twenty-seventh of the working frequency, followed by multiplier stages with preset tuning controls. Where telegraphy is required, this is of the keyed-tone constant-carrier type, necessitating automatic switching out of the transmitter "VOGAD" circuit when the remote-control unit is set to M.C.W. (A2).

With ever-increasing numbers of aircraft in service, it is general practice to restrict messages to the minimum in duration to avoid congestion; in course of time the use of still higher frequencies is inevitable for short-range work, leading to radical changes in transmitter-design technique, and perhaps widespread use of frequency modulation.

L. P. L.

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V.H.F. SCATTER COMMUNICATIONS

The use of the scattering effect, which has long been known to occur when radio signals pass through an ionized layer or are reflected by the ground, for practical beyond-the-horizon communication in the V.H.F. and U.H.F. ranges has advanced notably in recent years. As a result of the studies by E. C. S. Megaw in the United Kingdom and H. G. Booker and W. E. Gordon in the United States, it was shown that signals due to incoherent scattering caused by random fluctuations of the refractive index in: (1) the troposphere; and (2) the lower ionosphere; could be used for medium-distance propagation of V.H.F. signals, and this was confirmed in a series of experiments by the U.S. Bureau of Standards.

Ionospheric Scattering

Since the lower ionosphere is permanently turbulent, some degree of scatter can be detected at all times given suitable equipment; during periods of ionospheric disturbance which impair H.F. communications, V.H.F. scatter communications can be maintained, often at improved strengths.

With present equipment, the range of distance over which ionospheric scatter signals can be detected is about 600–1,250 miles from the transmitter. A typical equipment for multi-channel telegraphy operating at about 50 Mc/s requires a transmitter with some 50 kW output and with high-gain transmitting and receiving aerials (gain about 20 dB). Received signals fluctuate fairly rapidly, and there is also some seasonal and diurnal variation; some form of diversity reception is thus essential for reception of high-speed telegraphy. Error rates in reception tend to be high due to the presence of delayed signal components.

Tropospheric Scattering

Communication is also possible at distances up to a theoretical maximum of about 700 miles by the reception of energy scattered in the troposphere, using high-power transmitters of about 1–10 kW in the centimetric range; metric-range transmitters can also be used for this purpose but other forms of propagation usually account for some part of the signals received. An advantage of tropospheric scatter communication over ionospheric scatter is that the choice of frequency is much more flexible. The practical limit of tropospheric scatter transmission at present is of the order of 400 miles, and less than this where a number of telegraph channels must be accommodated.

Signals received by tropospheric scattering are subject to both rapid and slow fading; the frequency of the rapid fading (1–10 c/s) tends to increase with distance and with the radio-frequency. Generally frequencies within the limits 400–5,000 Mc/s can be utilized, with some falling off of the scattering process at the higher frequencies.

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8. V.H.F. TRANSMITTER–RECEIVER EQUIPMENT

8. V.H.F. TRANSMITTER–RECEIVER EQUIPMENT

V.H.F. RADIO SYSTEMS

In all radio systems the intelligence which is required to be conveyed to a distant point is used to modulate a carrier wave. According to the nature of the intelligence and the method of modulation, a certain frequency band-width is required. For example, to transmit a single speech channel, 300–3,000 c/s, using double-sideband amplitude modulation, a band-width of 6 kc/s is necessary, while a television service requires a band-width of the order of 3 Mc/s. For this reason the number of channels which can be accommodated in the frequency spectrum up to 30 Mc/s is limited. By comparison, the V.H.F. range—30–300 Mc/s—offers a considerably greater number of channels.

V.H.F. radio waves are attenuated rapidly beyond line of sight, so that the range of communication, depending on the aerial installation and terrain, is of the order of 40 miles between ground stations, under normal conditions. The same frequency can therefore be allocated to several systems and, provided that there is no overlap between the service areas, there will be no interaction between them.

The chief applications of V.H.F. are summarized as :

(*a*) Systems requiring wide band-width, such as television, multi-channel telephone links, telemetering and control links, and some radar equipments.

(*b*) Systems requiring comparatively short ranges, such as police, fire, ambulance, airport and harbour services, and private services for taxis, tugs, doctors, etc.

V.H.F. communication is particularly suitable for mobile applications, as these frequencies are not affected by atmospheric noise and, since efficient half-wave or quarter-wave aerials can be mounted on a vehicle, good communication can be obtained with low-powered equipment, causing a minimum of drain on the vehicle battery.

Channel Spacing and Frequency Control

For communication equipment, speech frequencies up to 3 kc/s have to be transmitted, and a band-width of 6 kc/s is thus required if amplitude modulation is used. Using frequency modulation, the band-width requirement increases to the order of 30 kc/s. The frequency spacing between adjacent channels is governed, not only by the band-width required for modulation, but also by the tolerance on the nominal frequency of the carrier. For a tolerance of 0·01 per cent, at 100 Mc/s, the carrier may be ± 10 kc/s relative to its nominal value. Frequency channels are often spaced at 50 kc/s intervals below 100 Mc/s and at 100 kc/s above 100 Mc/s, but recent equipment may be required to operate with channels of half these widths.

To obtain a frequency tolerance of 0·01 per cent, or better, it is necessary to employ quartz-crystal oscillators to control the frequency of both transmitters and receivers.

Choice of Modulation

For communication equipment the choice is between amplitude modulation (A.M.) and frequency modulation (F.M.), and there are many factors for and against each method.

For reasons to be mentioned later, F.M. transmitters use a phase modulator, and an audio-frequency correcting network is used to give virtually frequency modulation. Even with a phase variation of $\pm 90°$, it is necessary to multiply the crystal frequency thirty-two times to obtain 15 kc/s deviation with an audio-frequency input at 300 c/s; therefore, although the F.M. transmitter saves the cost of a high-powered anode modulator, it introduces the cost of a number of low-powered frequency-multiplier stages. These factors must be taken into account, together with the power of the transmitter. For low-power transmitters, A.M. equipment is cheaper, while for high powers F.M. is cheaper. There is little difference in the cost of A.M. and F.M. receivers, but F.M. receivers need closer frequency control if the advantages of F.M. operation are to be obtained.

Theoretically, there is a signal-to-noise ratio improvement in using F.M. rather than A.M., and this can be demonstrated in the laboratory; but, since this improvement depends on the carrier being received precisely in the centre of the receiver passband and on the crossover point of the discriminator, it is not always obtained under service conditions, due to the crystal frequencies varying with temperature and de-tuning of circuits by valve changes, ageing, etc. In general, F.M. equipment requires maintenance by more skilled personnel than are needed for A.M. equipment, and the equipment needs checking more often if optimum results are to be obtained.

It should be noted that the above remarks apply particularly to narrow-band F.M. communication equipment and not to F.M. broadcasting services, where a deviation of 50 kc/s or more is used, and the band-width of the receiver is much greater.

TRANSMITTER DESIGN

The complete transmitter comprises a high-stability crystal oscillator, frequency-multiplier stages to obtain the required radiated frequency,

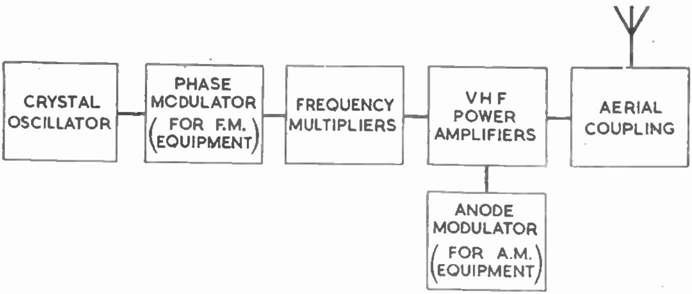


FIG. 1.—BLOCK DIAGRAM OF A V.H.F. TRANSMITTER.

·es giving the required output power, and aerial coupling her with some means for modulating the carrier. The · is normally delivered to a 50- or 70-ohm feeder cable for .on to a distant aerial, but may, in the case of walkie-talkie .ilar equipments, energize the aerial directly.

.onsiderable care must be taken throughout the transmitter to ensure .nat the spurious emissions are the minimum possible. One milliwatt, for example, can produce an interfering signal of the order of millivolts in an adjacent receiver working with a normal input of a few microvolts.

The highest possible crystal frequency should therefore be used, since the spurious emissions are spaced about the carrier frequency by the crystal freq..ency and its multiples, and can most easily be eliminated when this spacing is large.

A block diagram of a complete transmitter is shown in Fig. 1.

·The Crystal Oscillator

Crystals are now available which operate on their third overtone at frequencies up to 60 Mc/s, and these are attractive for use on straightforward A.M. equipment. For F.M. equipment—and for some A.M. equipment employing crystal switching to give alternative channels—fundamental crystals are used, usually in untuned circuits.

Temperature Variation

To cater for variations in ambient temperature as well as the temperature rise experienced in the equipment, the oscillator must keep within frequency tolerance with a range of crystal temperature between about 0° and +70° C.

Graphs of frequency error versus temperature, for A.T. and B.T. cut crystals, are shown in Fig. 2.

Referring to Fig. 2 (*a*) it is seen that—for B.T. cut crystals—the curve is approximately a parabola; the crystals can be adjusted in manufacture to bring the apex of the parabola to a specified temperature. Over a reasonable range of temperature near the apex of the curve the frequency error is comparatively small. For normal applications crystals having the apex of the parabola at about 40° C. are used. For applications where close frequency tolerance is required—such as

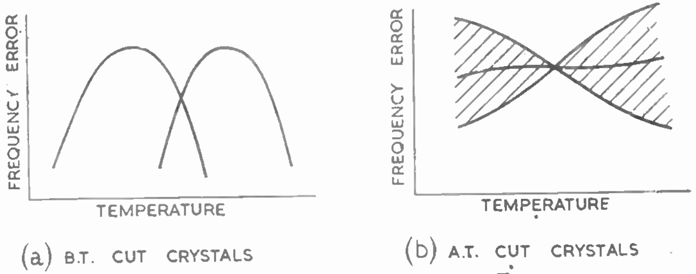


FIG. 2.—CRYSTAL FREQUENCY ERROR VERSUS TEMPERATURE.

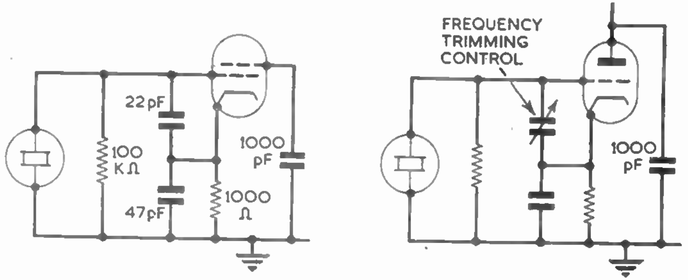


Fig. 3.—Fundamental Crystal Oscillator.

F.M. equipment and A.M. diversity equipment—the apex is often adjusted for 75° C. and the crystal is enclosed in an oven, controlled by a thermostat.

For A.T. cut crystals, the frequency error versus temperature curve lies between the limits shown shaded in Fig. 2 (*b*). The slope of the curve is determined by the angle of the crystal cut. Provided that care is taken in manufacture and the angle of cut held to within a few seconds of arc, the frequency error can be kept very small over a considerable range of temperature. A tolerance of 0·002 per cent can be obtained with a temperature variation between 0° and +70° C.

Apart from errors due to temperature variations, the oscillator frequency is also dependent on the capacitance across the crystal. This fact can be utilized by providing a trimmer capacitor across the crystal, and then adjusting the parallel capacitance to take up the manufacturing tolerance.

Crystals are often mounted in evacuated B7G valve envelopes. The vacuum reduces air damping on the crystal, leading to higher *Q*, and also helps to reduce change of frequency with ageing. For special applications, several crystals can be fitted in one B7G envelope.

For optimum stability, care must be taken not to exceed the manufacturers' power and voltage ratings for the crystal.

Typical Circuits

Some useful oscillator circuits are given in Figs. 3 and 4.

Fig. 3 shows a circuit suitable for fundamental crystals in the range 2–20 Mc/s. This is a very stable circuit, no tuning is needed, and crystal frequency trimming may be accomplished by a capacitor connected from grid to cathode (as shown), or alternatively, from grid to earth. This circuit is useful for applications where operation is required on a number of frequencies, the appropriate crystal (and trimmer, where fitted) being selected by means of a single-pole multi-way switch. Alternative connections for the crystal are between grid and anode, or grid and screen.

A circuit for overtone crystals, between 20 and 40 Mc/s, is shown in Fig. 4. *L*, *C* forms the resonant maintaining circuit. Either *L* or *C* can be made adjustable. The position of the tap on *L* should be checked

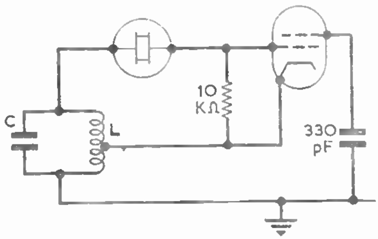


FIG. 4.—OVERTONE CRYSTAL OSCILLATOR.

for stable operation, as, if it is too low, the circuit may oscillate at a frequency determined by the resonance of L–C, the crystal merely acting as a coupling capacitor.

Frequency Multipliers

A frequency-multiplier stage is essentially a class-C amplifier in which the anode circuit is tuned to a harmonic of the input frequency. A high-Q anode circuit is necessary to reduce any component of the fundamental frequency in the output. For optimum efficiency the angle of flow of the anode current should be $90°/N$, where N is the frequency multiplication; the power output will be approximately $1/N$ times the output of the valve when used as a normal power amplifier with the same operating conditions.

Since the input and output circuits are tuned to different frequencies, there is little difficulty in achieving stability; neutralizing is not required, and triode valves can be used if desired.

The question of stability in the transmitter drive stages is eased appreciably if several doubler or trebler stages are used, each giving a reasonable amount of power amplification, instead of using one or two high-order multipliers followed by normal power-amplifiers.

To obtain the required small angle of anode current flow, the grid bias must be made higher than is usual for power amplifiers, and, in the case of pentode valves, the screen grid is run at a fairly low voltage. Grid current bias is general, the value of grid leak being fairly high.

The circuit of a typical frequency multiplier is shown in Fig. 5, and the waveforms associated with the circuit in Fig. 6.

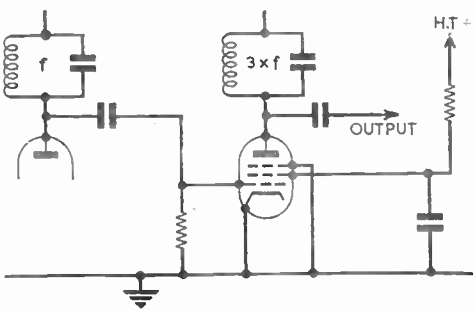


FIG. 5.—FREQUENCY-MULTIPLIER CIRCUIT.

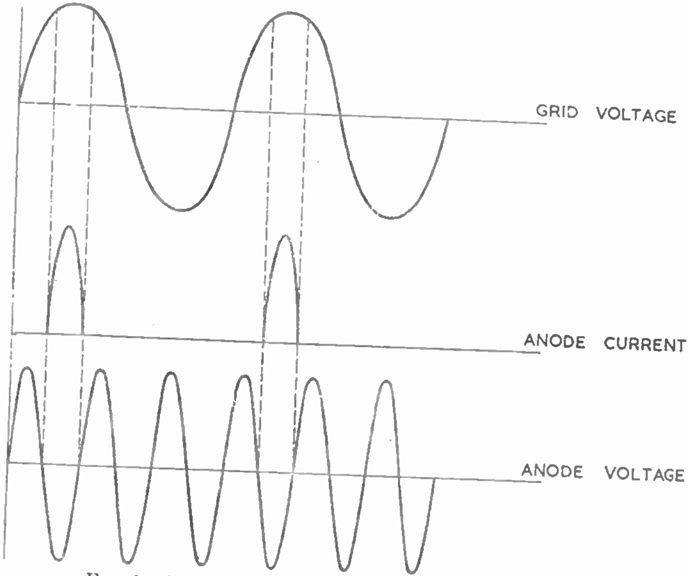


FIG. 6.—WAVEFORMS IN THE FREQUENCY MULTIPLIER.

The Power Amplifier

Owing to the propagation characteristics of V.H.F. radio waves, raising the power of a V.H.F. transmitter does not increase the range appreciably beyond the horizon; so that the power of the transmitter is chosen to give a reasonable signal-to-noise ratio within the required service area. For mobile radio systems, where battery consumption on vehicles is limited, 3–10 watts is the normal figure for A.M. equipment and 10–30 watts for F.M. Higher power is usual for fixed stations, although 25 watts is the maximum power permitted by the G.P.O. for most private systems. Some police services use powers up to 250 watts and, for many systems abroad, powers of 50–100 watts are common. The effective radiated power of fixed stations may easily be raised by using fairly simple, high-gain aerials.

Input Impedance

The input impedance of a valve falls as the frequency is raised, and the effect of inter-electrode capacitances becomes more important. The values of inductance and capacitance for resonance are very small, and the inductance of connecting leads is no longer a negligible factor at V.H.F. Point-to-point connections must be kept as short as possible. De-coupling capacitors must be chosen with care, since de-coupling will be quite ineffective if the impedance due to the inductance of the

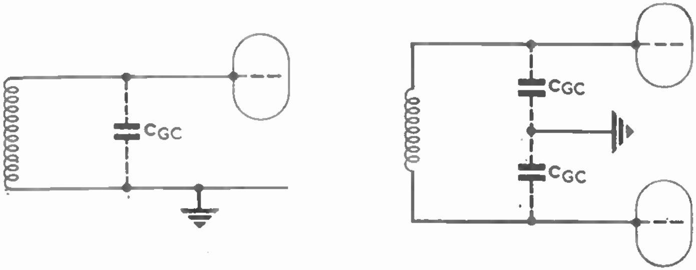


FIG. 7.—EFFECT OF INTER-ELECTRODE CAPACITANCE ON TUNED CIRCUITS.

capacitors becomes appreciable. Where good de-coupling is essential, a series-tuned *L–C* circuit is often used, and gives better results than more orthodox de-coupling. Fortunately, components and valves specially designed for V.H.F. applications are now more readily available, and their range is constantly being extended.

Push–pull operation, using beam tetrode or pentode valves, is usual–

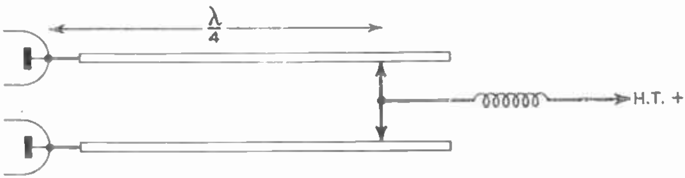


FIG. 8.—QUARTER-WAVE PARALLEL-LINE SECTION USED AS A CIRCUIT ELEMENT.

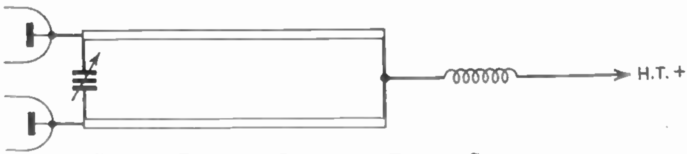


FIG. 9.—PARALLEL LINE WITH TUNING CAPACITANCE.

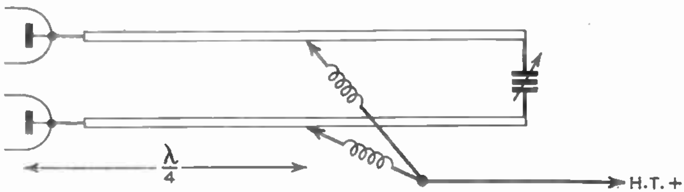


FIG. 10.—HALF-WAVE LINE USED AS A CIRCUIT ELEMENT.

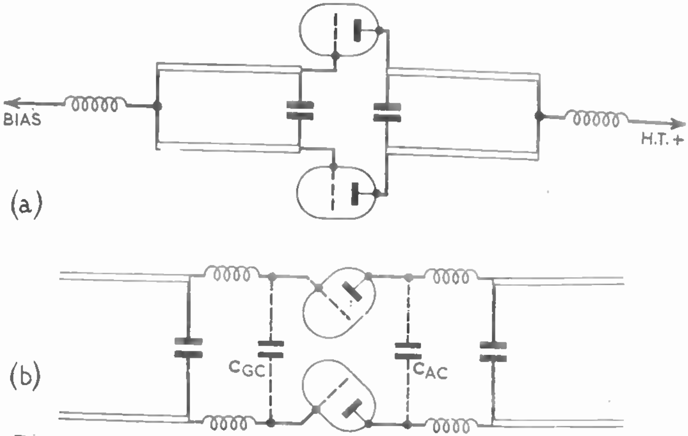


FIG. 11.—AMPLIFIER STAGE SHOWING INDUCTANCE OF THE CONNECTING LEADS AND THE INTER-ELECTRODE CAPACITANCES.

at least for the final stage—since the input impedances of the two valves are then effectively in series across the tuned circuit; the capacitances across grid and anode tuned circuits, due to inter-electrode capacitances, are thus halved, as indicated in Fig. 7. With single-ended stages, care must be taken to return one side of the input and output circuits to cathode, via a low-inductance path.

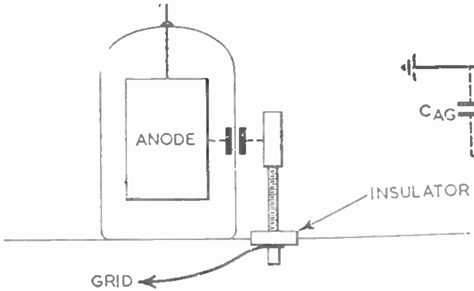
Parallel-line Tuning Elements

Up to about 200 Mc/s normal inductance coils can be used, but at higher frequencies orthodox circuits are usually replaced by parallel-line sections. Quarter-wave line sections, short-circuited at one end, are frequently used. Tuning may be accomplished by varying the position of the short-circuit, as illustrated in Fig. 8, or, more conveniently, by varying the effective length of the line by a variable capacitor across the open-circuit end, as shown in Fig. 9. The split stator type of capacitor should always be used, and best results are usually obtained if the moving vane is free from earth.

Fig. 10 illustrates a circuit using a half-wave line, which is more convenient for the higher frequencies, where the length of a quarter-wave line external to the valve may become small. It should be noted that the internal valve connections contribute to the total length of the line. The position of the D.C. feed point varies with frequency and, for optimum results, should be made adjustable.

Neutralization

Instability and spurious oscillations may occur due to anode-grid capacitance, and, even when tetrode or pentode valves are used, neutralizing is usually necessary above 100 Mc/s. A common cause of



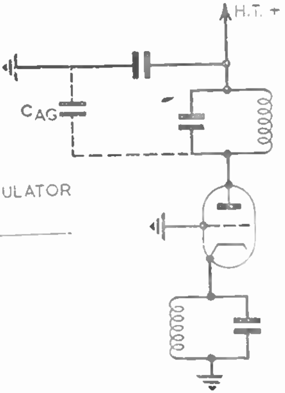


FIG. 12 (*left*).—METHOD OF OBTAINING SMALL NEUTRALIZING CAPACITANCE.

FIG. 13 (*right*).—GROUNDED GRID AMPLIFIER.

oscillation is described below to illustrate the type of problem which may be encountered. Fig. 11 (*a*) shows an amplifier stage, and Fig. 11 (*b*) shows the same circuit including the lead inductances and inter-electrode capacitances. It is seen that at some particular frequency above the working frequency—depending on the component values—the stage may oscillate as a tuned-grid, tuned-anode circuit.

Since the neutralizing capacitance required is only a fraction of a picafarad, a normal variable capacitor cannot be used. One solution is to use the direct capacitance between the valve anode and a screw which is mounted near the valve envelope and connected to the opposite grid. This arrangement is illustrated in Fig. 12.

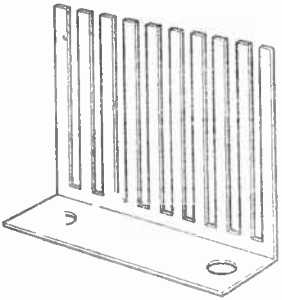
As an alternative to neutralizing, grounded-grid circuits can be used to overcome oscillation difficulties. A typical circuit is shown in Fig. 13. With this arrangement the cathode driving power appears in the output circuit, and in the case of A.M. the driving stage must therefore be modulated in addition to the final stage. Because of this disadvantage grounded-grid stages are usually confined to relatively high-powered F.M. transmitters.

Choice of Valves

Operation with triode valves is possible up to about 100 Mc/s, but due to the high drive power required, the tendency is to use tetrodes or pentodes.

For applications involving anode dissipations up to about 40 watts, valves are now available containing two tetrode assemblies in one envelope, with a common cathode and screen, and containing built-in anode-grid neutralizing. Some valves include the screen de-coupling capacitor inside the envelope to minimize the effect of lead inductance.

Circuit elements are usually silver-plated, and often polished, in order to reduce circuit losses due to skin effect. Depending on frequency and valves used, anode efficiencies of about 40–70 per cent can usually be obtained.



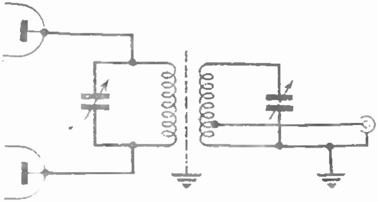


FIG. 14 (*left*).—FARADAY SCREEN.
FIG. 15 (*right*).—AERIAL COUPLING CIRCUIT.

Aerial Coupling

Attenuation of Spurious Radiations

V.H.F. transmitters are almost always coupled to the aerial via unbalanced co-axial feeder cable. Capacitive coupling, between the output circuit and the feeder, should be avoided owing to the low harmonic attenuation obtained by this method. Even when loop coupling is used there will be an appreciable amount of capacitive coupling between the circuits, giving appreciable harmonic output, unless precautions are taken. The aerial coil and final stage tuning coil should be separated by a Faraday screen, a convenient form being shown in Fig. 14. Fig. 15 shows the electrical circuit of a typical aerial coupling assembly.

Loop coupling should also be used, where space permits, between the multiplier stages of the transmitter, to keep down spurious radiations in the output, due to crystal harmonics.

If very low spurious output is essential, bandpass filter sections or resonant cavity filters may be inserted in series with the aerial feeder.

The Amplitude Modulator

Anode modulation is almost universally adopted for amplitude-modulated V.H.F. equipment, and the design follows standard practice. The final power amplifier is usually a push-pull stage operating in class AB, the output being coupled to the final radio-frequency power amplifier by a transformer. Combined anode and screen modulation is generally used, and this is sometimes accomplished by means of two windings on the modulation transformer. The usual arrangements are shown in Fig. 16.

Care must be taken in the design of the modulation transformer to avoid saturation of the core, due to D.C. polarizing current flowing in the secondary windings.

In communication systems intelligibility at low signal strengths, giving optimum range, is of more importance than high fidelity and, where transmitter power is limited—as, for example, in mobile installations—refinements in the modulation circuits can give considerable improvement in intelligibility under noisy conditions. Attention should be given to the following points:

(*a*) automatic gain-control of modulator;
(*b*) peak-clipping.

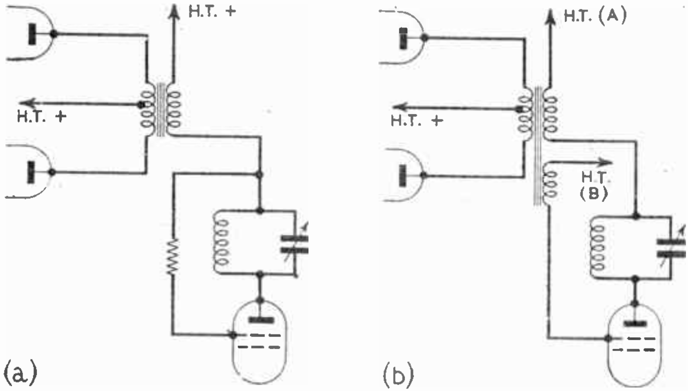


FIG. 16.—ANODE-MODULATION CIRCUITS.

Automatic-gain-control

Owing to variations in level of different voices, the peak modulation level may vary between 10 and 100 per cent, depending on the operator. Since peaks of noise in the associated receiver are equivalent to 100 per cent modulation when a low signal is being received, 10 per cent speech modulation will give relatively poor intelligibility. Automatic-gain-control circuits can be incorporated in the modulator to overcome this difficulty, and an effective arrangement is shown in Fig. 17.

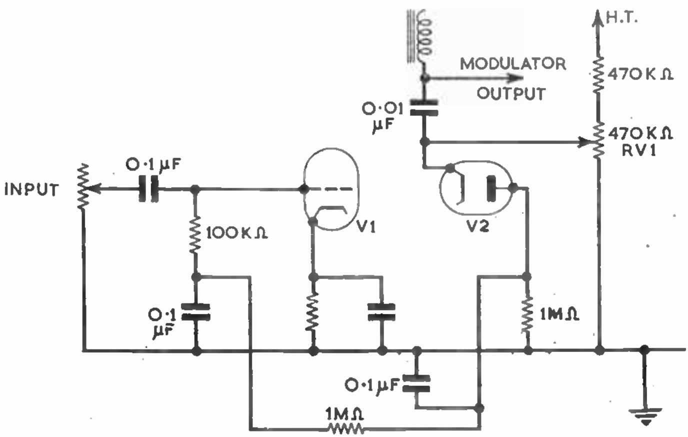


FIG. 17.—MODULATOR A.G.C. CIRCUIT.

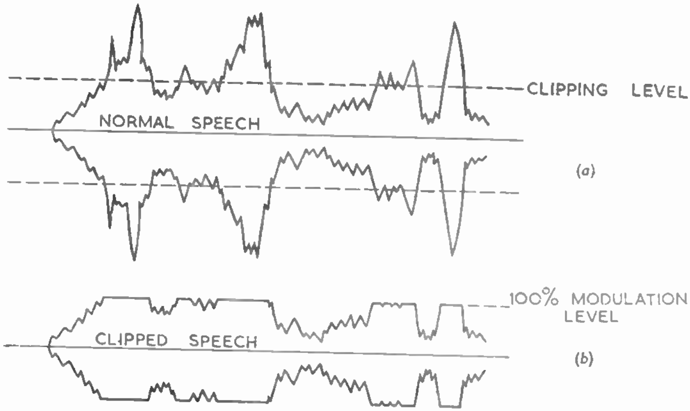


FIG. 18.—SPEECH WAVEFORMS SHOWING EFFECT OF CLIPPING.

This circuit uses a diode, V2, biased by potentiometer, RV1, which determines the input level at which the diode conducts. Part of the modulator output voltage is coupled to V2 and, on rising to a certain voltage, causes V2 to rectify and apply negative bias to a variable slope valve, V1, in the early stages of the modulator.

It is quite easy with such an arrangement to obtain a variation in output level of less than 2 db for an input-level change of 20 db. The circuit is also useful as a limiter for preventing over-modulation.

Peak-clipping

Examination of the waveform of normal speech indicates that it is of a very peaky nature, as indicated in Fig. 18 (*a*); the mean level is of the order of 30 per cent or less of peak value. If a clipper circuit is used to remove peaks of speech above a level corresponding to, say, 25 per cent of peak value, the waveform of the clipped signal will be as shown in Fig. 18 (*b*). The average modulation level is seen to be much greater than in the original waveform. The clipper circuit naturally introduces distortion, but comparative tests show a considerable improvement in readability when clipping is used in areas where the signal strength is low.

A variety of circuits can be used, the simplest being those incorporating a pair of diodes connected either in series or in parallel. Fig. 19 shows one variation of each. The series arrangement is generally preferred, since biasing then presents no problem. The operation is as follows :

Two diodes are used, one to clip positive peaks and one for negative peaks. The diodes are biased, so that although both are normally conducting, and therefore passing the audio signal to the subsequent stages, one is cut off by positive peaks and ceases to conduct, while the other cuts off on negative peaks, thus clipping the signal.

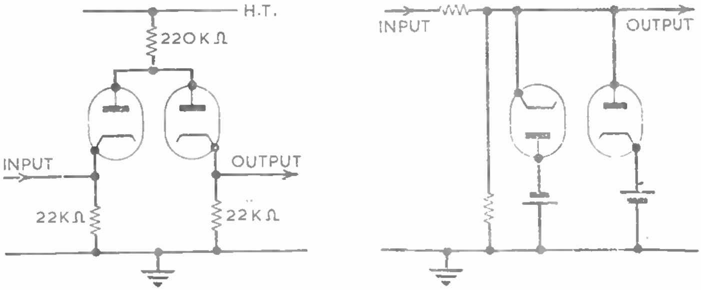


FIG. 19.—CLIPPING CIRCUITS.

The output of the clipper is often passed via a low-pass filter to eliminate the higher-order harmonics, which are produced. The circuit also acts as a limiter preventing over-modulation.

The complete modulator in block-diagram form is shown in Fig. 20.

It should be noted that the control voltage for the automatic-gain-control circuit must be taken before the clipper, if fitted, to be effective and an automatic-gain-control amplifier is sometimes necessary.

Microphones

Carbon microphones are commonly used on mobile equipments, since, owing to their high voltage output, a minimum of amplification is needed. Although relatively insensitive, moving-coil microphones are used on some fixed-station modulators when a better frequency response than can be obtained with carbon microphones is required.

The Frequency Modulator

Direct Frequency-modulation Circuits

Direct frequency modulation of a crystal-controlled transmitter can be achieved by connecting the equivalent of a capacitor or inductor into the oscillator circuit and varying the reactance in accordance with the modulating intelligence. The reactance valve circuit shown in Fig. 21 (*a*) is generally used to obtain this varying reactance.

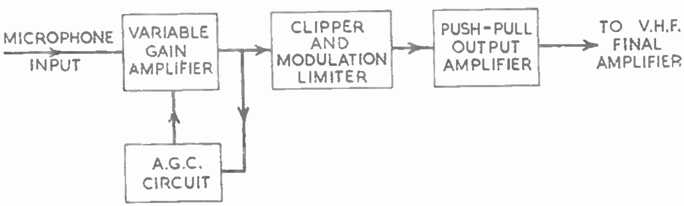


FIG. 20.—BLOCK DIAGRAM OF AN AMPLITUDE MODULATOR.

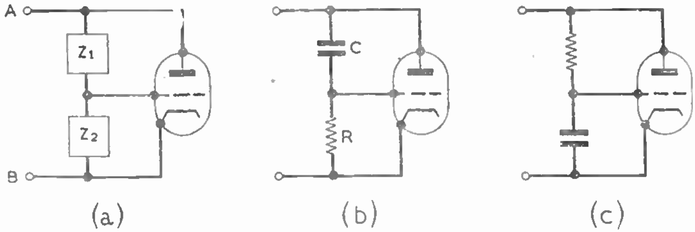


FIG. 21.—REACTANCE VALVE CIRCUITS.

The impedance across the terminals AB is given approximately by the formula :

$$Z_{AB} = \frac{Z_1}{g_m \times Z_2}$$

where g_m is the mutual conductance of the valve.

The circuit is most often used with capacitance between anode and grid, and resistance between grid and cathode, as illustrated in Fig. 21 (*b*). In this case :

$$Z_{AB} = \frac{1}{j\omega C g_m R}$$

and the impedance across AB is therefore equivalent to that of a capacitance of value $g_m RC$. If the mutual conductance of the valve is now varied in sympathy with the modulating signal, the capacitance across AB also varies, and will modulate the frequency of the oscillator.

The circuit shown in Fig. 21 (*c*) is an alternative arrangement equivalent to a varying inductance. Other combinations are possible using inductors in place of the capacitors shown.

The chief disadvantage of this arrangement for direct frequency modulation is that the centre frequency of the oscillator—in the no-modulation condition—is governed by the mutual conductance of the valve, and this will vary with ageing, supply voltages, change of valve and the like. Although these factors can fairly easily be controlled in the case of broadcast transmitters, direct frequency modulation of the crystal is inconvenient for simple communication equipment, which often has to be serviced by comparatively unskilled personnel.

If, however, the varying capacitance, due to the reactance valve, is connected across a tuned circuit in one of the drive stages of the transmitter (subsequent to the oscillator), the varying reactance—in altering the tune of the circuit—will alter the phase of the signal, thus producing phase modulation. Some amplitude modulation will also be introduced if the phase swing is large.

Deviation

In the case of F.M., the frequency variation, or deviation, is proportional to the amplitude of the modulating signal, while for phase modulation the phase variation is proportional to input level, and the resultant deviation therefore varies with modulating frequency. It is therefore necessary to introduce a 6-db/octave bass-boosting network between the microphone and reactance valve to obtain F.M. using a phase modulator. This network can be omitted if pre-emphasis is required, and de-emphasis is then provided on the associated receiver.

The deviation obtainable from the phase modulator is equal to the product of the modulating frequency and the phase swing in radians. Since the maximum phase swing is 90° the maximum deviation is:

$$\frac{90}{57} \times f$$

where f is the modulating frequency. If the lowest modulating frequency is taken as 300 c/s, the deviation is approximately 500 c/s. The deviation is independent of the carrier frequency. The carrier frequency and deviation must therefore be multiplied approximately thirty-two times to give 15 kc/s deviation on the radiated frequency. In practice, a phase swing of 90° cannot be obtained with this circuit, owing to the distortion produced, which rises rapidly with increase of phase swing, and the crystal frequency has to be multiplied more than thirty-two times before the output stage, unless a loss of low-frequency response is tolerated. The low-frequency response is usually adjusted so that a maximum phase variation of about 20° is used. As the modulating frequency is raised the phase swing is correspondingly reduced to give correct deviation, so that distortion is less at the higher frequencies.

The circuit of a simple reactance valve phase modulator is shown in Fig. 22.

Automatic-gain-control, limiter and clipper circuits, as described for the amplitude modulator, can also be used for frequency modulators. It is particularly important in F.M. systems to include limiter circuits to prevent over-deviation, since there is no automatic limiting action such as occurs in amplitude modulators at the 100 per cent modulation level.

RECEIVER DESIGN

It is generally considered that a signal-to-noise ratio of 10 db is the minimum requirement for intelligible speech reception. According to the modulating system used and the audio-frequency band-width of the receiver, a signal between 0·5 μV and 2μV is needed to give this signal-to-noise ratio with the valves at present available. A level of about 5–10 volts is desirable at the demodulator for efficient operation, and this sets the overall gain requirement. It is difficult to achieve large stage gains at V.H.F. and, since the gain from aerial to demodulator may be of the order of 140 db, superheterodyne circuits are always used. Double-superheterodyne circuits, using two intermediate frequencies, are common.

The receiver must also include sufficient selectivity to eliminate unwanted carriers, on adjacent channels, which may be simultaneously received at much higher levels. Unless care is taken in the design of the receiver, these large unwanted signals can give rise to blocking and cross-modulation of the required signal. At the same time the bandwidth of the receiver must be sufficient to pass the sidebands of the wanted signal and also to allow for frequency drifts due to temperature

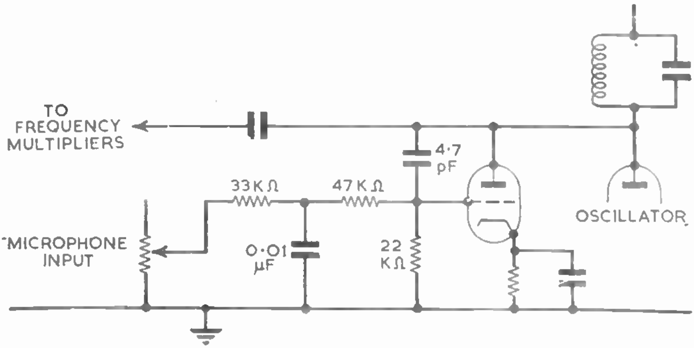


FIG. 22.—PHASE MODULATOR USING REACTANCE VALVE.

changes of the crystal oscillators in both the receiver and the associated transmitter. The band-width will therefore depend on: (1) crystal stability; (2) method of modulation, i.e., A.M. or F.M.; and (3) the required speech band. In practice, band-widths of 20–50 kc/s are usually required.

Good automatic-gain-control performance is essential for A.M. receivers used in mobile systems, since the input signal may vary by 100 db or so, according to the range from the H.Q. transmitter. Noise limiters must be included in A.M. receivers to reduce or eliminate impulsive interference, particularly from car ignition systems.

Spurious responses to unwanted carriers must be as low as possible and care must be taken to guard against radiation from the local oscillator.

To reduce blocking and cross-modulation, the gain should be kept to a minimum before the main intermediate-frequency amplifier circuits, since the radio-frequency tuned circuits cannot normally be made selective enough to give any appreciable rejection of unwanted carriers within a few Mc/s of the required signal. In this way the signal level is kept low at the mixer grids, where cross-modulation is most likely to occur. Cross-modulation is liable to occur in any component, or valve, having a non-linear characteristic, which gives rise to mixing action. At least one radio-frequency stage is, however, necessary, in order to give a reasonable input to the mixer and to isolate the local oscillator from the aerial circuit.

If the radio-frequency gain is kept low, the gain of the intermediate-frequency amplifier must be considerable, and difficulties may then arise in the intermediate-frequency stage due to instability. This difficulty is often overcome by using double mixing and two different intermediate frequencies. Double mixing also helps to reduce the level of the image response.

Crystal control is used for receiver oscillators in order to cater for close channel working, and the oscillator circuits employed are similar to those used in the transmitter.

Effective limiting is necessary in F.M. receivers if the theoretical advantages which this system claims over A.M. are to be obtained,

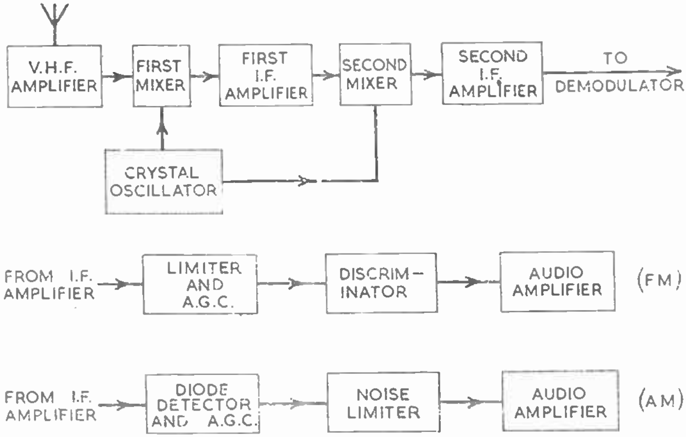


Fig. 23.—Block Diagram of a V.H.F. Receiver.

and at least two limiter stages should precede the discriminator. A variety of discriminator circuits is available. Automatic-gain-control is not always used in F.M. receivers, but may be included in some designs to prevent over-loading of the mixer on strong signals.

Diode detectors are used for A.M. receivers, and noise limiters are almost always incorporated. The use of efficient noise-suppression circuits on A.M. receivers can do much to make A.M. and F.M. systems comparable for communication services. The noise-limiter circuit follows immediately after the detector.

Signal Frequency Amplifiers

The noise factor of the receiver is determined principally by the noise produced by the first radio-frequency valve. Short-grid-base pentodes, or grounded-grid triodes, are principally used at the present time, and noise factors of the order of 6 db are obtained at 100 Mc/s. High-Q anode circuits are required, but, since appreciable amplification is undesirable, the coupling to later stages is frequently tapped down the circuit, low-value coupling capacitors being used. This has the further advantage that the low input-impedance of the following stage does not produce too much damping across the circuit.

Slug-tuned inductance coils take up less space than circuits using variable capacitors, and, although their Q is generally lower, they are frequently used in mobile equipments.

For the higher frequencies, or, where high-Q is essential to obtain good selectivity at signal frequency, conventional L–C circuits may be replaced by line sections. Tapped co-axial quarter-wave lines are normally used for these applications.

Fig. 24 shows a typical radio-frequency circuit for 100–200 Mc/s. An alternative cascode circuit is shown on page 8-25.

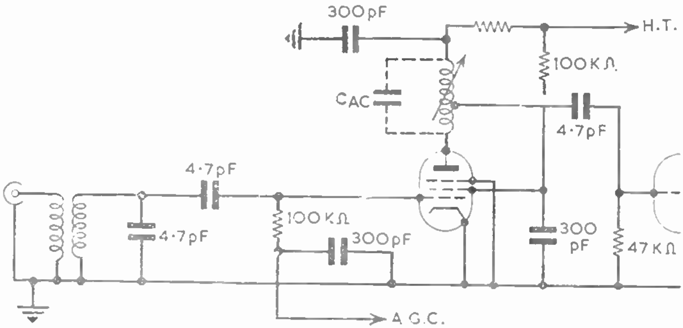


FIG. 24.—V.H.F. AMPLIFIER CIRCUIT.

Mixing

Generally, the best and simplest mixer at V.H.F. is a pentode or triode valve with both signal-input and local-oscillator-output injected on the first grid; although, for "walkie-talkie" equipment, crystal mixing is often used, since available valves with 1·4-volt filaments do not usually give good results.

When double mixing is required, two separate local oscillators can be used, and although this involves the provision of two crystals, one of them can be cut for a fixed frequency, regardless of signal frequency. In this case the second oscillator must be carefully screened from the radio-frequency and first-oscillator circuits, since harmonics of the two oscillators may beat to produce frequencies at, or near, the signal or intermediate frequencies. An alternative method is to use a single oscillator and either inject the same frequency twice or inject different harmonics of the oscillator at the two mixers.

The first method is illustrated in block form in Fig. 25. For example, taking a signal frequency of 100 Mc/s and a second intermediate frequency of 2 Mc/s, the local oscillator injects a signal of 49 Mc/s to the mixer-valve grid. This beats with the 100-Mc/s carrier to give the first intermediate frequency of 51 Mc/s. The 49-Mc/s local-oscillator frequency again beats with the 51-Mc/s first intermediate frequency in the second mixer, to give the second intermediate frequency of 2 Mc/s. In the arrangement shown, the pass-band of the first intermediate-

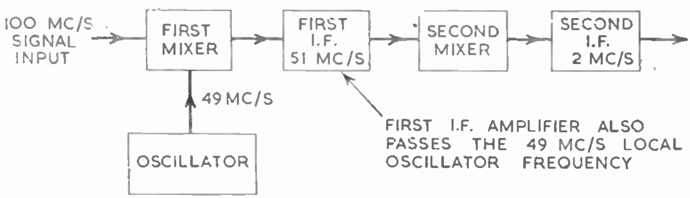


FIG. 25.—BLOCK DIAGRAM OF A DOUBLE MIXER.

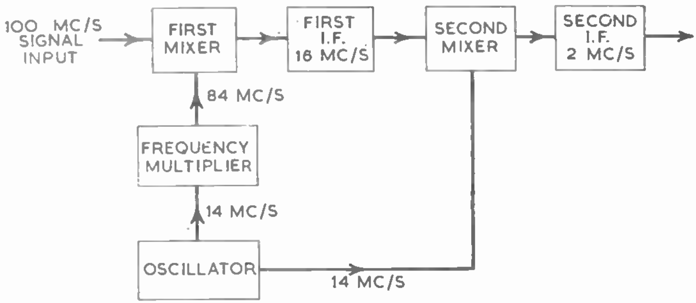


FIG. 26.—ALTERNATIVE METHOD OF DOUBLE MIXING.

frequency amplifier is made wide enough to pass both the 49- and 51-Mc/s signals, thus making a connection from the oscillator to the second mixer unnecessary.

Fig. 26 shows an alternative arrangement, which is more flexible in that it allows a choice of the first intermediate frequency, and the first-intermediate-frequency amplifier can be made more selective. Again, taking an intermediate frequency of 2 Mc/s and a signal frequency of 100 Mc/s, a local-oscillator-frequency of 14 Mc/s could be used. The oscillator output is passed through a sixth harmonic generator giving an output of 84 Mc/s, which beats with the carrier to give a first intermediate frequency of 16 Mc/s. The 14-Mc/s fundamental output of the oscillator is then injected to the second mixer to give the final intermediate frequency of 2 Mc/s.

The oscillator and frequency multiplier can be combined in one stage by using a grid–cathode or grid–screen oscillator circuit, and tuning the anode circuit to the required harmonic. A typical circuit is shown in Fig. 27.

A high-Q anode circuit is essential to keep unwanted crystal harmonics from the mixer, since this would cause the receiver to have poor discrimination against unwanted frequencies, spaced by multiples of the crystal frequency from the carrier. The oscillator is usually coupled to the mixer by a low-value capacitor of the order of a few picafarads in order to minimize pulling between the circuits.

Intermediate-frequency Amplifiers

Choosing the value of the final intermediate frequency is governed by a number of factors. A low frequency will give good stage gain and selectivity, but difficulties may arise in obtaining sufficient band-width. A higher frequency will generally simplify the problem of obtaining good image rejection.

Intermediate frequencies in use at present vary between 450 kc/s and 10 Mc/s. In order to obtain an intermediate-frequency response close to the ideal, where the gain is level over the required pass-band and then falls immediately to a very low level, band-pass circuits, using transformer coupling, are usually employed. An alternative arrangement is described later.

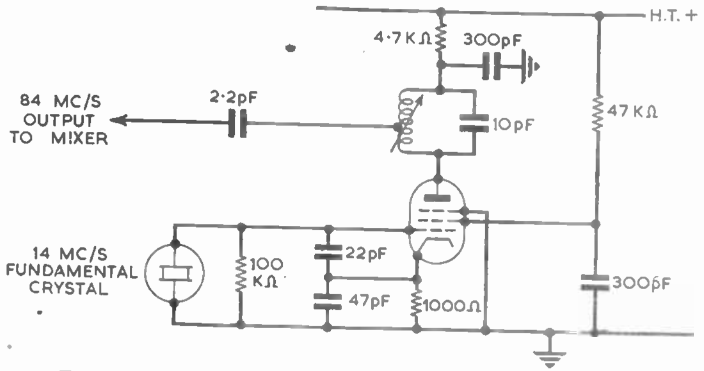


FIG. 27.—COMBINED OSCILLATOR AND FREQUENCY MULTIPLIER.

Good temperature characteristics are required to prevent variation of the centre frequency of the intermediate-frequency stage, since this will reduce the rejection of an unwanted carrier on the adjacent channel 50 or 100 kc/s away. In F.M. receivers variation of the intermediate frequency can produce considerable noise and distortion. Judicious choice of positive and negative temperature-coefficient capacitors is necessary for all tuned circuits, and care should be taken, if capacitors are changed when servicing, to replace them with new ones of appropriate temperature characteristics. The use of a low intermediate frequency is an advantage, since temperature drift is proportional to the working frequency.

Although, due to the selectivity of each of the tuned circuits in the amplifier, the ratio between unwanted and wanted signal is reduced, there may be some stage gain at the unwanted frequency if the frequency separation between the two carriers is small; so that, if the unwanted signal is much higher in level than the working carrier, blocking and cross-modulation may occur somewhere in the intermediate-frequency chain. In some receivers this problem is overcome by providing all the required selectivity in a multi-section band-pass filter immediately following the last mixer stage. Inexpensive resistance–capacitance-coupled, or choke–capacitance-coupled amplifiers can then be used to give the necessary stage gain. Filters for this purpose are usually aligned and sealed before despatch from the factory. With this arrangement, valve changes, when servicing, do not cause de-tuning, which can have undesirable effects in conventional intermediate-frequency amplifiers.

A.M. Demodulation and Noise Limiting

Diode detectors are almost invariably used for A.M. receivers. The circuits are simple, and no complicated lining up procedure is involved.

A.M. noise limiters operate on the principle of shorting-out or presenting an open-circuit to any narrow, steep-fronted pulses in the audio waveform, and are set to operate at a level corresponding to a certain

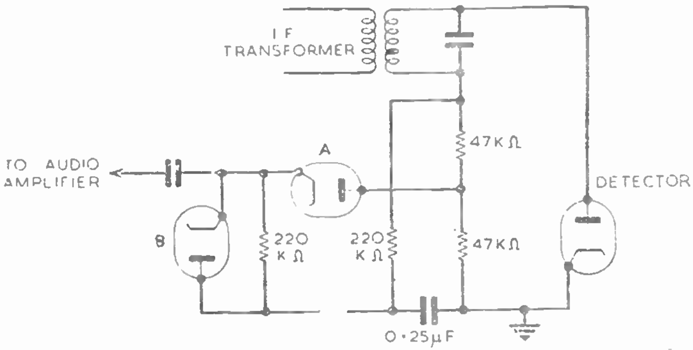


FIG. 28.—PEAK NOISE LIMITER FOR AN A.M. RECEIVER.

modulation depth. Series or shunt diodes are usually employed, or a combination of both.

One form of peak noise limiter using both series and shunt diodes, or crystal rectifiers, is illustrated in Fig. 28. While a carrier is being received, diode A is conducting while diode B is cut off. A peak of noise will increase the negative voltage of the anode of diode A, while provided that the time constant of the RC components is chosen correctly, the cathode voltage remains unchanged. Diode A is therefore cut off momentarily, preventing passage of the noise peak to the audio amplifier. If diode A is not cut off completely and still conducts, it will permit the cathode of diode B to be driven negative relative to its anode. Diode B will then conduct and short-circuit the noise to earth via the capacitor.

The point at which limiting occurs is usually adjusted to correspond to 50 per cent modulation; but, when clipping is used on the associated transmitter modulator, the limiting level should be increased to correspond to a higher level of modulation, depending on the degree of clipping used.

F.M. Limiters and Demodulators

The limiting stage immediately preceding the demodulator is used to remove any variations in amplitude of the carrier, including noise peaks, which may be superimposed on the signal.

The usual limiter is a saturated pentode with low anode- and screen-voltage, and operating with high grid-current bias to give class-C operation. As the amplitude of the input wave varies, the grid-bias of the valve must vary in sympathy to alter the stage gain, thus keeping the output voltage constant. The time constant of the self-bias circuit is therefore critical, and must be chosen to suit the particular type of amplitude modulation which is encountered. If the unwanted amplitude-modulation is caused by car-ignition noise, and a single limiter is used, a time constant of about 10 microseconds is common. For optimum results, however, two or even more limiters should be used with different time-constants, for example 2 microseconds and 20 microseconds.

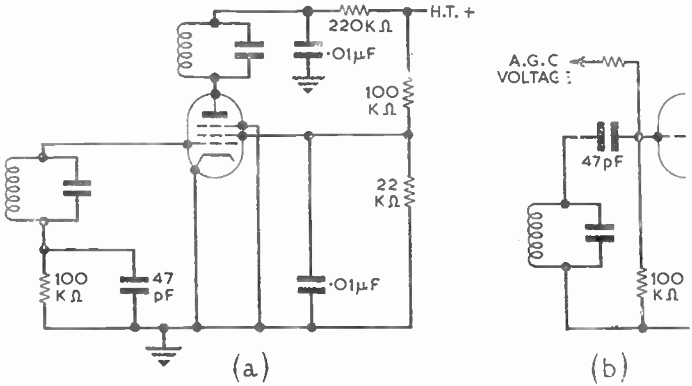


FIG. 29.—F.M. LIMITER CIRCUIT.

Automatic-gain-control voltage for F.M. receivers is usually derived from the grid of the first limiter stage, since this point becomes more negative with increasing input.

To ensure that the first limiter stage is saturated at the weakest input level, the aerial to demodulator gain of F.M. receivers is usually higher than that of a comparable A.M. receiver.

The circuit of a typical limiter is shown in Fig. 29 (*a*). Fig. 29 (*b*) shows an alternative biasing arrangement. The time constant of the self-bias circuit shown, using a 47-pF capacitor and 100,000-ohm resistor, will be 4·7 micro-seconds.

Discriminators

The commonest form of discriminator used in F.M. receivers is illustrated in Fig. 30. Many variations in detail of this basic circuit may be encountered. There is a phase difference of 90° across the transformer primary and secondary at resonance, and equal and opposite signals are applied to each of the diodes. Since the outputs

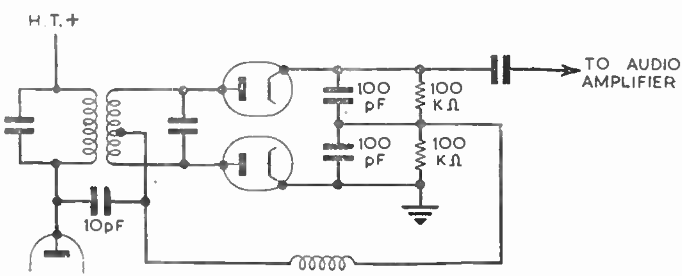


FIG. 30.—DISCRIMINATOR FOR F.M. RECEIVERS.

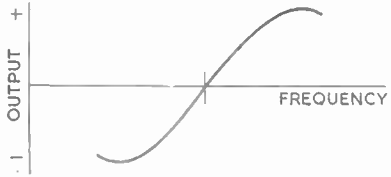


FIG. 31.—DISCRIMINATOR RESPONSE.

of the diodes are in series, the net output at resonance is zero. Off resonance, the phase difference across the transformer varies, causing different voltage levels to be applied to the two diodes. There is, therefore, a D.C. voltage produced across the diode loads. Fig. 31 shows the output versus frequency response of the circuit. If the input to the circuit is varied in frequency at an audio-frequency rate, this audio frequency will appear in the output circuit.

From the response curve, it is seen that for optimum operation, with minimum distortion, the unmodulated carrier must fall at the crossover point corresponding to no D.C. output, and the frequency separation between peaks must be at least twice the deviation. The D.C. current in the output resistor can be monitored to assist in setting up the circuit. Distortion rises rapidly if the input frequency extends beyond the peaks; it is therefore important that the circuit should not be detuned by changes in temperature, étc.

Audio Amplifiers

Simple, conventional audio amplifiers are used for V.H.F. receivers. A low-pass filter is usually provided, cutting off at about 3 kc/s; this is to eliminate noise above the speech band and thus improve the overall signal-to-noise ratio. Power outputs of 1 or 2 watts are usual but up to 10 watts may be required for installations where the ambient noise is high. On simplex equipment, where the transmitter is not energized while receiving, the modulator can be used for this purpose.

Muting

Because of the high gain of V.H.F. receivers, the background noise in the loudspeaker in the absence of signal is considerable, and some method of muting this noise is usually provided. Methods using a relay are preferred, since a contact on the relay can then be used for signalling purposes. Differential circuits can be used, the simultaneous reduction of noise and increase of signal altering the operating conditions of a two-valve, flip-flop type of circuit. One of these valves can be one of the audio-frequency voltage amplifiers, and the other the relay valve. The noise component of the audio signal, for use in this arrangement, is usually coupled via a high-pass filter to prevent muting due to modulation. The noise is then amplified and rectified to provide a voltage tending to prevent the flip-flop circuit from operating. Variation of the gain of the noise amplifier can be used to control the operating level. The audio output is shorted to earth via a contact on the muting relay until the required signal is received.

Cascode Signal Frequency Amplifier

The input impedance of a grounded-grid stage is low, and the grounded-grid triode may be preceded by a grounded-cathode triode to reduce damping across the input circuit. Such an arrangement is known as the cascode amplifier. The D.C.-coupled version is usually preferred, and is shown in Fig. 33. The gain of the first valve is approximately unity, and neutralizing is not therefore required for stability. The first valve should, however, be neutralized to achieve optimum noise factor, and to reduce the effects of de-tuning, due to change of A.G.C. voltage.

AERIALS

A quarter-wave rod, mounted on a flexible insulator and coupled to the equipment via 50-ohm co-axial feeder cable, is the commonest fitting for mobile installations. The roof of the car is used as the ground plane.

Quarter-wave aerials with ground planes can also be used for fixed stations. An alternative is the omni-directional, centre-fed, half-wave dipole. With this aerial the 70-ohm feeder is run inside the co-axial lower section to correct for the balanced-to-unbalanced connection and to eliminate obstruction round the radiating elements. The arrangement is illustrated in Fig. 32. Stacked dipoles can be used to obtain aerial gain while preserving the omni-directional radiation pattern.

Yagi aerials, with quarter-wave line sections to match to 50- or 70-ohm feeder cable, are used when a high-gain directional characteristic is required.

Vertically polarized aerials must be used for mobile systems, but horizontally polarized Yagi aerials are usually provided for fixed point-

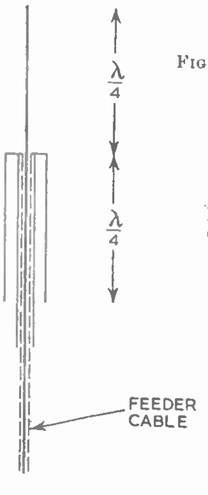


FIG. 32 (*left*).—CO-AXIAL HALF-WAVE DIPOLE AERIAL.

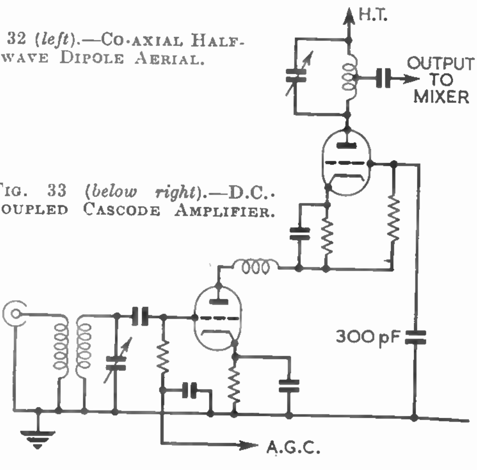


FIG. 33 (*below right*).—D.C.-COUPLED CASCODE AMPLIFIER.

to-point applications, as electrical interference at V.H.F. is generally vertically polarized.

Co-axial cavity filters are sometimes used to permit a number of transmitters or receivers to be operated simultaneously with a common aerial. When a transmitter and receiver are being operated simultaneously with close physical spacing between aerials, blocking may be experienced in the receiver, due to grid current being produced in the first radio-frequency valve. Cavity filters can be connected in series with the receiver feeder to reduce this effect.

For maximum range the aerial must be installed at as high a point as possible.

POWER SUPPLIES

For fixed station equipment, normal A.C. rectifier circuits are used. For mobile equipment operating with D.C. supplies, the choice is between vibrators and rotary transformers. Vibrators are more efficient than machines, but are rather more expensive, and effective suppression of hash is sometimes tricky. The power-handling capability of vibrators is not very big, and they are chiefly used for low-power equipment when low-battery drain is important.

Further information on this subject will be found in Section 14 (Car Radio Receivers) Section 36 (Batteries and Vibrators), and Section 41 (Mobile Installations).

MECHANICAL FEATURES

Fixed station equipment is often mounted on 19-in. panels for use in standard racks or cabinets. Flexible arrangements to suit individual customer's requirements are then more easily catered for.

For mobile equipment it is convenient to provide a simple shock-absorbed rack, complete with battery and control connections, permanently fixed to the vehicle, the transmitter/receiver unit, or units, connecting to the rack by plugs and sockets and secured by some quick-release mechanism.

For the cheaper, low-power equipment a common chassis may be used for transmitter, receiver and power supplies. From the servicing aspect, however, it is more convenient to supply separate units for each, preferably with these units further broken down into sub-assemblies which can be replaced rapidly. The defective sub-assembly can then be serviced at the depot.

OPERATION OF COMMUNICATION SYSTEMS

Both simplex and duplex systems are used. In the former, transmitter and receiver work alternately, controlled by a send/receive switch, and for simplex equipment some stages of the transmitter and receiver, such as the crystal oscillator, audio amplifier and power supplies, may be common.

For duplex equipment, where transmitter and receiver are operated simultaneously, two-frequency working is, of course, essential. To reduce power drain in duplex equipment, the transmitter is normally kept in the stand-by condition, with valve heaters energized, and is switched on only when required for use.

Two-frequency operation is also usual for simplex equipment in mobile schemes, although this theoretically halves the number of available channels. In practice, operation ~~on~~ the remaining channels is improved, since H.Q. transmitters are grouped together in one band and receivers in another. Adjacent mobile equipments then cause little interference with each other, since there is a frequency spacing of about 10 Mc/s between wanted and unwanted signals.

In large simplex systems the disorganization caused by several out-stations talking to each other at the same time is avoided if two-frequency working is employed. Out-stations are then unable to communicate with each other directly, and when this is necessary, as is often the case on tug installations, for example, the H.Q. equipment can be switched as a repeater, the H.Q. operator monitoring both sides of the conversation on the loudspeaker and re-connecting the equipment for normal operation at the end of the call.

Diversity Systems

The range of V.H.F. equipment being limited, a large area cannot be covered by a single equipment.

This difficulty can be overcome in the case of A.M. systems by providing a number of equipments, each operating on a slightly different frequency about the nominal frequency of the scheme.

The arrangement is illustrated in Fig. 34. The service areas of the individual equipments have a small overlap. A common frequency is not used for all equipments, owing to the difficulty of obtaining correct phasing of signals received simultaneously from two or more transmitters in the overlap area. It is relatively simple to ensure that the inputs to the three transmitter modulators are in phase. The receiver pass-band is made wide enough to accommodate all three carriers, and a low-pass filter eliminates the beat notes produced by reception of more than one carrier. For successful operation the frequencies of the main station equipments must be rigidly controlled, and crystal ovens are essential.

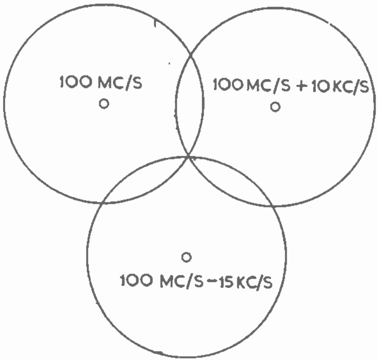


FIG. 34.—DIVERSITY SYSTEM FOR LARGE AREA COVERAGE.

The above system cannot be used with F.M. equipment owing to the nature of the F.M. demodulator, and, to date, similar systems using synchronized F.M. carriers have not given such good results in the overlap area.

CONTROL ARRANGEMENTS

Fixed stations are often installed at a site remote from the control point, where electrical interference is low and good aerial elevation can be obtained. Control signals are then sent over G.P.O. lines or a further short-distance V.H.F. radio link. Where the equipment is in the same building as the control point, it is desirable that it should be mounted as close as possible to the aerial in order to reduce feeder losses; extended control arrangements, up to about 100 yd., are then required, using multicore cable.

Controls are very simple. In a typical duplex arrangement lifting the handset on a G.P.O. type desk unit brings on the carrier, and receipt of a call is indicated by a bell, operated by the muting relay. For simplex schemes, a press-to-talk switch in the handset is used for send/receive operation.

On mobile installations the transmitter–receiver is often mounted in the boot of a car, and the controls are then extended over about 15 ft. of multi-core cable to a control unit on the dashboard. On/off, volume and send-receive controls are required. A standby switch is sometimes incorporated to reduce battery drain by switching off transmitter valve heaters when on receiver watch. Adjustment of muting operating level is sometimes brought out to the control unit, since the setting of the relay circuit can be affected by the state of charge of the vehicle battery.

For police and similar services public-address facility is often provided. On A.M. equipment the transmitter modulator is switched for this purpose. In the case of F.M. equipments, where there is no high-level modulator, the final radio-frequency stage can be used, but the switching requirements are more complicated.

MULTI-CHANNEL SYSTEMS

An important application of V.H.F. radio is its use to replace telephone cables in undeveloped regions or over difficult terrain, such as rivers, valleys or forests. Several speech channels, up to forty-eight in number, can be accommodated on one radio circuit. The separate speech signals are first combined by means of channelling equipment to give a traffic channel, which requires a band-width of 12–204 kc/s to accommodate forty-eight channels to C.C.I.F. standards.

The simplest multi-channel system consists of two terminal equipments, linking two telephone exchanges. When a long transmission path is involved, repeaters must be used. Frequency modulation is invariably used for multi-channel equipment in the V.H.F. band, and at repeaters the F.M. carrier can be frequency changed to an intermediate frequency, amplified and frequency changed again to the new transmitter frequency, without de-modulation. This system avoids cross-modulation between speech channels, introduced by modulator and de-modulator circuits at repeaters. For some systems however, when the number of repeaters is small, the repeaters may take the form of two terminal equipments with through connections.

The signal input required at each receiver in a system depends on the number of telephone channels, the signal-to-noise ratio required for each telephone channel, and the number of repeaters used in the system. The maximum distance between repeaters is very little more than optical range, so that careful siting of stations is required.

Transmitter powers are usually about 20 watts, but powers up to 100 watts are occasionally used. The aerial arrays are generally hori-

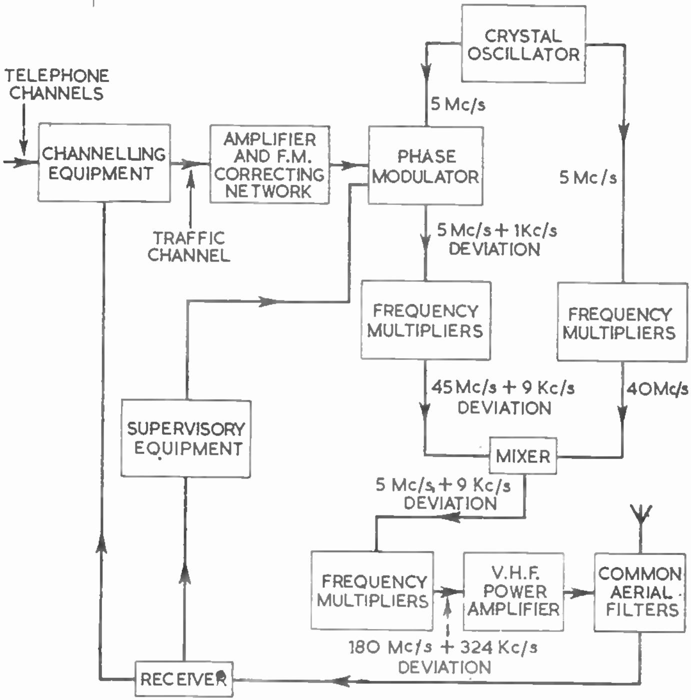


FIG. 35.—MULTI-CHANNEL TERMINAL TRANSMITTER AND RECEIVER.

zontally polarized Yagis. Four frequencies are required at each repeater, and common aerial working is usually employed to reduce the number of aerials and masts, the transmitter and receiver operating in a given direction sharing the same aerial.

Since repeaters often operate unattended for long periods between inspection visits, reliability of the equipment is of prime importance. Facilities are provided to enable the repeaters to be checked from the terminal stations by loop tests, and facilities are sometimes provided for automatic change-over to stand-by equipment in the event of a fault developing at a repeater.

Terminal Transmitters

Modulation may be direct F.M., or by means of a phase modulator with a correcting network to give effective frequency modulation. If a phase modulator is used, the initial phase deviation must be kept very low, since distortion in the modulating process will produce cross-modulation between speech channels. The deviation corresponding to the necessarily small phase swing is consequently also very low, of the order of 1,000 c/s. As a deviation of about 300 kc/s is required at the radiated frequency, there must be considerable multiplication between the output of the modulator and the final power-amplifying stages. To avoid the use of an unduly low crystal frequency, different harmonics of the crystal, one modulated and one unmodulated, are mixed to produce a signal at crystal frequency with increased deviation. Referring to Fig. 35, which shows a typical terminal transmitter and receiver in block form, the 5-Mc/s output from the crystal oscillator is modulated with a deviation of 1 kc/s. After frequency multiplication (nine times) the 45-Mc/s signal (deviation 9 kc/s) is mixed with the unmodulated eighth harmonic of the crystal to give a signal at 5 Mc/s with 9 kc/s deviation.

Supervisory facilities are included to permit maintenance engineers at the terminals and repeaters to intercommunicate on a common-party-line basis. The frequency channel 300–3,000 c/s, which is below the frequency of the main traffic channel, is used for this purpose. Common ringing facilities are provided, operated by means of a tone in the supervisory frequency band. Receipt of this tone at the other terminal and repeaters operates a relay to connect a call bell. The circuit may usually be extended, when required, on a two-wire junction with D.C. signalling, to a local exchange.

Terminal Receivers

The best possible noise factor is essential, and a cascode signal-frequency amplifier is usually preferred. The design of the wide-band intermediate-frequency amplifier requires care, since any phase shift before the de-modulator will have adverse effects on the channel cross-modulation performance.

Repeaters

A block diagram of a non-de-modulating repeater is shown in Fig. 36. The arrangement shown is duplicated for traffic in the opposite direction.

After signal-frequency amplification the carrier is reduced to intermediate frequency by means of a mixer and local oscillator. After further amplification at intermediate frequency, another mixer and local oscillator are then used to give the new radiated frequency. To

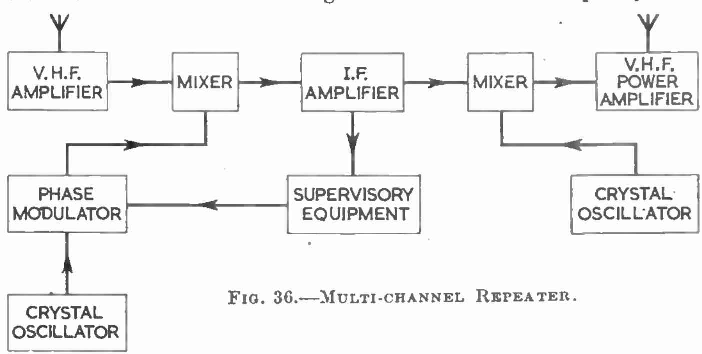


Fig. 36.—Multi-channel Repeater.

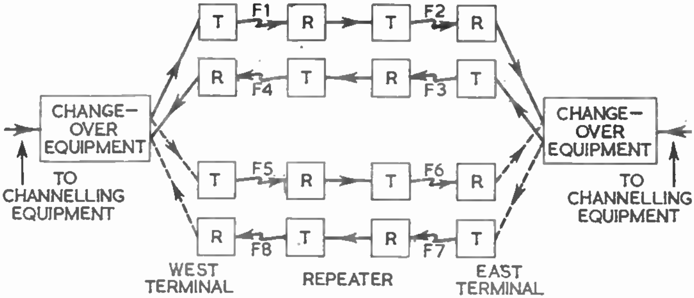


FIG. 37.—MULTI-CHANNEL SYSTEM WITH AUTOMATIC CHANGE-OVER FACILITIES.

minimize the effects of the frequency tolerance being degraded at each repeater, the two local oscillators are sometimes controlled from a common crystal oscillator.

To obtain the supervisory facilities, part of the intermediate-frequency output is taken via a local discriminator to extract the supervisory channel for the maintenance engineers' intercommunication facility, and for calling and test tones. Supervisory information is put into the system by phase modulating the local oscillator input to the first mixer.

Test Facilities

Provision is usually made for the unattended repeater equipments to be tested from the main terminals at the ends of the link. The tests are made by use of tones, different for each repeater, in the supervisory band. Receipt of the correct tone at a repeater operates relays, which loop part of the boosted transmitter output back to the receiver of the opposite direction repeater, via a frequency-changing mixer and local oscillator. The overall performance of the terminal transmitter/receiver, the repeater transmitter/receiver and any intervening repeaters may then be measured.

Circuits are often provided at unattended repeaters to send warnings in the event of a given degradation of signal-noise ratio, or if the transmitter power output falls below a pre-determined level. The warnings are transmitted by modulating tones in the supervisory channel.

Automatic Change-over to Stand-by Equipment

There are a number of ways in which the automatic change-over facility can be used in a multi-channel link, and one is illustrated in Fig. 37.

In this arrangement the entire link, shown operating on frequencies F1–F4, is duplicated by similar equipment operating on frequencies F5–F8. The traffic over the system is normally connected to one of the links, F1–F4 in the diagram. The other link is fully operational, and loop tests, as mentioned above, may be carried out.

To follow the operation of the system, let us assume that a fault occurs in say the path F1–F2. The signal-to-noise ratio at receiver F2 will fall, causing a change-over relay at the east terminal to operate. The operation of the change-over relay rings an alarm, switches off transmitter F3 and changes the traffic over to the stand-by link F5–F8, as indicated by the dotted lines. Transmitter F3 going off causes receiver F3 to switch off transmitter F4, and transmitter F4 going off operates the change-over circuit at receiver F4. The change-over equipment at the west terminal then transfers traffic to the stand-by link F5–F8.

Operation of the alarm circuits at the terminals draws the attention of the maintenance engineers, who restore H.T. to the terminal transmitters, bringing on the system and permitting loop tests to be made to find the faulty repeater. If the stand-by link should develop a fault first, only the alarm circuits operate.

Further information on V.H.F. communications systems and equipment is given in Sections 7, 17 and 41.

P. R. K.

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P. R. Keller, B.Sc., A.M.I.E.E., "V.H.F. Radio Manual", *George Newnes, Ltd.*

9. AMATEUR RADIO EQUIPMENT

9. AMATEUR RADIO EQUIPMENT

The Amateur Radio Service has been defined, internationally, as: "A service of self-training, intercommunication and technical investigations carried on by amateurs, that is, by duly authorized persons interested in radio technique solely with a personal aim and without pecuniary interest." There are in the United Kingdom more than 8,000 licensed amateurs, and throughout the world more than 250,000 such stations. While many of these stations are operated purely for the pleasure to be derived from two-way communication with like-minded persons in other lands, it must be recognized that the amateur plays a valuable role in the practical application of new techniques in the general field of radio communication, while the competitive nature of such operating, the extremely crowded amateur frequency bands, the frequent absence of elaborate test equipment and the difficulties of operating radio transmitters from private houses surrounded by domestic radio and television receivers have all tended to bring about the development of equipment specially designed to meet the needs of the amateur.

Licences

To operate an amateur station in the United Kingdom it is necessary to obtain a licence from the Postmaster General (Radio Services Department, Radio Branch, General Post Office, London, E.C.1). Applicants must be over fourteen years of age, furnish proof of nationality, and be prepared to take technical and Morse code examinations. Apart from a knowledge of fundamental radio transmitting and receiving theory, the questions require a knowledge of the amateur licence regulations and the ability to send and receive plain language in Morse code at a speed of twelve words per minute. There is a fee of £2 per annum (additional fees are payable for mobile or amateur television facilities). A pamphlet "How to Become a Radio Amateur" can be obtained from the above address.

Amateur Frequencies

The frequencies on which amateurs are permitted to operate in the United Kingdom are as given in Table 1.

The use made of these bands may be summarized as follows:

1·8 Mc/s. Used for semi-local (up to about 100 miles) telegraphy and telephony during daylight, and for inter-British Isles working during darkness. In favourable conditions during the winter nights long-distance and Transatlantic contacts are made occasionally, despite the limitation of power.

3·5 Mc/s. Widely used for inter-British Isles and European communication, and for long-distance communication during favourable propagation conditions. Telegraphy stations generally use from about 15 to 100 watts input. Telephony stations tend to use from 50 to 150 watts.

TABLE 1.—AMATEUR FREQUENCIES

<table>
<tr><th>Frequency (Mc/s)</th><th>Max. D.C. Input (Watts)</th><th colspan="2">Types of Emission Permitted</th></tr>
<tr><td>1·8–2·0 *</td><td>10</td><td>A1, A2, A3, A3a</td><td rowspan="5">Narrow band F.M. (F1, F2, F3) with maximum deviation ·2·5 kc/s and guard band of 10 kc/s. Maximum modulating frequency 4 kc/s.</td></tr>
<tr><td>3·5–3·8 *</td><td>150</td><td>A1, A2, A3, A3a</td></tr>
<tr><td>7·0–7·15 †</td><td>150</td><td>A1, A2, A3, A3a</td></tr>
<tr><td>14·0–14·35</td><td>150</td><td>A1, A2, A3, A3a</td></tr>
<tr><td>21·0–21·45</td><td>150</td><td>A1, A2, A3, A3a</td></tr>
<tr><td>28·0–30·0</td><td>150</td><td colspan="2">A1, A2, A3, A3a, F1, F2, F3</td></tr>
<tr><td>70·2–70·4 * ||</td><td>50</td><td colspan="2">A1, A2, A3, A3a</td></tr>
<tr><td>144–146 §</td><td>150</td><td colspan="2">A1, A2, A3, A3a (144·5–145·5 Mc/s: F1, F2, F3)</td></tr>
<tr><td>420–460 *</td><td>150</td><td colspan="2">A1, A2, A3, A3a, F1, F2, F3 (425–455 Mc/s: A5.‡ F5 ‡)</td></tr>
<tr><td>1,215–1,325 *</td><td>150</td><td colspan="2">A1, A2, A3, A3a, F1, F2, F3 (1,225–1,290 Mc/s: A5, F5)</td></tr>
<tr><td>2,300–2,450</td><td>150</td><td colspan="2">A1, A2, A3, A3a, A5,‡ F1, F2, F3, F5 ‡ (2,350–2,400 Mc/s: 25 watts mean, 2·5 kW peak, P1, P2d, P2e, P3d, P3e)</td></tr>
<tr><td>5,650–5,850</td><td>150</td><td colspan="2">A1, A2, A3, A3a, A5,‡ F1, F2, F3, F5 ‡ (5,700–5,800 Mc/s: 25 watts mean, 2·5 kW peak, P1, P2d, P2e, P3d, P3e)</td></tr>
<tr><td>10,000–10,500</td><td>150</td><td colspan="2">A1, A2, A3, A3a, A5,‡ F1, F2, F3, F5 ‡ (10,050–10,450 Mc/s: 25 watts mean, 2·5 kW peak, P1, P2d, P2e, P3d, P3e)</td></tr>
</table>

* Shared with other Services.
† 7·10–7·15 Mc/s shared with other Services.
‡ An additional licence is required for Amateur Television transmission.
§ 144–145·5 Mc/s shared with other Services.
|| Not to be used within a 50-mile radius of Jodrell Bank Observatory.

7 Mc/s. Inter-British Isles communication during periods of peak sunspot activity, inter-European and long-distance communication at other times. Considerable interference is caused to amateurs by broadcast stations operating within this band.

14 Mc/s. The most reliable long-distance band; considerable inter-European working also takes place.

21 Mc/s and 28 Mc/s. These bands are used for long-distance daylight communication during periods when the maximum usable frequency is high, but activity tends to fall sharply at other times.

144 Mc/s. Used for semi-local communication, and for longer distances (100–500 miles) during periods of tropospheric bending.

420 Mc/s. Used for U.H.F. and amateur television experiments, with occasionally contacts over several hundred miles when conditions favourable. Used by about 100 British stations with power inputs usually below about 40 watts.

Above 420 Mc/s. These bands are used for experimental work, and little or no "general" communication occurs.

Division of Equipment

It will thus be seen that the pattern of equipment in any particular station is set largely by the type of communication—local, semi-local or long-distance—and the mode of operation—telegraphy or telephony —in which the owner is most interested. Certain general tendencies in design may, however, be distinguished.

For example, on 1·8 and 3·5 Mc/s one small general-purpose transmitter for telegraphy (A1) and amplitude-modulated telephony (A3) with a maximum input of the order of 25–40 watts is often used in conjunction with an end-fed "long-wire" (exceeding 100 ft.) aerial; a half-wave aerial on 3·5 Mc/s (about 135 ft.) is a popular arrangement.

The coverage of all bands from 3·5 to 28 Mc/s is usually regarded as being within the scope of one transmitter, with variable frequency control, and preferably with some form of band-switching. Maximum input will usually be of the order of 60–150 watts. Popular H.F. aerials include dipoles fed by co-axial feeder or by resonant lines, folded dipoles fed by 300-ohm ribbon feeder, "Zepp" multi-band aerials and 132-ft. "long-wire" aerials. For omni-directional, low-angle radiation the "ground plane" is a most effective aerial. The above types are similar to those described in Section 21 for receiving. Recent trends have been towards greater use of rotating directional Yagi arrays on 14 Mc/s and above, often with elements shortened by electrical loading. Another popular parasitically excited array is the "cubical quad", in which each element is an electrical wavelength long, folded in the form of a square. Multi-band aerials with paralleled dipoles or with "traps" are also widely used.

A separate transmitter, in so far as the radio-frequency section is concerned, is normally employed for 144 Mc/s, usually crystal-controlled from an 8-Mc/s crystal, then employing a series of frequency multipliers to 144 Mc/s, followed by a double-tetrode valve in a power-amplifier stage. Inputs vary according to the type of output valve employed between about 25 and 150 watts. Multi-element stacked or Yagi aerial arrays are most popular.

For 420 Mc/s, it is common practice to use the 144-Mc/s transmitter followed by a power-tripler stage or alternatively by a frequency-tripler stage and a power amplifier. A few single-stage self-excited transmitters are also in operation. Stacked aerial arrays of up to about forty-eight elements are used.

Above 420 Mc/s most of the equipment so far employed has been modified forms of equipment designed originally for the Services.

There is increasing interest in mobile operation, usually with telephony on 1·8, 28 or 144 Mc/s. In the U.S.A. increasing use is being made of single-sideband transmissions for this application.

HIGH-FREQUENCY TRANSMITTER DESIGN

The transmitter may be considered in three main sections: oscillators; exciters; power amplifiers. An important factor in the design of amateur transmitters, as opposed to those for commercial purposes, is that the amateur frequency bands are approximately in harmonic relationship, and high efficiency at these particular portions of the spectrum is required rather than wide continuous coverage.

Before the 1939–45 war, the vast majority of British amateurs used crystal-controlled transmitters—a form of frequency control that the amateurs had indeed done much to popularize. But in the post-war years the crowded state of the various high-frequency bands and the general practice of "netting"—that is to say the use of the same frequency by the two or more stations in communication—have led to almost universal reversion to the master oscillator, generally termed by amateurs "variable-frequency oscillator". But in doing so, the extreme selectivity of receivers, and the general desire, partly from a prestige viewpoint, to possess a "clean" note, free from frequency drift and keying chirp, has led the amateur to pay considerable attention to the problem of obtaining, with relatively simple and economical circuits, a note comparable to that previously obtained from crystal-controlled equipment, now mainly restricted to use on V.H.F. or by beginners or for simple portable apparatus.

The amateur variable-frequency oscillator is often built separately from the main transmitter, unless this be of the compact table-top type, with the oscillator placed directly upon the operating table, so that the frequency may be readily changed within a particular band without the operator having to move from his chair. In practice, this unit generally comprises a low-power receiving-type valve oscillator with the fundamental tuned circuit on 3·5 Mc/s or below, fed from a stabilized power source, and followed by at least one isolating stage operating in Class A, the unit providing a radio-frequency output of the order of 1 watt.

Fig. 1 shows four popular oscillator circuits. Fig. 1 (*a*) is the electron-coupled oscillator (E.C.O.) multiplier developed from the Dow oscillator

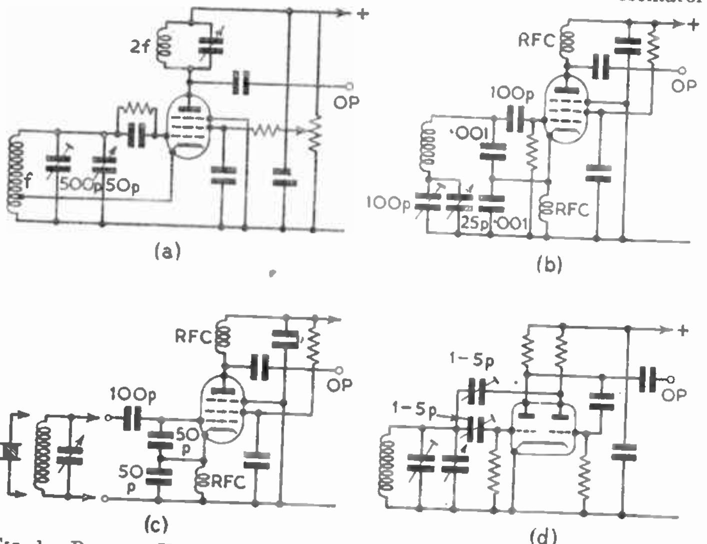

FIG. 1.—POPULAR VARIABLE FREQUENCY OSCILLATOR CIRCUITS: (*a*) E.C.O.; (*b*) "CLAPP"; (*c*) "COMO"; (*d*) "FRANKLIN".
(In (*c*) a resistor should be included between grid and chassis.)

circuit. Provided that care is taken in the choice of the L–C ratio (high-C is essential) and the cathode tap is closed to the "earthy" end of the coil, a reasonably stable output can be obtained with the frequency varying little for relatively large changes of supply voltage and load conditions. As with all self-excited oscillator circuits, mechanical stability of the frequency-control circuit is of great importance; as is the avoidance of undue heating of this circuit, unless provision is made for the use of two capacitors of opposing temperature coefficients. The valve should possess good internal shielding; audio-beam tetrodes in which the beam-forming plates are connected to the cathode or pentodes with the suppressor grid connected internally to the cathode are unsuitable.

The popular "Clapp"—a series-tuned modified Colpitts oscillator circuit is shown in Fig. 1 (*b*). This circuit has the advantage that changes of valve-input capacitance and anode loading have comparatively little effect on the frequency; the circuit also permits the use of a series-resonant tuned circuit with a high L–C ratio and high Q. A further advantage of this arrangement is that the frequency-determining tuned circuit may be physically separated from the oscillator valve, to which it can be coupled by co-axial cable; this permits the tuned circuit alone to be placed on the operating table, completely unaffected by the heat of the valve, which may be in the main transmitter chassis.

The "COMO" (Crystal Oscillator or Master Oscillator) circuit shown in Fig. 1 (*c*) has the advantage that the variable-frequency tuned circuit may be readily replaced by a quartz crystal. When operating as a master oscillator, the circuit forms a modified Colpitts oscillator.

The "Franklin" oscillator (Fig. 1 (*d*)) has the feature that the coupling to the frequency-determining circuit may be made very small, so that valve-impedance changes, etc., have little effect on the frequency.

With care, any of these arrangements will provide a stable source of radio frequency. Receiving-type radio-frequency pentodes, such as the 6AM6, SP6, Z77, 6F12, EF91, 6SK7, 6SJ7, EF50, are suitable.

Exciters

In recent years the function of the intermediate stages between the oscillator unit (including any isolating stages that may be regarded as a part of this) and the amplifier which delivers power to the aerial has changed. Formerly the main function of these stages was—with each successive valve—to build up the radio-frequency power to whatever was necessary to drive a triode power amplifier; frequency multiplication, where necessary, was only a secondary function of these stages. Today the relative importance of these functions is reversed. The widespread use of power tetrodes in the power-amplifier stage means that only a few watts (2–10) of radio-frequency power are required to drive such a stage to the full 150 watts permitted. On the other hand, the growing desire that the transmitter should be capable of being switched immediately from one band to another, without manual coil changing, and without extensive re-adjustment, means that greater attention has to be paid to the use of frequency doublers that can be switched in and out of operation, and—where the exciter unit is physically separated from the power-amplifier stage—that deliver a low impedance output for matching to co-axial line.

Incidentally, when planning an exciter it should be remembered that whilst over-driving of beam tetrodes in the power amplifier must be avoided in order to minimize harmonic output, nevertheless, roughly twice the grid driving power specified by the valve manufacturer should preferably be available from the exciter in order to allow for circuit losses and to provide good regulation.

To secure " wide-band " characteristics for each stage—i.e., so that the stage will function without loss of efficiency from the high to the low limits of the band—it is now common practice for " wide-band couplers " to be incorporated as intervalve coupling units. These couplers resemble basically intermediate-frequency transformers, with the desired band pass characteristics obtained by over-coupling, or by choice of L–C ratio. Occasionally link-coupling between the two resonant circuits is used, and this facilitates the provision of a low-impedance output where required.

Since the power required from an exciter is usually of the order of 3 watts, receiving-type audio-output valves, such as the 6AQ5, 6V6, KT66, 6L6, etc., are frequently employed; the slight loss in efficiency as compared with that which can be obtained with valves specifically designed for radio-frequency amplification being usually considered less important than economy. Where greater efficiency is required, valve types such as the 5763, QV04/12, etc., may be used.

Power Amplifiers

The basic design of power-amplifier stages for amateur requirements differs little from that followed in commercial practice. The most popular arrangements are: (1) single-ended beam tetrode; (2) two beam tetrodes in push–pull; (3) two beam tetrodes in parallel. Triodes and radio-frequency pentodes are also encountered. For efficiency, the valves are invariably operated in either Class B or Class C condition.

For low- and medium-power transmitters, a single 807. the similar QV05/20, or the 6146 (QV06/20) is extremely popular, whilst for higher powers two of these valves may be used, or a single 813.

Instability in Amplifiers

The use of high-slope valves, often without neutralization, has led to considerable attention having to be paid to the prevention of self-oscillation and parasitic oscillation in amplifiers. Such instability may cause a roughening of the note, or a marked decrease in the efficiency of the stage, apart from the possibility of radiation occurring c : frequencies outside the amateur bands.

Self-oscillation at frequencies close to that of the transmitter is caused by positive feedback from the output to the input circuits of the stage, either through valve anode-to-grid electrode capacitance or externally through stray capacitances or coupling, so that the stage forms a tuned-anode–tuned-grid oscillator. In operation this effect may be masked by the effect of the bias voltage: a critical test of the stability of a Class B amplifier stage can be made by removing the drive and the output loading and reducing bias until the stage is drawing a fair current under Class A conditions; the anode and grid tuning capacitors are then alternatively varied and a watch kept for any sudden change in anode current that would denote self-oscillation. The anode-to-grid capacitance of an 807-type of valve is of the order of 0·2 pF, and if the

TABLE 2.—TETRODE R.F. POWER AMPLIFYING VALVES COMMONLY USED IN AMATEUR TRANSMITTERS

Type	*Mode*	*Max. D.C. Input (Watts)*	*Max. Anode Diss. (Watts)*	*Max. Anode Voltage (Volts)*	*Max. Anode Current (mA)*	*Max. Screen Voltage (Volts)*	*Max. Screen Diss. (Watts)*	*Max. Freq. at Full Ratings (Mc/s)*	*Approx. Max. Output (Watts)*	*Grid-anode Capacitance (pF)*	*Heater Rating*	
											Volts	*Amps*
6AG7	A1	11·25	9	375	30	250	1·5	10	7·5	0·06	6·3	0·65
6AQ5, EL90	A1	16·5	8	350	47	250	2	54	11	0·35	6·3	0·45
6V6GT	A1	16·5	8·0	350	47	250	2·0	10	11	0·7 6L6G = 0·9	6·3	0·45
6L6, KT66	A1	40	21·0	400	100	300	3·5	10	28	0·4	6·3	0·9
	A3	21	—	325	65	250	—	—	11	—	—	—
807, QVO5–25	A1	75	30·0	750	100	300	3·5	60	50	0·2	6·3	0·9
	A3	60	—	600	100	275	—	—	42·5	—	—	—
5763, QVO3–12	A1	15	12·0	300	50	250	2·0	175	8·0	0·3	6·0	0·75
6146, QVO6–20	A1	90	25	750	120	250	3	60	69	0·22	6·3	1·25
	A3	67	—	600	112	150	—	—	52	—	—	—
832	A1	35	15·0	500	72	250	5·0	200	26·0	0·05	6·3 12·6	1·6 0·8
832	A3	21	—	425	52	200	—	—	16·0	—	—	—
832A	A1	35	15·0	750	48	250	5·0	200	26·0	0·05	6·3 12·6	1·6 0·8
829	A1	120	40·0	500	240	225	7·0	200	83·0	0·1	6·3 12·6	2·25 1·12
	A3	90	—	425	212	200	—	—	63·0	—	—	—
829A	A1	120	40·0	750	160	240	7·0	200	87·0	0·1	6·3 12·6	2·25 1·12
813, QY2/100	A1	500	100·0	2,250	220	400	22·0	30	375·0	0·2	10·0	5·0
	A3	400	—	2,000	200	350	—	—	300·0	—	—	—
4–65A, QY3–65	A1	345 *	65	3,000	115	400	10	160	280	0·08	6·0	3·5
	A3	325 *	—	2,500	108	400	—	—	225	—	—	—

* Forced Air Cooling.

effect of this capacitance is appreciably increased by external coupling, t.a.t.g. oscillation will almost certainly occur. For this reason the physical layout of the stage, and the electrical separation of the input and output circuits are most important if the added complication of neutralizing is to be avoided. Bridge neutralization (see Section 15) is commonly used.

Parasitic oscillation may occur at much higher or much lower frequencies than that at which the stage is intended to operate. The basic cause is that the conditions for t.a.t g. oscillation may arise unintentionally in the input and output circuits. For example, V.H.F. parasitics—usually in the range 100–200 Mc/s—may be the result of the leads to the anode and grid tank circuits themselves forming series-resonant circuits within this range. Low-frequency parasitics, on the other hand, may be due to t.a.t.g. oscillatory circuits being set up by the resonance of two radio-frequency chokes.

To eliminate V.H.F. parasitics, a choke formed of fifteen turns, $\frac{1}{4}$-in.-diameter, heavy-gauge wire may be inserted in the grid lead close to the valve base, and a second choke of eight turns in the anode lead(s). Alternatively, carbon resistors of 10–22 ohms may be used. Parasitic chokes of this type may cause a slight loss in drive power above about 14 Mc/s. The use of parasitic chokes in the screen-grid leads is not recommended.

Other methods of improving stability include :

(1) The use of more effective by-passing to earth of screen-grid, filament, grid return and anode return.

(2) Making the grid leads much longer than the anode leads or vice versa (this arrangement is often most effective in stages operating on 3·5 Mc/s or less).

(3) The use of iron or nichrome wire for grid, anode or neutralizing leads.

(4) Neutralizing tetrode valves.

(5) Carefully balancing the drive to both valves in push–pull circuits (a good check is to compare screen-grid currents).

(6) Avoiding over-driving of the stage.

Tank Circuits

The amateur has a distinct advantage over the commercial designer in that he does not require complete frequency coverage, and his bands are roughly in harmonic relationship. Since, however, his power input is limited by licence regulations, the efficiency of the power-amplifier–anode-tank circuit is of considerable importance. For this reason, until comparatively recently, it was the practice to use separate plug-in coils, designed to have the correct L–C ratio for optimum output on each band. The desire for more rapid and easier band-changing, and the disappearance of open, unshielded, "bread-board" layouts, have brought into popularity a number of other arrangements. These include the rotary-turret—electrically similar to the plug-in system—and a number of multi-band devices. Fig. 2 shows a popular multi-band circuit, devised by G2HDU, and based on the pi-matching network; this circuit provides matching to aerials of any impedance over the range 3·5–30 Mc/s. Fig. 3 is a simplified arrangement for low-impedance output which dispenses with one tuning capacitor.

A different approach to the problem is a reasonably high-Q multiple

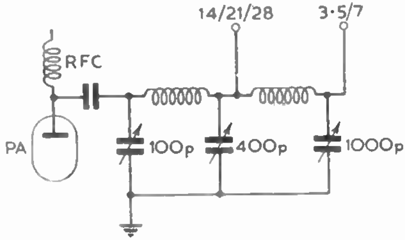

FIG. 2.—MULTI-BAND PI-COUPLING OUTPUT CIRCUIT.

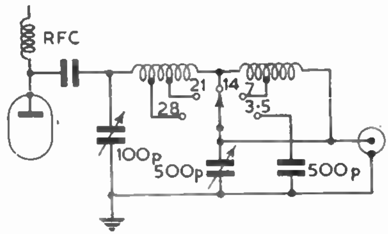

FIG. 3.—MULTI-BAND LOW-IMPEDANCE OUTPUT CIRCUIT.

system which can tune to any amateur band between 3·5 and 30 Mc/s without switching or coil changing. The basic idea, developed in the United States, is shown in Fig. 4. L1 is an inductance considerably smaller in value than L2. On the lower frequencies L1 acts, in effect, as a long lead to C1 so that this capacitor is connected in parallel with C2. On the higher frequencies L2 acts as a radio-frequency choke, and L1 determines the resonant frequency. Thus between the points A and B, the arrangement tunes simultaneously to two widely different frequencies: by careful arrangement of the relative values of the two inductances, no two amateur bands need appear at the same setting of the ganged tuning capacitors. In practice, the push–pull version of this circuit, shown in Fig. 5, is the more common.

Drive and Bias

In amateur radio-frequency power amplifiers, the bias required for Class C operation of power tetrodes is of the order of 60–100 volts.

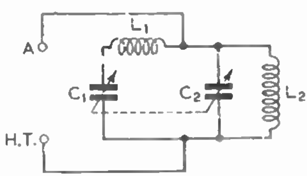

FIG. 4.—MULTIPLE TUNING ARRANGEMENT FOR SINGLE-ENDED OUTPUT.

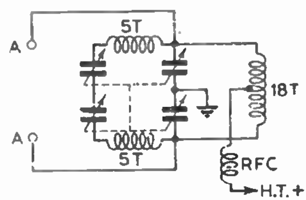

FIG. 5.—MULTIPLE TUNING ARRANGEMENT FOR PUSH-PULL OUTPUT STAGE.

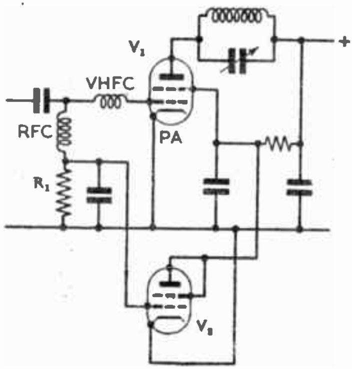

FIG. 6.—CLAMP-VALVE ARRANGEMENT FOR THE PROTECTION OF A POWER AMPLIFIER VALVE.

While an H.T. battery or a small, well-smoothed power pack may be used for this purpose, biasing voltages—on the grounds of simplicity and economy—are more often derived from biasing resistors inserted in the grid return lead. Since grid current flows only when drive is applied to the stage, some form of valve protection may be incorporated to prevent the anode and screen current from rising to a dangorous value in the absence of excitation, or—for telegraphy operation—where it is required to key a stage other than the power amplifier. The simplest form of protection is the use of a cathode bias resistor, in addition to the grid resistor, but this has the disadvantage for telegraphy operation that the standing current in the power-amplifier stage will be considerable, and some noise radiation will occur at roughly the frequency to which the tank circuit is tuned.

The "clamp valve" protection arrangement shown in Fig. 6 is becoming increasingly popular. In the absence of drive, V_2 draws current and this has the effect of reducing the voltage applied to the screen of the power-amplifier valve to a comparatively low figure, and thus limiting the current flowing through V_1. When radio-frequency drive is applied to V_1, the grid current flowing in R_1 will bias V_2 to cut-off (the radio-frequency choke prevents radio-frequency voltages from being applied to V_2) so that the normal screen-grid voltage is applied to V_1.

Television Interference

The major problem that has faced amateurs in the post-war period has been the operation of transmitters in close proximity to domestic television receivers. Although amateur transmitters are the cause of only a very small percentage of the total number of complaints of television interference as investigated by the G.P.O., the problem facing the individual amateur should not be underestimated: it is made more difficult from the amateur viewpoint by the fact that the third harmonic of the amateur 14-Mc/s band lies within the London television band (Channel 1); and the second harmonic of the 28-Mc/s band may affect either the Kirk o' Shotts or Sutton Coldfield television bands (Channels

3 or 4). Furthermore, the intermediate and image frequencies of many older television receivers are such that nearby amateur transmitters, through no fault of their equipment, are liable to break through. As with all forms of interference, the problem becomes more acute when the signal strength of the television station is low.

Television interference may be caused by:

(1) The radiation from the amateur transmitter of harmonic or spurious signals falling within the television channel concerned.

(2) Direct break-through of transmissions at or near the intermediate frequency of the receiver.

(3) A spurious response of the television receiver (image, beating with harmonic of local oscillator, etc.)

(4) Over-loading of the "front-end" of the television receiver causing cross-modulation.

(5) Cross-modulation caused by the external rectification of the amateur signals by a faulty joint in the television receiver aerial, or nearby metalwork.

(6) Parasitic oscillation in the audio section of the amateur transmitter.

(7) Harmonic radiation of the local oscillator of the amateur receiver.

(8) Key clicks or telephony splatter from the amateur transmitter.

(9) Direct pick-up of the amateur signals in the video amplifier or audio stages of the television receiver.

Of these nine items, only (1), (6), (7) and (8) are the direct responsibility of the amateur, though he will generally do whatever he can to help eliminate interference arising from the other causes, which, basically, are often due to defective television-receiver design.

It is towards the prevention of any harmful radiation of harmonic signals at television frequencies that the main activities of the amateur has been concentrated. Since, in urban conditions, the neighbour's television aerial may be within a few feet of the transmitting aerial, and a few milliwatts of radio-frequency energy may be more than sufficient to mar reception, it will be appreciated that harmonic suppression of a very high order is required.

However, many modern stations may be operated on all amateur frequencies without causing interference to television receivers, even when these are in the same room or building. Generally speaking, the most successful method so far developed lies in the screening and filtering of the transmitter to a far greater extent than was previously normal amateur practice, combined with careful checking of the harmonic output. By keeping the harmonic energy developed in each stage to a low level, completely screening the units, filtering all leads emerging from the units and inserting a three- or four-stage low-pass filter in a low-impedance co-axial line to the aerial tuning unit, harmonic output can usually be kept to manageable proportions, even in television fringe areas.

Transmitter design factors that are of value in reducing harmonic output may be summarized as follows:

(1) The use of adequate screening, completely enclosing the radio-frequency sections, and preferably including the power

supplies, taking into account meter apertures, ventilation louvres and loosely fitting lids. To provide ventilation, fine copper mesh may be employed, but on no account should slots be left unshielded.

(2) The operation of all stages other than the power amplifier (and penultimate amplifier, where more drive is required) at the lowest possible input, keeping drive and grid current in each stage to the minimum practicable.

(3) The operation of stages under Class B rather than Class C conditions, with high-C anode tank circuits, and link-coupling or "wide-band" couplers between each stage and a pi-coupling output network.

(4) The provision of efficient radio-frequency by-passing, with all earth leads for each stage returned directly to the cathode valve pin with short, heavy-gauge wire or copper strip, and good-quality capacitors.

(5) The use of cables made up of individually screened wires for inter-unit connections, with adequate radio-frequency filtering of all wires immediately before they emerge from the screened units.

(6) The aerial should not be coupled directly to the power-amplifier tank circuit but to a separate aerial coupling unit, connected to the power amplifier by low-impedance screened co-axial line with a low-pass filter interposed between the tank circuit and the aerial coupler.

(7) The use of a screened mains filter in the main power cable close to the transmitter

(8) The avoidance of feedback in all radio-frequency stages by: (*a*) use of multi-grid rather than triode valves; (*b*) careful layout; (*c*) use of screened coils.

(9) Use of parallel-tuned traps in the anode circuits, tuned to reject the offending harmonic.

When investigating television interference caused by harmonic radiation it is important to determine from which stage(s) the offending radiation is taking place: it is, for example, no use installing low-pass filters in the aerial feeders or redesigning the power amplifier if the harmonic signal is reaching the television receiver directly (or via the mains supply) from a frequency-doubler stage, or even from the oscillator unit.

For new equipment, the greater ease of providing adequate screening for one unit, rather than for a series of rack-mounted or separate units, has brought about a trend towards the compactly built "table-top" transmitter.

Modulation and Keying

All British amateurs may use either telegraphy (A1) or telephony (A3, A3a). In recent years there has been a trend towards more telephony operation but there is still considerable use made of telegraphy. In order to minimize interference between stations using these different modes, the Radio Society of Great Britain has requested British amateurs to adhere voluntarily to the following division of the high-frequency bands given in Table 3.

TABLE 3.—DIVISION OF HIGH-FREQUENCY BANDS

kc/s	Mode	kc/s	Mode
3,500–3,600	Telegraphy only	3,600–3,800	Telephony only
7,000–7,050	Telegraphy only	7,050–7,150	Telephony and telegraphy
14,000–14,100	Telegraphy only	14,100–14,350	Telephony and telegraphy
21,000–21,150	Telegraphy only	21,150–21,450	Telephony and telegraphy
28,000–28,200	Telegraphy only	28,200–30,000	Telephony and telegraphy

It will thus be seen that the frequencies available, in relation to the large number of stations involved, are extremely restricted, and for this reason considerable efforts have been directed at the reduction of the band-width of telephony transmissions in order to allow a greater number of stations to operate without mutual interference. The International Amateur Radio Union (Region 1 Bureau) has recommended that amateur telephony transmissions should have an attenuation at 4000 c/s of 26 db with reference to the response at 1000 c/s; such characteristics may be achieved by cut-off filters, used in conjunction with splatter suppression. To reduce the side-bands and to increase the average power output of telephony transmitters, the frequency range for speech communication may be restricted to about 500–2,500 c/s without effect on the intelligibility; average power output may be further increased by volume compression and by speech clipping and filtering. The crowded state of the amateur bands, and the considerably greater power efficiency of the system, has also turned the attention of a number of amateurs to single-side-band-suppressed-carrier (S.S.B.) transmission, and it seems likely that this system will be used increasingly in the future, especially when the problems associated with the initial setting up of S.S.B. transmitters without elaborate test equipment have been overcome.

Most amateur stations use either the filter or phasing method of S.S.B. generation, though interest is growing in the "third method" (see Section 7). Currently, a filter system based on half-lattice crystal

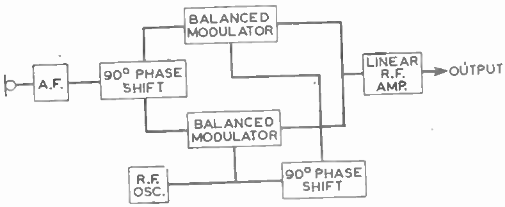

FIG. 7.—BLOCK DIAGRAM OF S.S.B. GENERATOR USING PHASING SYSTEM.

networks, usually with four crystals at about 450 kc/s, is popular: a mechanical filter (see Section 47) may be used instead of crystals. The phasing system, Fig. 7, though basically cheaper, uses close-tolerance components of uncommon values in the 90-degree A.F. phase shift and tends to require readjustment more often; however, less stages of frequency conversion are needed than with filters. With the filter system, a low-level D.S.B. signal is obtained from a balanced modulator. This is then passed through the filter to slice off one sideband and afterwards converted to the desired frequency in one or more heterodyne mixer stages. The output from a low-level S.S.B. exciter is usually fed to a linear amplifier.

The simpler double-sideband (D.S.B.) suppressed carrier system is sometimes used, and can be effective when received on a set having sufficient selectivity to filter out one of the sidebands (otherwise the re-injected carrier must be in correct phase as well as at the correct frequency). However, since for simplicity such transmitters usually have a high-level balanced modulator with the A.F. applied to the screen-grids, it is difficult to avoid some distortion, and the effective power gain will be less than that of S.S.B. unless special receiver circuits are used.

Conventional plate-and-screen modulation of tetrode power amplifiers continues to be widely used by amateurs, but the relatively high audio power required for 100 per cent modulation (approximately 50 per cent of the D.C. input to the power amplifier) has led to the development of alternative systems such as S.S.B., narrow-band-frequency-modulation (n.b.f.m.), and various forms of high-efficiency grid modulation. Limited use is also made of screen-grid, suppressor-grid and cathode modulation.

Narrow-band-frequency-modulation

Narrow-band-frequency-modulation with a maximum deviation of 2·5 kc/s (compared with 30–75 kc/s for normal frequency-modulation systems) and a maximum modulating frequency of 4,000 c/s is now permitted on all amateur bands. Although the effectiveness of an n.b.f.m. transmitter is of the order of only one-quarter that of a 100 per cent amplitude-modulated transmitter and, unless the higher audio frequencies are sharply attenuated, may cause side-band splatter, it possesses the advantages that the modulation equipment need consist of only a single-stage audio voltage amplifier and simple reactance modulator, and that it is less likely to cause interference to local broadcast and television receivers. The simplicity of the modulation equipment makes this system particularly suitable for portable and mobile operation. Unlike normal frequency-modulated transmissions, the signals may be received satisfactorily on a standard amplitude-modulation-type receiver, by tuning to one side of the carrier frequency and switching off the automatic gain control. It should be noted, however, that the use of a low modulation index means that the advantage of the improved signal/noise ratio possible with wide-band-frequency-modulation is lost.

Keying

While the adjustment of transmitters for telegraphy operation presents fewer complications than for telephony transmission, precautions

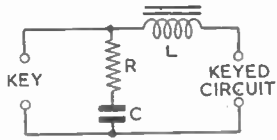

FIG. 8.—LAG FILTER.

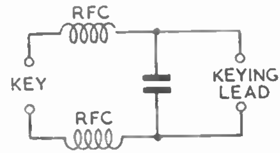

FIG. 9.—R.F. KEYING FILTER.

must be taken to prevent the radiation of key-clicks, whilst communication is greatly aided by the provision of " break-in " facilities by which the transmitting station can " listen-through " his own outgoing signals in order that the other operator may interrupt the transmissions immediately should interference be experienced.

Since in an amateur station there is seldom any appreciable physical separation between the transmitter and receiver or between incoming and outgoing frequencies, means must be found of preventing the receiver from being blocked by local pick-up, or the receiver input circuit from being damaged by local radio-frequency energy, and also to minimize any " thumps " in the receiver when the transmitter is keyed.

Key-clicks, extending many kilocycles on either side of the transmission frequency, are produced when the transmitter output has too short a rise or decay time when the transmitter is keyed; additionally, the minute spark that occurs at the key contacts may interfere with local broadcast and television reception unless there is effective radio-frequency filtering close to the Morse key. Furthermore, keying cannot be considered satisfactory if it produces any " chirp " (frequency variation) on the radiated signal.

Where break-in facilities are not required, and keying takes place in one of the later stages of the transmitter, the required lengthening of the rise and decay time may be achieved simply by the use of a lag filter in the keying line (see Fig. 8). In this circuit the inductance L serves to

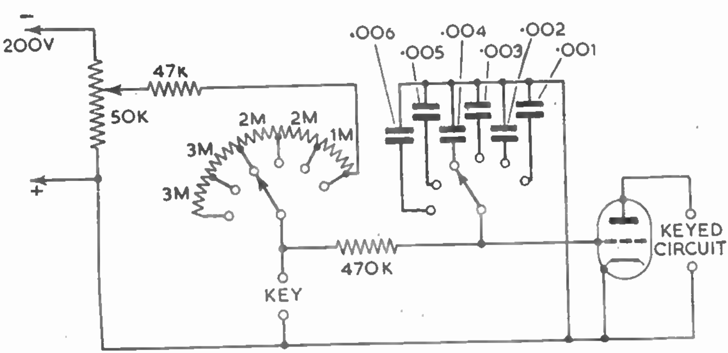

FIG. 10.—VALVE-KEYER WITH ADJUSTABLE LAG CHARACTERISTICS.

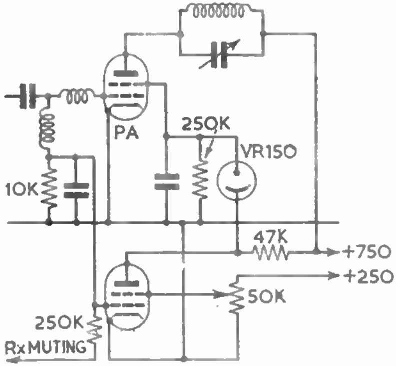

FIG. 11.—VOLTAGE REGULATOR BREAK-IN KEYING ARRANGEMENT.

The receiver muting voltage is used to provide biasing for the early stages of the receiver by means of the A.V.C. Line.

delay the rise in output for a fraction of a second, when the key is depressed, while C delays the fall in output by a roughly similar amount when the key contacts are opened; R—sometimes omitted—helps, in conjunction with C, to reduce sparking at the contacts. The values for these components will vary according to the current being broken by the key, and by the average speed of keying. A radio-frequency filter for installation close to the contacts or the key or the keying relay is shown in Fig. 9.

To provide more readily adjustable lag characteristics valve-keyers are occasionally used. In this system the action of the key removes bias from one or more "keyer" valves which are in series with the transmitter stage to be keyed. The keying characteristics may then be varied as desired by variation of the time-constants of the grid circuit. A typical valve-keyer circuit is shown in Fig. 10.

Break-in Operation

For break-in operation on the same frequency as the local transmitter, it is necessary to key the oscillator stage, unless this is so efficiently screened as to eliminate completely any pick-up by the receiver; in the latter case the first stage from which any signals are picked up must be keyed. Keying an oscillator stage tends to produce some frequency variation whenever the valve just goes into, or is just about to cease, oscillation; though this frequency variation can be much reduced by careful attention to the operating conditions of the oscillator valve and the time-lag produced by screen by-pass capacitors and the like, and by the use of stabilized power supplies. One of the most satisfactory ways of overcoming this difficulty is by "ganged" keying of the oscillator with a later stage of the transmitter in such a way that the oscillator is switched on a fraction of a second before, and switched off a fraction of a second later, than the amplifier stage.

A simple electronic means of achieving this condition is that shown in Fig. 11, which will be recognized as a modified form of the "clamp" circuit already described (Fig. 6). In the modified form a neon voltage regulator valve is inserted between the screen-grid of the amplifier stage and the anode of the clamp-valve. When the latter valve is drawing current, the voltage across the regulator tube is arranged to be less than its striking voltage; so that, in effect, no positive screen-grid voltage is applied to the amplifier valve. When the oscillator is switched on—by the Morse key—and radio-frequency drive applied to the amplifier, the clamp valve is biased back, the voltage at its anode rises, and the regulator valve strikes, thus applying voltage to the screen-grid of the amplifier. The slight delay in the striking of the neon regulator valve affords the necessary delay and prevents any variation of the oscillator frequency—and, incidentally, any key-clicks—from being radiated. This circuit can also be used to help desensitize the receiver automatically during transmission periods by providing a biasing voltage to the early stages of the receiver, taken through a high-value resistor to the A.G.C.-line of the receiver. To complete the elimination of offending thumps in the receiver, audio limitation of the output of the receiver may be necessary; this can be done by reducing the screen-grid potential on the audio-frequency amplifier to a low figure, so that the valve operates under saturated conditions and/or by the inclusion of an audio peak-clipping circuit.

HIGH-FREQUENCY RECEIVERS

Basically, the design of high-frequency receivers for amateur operation is similar to those for general-purpose high-frequency communications work, see Section 22, and it is therefore not intended to discuss such design in any detail in this Section.

Amateur Requirements

Compared with receivers for commercial or military purposes, however, the price factor assumes a greater importance, and provision for either electrical or mechanical bandspread on the amateur bands is essential. The other main requirements can be summarized as follows:

(1) Coverage 1·8–30 Mc/s, with good sensitivity and signal-to-noise ratio throughout, permitting the reception of extremely weak signals.

(2) A high order of selectivity, preferably with a crystal filter.

(3) Freedom from spurious responses. Since high-power commercial and broadcast stations will generally be operating on the image-response frequencies, a high order of second-channel suppression is essential.

(4) Electrical and mechanical stability. It is important that the operation of the gain or the send–receive switch controls do not affect the local oscillator frequency.

(5) Freedom from mains hum. Smoothing must be sufficient to eliminate mains hum on headphones, and care must be taken to ensure that modulation hum does not occur in the early stages of the receiver.

(6) Reasonably accurate calibration and re-set accuracy of the tuning mechanism.

(7) Convenient positioning of controls.

Refinements that are often incorporated include a noise limiter for protection against impulse interference, voltage regulation for the local oscillator, a signal-strength meter and an audio filter.

The double-conversion superheterodyne receiver enjoys a fair degree of popularity: with this arrangement a low second intermediate frequency of the order of 50–100 kc/s enables high selectivity to be obtained in this section of the receiver; whilst the first intermediate frequency of 465 or 1,600 kc/s assists image rejection. Good screening of the second local oscillator, however, is essential if spurious responses caused by beats of harmonics of this oscillator are to be avoided. In a number of instances conventional single-conversion receivers have been converted to double-conversion arrangements by feeding the output from the final intermediate-frequency transformer to a second superheterodyne receiver with a low intermediate frequency: such a unit is generally referred to as a " Q5-er ".

Two new aids to selectivity which have become popular recently are the "*Q*-multiplier" and the mechanical filter. The *Q*-multiplier utilizes the apparent increase of *Q* of an inductor in a circuit near the point of oscillation. By this means, the *Q* of an inductor can be made to approach that of a crystal, and if such a circuit is coupled to the receiver I.F. it can be arranged to provide either a sharp null or peak tunable through the receiver I.F. pass band. With simple communication receivers, an add-on *Q*-multiplier unit can bestow many of the benefits of a crystal filter. Alternatively, a bridged-T filter, sometimes in conjunction with a *Q*-multiplier valve, may be used to provide a sharp rejection notch. The mechanical filter, which can provide an excellent I.F. response curve, is popular for this application in the U.S.A.: see Section 47.

Crystal Filters

Although audio filters and audio-phasing devices may be of considerable assistance in the reception of telegraphy signals in a crowded band, and can give to a receiver most impressive selectivity characteristics on paper, it is important to remember that in practice a high degree of selectivity should be achieved fairly early in the receiver, otherwise strong interfering signals may overload one or more stages and cause cross-modulation. Since cross-modulation occurs readily in mixer stages, high gain at signal frequencies is not generally desirable, so long as a good noise figure is obtained. A crystal (or mechanical) filter in the first intermediate-frequency stage remains one of the best means of rejecting strong adjacent channel interference, provided that the intermediate-frequency transformers are correctly aligned so that the centre of the intermediate-frequency response curve coincides with the crystal frequency, and the user is fully conversant with the correct operating procedure.

As such knowledge is by no means universal among either amateur or commercial operators, the following brief notes may be of assistance.

For telegraphy reception, the beat-frequency oscillator of the receiver should always be set in advance between 500 and 1,000 c/s higher or lower than the crystal frequency. This may be done most easily with the aid of the signal-strength meter. First, with the beat-frequency oscillator switched off and the phasing control at mid-position, a steady signal is carefully tuned to give maximum reading on the signal-strength meter (or maximum audio output if there is no such meter fitted). The beat-frequency oscillator is then switched on, and adjusted to give the desired audio heterodyne note—generally between 500 and 1,000 c/s, depending upon the operator concerned—reducing the radio-frequency gain control as necessary to prevent overloading. From this point onwards, the beat-frequency oscillator control should not be touched. In operation, the desired station should be tuned for peak audio output, and an interfering signal can then be eliminated by adjusting the crystal phasing control, which changes the position of the crystal-rejection notch relative to the crystal frequency. Provided that the required signal has been correctly tuned, the insertion of the crystal filter, or the adjustment of the phasing control, should not seriously affect the strength of the desired station.

For telephony reception, band-pass filters using two crystals of slightly differing frequencies are sometimes used; a further development in recent years is a receiver having switched sideband selection.

This is usually achieved in double-conversion receivers by having a crystal-controlled second oscillator (or in some cases, where there is a tunable 1st I.F., the first oscillator) with two crystals one above and one below the signal frequency.

Single-sideband-suppressed-carrier Reception

The reception of single-side-band-suppressed-carrier telephony transmissions requires the use of a carrier insertion oscillator. In practice, the beat frequency oscillator of the conventional receiver is often used, provided that it has a high order of stability and can be adjusted conveniently to within about ± 20 c/s of the required frequency; alternatively, a crystal-controlled oscillator at the intermediate frequency may be employed to improve long-term stability. Without the use of a carrier-insertion oscillator, single-sideband-suppressed-carrier transmissions are completely unintelligible, whilst a slight drift of the frequency of either the local oscillator or the carrier insertion oscillator will rapidly cause distortion.

The stability and band-spread requirements of receivers for S.S.B. reception are considerably more stringent than those for normal A3 reception, and this is reflected in modern designs. Apart from greater stability of the various oscillators, and a slower tuning rate, the band-width may with advantage be reduced to about one-half of that for A3 reception. To reduce intermodulation effects at the detector, it is now common practice to fit both a conventional detector circuit for A3 and a product detector for use on S.S.B. and telegraphy.

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J. P. H.

10. TELEVISION TRANSMITTERS

10. TELEVISION TRANSMITTERS

Introduction

In order to conserve channel space it is now usual to adopt vestigial sideband operation for television transmitters.

Two practical systems are possible, namely, the transmitter attenuation (t.a.) system and the receiver attenuation system (r.a.). In the former the carrier-frequency response is attenuated 6 db, i.e., to half amplitude, and the shaping of the response curve is such that the vector addition of the two sideband components that are transmitted near the carrier frequency results in a flat response over the entire pass-band of a double-sideband receiver. This is illustrated in Fig. 1.

In the alternative system (r.a.) the attenuation is inserted at the receiver and the transmitter may have a double-sideband characteristic or an asymmetric frequency characteristic provided that it extends without attenuation over the whole of the receiver characteristic: see Fig. 2.

The latter (r.a.) system has been adopted exclusively so far, as it leads to a cheaper receiver due to the narrower band-width required, and the fact that a receiver tuned to a frequency displaced from the carrier so that the carrier is at the −6-db response will receive satisfactory pictures without undue complications.

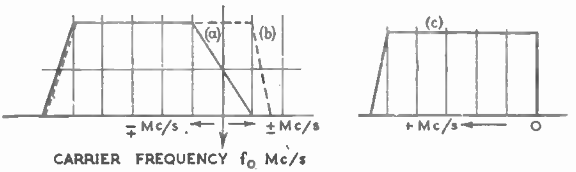


FIG. 1.—IDEALIZED T.A. SYSTEM OF V.S.B. TRANSMISSION AND RECEPTION.

a) Idealized transmitter response; (*b*) idealized receiver response; (*c*) idealized equivalent low pass (demodulated) response of system.

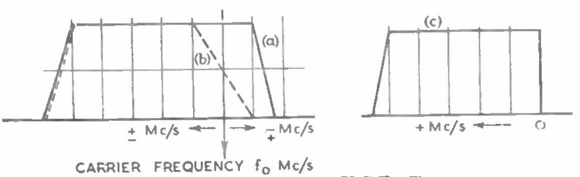


FIG. 2.—IDEALIZED R.A. SYSTEM OF V.S.B. TRANSMISSION AND RECEPTION.

(*a*) Idealized transmitter response. (*b*) idealized receiver response: (*c*) idealized equivalent low pass (demodulated) response of system.

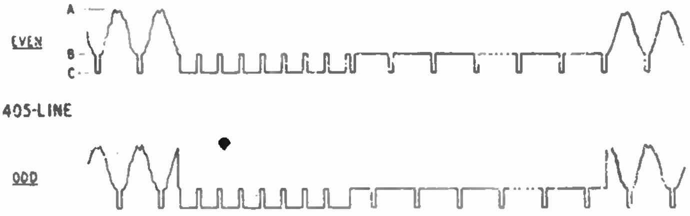


Fig. 3.—British Standard 405-line System Waveform Characteristics.

Performance Specifications

The standard television waveform for the United Kingdom is shown in Fig. 3, and performance specifications for transmitting equipment are as follows :

Input Signal

Overall amplitude	1-volt d.a.p. subject to ±3 db variation
Picture-signal/synchronizing-pulse signal voltage ratio	Normally 70/30, but can change within the range 80/20 and 60/40
Picture signal polarity	Positive with respect to black level
Synchronizing pulse polarity	Negative with respect to black level
Input impedance to match a feeder	75 ohms ± 5 per cent

Distortions that may occur in the input signal :

(1) Bad time constant of A.C. coupling	Not to affect performance of transmitter
(2) Superimposed noise signal up to about 1 microsecond in width	Not to affect performance of transmitter
(3) Superimposed hum voltages	To be removed by transmitter

Output Signal

Carrier frequency stability	0·005 per cent
Peak white signal (A)	100 per cent max. carrier amplitude
Black-level signal (B)	30 per cent max. carrier amplitude
Synchronizing signal (C)	Not to exceed 2 per cent max. carrier amplitude
Medium and low-frequency response	Response to step waveform in picture-signal amplitude range shall after 1 microsecond be constant within ±1 per cent up to 0·1 second later and be constant ±2 per cent at any subsequent time
High-frequency response (0·5–2·75 Mc/s)	Rise time in response to step function not to exceed 0·2 microsecond with less than 5 per cent overshoot decaying to less than 1 per cent in less than 1 microsecond

Output Signal

Black-level stability	±2 per cent
Amplitude/linearity characteristics	With an input signal of saw-tooth waveform the departure from linearity of the output signal to be less than 4 per cent
Hum modulation	Any spurious low-frequency noise or hum to be at least 52 db below that of the picture signal at 300 c/s, and should not rise higher than −45 db at 50 c/s; double-amplitude peak values being considered in each case

A summary of responses for various systems in the radio-frequency spectrum is shown below in Table 1.

TABLE 1.—PERMITTED TOLERANCES FROM IDEAL CHARACTERISTIC

Frequency Scale in Wanted Sideband (Mc/s from carrier) (Mc/s)	*American 525-line Standard* F.C.C. (db)	R.M.A. (db)	*Suggested C.C.I.R. 625-line Standard* (db)	*B.B.C. 405-line Standard* (db)	*625-line Standard Based on Equal Vertical and Horizontal Definition* * (db)
1·25	2	2	2	0·5	0·5
2·0	3	3	2½	0·5	0·5
2·75	5	3	3	1·0	—
3·0	6	3	3¼	3·0	—
3·5	12	4	3½	—	—
4·0	—	6	4	—	—
4·5	—	—	5	—	—
5·0	—	—	6	—	—
6·0	—	—	—	—	1
7·0	—	—	—	—	3

* These figures are approximate, and have been obtained from a curve drawn through the permitted tolerances.

Special Problems Arising from the Specifications

Generally speaking, it may be said that this specification for a 405-line system is a difficult one to meet, and is very close by comparison with that now currently adopted by the users of 525-line standards.

The implications of these standards on design are dealt with briefly below before proceeding to more detailed discussion.

Low-frequency Response. Even if perfect line-by-line clamping is possible, so that the black level is corrected accurately at the commencement of each line, the requirement of constancy within ± 1 per cent from 1 microsecond onwards means that the whole of the transmitting system must not introduce an overall effective A.C.-coupling time-constant smaller than about 8·5 milliseconds. Since the system will in fact have many A.C. couplings in cascade, each stage must be designed for very low-frequency response.

Also in order to prevent "hunting" effects, the low-frequency response must be critically damped.

Furthermore, the long-term black-level stability requirements depend on low-frequency couplings, supplies, black-level stability circuits; and, on the whole, require much greater attention to detailed design than do high-frequency response requirements for a given standard of picture quality.

High-frequency Response. A rise time of 0·2 microseconds with less than 5 per cent overshoot is well within the capabilities of economic design, but it is doubtful if such a specification should be based on response to an infinite frequency spectrum waveform.

The "Heaviside Unit Step", or its counterpart the infinite rate of rise pulse, both represent waveforms with frequency spectra extending to infinity. The camera channel will produce only limited spectra information, and a more realistic test would use limited spectra waveforms to give a practical answer directly without mathematical interpretation.

The "sine squared" pulse [1,2,3] and the sine squared step transition [3] give an approach to practical test conditions.

The use of a vestigial sideband characteristic also affects the performance of the system due to the inherent quadrature distortion.[3,4]

Linearity. Non-linearity in a transmitter is determined by:

(1) the VA capacity of the system being adequate for all demands;
(2) valve characteristics;
(3) valve and circuit impedances and the effects of non-linear loading in radio-frequency and video-frequency channels;
(4) power supplies.

These factors are usually insufficiently good—in an economic design—to meet the specification. Some predistortion is therefore employed. Given a predistorting system, linearity within 1 or 2 per cent is often achieved.[6,8,9]

Stability of Synchronizing Pulse Amplitude. Given a stable black level, the performance of the average receiver will depend on a stable synchronizing pulse amplitude.

The tolerances permitted in the specification allow ± 3 per cent variation in the synchronizing pulse amplitude.

The transmitter must maintain this performance, even if the input signal varies in picture-signal amplitude to synchronizing-pulse amplitude ratio. Means are therefore employed to stabilize the pulse amplitude.

Transmitter Noise. The greatest trouble likely to arise in practice is due to unsynchronized 50 c/s or harmonic components appearing on the receiver picture as dark bands moving vertically across the screen.

With A.C.-heated transmitter valves, it is difficult to achieve the -50 db necessary on these low-frequency noise components. Overall feedback methods are employed where necessary to solve this problem.

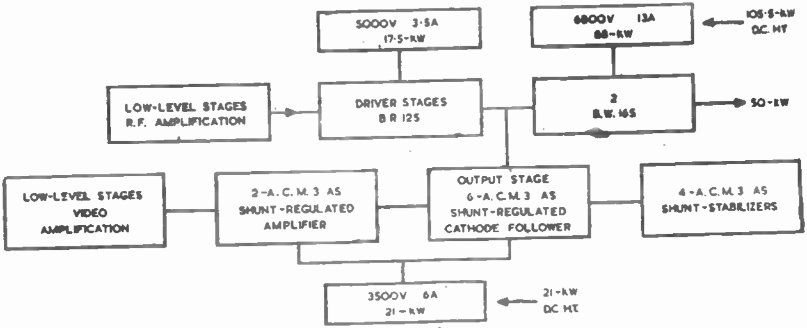


FIG. 4.—SCHEME FOR 50-kW HIGH-LEVEL MODULATED TRANSMITTER.
(*Courtesy of the I.E.E.; reference* 8.)

MODULATION METHODS

It is usual for broadcast transmitters to employ amplitude modulation, but frequency modulation is often used for point-to-point relay systems.

Owing to the wide band of modulation frequencies, it has not been found practicable so far to use transformer coupling in the video-frequency stages, so the methods of amplification and modulation, familiar in sound transmitters, are not available to the designer of television transmitters.

The design of a television transmitter is dictated predominantly by the method by which modulation is applied. At the present time the vast majority of transmitters employ grid modulation, which owes its popularity to the fact that, for a given radio-frequency peak power, the VA required from the modulator is a minimum. Grid modulation can be designed to operate at any stage in the chain of radio-frequency amplifiers.

Low-level Modulation

This is usually understood to mean modulating a radio-frequency stage at a low power level in the transmitter, and following the modulated stage with one or more stages of linear amplification. At least one author [5] on this subject subdivides further and introduces the term "mid-level modulation" to imply a modulated radio-frequency stage followed by a single radio-frequency linear amplifier, and uses the term "low-level modulation" for systems employing more than one linear amplifier following the modulated stage. Examples of low-level modulated transmitters are described in references 5 and 6.

High-level Modulation

This is usually understood to imply modulation at the grid of the final radio-frequency amplifier, but may also include any form of modulation at the last radio-frequency amplifier or following this amplifier, such as absorption modulation at the feeder.

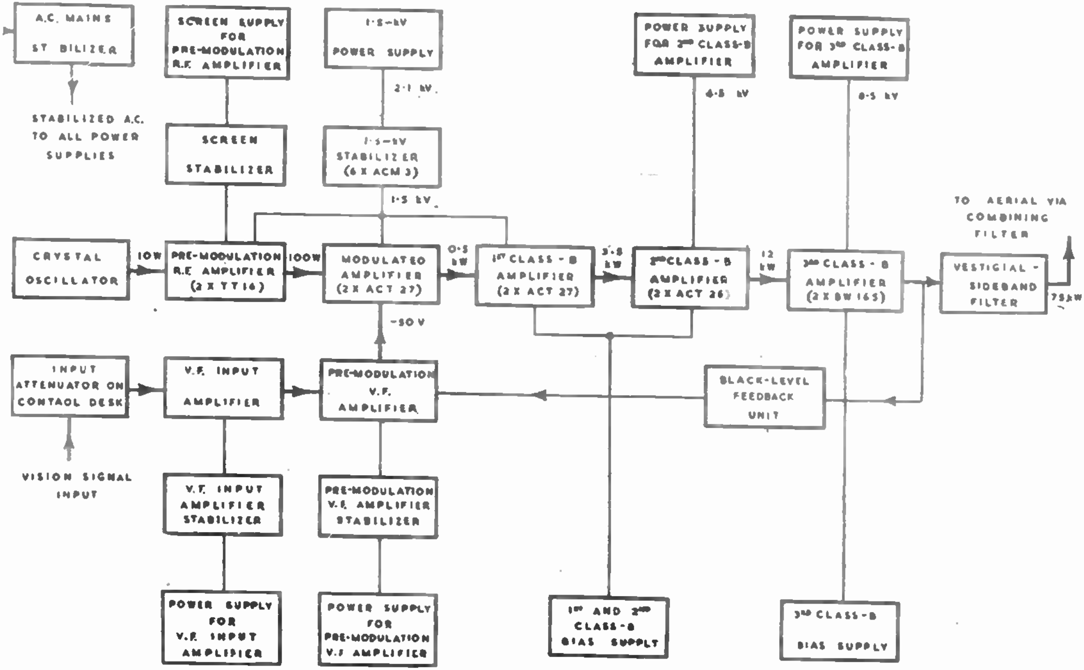


FIG. 5.—BLOCK DIAGRAM OF LOW-LEVEL MODULATED TRANSMITTER.

(Courtesy of the I.E.E.; reference 6.)

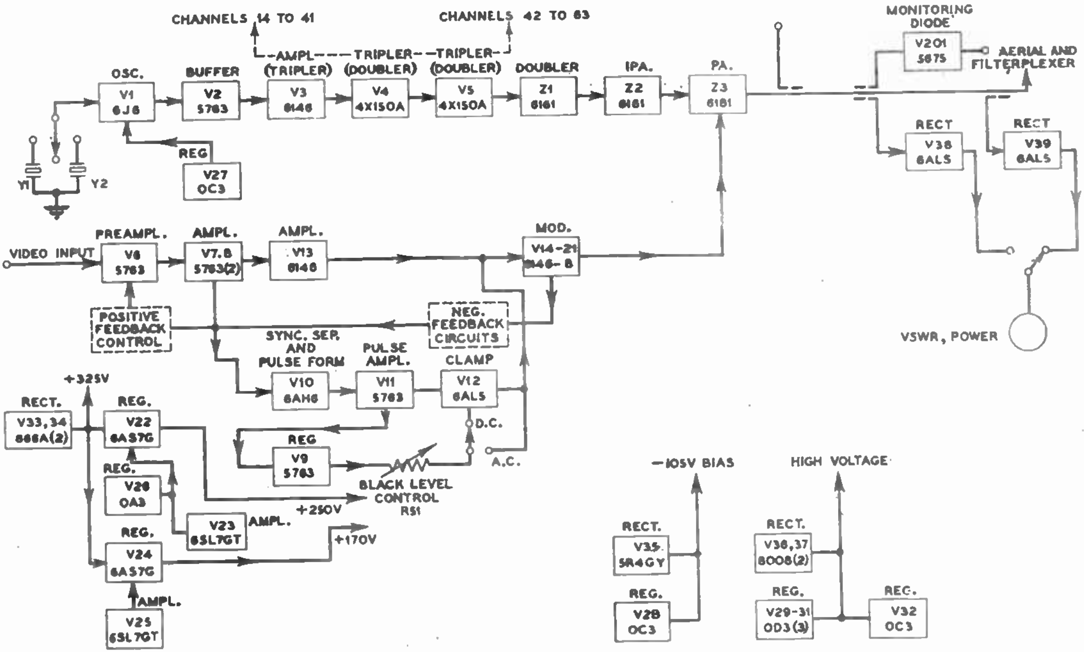


Fig. 6 (a).—Schematic Diagram of a Cathode-modulated 1-kW U.H.F. Transmitter.

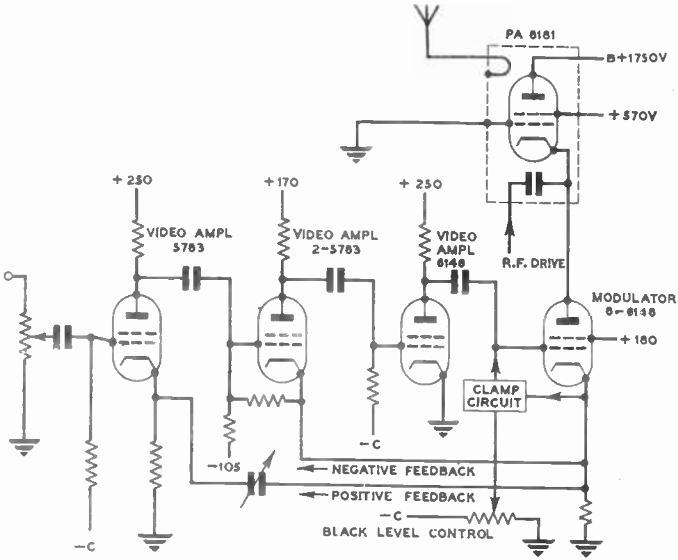


FIG. 6 (*b*).—MODULATOR ARRANGEMENT FOR CATHODE-MODULATED 1-kW TRANSMITTER.

The matter of choice between the two broad methods is still a subject for considerable discussion,[6,7,8] and in practice is often decided by the restricted availability of valves rather than by aspects of performance, operational reliability, ease of adjustment and other similar criteria.

Fig. 4 shows a typical schematic arrangement for a grid-modulated transmitter employing push–pull radio-frequency stages and high-level modulation. Fig. 5 shows a typical grid-modulated transmitter employing push–pull stages and low-level modulation. Other examples of grid-modulated transmitters are described in references 9, 10, 11, 12, 13 and 14.

Cathode Modulation

Of the other methods of modulation possible, cathode modulation is the most attractive at the present time, and is used currently on U.H.F. and V.H.F. systems.[15,16] Fig. 6 shows a typical arrangement in use in the U.S.A. on the U.H.F. service.

Anode Modulation

Generally speaking, anode modulation is not economically possible for television transmitters. Theoretically, Heising type modulation is possible using a high-impedance constant-resistance network as the

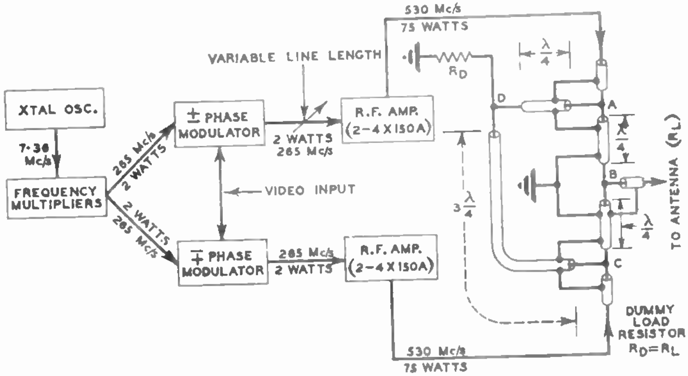


Fig. 7.—Phase-to-amplitude Modulation System Used at U.H.F.

"choke", but would need the development of high-power valves suitable for the video-frequency modulator. Series anode-modulation is also a theoretical possibility, given the high-power video-frequency modulating valves.

Amplitude-phase Modulation

This has been used in the U.S.A. at U.H.F.,[17] and is illustrated in Fig. 7. In this system the output stages are combined in a bridge network with a radiating load and a non-radiating load, so that each stage operates independently of the other. At the trough of modulation all the power is dissipated in the non-radiating load. For the peak of modulation the phase of each radio-frequency system is rotated through 90 degrees, and all the power goes to the radiating aerial load. Linearity for this system is such that, for a linear input, the output is sinusoidal.

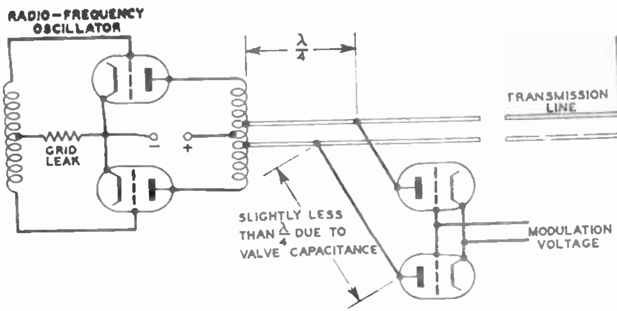


Fig. 8 (a).—Arrangement for Absorption Modulation.

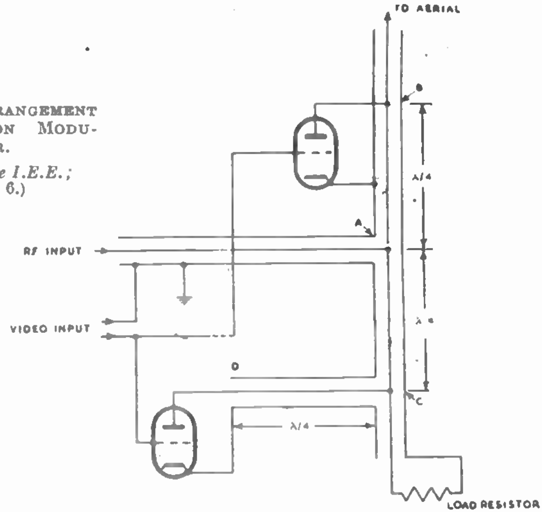


FIG. 8 (b).—ARRANGEMENT OF ABSORPTION MODULATOR.

(*Courtesy of the I.E.E.; reference* 6.)

In addition to the poor linearity, the system suffers from low efficiency. Against these disadvantages must be weighed the saving in power-supply complications, since each transmitter chain may operate at constant load during modulation.

Absorption Modulation

This is probably the most simple of all modulation methods in theory, having the advantage over all other systems of keeping the radio-frequency and video-frequency functions of the transmitter completely segregated until after all power amplification has taken place.[18, 19] Difficulties that have arisen in designing the correct modulation impedances with available valves and achieving good depth of modulation have prevented practical adoption of this method. Two possibilities are illustrated in Fig. 8, (a) and (b) the first due to Parker [18] and the other due to Blumlein and Nind.[6]

GRID-MODULATED TELEVISION TRANSMITTERS

The crystal oscillator, frequency-multiplying stages and the earlier power amplifiers can be of conventional design.

The Modulated Amplifier Driving Stage

The amplifier stage, driving the modulated amplifier, calls for special consideration. First, the modulated stage, whether grounded cathode or grounded grid, will present a non-linear load to the driving valve.

If amplitude linearity during modulation is to be preserved, the radio-frequency voltage excitation on the modulated stage must be kept substantially constant despite the varying load. This requirement demands a low-impedance source of radio-frequency excitation. The low-impedance source has been achieved in a variety of ways.

The Heavily Damped Driving Stage

This was the method chosen for the Alexandra Palace transmitter installed in 1936,[12] and was economically possible because the valves then available and suitable for the modulated amplifier (type CAT 9) required very small grid current and therefore presented a much less non-linear loading than designers are accustomed to meet at the present time with more modern and efficient valves. In this method a conventional Class B or Class C amplifier is loaded by resistors at the grid circuit of the modulated valves, so that the change of loading, due to grid current and transit-time effects in the modulated valve, are negligible in comparison with the steady load.

The Radio-frequency Cathode Follower Drive

This inherently low-impedance driving source, has been used successfully in a 50-kW transmitter,[8] to achieve a very low effective driving impedance. With careful design, this method can give highly satisfactory performance, but presents a number of difficulties in design. The grid-cathode excitation of a cathode follower is a relatively small voltage compared with the cathode–anode voltage swing, and in practice is determined by the vector difference between two large voltage swings, one cathode-ground, and the other grid-ground. It is therefore necessary in design to make special provision for protecting the valve against excessive excitation due to relatively small misadjustment of tuning of the input and output circuits, which therefore become somewhat more critical in operation.

In one existing design [8] of a high-power transmitter, the inherent low-output impedance of the valves connected as radio-frequency cathode-followers, combined with a capacitative potentiometer into the modulated grids, produces an effective driving impedance of about 10 ohms.

The Tetrode or High-impedance Driving Stage with Impedance Transformation

It is well known that by the use of a quarter-wave network coupling [20] following a high-impedance driving source, the effective driving impedance as seen by the succeeding stage can be made very low. This has been used in several designs, and is attractive where suitable tetrodes are available as driving valves, since the properties of an ideal quarter-wave network can be simulated satisfactorily in practice by a simple double-tuned single-coupled circuit.[20]

The Hard-driven Class C Driver Stage

The driving impedance obtainable by this means with present-day available valves cannot be made as low as can be achieved with cathode-followers, but if followed by impedance transformation in the coupling circuits can be made satisfactory with simplicity in adjustment and operation.[9]

Band-width Requirements of the Driver and Coupling Network

The overall problem of the modulated-stage driving impedance can be subdivided into two problems :

(1) The problem of linearity caused by the varying load on the driver stage due to grid current at low modulation frequency.
(2) The frequency response of the driver stage impedance.

The first problem is covered by the previous section, and we are left with a consideration of the band-width requirements in the driving impedance. The problem arises because any change of load condition on the driver stage during modulation causes a varying transfer through the driver output impedance, and in the modulated stage coupling. If these impedances are not effectively aperiodic over the required band-width, then the effective driving impedance will change with an effect on linearity and apparent frequency response of the modulated stage. One solution used at present is to make the driving stage and coupling network wide-band so that the effective driving impedance remains substantially constant over the required bandwidth. This often leads to an inefficient driver stage.

Probably the simplest arrangement is to make the driving stage have a critically damped narrow band-width and to couple this to the input of the modulated stage with negligible impedance. This will give a falling driving impedance with increasing frequency, and will tend to lift the apparent frequency response. A compromise must be sought between level response and driver efficiency.

The Grid-modulated Stage

For a Class B, push–pull radio-frequency stage using valves with peak useful emission of I amperes, it can be shown [12] that the maximum power available is given by

$$\text{Peak Power} = \frac{I^2 k}{16\pi C \Delta f}$$

where C is total circuit capacitance, Δf is the 3-db band-width of the circuit and k equals 1 for a single circuit.

Improvements in the radiated band-width for the same power can be made by the use of double-tuned, maximally single-coupled circuits or by a triple-tuned maximally double-coupled circuit.[7] In this case the factor k is $\sqrt{2}$ for double circuits and 1·5 for triple circuits.

The profit of triple circuits over double circuits is more marked if flatness of response within the pass-band is important,[7] or if band-width is defined between ½- or 1-db points, as is becoming more common in television transmitter specifications. Figs. 9 (a) and 9 (b) illustrate the comparison of response for various circuit conditions.

Difficulties arise because in this modulated amplifier all the impedances associated with the stage, including supply impedances, must handle currents at all video frequencies, and consideration of all

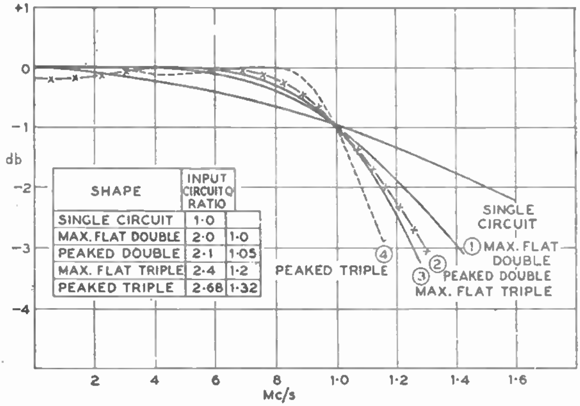


FIG. 9 (*a*).—COUPLED CIRCUITS ON BASIS OF EQUAL 1-db BAND-WIDTH.

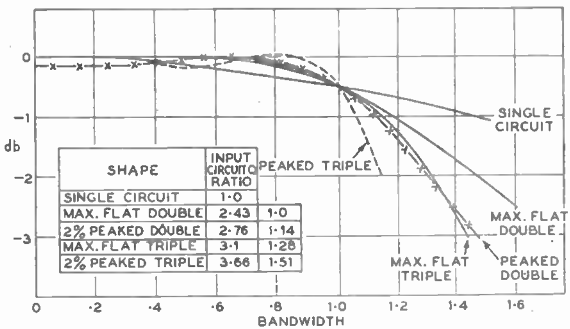


FIG. 9 (*b*).—COUPLED CIRCUITS ON BASIS OF EQUAL ½-db BAND-WIDTH.

possible current paths must be made in order to design for adequate frequency response.

Fig. 10 illustrates the impedances in the video-frequency current path that may affect frequency response.

Grounded-cathode Operation

This gives the simpler radio-frequency driving condition whether tetrodes or neutralized triodes are used, but gives increased input and output capacitances compared with grounded-grid operation.

For the case illustrated in Fig. 10, the input impedance of the modulated stage, as seen by the modulator, is illustrated in Fig. 11.

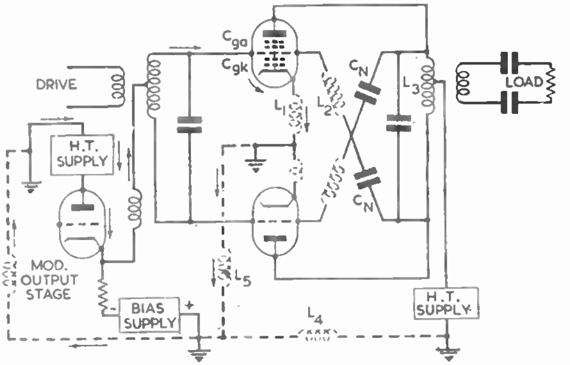


FIG. 10 (*a*).—SHOWING PATH OF TRANSMITTER GRID CURRENT (FOR ONE VALVE ONLY) AND THE IMPEDANCES IN THE PATH OF THE VIDEO FREQUENCY CURRENT.

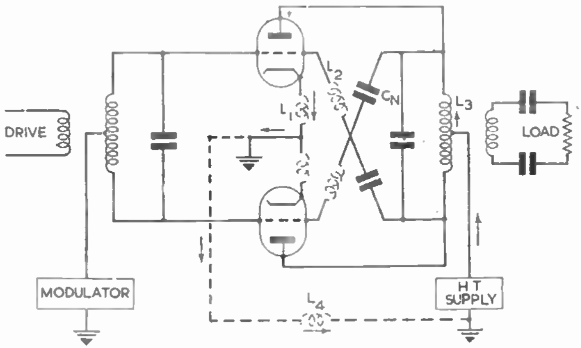


FIG. 10 (*b*).—SHOWING THE PATH OF VIDEO-FREQUENCY CHANGES OF ANODE CURRENT.

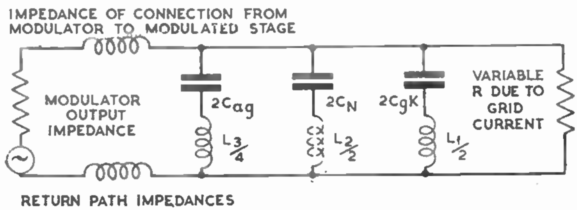


FIG. 11.—INPUT IMPEDANCE OF THE MODULATED STAGE.

Grounded-grid Operation

Here a saving of 30 per cent or more on capacitance is possible in practice,[21, 22] but throws a further problem back into the penultimate stage.[9]

Both methods are in current use at all power levels, and neither can be said to be predominantly superior. The simplest arrangement is the grounded-cathode tetrode, which is likely to emerge as the most modern choice for reasons of simplicity and efficiency.

THE LINEAR AMPLIFIER

In cases where it is desirable or necessary to follow the modulated radio-frequency amplifier by further power amplification, use can be made of the Class B push–pull linear amplifier.

This may take the form of a grounded-cathode stage using triodes, tetrodes or pentodes where they exist, or may take the more popular form of a grounded grid–cathode driven stage.[21, 22]

It is desirable to keep the input resistance to such a stage as nearly constant as possible over the range of amplitude corresponding to the picture signals, from two points of view. First, the linearity of the system will be dependent on the regulation of the driving-valve impedance, and any non-linearity of the input resistance of the linear amplifier will cause a non-linearity of the transmitted signal. Secondly, the input resistance of the linear amplifier is a predominant factor producing circuit damping, and will determine the band-width of the coupling circuit.

Unfortunately the flow of grid current in the linear amplifier causes both the stage amplitude linearity and the frequency response to change over the modulation cycle. Series resistance in the grid circuit helps to maintain the response equality,[6] but only at the expense of linearity; so some compromise must be found.

All the circuits associated with the linear amplifier carry currents, the amplitudes of which vary with modulation frequency. It is thus necessary to ensure that all neutralizing impedances and current return paths from anodes to cathodes are of sufficiently low impedance at all modulation frequencies, or that they are constant impedances at all frequencies.

Output-coupling Systems

Coupling the low-impedance aerial feeder or the low impedance of the input of a grounded-grid linear amplifier on to the anode circuit of the modulated amplifier can be achieved in a number of ways.

By far the most popular method is to employ coupled circuits, either single-coupled double-tuned circuits or double-coupled triple-tuned circuits.

As already indicated during the consideration of the grid-modulated stage, the use of coupled circuits can give an increased power–band-width product.

Double-tuned Coupled Circuit

If critically coupled for maximal flatness, this arrangement gives a power band-width product increase of $\sqrt{2}$ at the 3-db points in the response curve over that obtainable from a single circuit.

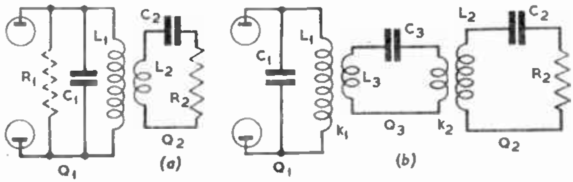


FIG. 12.—(a) DOUBLE-TUNED COUPLED CIRCUITS; (b) TRIPLE-TUNED CIRCUITS.

For maximal flatness the following conditions must be fulfilled. Reference is made to Fig. 12 (a).

$$Q_0 = \frac{f_0}{f} \text{ (3 db)}$$

$$Q_1 = 1{\cdot}414\ Q_0$$

$$Q_2 = 0{\cdot}707\ Q_0$$

$$k = \frac{1}{Q_0}$$

Triple-tuned Circuit

For maximal flatness the triple-tuned circuit, as in Fig. 12 (b), shows an improvement over a single circuit at the 3-db point of 1·5, and circuit requirements are:

$$Q_0 = \frac{f_0}{f} \text{ (3 db)}$$

$$Q_1 = 1{\cdot}5\ Q_0$$

$$Q_2 = 0{\cdot}5\ Q_0$$

$$k_1 = \frac{1}{\sqrt{2}} \times \frac{1}{Q_0}$$

$$k_2 = \sqrt{\frac{3}{2}} \times \frac{1}{Q_0}$$

To illustrate the comparative conditions, Table 2 shows the circuit values for a push–pull stage with 50-pF shunt capacitance working into a 50-ohm load at 50 Mc/s with a 3-db response at 4 Mc/s.

In Table 2, column 3 shows that at the 1 db response band-width, the power–band-width product is improved by a factor of 1·2 by using a triple circuit instead of a double circuit. The intermediate circuit of a "triple" can be made any convenient L–C ratio without affecting the performance.

The design of "double" and "triple" circuits is straightforward, and can be made to achieve responses of the maximally flat type or of the "Chebyshev" type [23] in which there are departures from flatness in the form of a "ripple" in the response curve within the pass band.

TABLE 2.—COMPARATIVE CONDITIONS OF TRIPLE-TUNED CIRCUITS

	Double	*Triple*	*Triple*
3 db band-width	4	4	3·53
Q_0	12·5	12·5	14·15
Q_1	17·7	18·75	21·2
Q_2	8·85	6·25	7·1
L_1	0·2 μH	0·2 μH	0·2 μH
C_1	50 pF	50 pF	50 pF
L_2	1·42 μH	0·95 μH	1·13 μH
C_2	7·15 pF	10·7 pF	9·0 pF
R_2	50 ohms	50 ohms	50 ohms
k_1	8·0%	5·66%	5%
k_2	—	9·8%	8·7%
Power output for valve emission, I amperes	141 I^2 watts	150 I^2	170 I^2
Band-width for 1 db.	2·82 Mc/s	3·2 Mc/s	2·82 Mc/s
Resistance load line per valve	280 ohms	298 ohms	337 ohms

The Modulator–Radio-frequency Transmitter Connection

This needs special consideration, as any inductance in this connection will cause a signal at the modulated valve which rises with frequency and will impose a reactive load on the modulator which rises with frequency.

It is also desirable to keep the radio-frequency signals from the modulator. The connection must therefore be of as low an inductance as possible and made to a point in the radio-frequency circuit which has the minimum of radio-frequency signals present.

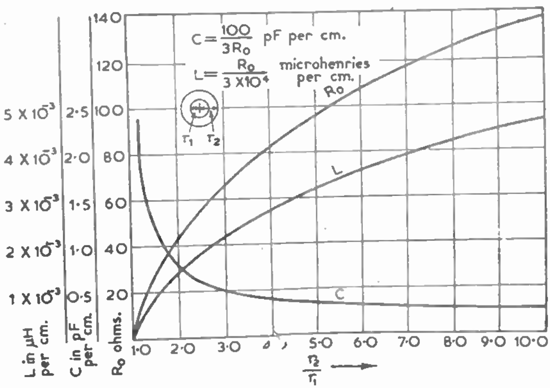


FIG. 13.—CONCENTRIC FEEDER DATA.
Surge impedance $R_0 = 60 \log_e r_2/r_1$ ohms.

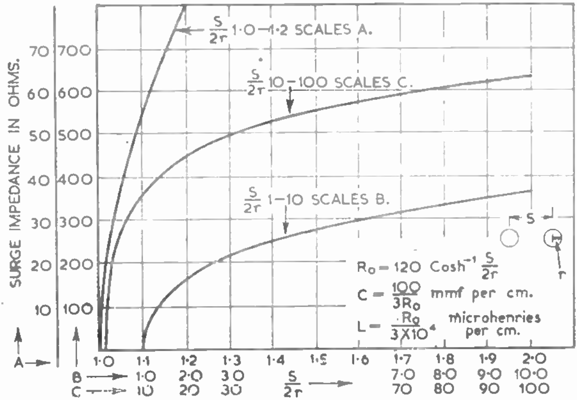


Fig. 14.—Surge Impedance of Parallel Cylinders—Any Spacing.

The mid-point of a push–pull circuit is the optimum practical point. The use of additional shunt chokes or equivalent lines is often adopted in practice for purely physical reasons. Lumped-circuit-type " chokes " produce the minimum additional shunt capacitance at low frequencies, but produce a rising effective capacitance at high frequencies.

The low-impedance line-type choke can give a flatter reactance characteristic over the band at the expense of heavier capacitance loading at all frequencies.

Care must therefore be exercised to get a satisfactory compromise.

Assistance in designing is provided by Figs. 13, 14 and 15, which give the properties of unbalanced and balanced lines, and by Fig. 16, which gives design data for self-resonant chokes.

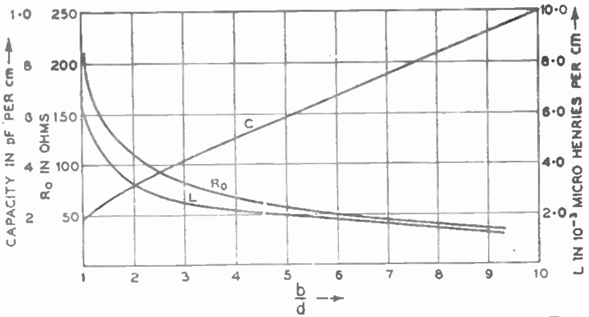


Fig. 15.—Inductance, Capacitance and Surge Impedance of Parallel Tapes.

b = width; d = spacing.

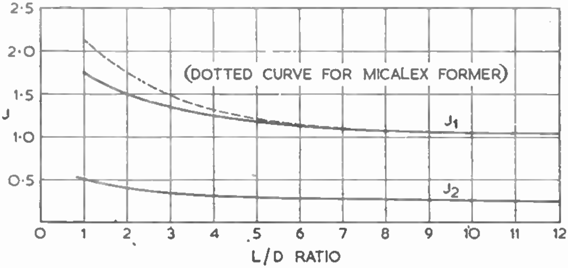


FIG. 16.—ACCEPTOR AND REJECTOR RESONANCES ON SINGLE LAYER COILS.

These curves are mean curves which will give an accuracy within ±10 per cent for close-wound coils using any gauge of wire between 36 and 16 S.W.G. D.S.C. or D.C.C. on formers of low dielectric constant.

Notes: (1) For any particular acceptor frequency the highest inductance coil will be obtained by using the maximum L/D ratio possible, and vice-versa.

(2) Where high high-frequency voltage occurs across the coil and it is necessary to keep up the "Q" value to minimize watts dissipation at the lowest working frequency (e.g., in wide-band choke applications) an L/D ratio = 4–6 is recommended.

To Use the Curves:

Acceptor wavelength ($\frac{1}{4}\lambda$) = J_1L cm.
Rejector wavelength ($\frac{1}{4}\lambda$) = $10J_2L$ cm.

where L = length of wire in cm.;
J = form factor of no dimensions.

THE VIDEO-FREQUENCY MODULATOR

The Output Stage

The video-frequency modulator output stage must have a wide frequency response and must be capable of maintaining this response across a load determined by the input impedance of the modulated stage, and consisting, apart from lead inductances, of a shunt capacitance and shunt resistance which, due to the non-linear flow of grid current, will vary from a high value to a low value with the amplitude of signal.

Also, whereas the reactance loading will increase directly with modulation frequency, the grid-current loading is proportional to modulation signal amplitude, and can be a maximum at any frequency at which full amplitude signals are present.

The combined problem results in the necessity to make the output impedance of the modulator negligibly low in comparison with the load impedance.

The low impedance necessary may be obtained in a number of ways, viz:

(*a*) conventional amplifier working into a low resistance;
(*b*) cathode follower;
(*c*) feedback amplifier system;
(*d*) shunt-regulated amplifier system.

Of these, the simple cathode follower and the shunt-regulated cathode follower are used most widely at the present time.

The properties of these are compared under "Shunt Regulated Amplifiers".

VIDEO-FREQUENCY AMPLIFIERS

These amplifiers raise the level of the standard input level of about 1 volt to the level of perhaps several hundreds of volts required to drive the grid-modulated stage of the transmitter.

A number of methods of achieving the required performance are available:

(1) conventional reactance compensated R–C coupled amplifiers;
(2) feedback amplifiers;
(3) shunt regulated amplifiers.

It frequently happens that a television modulator chain of amplifiers will embody types of various forms, and each of the above types have certain attractive features for particular applications. Each type is therefore dealt with separately, following general considerations of performance.

Performance of Video Amplifiers

High-frequency Response

Television requires the amplification of video-frequency signals which extend in the frequency scale out to many megacycles. The nature of the signal is also such that the amplifier must deal with sudden change of level without significant distortion. This means that attention must be given in design to the transient response as well as the steady-state frequency response. For good transient response it is necessary for the phase/frequency response law to be linear, and it is unfortunate that it is difficult to achieve both flat frequency response and linear phase response at the same time.

If the frequency response of an R–C network is extended by simple reactance compensation so that the 3-db response is improved from a frequency f_1 to a frequency f_2, the linearity of the phase response is not correspondingly extended to the same degree.

Similarly, as more complicated reactance networks are employed to extend the amplitude frequency response still further, the useful range of linear phase response is improved by only a relatively small amount.

This means that as the amplitude response curve is extended, and given a sharper fall away outside the pass-band, the indicial rise time will fall but the overshoot will increase.

In practice, however, we are not called upon to deal faithfully with the transmission of "Heaviside unit steps" (change of level at infinite rate), which, for mathematical convenience, are used to evaluate transient response. In practice we shall have to deal only with a limited range of frequencies, and provided an amplifier system will deal reasonably faithfully with this limited range of components, both in amplitude and phase, then no significant overshoot will occur.

This practical fact means that although maximally flat amplifiers, and any amplifier in which the response has been extended beyond the critically damped condition, overshoot in their indicial response, they may be designed to deal quite faithfully with television signals with their limited frequency spectra.

Low-frequency Response

In addition to this high video-frequency aspect, there is also the requirement of preserving the low-frequency components in the television signal, and these extend virtually to the lowest visible frequencies that are observable during the fading of one picture into the next, i.e. down to fractional frequencies, say $\frac{1}{5}$ c/s.

The effects at high frequencies due to poor phase response giving rise to overshoot have their counterpart at low frequencies, and result in "hunting", i.e. an undamped settling-down process that can be more annoying visually than the "ringing" at high frequencies. It is customary in testing high-performance television transmitters in this country to check the absence of this hunting effect at all rates of change of level down to, say, $\frac{1}{4}$ c/s to ensure that slow fades are faithfully transmitted.

Where possible at high levels amplifiers are direct connected to eliminate these effects, but it is uneconomic to design the whole chain of amplifiers from camera to radio-frequency transmitter on this basis, so advantage is taken of the existence in the transmitted waveforms of the synchronizing pulse and the black-level period following the synchronizing pulse and preceding the picture signal, line by line, to use D.C. reinsertion methods. These are dealt with under a separate section.

Further design features relating to this low-frequency response requirement are given under the heading "Direct-coupled Amplifiers".

Linearity

The modulation of picture signals from black to white must be transmitted faithfully to avoid the crushing of any part of the scale of brightness. In this country, if a test signal comprising a linear sawtooth extending across a line of picture with amplitude changing linearly from black to white is applied at the input terminals, the transmitted waveform is expected to maintain this linearity within 4 per cent. This degree of linearity usually means introducing some pre-distortion to counteract the curvature of the characteristics of the grid-modulated valve and any subsequent amplifiers, and with this predistortion adjusted, the radiated amplitude response can be linear within 1 per cent.

The design of such pre-distorting amplifiers is given under the title "Predistortion".

One further correction system is also incorporated to compensate for the curvature of the fully exploited radio-frequency modulated valves which tend to crush the amplitude of the synchronizing pulses. This correction takes the form of pulse stretching, and is also described in the section on predistortion and waveform correction.

Conventional Reactance Compensated R–C Amplifiers

The reactance compensation of *R–C* coupled amplifiers is now well established, and no attempt is made here to analyse the reactance compensation process. Results obtained by various workers in the field

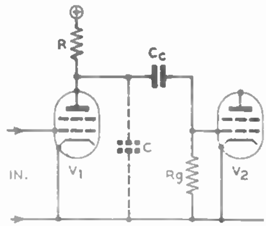


FIG. 17.—SIMPLE FORM OF TWO-TERMINAL COUPLING NETWORK.

are quoted, and where possible curves and tables are used to summarize the results.

In practice the application of these methods gives rise to some difficulty owing to the impossibility of assessing all the stray circuit and component capacitances with accuracy. With simple shunt compensation or with other two-terminal network-type compensation, much less work is entailed, as a simple R–C response establishes the total effective circuit capacitance, since the -3-db point in the amplitude response will occur at a frequency where $2\pi fCR = 1$. From this, the further elements of reactance in any of the two-terminal network-coupling systems can be assessed with some accuracy.

Broadly, reactance compensation methods fall into two categories, according to whether or not a significant reactance is included in the connection between the anode of the amplifier and the grid of the succeeding stage. If a direct connection or one with insignificant reactance is made, the system becomes equivalent to a two-terminal impedance network as the coupling impedance. With a significant reactance included in the connection the system becomes equivalent to a four-terminal coupling impedance.

Two-terminal Coupling Networks

The simplest of all two-terminal coupling networks is shown in Fig. 17, and is the R–C coupled amplifier in which the coupling impedance is a

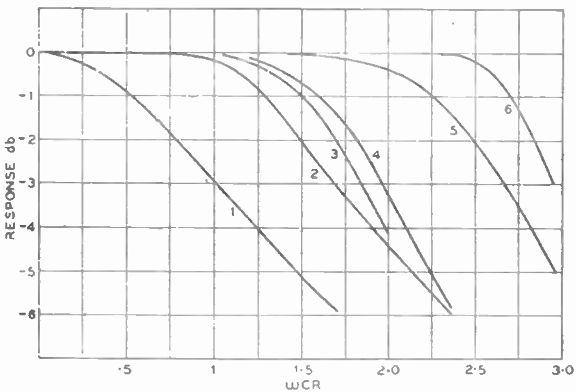


FIG. 18.—RESPONSE CURVES FOR: (1) RESISTANCE-CAPACITANCE CIRCUIT; (2) SIMPLE SHUNT COMPENSATION; (3) TUNED SHUNT COMPENSATION; (4) SERIES COMPENSATION; (5) SERIES SHUNT COMPENSATION; (6) TUNED SERIES/SHUNT COMPENSATION.

TABLE 3.—VIDEO AMPLIFIERS WITH TWO-TERMINAL COUPLING NETWORKS

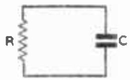
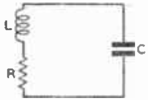
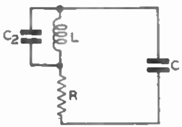
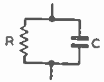
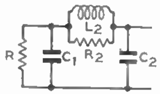
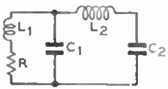
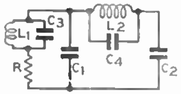
Circuits	*Approx. Band-width Improvements*				*Approx. Rise Time Improvements*	*Over-shoot in Indicial Response*	*L*	*C_2*	*Remarks*	*Fig. 18 Ref.*
	3 db	*2 db*	*1 db*	*½ db*						
	1·0	1·0	1·0	1·0	1·0	0	—	—	—	1
	1·5	1·6	1·7	1·8	1·5	0	$0{\cdot}3CR^2$	—	Critically damped condition	—
	1·7	1·9	2·2	2·7	1·8	2·5%	$0{\cdot}4CR^2$	—	Maximally flat condition	2
	1·8	2·1	2·7	3·7	2·0	6·5%	$0{\cdot}5CR^2$	—	3% rise in steady-state frequency response	—
	1·5	1·7	1·7	1·8	1·5	0	$0{\cdot}3CR^2$	$0{\cdot}125C$	Critically damped	—
	1·8	2·2	3·0	4·0	1·8	7%	$0{\cdot}4CR^2$	$0{\cdot}35C$	Maximally flat	3
	1·8	2·1	2·7	3·6	1·9	3·0%	$0{\cdot}43CR^2$	$0{\cdot}211C$	—	—

TABLE 4.—VIDEO AMPLIFIERS WITH FOUR-TERMINAL COUPLING NETWORKS

Circuits	*Approx. Band-width Improvements*				*Approx. Rise Time Improvements*	*Over-shoot in Indicial Response*	L_1	L_2	R_2	C_1	C_2	*Remarks*		*Fig. 18 Ref.*
	3 db	*2 db*	*1 db*	*½ db*										
	1	1	1	1	1	0	—	—	—	—	—	—	—	1
	1·5 2·0 1·3 1·4	1·6 2·4 1·5 1·5	1·7 3·1 1·7 1·8	1·7 4·5 2·0 2·1	1·6 1·9 1·4 1·4	0 9% 3% 3%	— — — —	$0{\cdot}375CR^2$ $0{\cdot}67CR^2$ $2{\cdot}51CR^2$ $0{\cdot}95CR^2$	— — $2{\cdot}88R$ $1{\cdot}88R$	$0{\cdot}111C$ $0{\cdot}25C$ $0{\cdot}8C$ $0{\cdot}5C$	$0{\cdot}888C$ $0{\cdot}75C$ $0{\cdot}2C$ $0{\cdot}5C$	Crit. damp. Max. flat — —		— 4 — —
	1·7 2·7 1·8	1·8 3·3 2·0	1·9 4·5 2·3	1·9 6·0 3·5	1·8 2·6 1·9	0 11% 3%	$0{\cdot}0625CR^2$ $0{\cdot}143CR^2$ $0{\cdot}125CR^2$	$0{\cdot}391CR^2$ $0{\cdot}583CR^2$ $0{\cdot}452CR^2$	— — —	$0{\cdot}2C$ $0{\cdot}404C$ $0{\cdot}2C$	$0{\cdot}8C$ $0{\cdot}596$ $0{\cdot}8C$	Crit. damp. Max. flat —		— 5 —
	2·4 2·9	2·7 3·5	3·2 4·9	3·9 7·9	2·5 2·8	0·3% 13%	$0{\cdot}133CR^2$ $0{\cdot}25CR^2$	$0{\cdot}466CR^2$ $0{\cdot}56CR^2$	— —	$0{\cdot}33C$ $0{\cdot}425C$	$0{\cdot}67C$ $0{\cdot}575C$	C_3 — $0{\cdot}307C$	C_4 — $0{\cdot}075C$	— 6

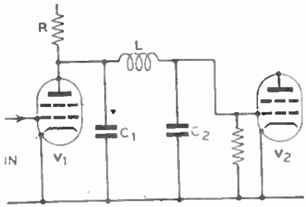


FIG. 19.—SIMPLE FORM OF FOUR-TERMINAL COUPLING NETWORK.

resistor R shunted by the capacitance C made up of the output capacitance of the amplifier valve V_1, the input capacitance of the valve V_2 and the stray circuit capacitances.

The grid leak R_g is an insignificant shunt on R, and the coupling capacitor C_c is an insignificant impedance at high frequencies. Basically then the two-terminal impedance is a resistor shunted by a capacitor. The form of response curve produced by this system is shown as curve 1 in Fig. 18.

It will be seen that when $\omega CR = 1$, i.e. when the reactance of C becomes numerically equal to the value of the resistance, the gain has fallen by 3 db.

The first stage of improvement can be obtained by including an inductance in series with the resistance and, by suitably adjusting its value, a smooth response curve extending beyond the basic R–C curve is possible.

This is shown as curve 2 in Fig. 18.

Further extended forms of two-terminal network are tabulated below in Table 3, with appropriate response curves given in Fig. 18.

Four-terminal Coupling Networks

The simplest four-terminal coupling network is the simple series compensation shown in Fig. 19. In this circuit the capacitances C_1 associated with the driving valve V_1 are separated from those C_2 associated with the driven valve V_2 by an inductance. The performance of this circuit depends very largely on the relative values of C_1 and C_2.

The flattest response is obtained when $C_2 = 3\times C_1$, in which case $L = 0{\cdot}67CR^2$. If, however, similar stages are to be cascaded, it is preferable to select values approaching those of the critically damped condition given by $C_2 = 8C_1$ and $L = 0{\cdot}375CR^2$.

Further more complicated types are tabulated in Table 4, and appropriate curves are given in Fig. 18. References 24–34 give detailed information of these compensated amplifiers.

FEEDBACK AMPLIFIERS

High-frequency Response

As is well known, when negative feedback is applied to an amplifier of two or more stages, the degree of feedback is severely limited unless peaks in the gain/frequency response curve are tolerable. This is due to the fact that as the natural response of the amplifier falls there is also a phase shift, so that the feedback voltage is out of phase with the input voltage and has therefore a smaller effective feedback value, giving rise to an initial increase of gain with increasing frequency. Staggered time constants in the coupling networks can be arranged to overcome this difficulty,[35] and further means have been described for shaping the response curve to give critical damping so that no overshoots occur when the amplifier is handling transients containing components of frequency well outside the passband.[36, 37, 38, 39]

It has also been shown [40] that by simple means it is possible to design a feedback amplifier with any desired shape of response curve. Thus:

(1) Aperiodic feedback produces a response curve of aperiodic flatness, and this result is characterized by the property that while the gain is changed by the feedback, there is no change of frequency response.

(2) Critical feedback produces a response curve of critical flatness and corresponds to the critically damped circuit, which has no overshoot in its indicial response.

(3) Maximal feedback produces a response curve of maximal flatness. This is the maximum extension of the amplitude response with frequency consistent with a smooth response curve without maxima or minima. It is characterized ideally by the formula

$$\text{Gain} = \frac{A}{\sqrt{1 + x^{2n}}}$$

the value of n being dependent on the number of circuits used.

(4) Optimal feedback produces optimal flatness, and is an extended form of response curve produced by frequency-conscious positive feedback.

Each of the above forms of feedback find application, the most popular at the present time being maximal feedback and optimal feedback.

Single-valve Amplifier

High-frequency Response

For a conventional high-impedance valve stage without feedback the gain may be expressed by the formula

$$A = \frac{g_m R}{1 + jx}$$

$$= \frac{A_0}{1 + jx}$$

where R is the anode load resistor;
$x = \omega CR$;
C is the total capacitance shunting the load;
$\omega = 2\pi f$;
g_m = mutual conductance of valve;
A_0 = gain at a frequency where reactances are insignificant.

If feedback is now applied between anode and grid the gain equation becomes

$$A_x = \frac{g_m R}{1 + jx + A_0\beta'} = \frac{A_0}{1 + jx + A_0\beta'}$$

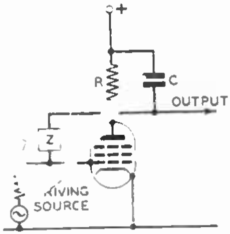


FIG. 20.—SINGLE STAGE WITH VOLTAGE FEED-BACK.

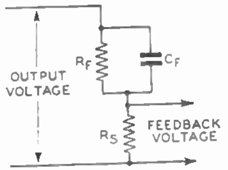


FIG. 21.—FEEDBACK NETWORK WITH CAPACITANCE.

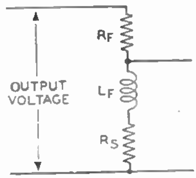


FIG. 22.—FEEDBACK NETWORK WITH INDUCTANCE.

where A_z is the "external gain", β' is the fractional feedback which may or may not be complex. This is illustrated in Fig. 20.

In order to achieve the various degrees of response-curve shaping defined above, the feedback fraction may be frequency and phase conscious, i.e. feedback is thus not necessarily applied by simple resistance networks but by impedance networks involving capacitance and/or inductance.

Thus the value β' is more generally expressed as $\delta + j\gamma$, where δ is the resistive part of the feedback and $j\gamma$ is the reactive part.

For example, if a feedback network is as Fig. 21, the feedback fraction is determined by the ratio

$$\frac{R_S}{R_S + (R_F \text{ in parallel with } C_F)}$$

and for cases where $R_S \ll R_F$ this becomes $\dfrac{R_S}{R_F}(1 + j\omega\, C_F R_F)$

i.e $\delta = \dfrac{R_S}{R_F}$ $\qquad \gamma = \omega C_F R_S$

Similarly for the network of Fig. 17, which is equivalent if $\dfrac{L_F}{R_S} = R_F C_F$ (Fig. 22). These two complex networks are used in single- and two-valve feedback amplifiers to produce the response shapes described above.

For the single-valve amplifier with voltage feedback we have

$$A_z = \frac{A_0}{1 + jx + A_0(\delta + j\gamma)}$$

Aperiodic Feedback

If we make $\gamma = \delta x$ we get

$$A_z = \frac{A_0}{1 + jx + A_0(\delta + j\delta x)}$$

$$= \frac{A_0}{1 + jx} \times \frac{1}{1 + A_0\delta}$$

from which it is seen that there is a gain change but no change of

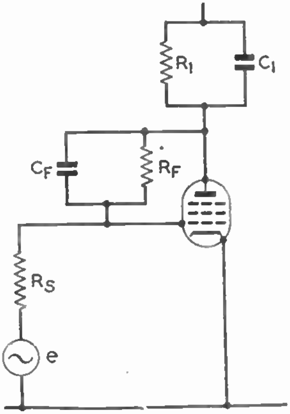


FIG. 23.—PRACTICAL FORM FOR APERIODIC FEEDBACK WHERE $R_S \ll R_F$.

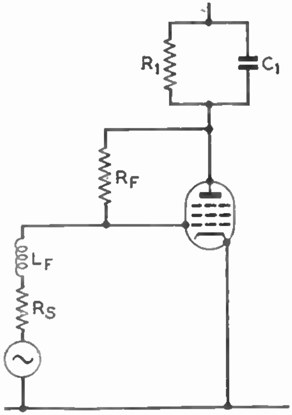


FIG. 24.—ALTERNATIVE FORM OF APERIODIC FEEDBACK WHERE $R_S \ll R_F$ AND L_F/R_S $R_F C_F$ (FIG. 23).

frequency response. In practice this is achieved as in Fig. 23 with the relations $R_F C_F = R_1 C_1$ or as Fig. 24 with $\dfrac{L_F}{R_S} = R_1 C_1$.

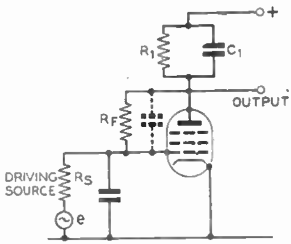
Critical Feedback

As before $A_x = \dfrac{A_0}{1 + jx + A_0(\partial + j\gamma)}$ and for this form the response must be extended by the feedback, but the response shape is unaltered. This is achieved by making $\gamma = 0$, in which case

$$\frac{A_0}{(1 + A_0\delta)\left(1 + \dfrac{jx}{1 + A_0\delta}\right)}$$

i.e. the gain and effective capacitance are reduced by the feedback factor $1 + A_0\delta$, and with $\gamma = 0$ the feedback is resistive. In practice this is achieved as in Fig. 25, in which the dotted capacitances indicate how

FIG. 25.—PRACTICAL FORM OF MAXIMAL OR CRITICAL FEEDBACK AMPLIFIER.



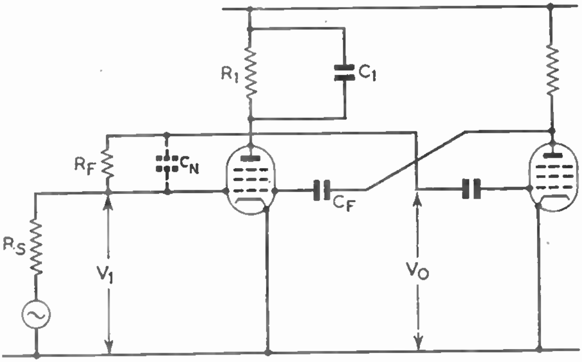


FIG. 26 (*a*).—CIRCUIT GIVING "OPTIMAL FEEDBACK".

the feedback fraction may be maintained constant at high frequency where the valve-input capacitance becomes significant.

Maximal Feedback

Again $$A_x = \frac{A_0}{1 + jx + A_0(\delta + j\gamma)}$$

rearranging we get

$$\frac{A_0}{A_x} = 1 + jx + A_0\delta + A_0 j\gamma$$

$$\left|\frac{A_0}{A_x}\right|^2 = (1 + A_0\delta)^2 + (x + A_0\gamma)^2$$

$$= (1 + A_0\delta)^2 + x^2 + 2A_0\gamma x + (A_0\gamma)^2$$

For maximal flatness, by definition, the terms

$$A_0\gamma x\, 2 + (A_0\gamma)^2 = 0$$

$$\therefore \quad \gamma = 0$$

and thus maximal flatness is the same as critical flatness for the single-valve amplifier.

Optimal Feedback

If again we take the external gain relation

$$A_x = \frac{A_0}{1 + jx + A_0\beta'}$$

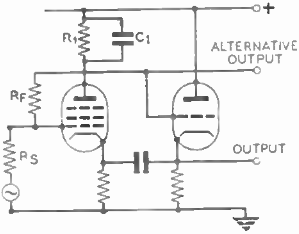


FIG. 26 (*b*).—ALTERNATIVE "OPTIMAL FEEDBACK" CIRCUIT.

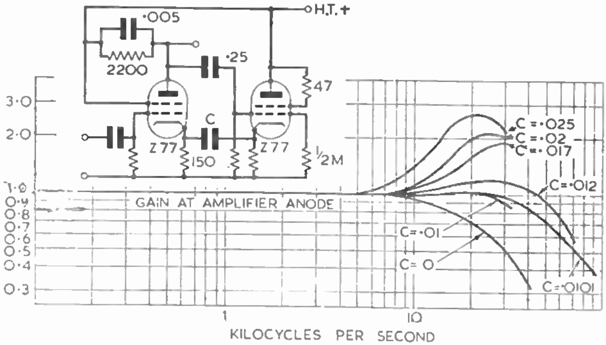


Fig. 27.—Experimental Form of "Optimal Feedback" Circuit.

we see that if $A_0\beta'$ is made equal to $-jx$ the frequency-conscious terms disappear, and we should get infinite frequency response.

This relation, however, of $A_0\beta' \doteq -jx$ involves including a negative capacitance in the feedback path.

This may be simulated by the circuits of Fig. 26. In Fig. 26 (a) the additional valve and couplings are equivalent over a restricted frequency range to a negative capacitance C_N shown dotted. Fig. 27 shows the profit that may be obtained in practice by this means in any case where an amplifier is followed by a cathode follower.

Single-valve Amplifiers with Cathode Degeneration

It has been shown [40] that, by suitable choice of the cathode impedance, the various categories of response curve shaping by feedback may also be obtained. Thus in Fig. 28 we get aperiodic feedback when Z_K is a resistance, critical feedback and maximal feedback when Z_K is a resistance and a capacitance in parallel with $R_K C_K = RC$.

Two-stage Amplifiers

The same approach is made to the design of two-stage amplifiers, and the circuit relations are similarly derived.

For a two-stage amplifier the gain equation is

$$A_x = \frac{A_0}{(1 + jx_1)(1 + jx_2) + A_0\beta'}$$

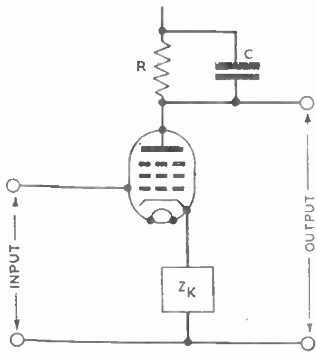


Fig. 28.—Single Stage with Cathode Degeneration.

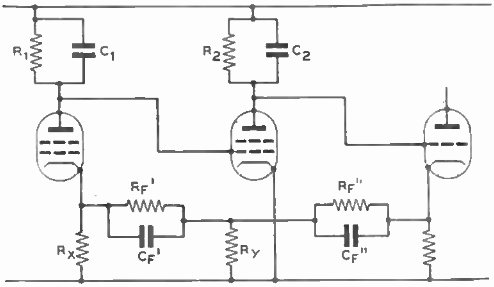


FIG. 29.—TWO-STAGE SIMPLIFIED AMPLIFIER WITH "APERIODIC FEEDBACK".

$R_x \ll R_F'$, R_F''. $R_y \ll R_F' . R_F''$

Aperiodic Feedback

For aperiodic flatness $A_0\beta'$ must be $(K - 1)(1 + jx_1)(1 + jx_2)$ and feedback must therefore be in two stages as in Fig. 29.

Critical Feedback

For critical feedback

$$A_0\beta' = A_0\delta + j[2\sqrt{x_1x_2N} - (x_1 + x_2)]$$

where N is the gain-reduction factor $(1 + A_0\delta)$ at low frequencies.

This is achieved by the circuit of Fig. 30.

Maximal Feedback

$A_0\beta' = A_0\delta + j[\sqrt{2x_1x_2N} - (x_1 + x_2)]$ and is again achieved by the circuit of Fig. 30 with a different value of feedback capacitance.

Using the simple feedback networks already illustrated and shown in Fig. 25 gives a good approximation to these forms of $A_0\beta'$.

Thus from the simple capacitance feedback network

$$\beta' = \frac{R_S}{R_F}(1 + j\omega C_F R_F)$$

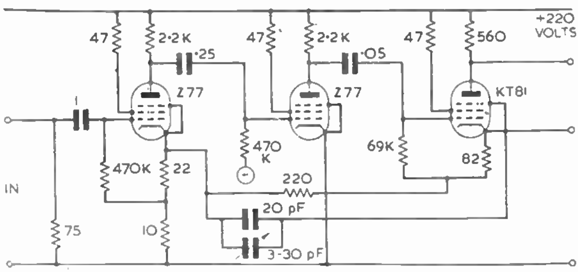


FIG. 30.—EXPERIMENTAL FORM OF TWO-STAGE MAXIMAL AND CRITICAL FEEDBACK AMPLIFIER.

and we require for maximal flatness

$$A_0\beta' = A_0\delta + j[\sqrt{2(A_0\delta + 1)x_1x_2} - (x_1 + x_2)]$$

so
$$\delta = \frac{R_S}{R_F}$$

and
$$j\omega C_F R_S = j\omega\left[\frac{\sqrt{2(A_0\delta + 1)R_1C_1R_2C_2} - [R_1C_1 + R_2C_2]}{A_0}\right]$$

and where $A_0\delta$ is large

$$C_F R_S = \frac{\sqrt{2A_0\delta R_1C_1R_2C_2}}{A_0} \text{ for maximal feedback}$$

and similarly

$$C_F R_S = \frac{\sqrt{A_0\delta R_1C_1R_2C_2}}{A} \text{ for critical feedback}$$

For the special case of $R_1C_1 = R_2C_2 = RC$ these relations simplify

$$C_F R_S = RC\sqrt{\frac{2\delta}{A_0}} \text{ for maximal flatness}$$

$$C_F R_S = RC\sqrt{\frac{4\delta}{A_0}} \text{ for critical flatness}$$

In practice the capacitance C_F is made variable to absorb tolerances of components and approximation in design.

Optimal Feedback

Optimal feedback across two stages has not so far been achieved, but interesting possibilities are illustrated by Figs. 31 and 32, which show

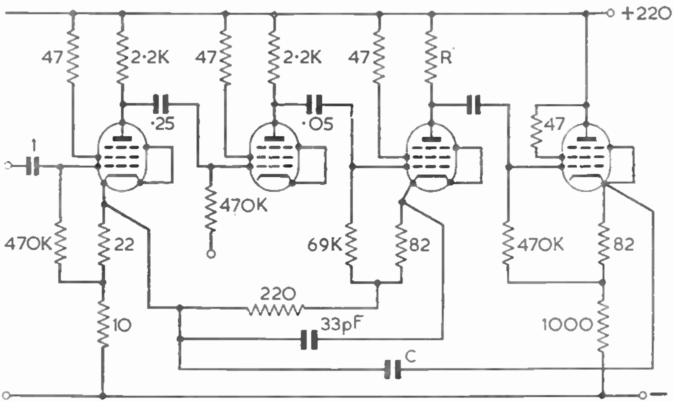


Fig. 31 (a).—Two-stage Maximal Feedback Amplifier Followed by One Stage with Overall Critical Feedback.

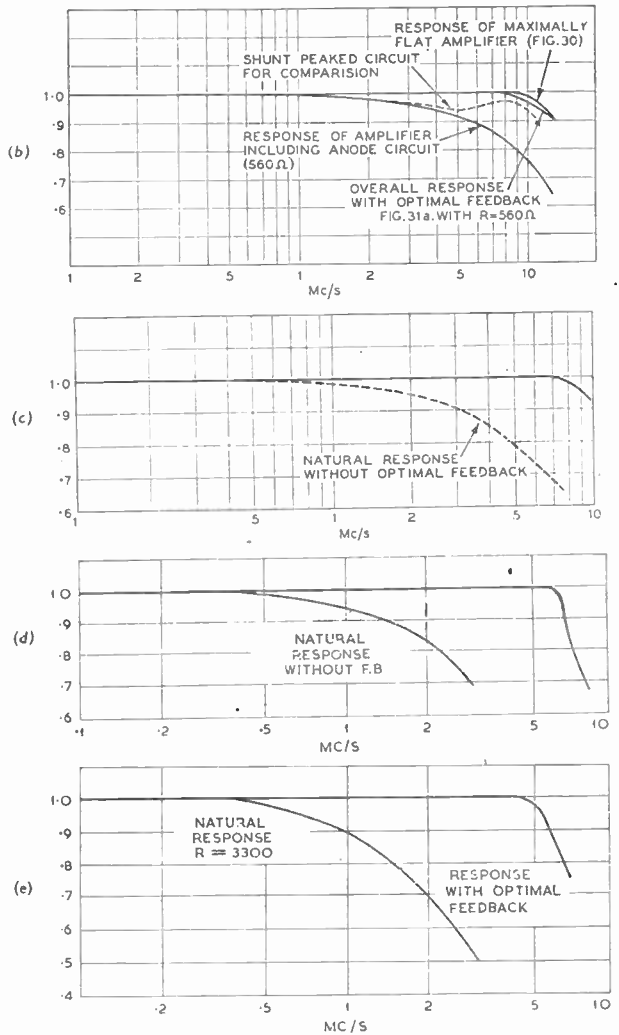


FIG. 31.—(*b*) RESPONSE CURVES FOR THE CIRCUIT GIVEN IN FIG. 31 (*a*). (*c*) RESPONSE WITH $R = 1{,}000$ OHMS, i.e., A GAIN INCREASE OF 1·8. (*d*) $R = 2{,}000$ OHMS, i.e., GAIN INCREASED BY 3·6. (*e*) $R = 3{,}300$ OHMS, i.e., GAIN INCREASED BY 5·9.

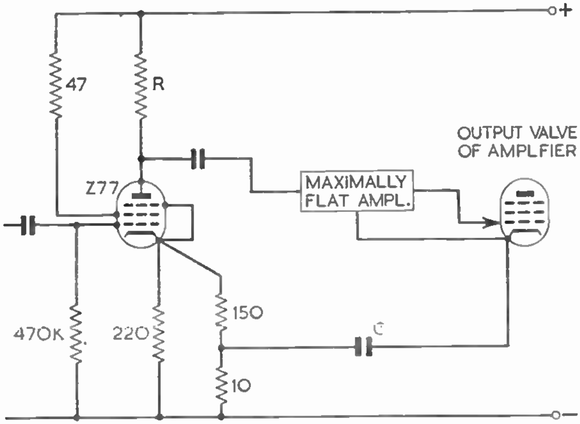


FIG. 32.—ALTERNATIVE ARRANGEMENT FOR SUPPLYING OVERALL OPTIMAL FEEDBACK.

single stages having optimal feedback applied via preceding or succeeding amplifier stages.

The application of this overall optimal feedback shows that, given an amplifier of wide frequency response flat to frequency f_1 and of gain A_1, associated with a poor response amplifier of gain A_2 and response f_2, then in combination the gain band-width may be approximately

$$A_1{}^2A_2f_2 \text{ (provided } A_1f_2 < f_1)$$

or

$$A_1A_2f \quad (\text{if } A_1f_2 > f_1)$$

A note of warning is necessary in the design of feedback amplifiers. Since the natural response is falling much more quickly than the overall effective response, certain valves in the loop will be handling much higher amplitudes of the higher-frequency components than for the amplifier without feedback. This condition will demand much higher reactive current for the valve or valves in question, and must be allowed for in the design, if the profit theoretically possible is to be achieved.

Staggered Time Constants

It has been shown [39] that in a two-stage circuit, maximal flatness is also achieved by making the feedback a simple resistance network and staggering the time constants of the amplifier.

In this solution the feedback factor $1 + A_0\beta$ is made equal to

$$\frac{\left(b + \frac{1}{b}\right)^2}{2}$$

where $$b = \sqrt{\frac{R_1C_1}{R_2C_2}} \text{ or } \sqrt{\frac{R_2C_2}{R_1C_1}}$$

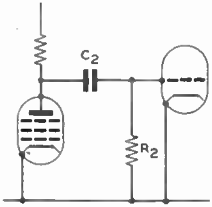


FIG. 33.—TYPICAL LOW-FREQUENCY RESPONSE-DETERMINING COUPLING ELEMENTS.

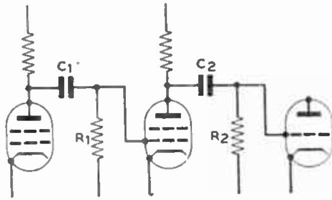


FIG. 34.—TWO-STAGE CIRCUIT WITH STAGGERED TIME CONSTANTS.

Low-frequency Response

All the above findings apply if we substitute y for x

where $$y \doteq -\frac{1}{C_2R_2}$$

where C_2 is the coupling capacitance and R_2 the grid resistor on the succeeding valve (Fig. 33).

In this case, however, the reciprocal feedback networks are not easily found, and it is usually preferable to adopt the staggered-time-constant method (Fig. 34) in order to get the response flat. In this case the feedback network is a simple resistance potentiometer, i.e. aperiodic, and the condition for maximal flatness becomes

$$y^2{}_1 + 2y_1y_2 + y^2{}_1 + 2y_1y_2 + y^2{}_2y_1 + y_2 + \sqrt{2Ny_1y_2}$$

with $$\frac{1}{y_1} = \omega C_1R_1$$

$$\frac{1}{y_2} = \omega C_2R_2$$

$$N = \frac{\left(b + \frac{1}{b}\right)^2}{2} = \text{the feedback factor with } b^2 = \frac{y_1}{y_2} = \frac{C_2R_2}{C_1R_1}$$

i.e., for feedback giving a gain reduction of 10/1

$$\frac{\left(b + \frac{1}{b}\right)}{2} = 10$$

$$b \doteq 4$$

SHUNT-REGULATED AMPLIFIERS

Simple Amplifier Type

This form of amplifier (or cathode follower) was developed to overcome some of the difficulties encountered, particularly at high level, in television modulator applications.[41]

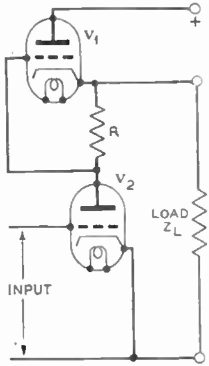


FIG. 35.—SIMPLE SHUNT-REGULATED AMPLIFIER.

Output impedance:

$$Z_0 = \frac{r_{a_1}(r_{a_2} + R)}{r_{a_2} + R(\mu_1 + 1) + r_{a_1}}$$

Gain:

$$\frac{V\ out}{V\ in} = \frac{\mu_2(r_{a_1} + \mu_1 R)}{r_{a_1} + r_{a_2} + (\mu_1 + 1)R}$$

It frequently arises that a voltage swing required from an amplifier causes distortion due to the non-linear characteristics of the valve towards cut-off. Operating the valve at higher H.T. supply voltage and/or with higher anode load resistance improves the condition, since it is then unnecessary to drive the valve towards low current to the same extent for the same anode voltage excursion. In Class A video-frequency amplifiers these are undesirable features from the point of view of economy and performance. Shunt-regulated amplifier technique provides means of simulating the desired conditions of high anode load resistance and high H.T. supply voltage while still maintaining a moderate H.T. supply voltage and providing improved frequency response, higher reactance-current handling capability, lower output impedance and higher conversion efficiency for a given performance.

Basically the circuit is shown in Fig. 35.

Valves V_1 and V_2 are in series as shown, with a resistor R connected between them. The grid of V_1 is returned to anode of V_2.

Signals are applied at the grid of V_2, and output is taken from the cathode of V_1. The connections shown neglect bias arrangements and show only signal connections.

Operation can be explained most simply by reference to the characteristics of the valves, which for simplicity are taken to be similar triodes. These are shown in Fig. 36.

Assuming a given instantaneous current flowing through the system with no load, and a given H.T. supply voltage, the grid-cathode excitation for V_1 is derived by taking the product of the current and the resistance R. This settles the voltage drop across V_1 and therefore the voltage across V_2. Current and voltage relations in V_2 are then known, and the voltage excursions for the two valves can be drawn.

This is demonstrated below, and curves are drawn in Fig. 36.

For this illustration large 2-kW dissipation triodes, type ACM3, are used, as these are in current use in 50-kW transmitter television stations in this country.

Assume H.T. voltage 2,500

take R = 100 ohms

then $V_2 = 2{,}500 - iR - V_1$

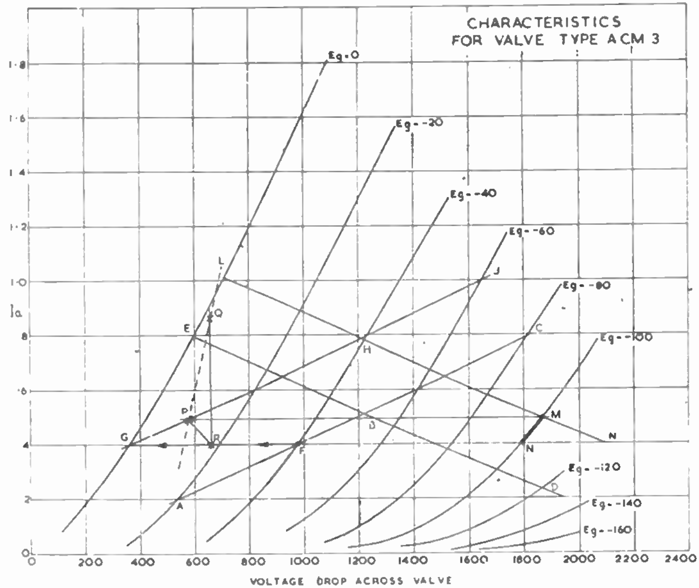


FIG. 36.—VOLTAGE EXCURSIONS FOR A SHUNT-REGULATED AMPLIFIER.

So by taking a range of values for i the two voltage excursions on the two valves can be plotted, viz. ABC for valve 1 and DBE for valve 2. It may be seen by inspection of Fig. 36 that so far the arrangement has replaced the conventional anode load resistor by a valve, and no improvement in linearity is apparent. It may be seen that the equivalent load presented by the valve is higher than its anode impedance, owing to the negative feedback due to R and is actually

$$r_a + (\mu + 1)R$$

If now the lower limit of the current excursion is decided—say point F on curve ABC, this value of current can be made to coincide for Class A operation with the $E_g = 0$ line of the characteristics of valve 1 (point G) by suitably biasing the valve. This transposes both excursion lines to the positions GHJ and KHL shown in Fig. 36, and is approximately equivalent to increasing the H.T. voltage by μ times the bias applied to effect the transposition.

In this condition, while the D.C. resistance of the valve and resistors is almost the same as the resistance of the valve itself, the A.C. impedance $r_a + (\mu + 1)R$ of the combination is very much higher and enables the greater linear voltage swing to be obtained. Further advantage accrues if instead of the arbitrary zero loading condition so far used to demonstrate the operation, a capacitative load is assumed.

The mean current through the two valves is higher for a given H.T. voltage. As the frequency of the signal is raised, an increasing demand of reactive current is made by the capacity load. This demand is met, for rising voltages, by an increase of current in V_1 and a decreasing current in V_2, i.e., the total reactive current available is the sum of the currents in both valves, i.e., twice the mean current.

The properties of this form of amplifier are summarized below :

$$\text{Output Impedance } Z_0 = \frac{r_{a1}(r_{a2} + R)}{r_{a1} + r_{a2} + (\mu_1 + 1)R}$$

Gain

$$\frac{V_0}{V_i} = -\frac{\mu_2(r_{a1} + \mu_1 R)}{r_{a1} + r_{a2} + (\mu_1 + 1)R}$$

For the ACM3 valves taken to illustrate the operation, appropriate values are :

$$\mu_1 = \mu_2 = 15$$
$$r_{a1} = r_{a2} = 600$$
$$\text{for } R = 100$$

$$\text{Output Impedance} = 120 \text{ ohms}$$

$$\text{Gain } -11{\cdot}25 \text{ (i.e. } \tfrac{3}{4}\mu)$$

The low-output impedance may be shown graphically, as in Fig. 36, by assuming a load change which causes a fall in current in valve 2 from

TABLE 5.—COMPARISON OF CONVENTIONAL AND SHUNT-REGULATED AMPLIFIERS

	Conventional	*Shunt regulated*
ACM3 Valves ($\mu = 15$; $r_a = 600$)		
Number of valves	2 in parallel	2 in series
Gain	11	11
Output voltage	1,000	1,000
Resistor	825 ohms	100 ohms
Watts in resistor	2·1 kW	64 W
Input capacitance	400 pF	200 pF
Departure from linearity in response to a saw-tooth waveform	Approximately 6%	Approximately 1%
H.T. required	2,900 V	2,500 V
Mean current from H.T. supply	1·6 A	0·8 A
H.T. power input	4,640 W	2,000 W
Output impedance	228 ohms	150 ohms
Current fluctuation in H.T. supply for 1,000 volts swing	1·2 A	0·6 A

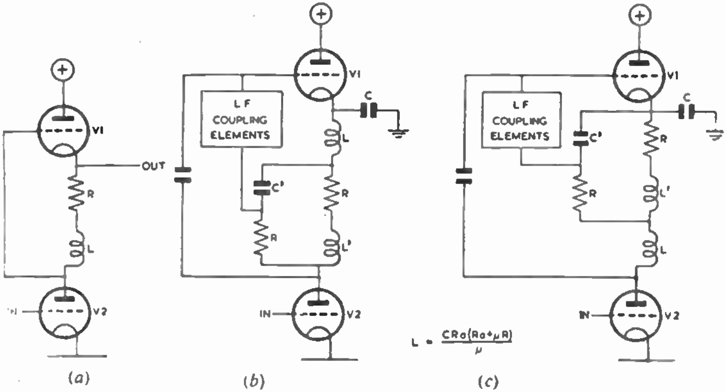


FIG. 37.—FREQUENCY COMPENSATION OF SHUNT-REGULATED AMPLIFIERS.

In (*a*), the load capacitance C normally appears between the output terminal and earth.

(*Courtesy of the I.E.E.; reference* 8.)

M to N. There is an increase in the anode voltage of V_2 and an increase in the current in V_1, due to the reduction of iR. The construction PRQ satisfies these changes and shows that the equivalent impedance is represented by the slope PQ, which is a much lower resistance than the characteristic resistance of the valve.

A comparison of the performance of a shunt-regulated amplifier and a conventional amplifier with the same gain and reactive current-handling capability is shown in Table 5, and illustrates the advantages obtained.

High-frequency Response

Reactive compensation to improve frequency response may be included in a variety of ways.

One particularly interesting property is that if an inductance L is included with the control resistor R as in Fig. 37, of value determined by the relation

$$L = \frac{Cr_a(r_a + \mu R)}{\mu}$$

Then the impedance presented to the anode of V_2 is

$$R(\mu + 1) + r_a + j\omega L$$

and for all practicable purposes removes all the load capacitance C from the amplifier V_2.

Various stages of improvement are shown in the accompanying diagrams. So far, no rigid analysis has been completed for these types

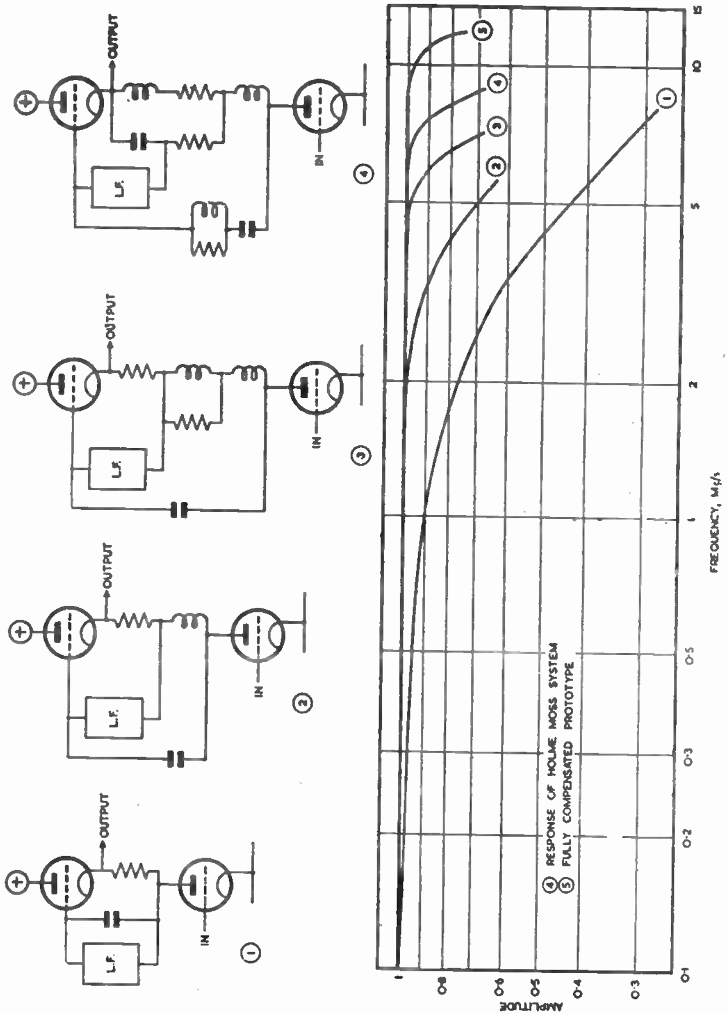


FIG. 38.—MEASURED RESPONSE CURVES OF A SIMPLE SHUNT-REGULATED AMPLIFIER WITH PROGRESSIVE COMPENSATION (FOR A GAIN OF APPROXIMATELY 10).

(*Courtesy of the I.E.E.; reference* 8.)

of compensation, and results are usually obtained by cut-and-try methods.

Several useful steps in an approach to good frequency compensation are, however, known, and are illustrated below.

In practice the bias arrangements and any additional compensation reactors add additional stray capacitance to the circuit, and the heater transformer for the valve V_1 introduces a further capacitance. The most critical part of the circuit from the point of view of frequency response is the anode of V_2, and it is necessary to remove as much of the reactance loadings as possible from this point.

The effect of the D.C. and low-frequency coupling system is removed by including the stray reactances in a constant-impedance network. This is shown in Fig. 37. The value of C' is chosen to be large compared with the reactances involved, and L' is then governed by the relation $R = \sqrt{(L'/C')}$.

In practice, a further improvement is obtained by rearranging the elements so far used in the form of Fig. 37, in order to remove as much as possible of the stray capacitance, associated with the constant-impedance network, from the valve V_2.

The high-frequency connection between V_1 and V_2 is made direct with a capacitor as shown also in Fig. 37. In this path is included a damped series inductance.

The progressive improvement with these compensating elements is seen in the curve of Fig. 38.

The use of shunt-regulated amplifiers is a relatively new technique, and, at the time of writing, there is little published data on the precise reactance relations to achieve the optimum frequency response.

As a guide, however, the practical amplifier of Fig. 39 produces the results indicated.

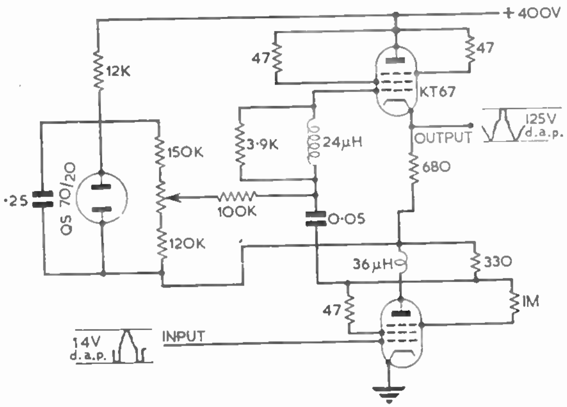


FIG. 39.—PRACTICAL SHUNT-REGULATED AMPLIFIER.

This has a frequency-response flat within ½ db to 7 Mc/s.

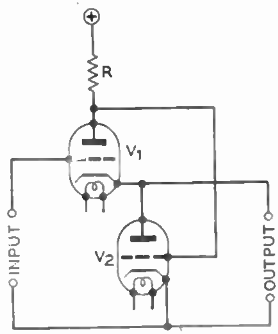


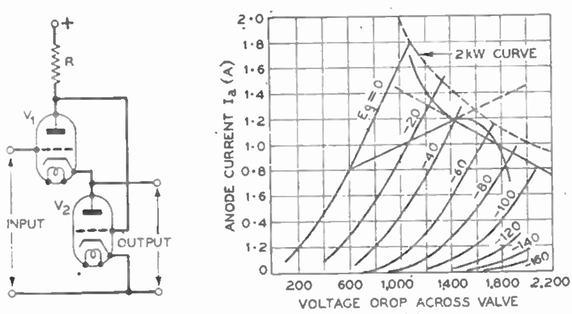
Fig. 40.—Simple Shunt-regulated Cathode Follower.

Simple Cathode-follower Type

With conventional cathode followers the upper frequency limit for a given voltage swing is usually determined by current cut-off during the negative-going voltage excursion due to capacitance in shunt with the load.

The simple type of shunt-regulated cathode follower minimizes this effect and offers considerable improvements in operation.

A particular case arises in the output stage of a television modulator feeding into a grid-modulated radio-frequency amplifier. Usually the load presented to the modulator is both reactive (due to circuit and valve capacitances) and resistive due to transmitter grid current. The orthodox solution to this problem is to use a conventional cathode follower with current-handling capacity adequate to supply the reactive current and the peak resistive current, and where conditions are particularly unfavourable it is also necessary to provide a current stabilizer to offset the effects of the non-linear grid current.



(a) (b)

Fig. 41.—(a) Simple Shunt-regulated Cathode Follower.

Output Impedance:

$$Z_o = \frac{(r_{a1} + R)r_{a2}}{(\mu_1 + 1)r_{a2} + r_{a1} + [(\mu_1 + 1)\mu_2 + 1]R}$$

Gain (no load):

$$\frac{V_o}{V_i} = \frac{\mu_1(r_{a2} + \mu_2 R)}{(\mu_1 + 1)r_{a2} + r_{a1} + [(\mu_1 + 1)\mu_2 + 1]R}$$

(b) Characteristics of ACM3 Valve with Curved Load Lines Due to Grid Current in the Following Stage; Shunt-regulated Cathode Follower.

Using a shunt-regulated cathode follower, V_2, Fig. 40, acts partially as a current stabilizing device, so that the amount of stabilization left to be done is reduced, and the mean current required from the H.T. source is approximately halved.

The voltage excursion curves for ACM3 valves used as a simple shunt-regulated cathode follower are shown in Fig. 41, and are derived in a manner similar to that described earlier for the amplifier-type arrangement.

The operating properties are as follows:

$$\text{Output Impedance } Z_0 = \frac{r_{a2}(r_{a1} + R)}{r_{a1} + R + (\mu_1 + 1)(r_{a2} + \mu_2 R)}$$

$$\text{Gain (Open-circuit) } \frac{V_0}{V_1} = \frac{\mu_1(r_{a2} + \mu_2 R)}{r_{a1} + R + (\mu_1 + 1)(r_{a2} + \mu_2 R)}.$$

which for the ACM3 valves illustrated with

$\mu = 15$ $\quad r_a = 600$ ohms $\quad$ R = 100 ohms

Output Impedance Z_0 = 12 ohms

Gain = 0·92

A comparison of the performance of a single shunt-regulated cathode follower with a conventional cathode follower for the same reactive current handling capability at the same voltage swing is given in Table 6.

TABLE 6.—COMPARISON OF CONVENTIONAL WITH SHUNT-REGULATED CATHODE FOLLOWER

	Conventional	*Shunt regulated*
ACM3 valves ($\mu = 15$; $r_a = 600$)		
Number of valves	2 in parallel	2 in series
Output voltage	1,000	1,000
Gain	0·92	0·92
Resistor	875 ohms	100 ohms
Watts dissipated in resistor	5 kW	140 W
Output impedance	18 ohms	12 ohms
Mean current	2·4 A	1·2 A
H.T. required	3,480 V	3,000 V
H.T. power required	8·3 kW	3·6 kW

Other Forms of Shunt-regulated Amplifiers

More complicated forms exist [41] and offer further advantages. These are illustrated in Figs. 42 and 43.

In these forms, zero output impedance and infinite input impedance are possible, and they also possess a form of impedance inversion property by virtue of which a load capacitance is reflected into the driving circuit as a negative capacitance. This is demonstrated by the

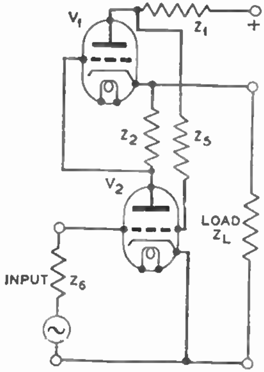


FIG. 42.—DOUBLE-SHUNT-REGULATED AMPLIFIER.

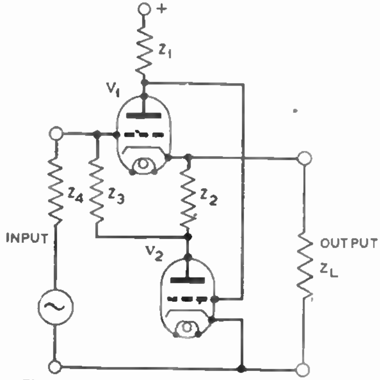


FIG. 43.—DOUBLE-SHUNT-REGULATED CATHODE FOLLOWER.

(*Formulæ for the circuits in Figs. 42 and 43 are given on page 10-66.*)

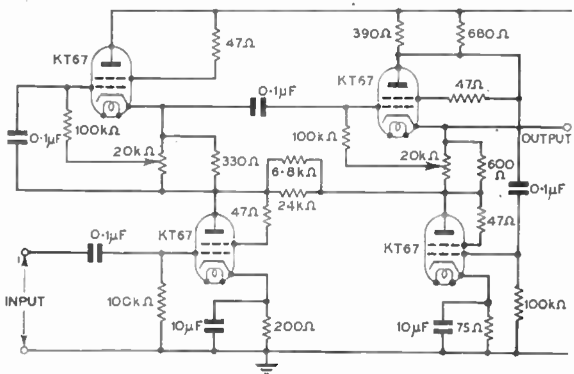


FIG. 44.—EXPERIMENTAL SHUNT-REGULATED AMPLIFIER.

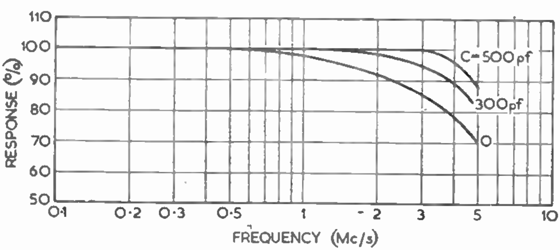


FIG. 45.—MEASURED RESPONSE CURVES OF CIRCUIT IN FIG. 44.

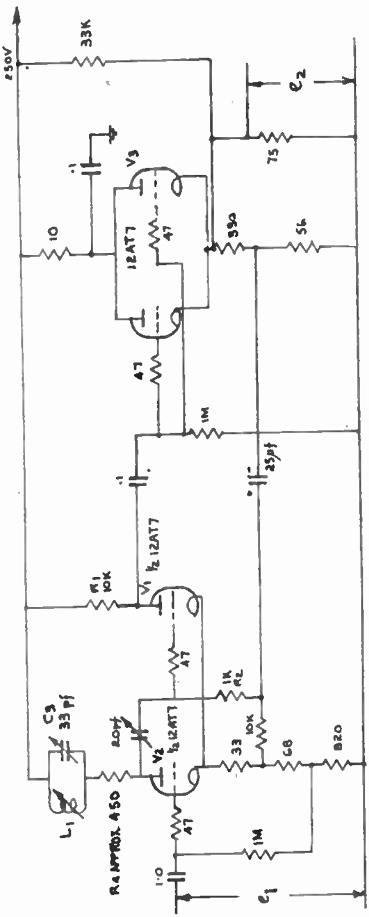


FIG. 46.—VIDEO-FREQUENCY AMPLIFIER WITH REACTANCE CONPENSATION AND FEEDBACK.

$G = e_2/e_1 = 3{\cdot}3$

Response level to 16 Mc/s

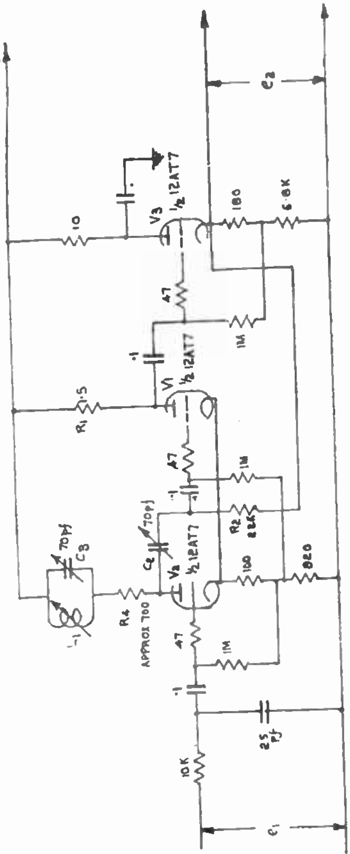


FIG. 47.—VIDEO AMPLIFIER WITH REACTANCE COMPENSATION AND FEEDBACK. IN THIS ARRANGEMENT THE AMPLIFIER CHARACTERISTIC COMPENSATES FOR THE INPUT TIME CONSTANT 10k, 25pF, AND IS INTENDED FOR "HIGH PEAKING" APPLICATIONS.

Response e_1 to e_2 level to 12 Mc/s.

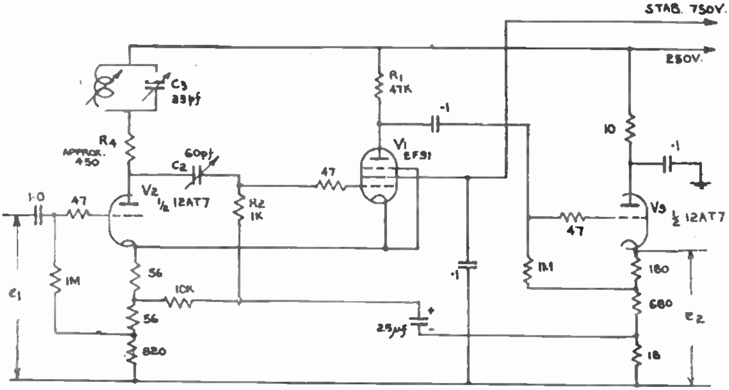


FIG. 48.—VIDEO AMPLIFIER WITH REACTANCE COMPENSATION AND FEEDBACK CIRCUITS.

$G = e_2/e_1 = 30$. Response level to 6 Mc/s: − 1 db, 8 Mc/s; − 3 db, 13 Mc/s.

practical circuit of Fig. 44, for which response curves are given in Fig. 45, which show that frequency response improves as capacitance loading is increased, due to the fact that in the circuit shown the output pair have a much better frequency response than the driving pair. As long as the output pair will deal adequately with the load as the capacitance is increased, then more and more negative capacitance is fed into the driving-valve coupling impedance, thus improving the performance of the amplifier.

This feature is, of course, as always, strictly dependent on all the valves being able to deal with the reactive current demands.

Hybrid Types

These may take the form of reactance-compensated amplifiers which also include feedback or any other combination of the techniques already described. Some interesting examples are shown in Figs. 46, 47 and 48, which also indicate performance.

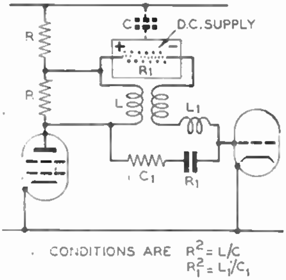
These amplifiers are due to E. Davies and are in current use.

DIRECT-COUPLED AMPLIFIERS

It is frequently desirable at high level, particularly, to include direct coupling in the video-frequency amplifier system. This arises from the difficulties of clamping at high voltage levels as against the cost of direct coupling.

Earlier designs of direct-coupled amplifiers used some form of D.C. supply included in a constant-impedance network " hold off " system, in order to couple anodes at high potential to grids at low potential.[12]

The method used is illustrated in Fig. 49, and shows that the large



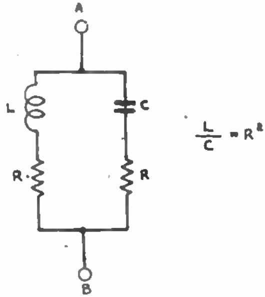


Fig. 49 (*top left*).—D.C. Supply Connected to Form Part of Constant-impedance Network for a D.C. Amplifier.

Fig. 50 (*top right*).—Constant-impedance Network.

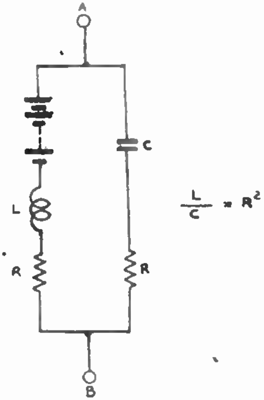
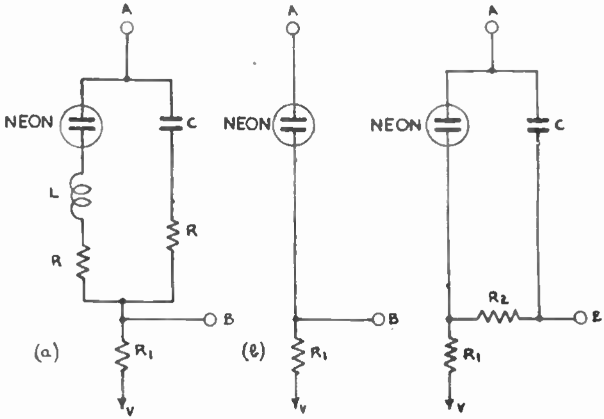
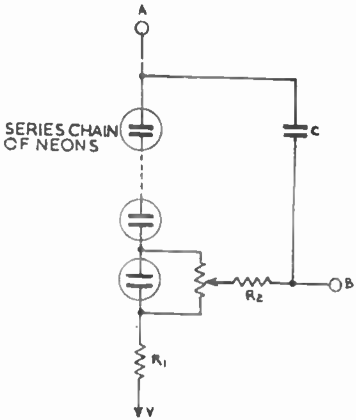


Fig. 51 (*left*).—Constant-impedance Network Including a Battery.

Fig. 52 (*below*).—Neon Coupling Networks.





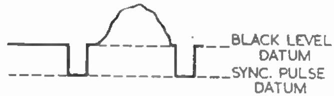


Fig. 54.—Datum Levels of the Television Waveform.

Fig. 53 (*left*).—Neon Coupling Network with Voltage Adjustment.

capacitance of the supply can be built out as a constant impedance R if sufficient care is taken.

More recently the neon coupler has been used successfully, and results in considerable simplification.[8]

Neon couplers are employed to connect two points between which it is necessary to establish a fixed D.C. potential, and are derived from the constant-impedance coupling network shown in Fig. 50. In this network, provided $L/C = R^2$, then the impedance through the L–R–C network is equal to R at all frequencies. If a battery is included in the low-frequency branch of the network, Fig. 51, the conditions required for a coupler with a potential difference are met, provided the inclusion of the " battery " does not add significant capacitance to the system.

In practice, the " battery " is replaced by one or more neon tubes, the internal resistance and inductance of the neon tube being absorbed into the designed values for the low-frequency branch of the network.

In designing these coupling networks care must be taken to ensure that over the signal excursion the neon current remains within the minimum and maximum values permitted.

This form of coupling has been used for potential differences from 70 volts to over 2,000 volts with satisfactory results.

A true constant-impedance neon coupler and simplified arrangements are shown in Figs. 52 and 53.

D.C. RESTORATION AND CLAMPING

The television waveform has two datum levels, one at the extreme of the synchronizing pulses and the other at " black level " or " pedestal level " as illustrated in Fig. 54.

These levels should be transmitted with their D.C. potential always in the same relation.

The effect of R–C couplings in the video-frequency amplifier system is to remove all the very low-frequency components and so disturb these datum levels, and the form of distortion that occurs is illustrated

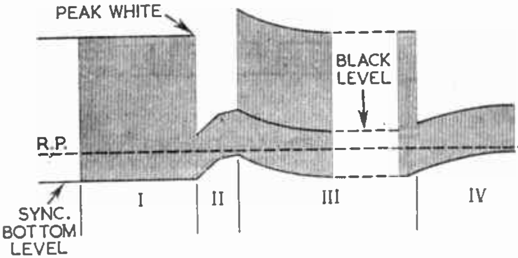


FIG. 55.—DISTORTION OF TELEVISION WAVEFORM.

in Fig. 55. Referring to Fig. 55, periods I and III represent a peak white signal immediately before and after a frame-suppression period II, while period IV represents a period of "black" with synchronizing pulses occurring by a sudden transition from the condition of period I.

It is necessary to re-establish these levels before transmission, and this is done by a process commonly called D.C. re-insertion.

The process may be accomplished in one of three ways :

(1) D.C. restoration ;
(2) synchronizing pulse peak level clamping ;
(3) black-level clamping.

D.C. Restoration

This is the simplest of the various methods, and is illustrated in Figs. 56 (*a*) and 56 (*b*). The diode D is arranged to conduct on the peak of the synchronizing pulses and restore the charge on the capacitor C to the reference potential RP, to which the diode is connected.

In between the periods of diode conduction the capacitor C discharges through the leak resistor R.

Owing to the finite resistance of the diode during conduction, which

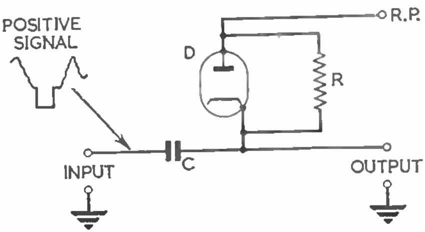


FIG. 56 (*a*).—SIMPLE D.C. RESTORATION CIRCUIT FOR POSITIVE PICTURE SIGNALS.

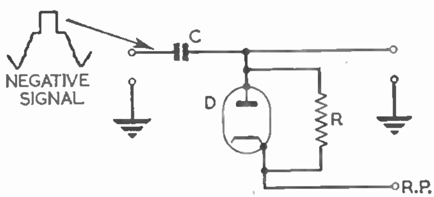


FIG. 56 (*b*).—SIMPLE D.C. RESTORATION CIRCUIT FOR NEGATIVE PICTURE SIGNALS.

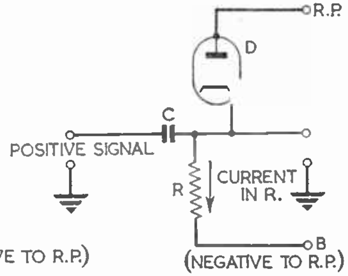


FIG. 57 (*a*).—(*right*) IMPROVED D.C. RESTORATION CIRCUIT FOR POSITIVE PICTURE SIGNALS.

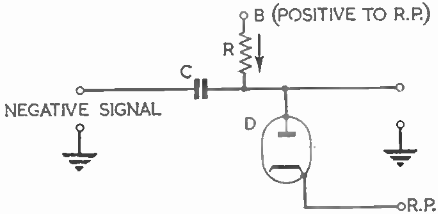


FIG. 57 (*b*).—(*left*) IMPROVED D.C. RESTORATION CIRCUIT FOR NEGATIVE PICTURE SIGNALS.

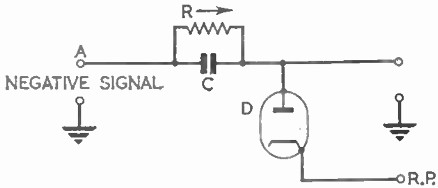


FIG. 57 (*c*).—(*right*) ALTERNATIVE TO CIRCUIT IN FIG. 57 (*b*).

This circuit is only suitable when A is positive to R.P.

can be high for small anode–cathode voltage, the capacitor potential is not restored absolutely to the reference potential, and so a finite marginal voltage is required above the reference potential at each synchronizing pulse in order that the system shall function. This marginal voltage is reduced by making R large. The magnitude of R is limited by the reverse grid current of the valve shunting R. Alternatively, if C is increased in value, making the CR product larger, the discharge during the line period will be less.

A form of distortion occurs when the picture content changes from, say, steady white to steady black. This is minimized by keeping the current in R as near constant as possible. The methods shown in Figs. 57 (*a*), (*b*) and (*c*) approach this constant-current condition.

A further source of trouble arises if the low-frequency distortion preceding the restorer is such that the synchronizing pulse tips tend to wander in potential away from the diode conduction potential. If this happens the diode is inoperative unless the restorer CR product is small, which is undesirable for the previous reasons. It is therefore preferable to keep the earlier couplings good, so that the rate of input signal variation is slow compared with the time constant CR.

Simple D.C. restoration is most effective when the signal voltage is large compared with the marginal voltage. As a practical guide a

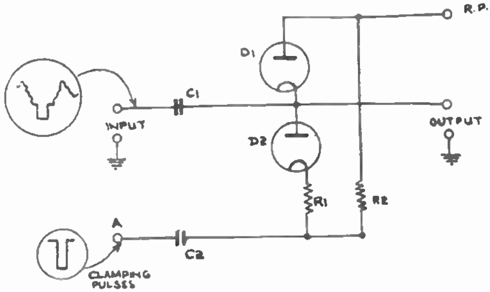


FIG. 58.—ONE FORM OF SYNCHRONIZING PULSE PEAK-LEVEL CLAMP.

signal of less than 10 volts is difficult to restore well with diodes currently available. Signals greater than 20 volts can be restored so that distortion in the waveform datum levels is less than 3 per cent.

Synchronizing Pulse Peak Level Clamping

This operates during the same part of the waveform as simple D.C. restoration, but can be made far more effective, since the diode-operating signals may be large and independent of signal levels and variations in signals.

One type is illustrated in Fig. 58.

Large-amplitude, separated-synchronizing pulses are applied to the diode D_1 to cause conduction during the synchronizing pulse period.

Advantages over simple restoration occur because :

(1) The period of diode conduction is constant for all conditions of the signal.

(2) The diode conduction impedance is low and substantially constant over the whole restoring period.

In the form shown in Fig. 58, the product R_2C_2 is made as large as practicable. R_1 increases the marginal voltage and is increased until

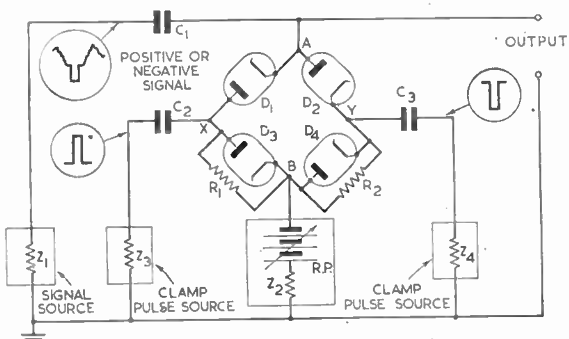


FIG. 59.—GENERALIZED CIRCUIT FOR A DIODE BRIDGE BLACK-LEVEL CLAMP.

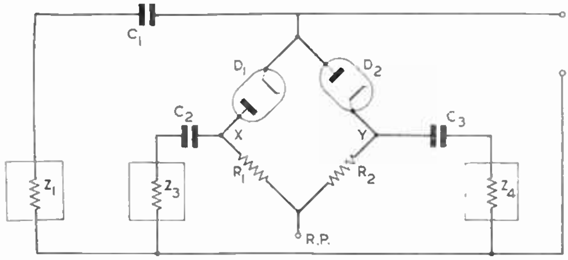


FIG. 60.—GENERALIZED TWO-DIODE CLAMP CIRCUIT.

C_1 = 0·005 μF.
C_2 = C_3 = 0·1 μF.
D_1, D_2 two halves of a D77.
R_1 = R_2 = 270k.
Z_1 = 1k.
Z_3 and Z_4 are 680 Ω sources of clamp pulses.
Pulse amplitude = 25 volts.
Signal amplitude = 10 volts picture; 8 volts sync.

the p.d. becomes many times the variation in p.d. which occurs during the frame period.

C_1 is made small, since there is now no need for the discharge resistor.

Synchronizing pulse peak level clamping may also be accomplished by using the form of clamps described under the heading "Black-level Clamping", the only difference then being that the diode gating signals are amplified synchronizing pulses instead of the very narrow black-level clamping pulses.

Black-level Clamping

This method restores the datum level following the synchronizing pulses and preceding the picture signals. Clamping pulses are generated in a clamp-pulse generator (described later), and are applied to the clamping diodes shown in Figs. 59 and 60.

The duration of the black-level period varies with the standards. For British standards the duration is approximately 5 microseconds, so it is necessary for the duration of the clamping pulses to be less than 5 microseconds and disposed in time so that they occur within the 5-microsecond period.

It is common practice to "slave" the clamp-pulse generator to the waveform so that the clamp pulse starts at a short time, say ½–1 microsecond, after the lagging edge of the synchronizing pulse, and to be anything from 1 to 3 microseconds in duration.

Two forms of diode clamp are commonly used, and appear to have superseded triode clamps These diode clamps are illustrated in Figs. 59 and 60.

Four-diode Bridge Clamp

This is illustrated in Fig. 59. Clamp pulses are applied with polarity as shown. The diodes are caused to conduct, current flows round the bridge system, and the capacitor C_1 is restored to the reference potential.

The charges which accumulate on coupling capacitors C_2 and C_3 during this process are removed by the discharge resistors R_1 and R_2.

The criterion for high-speed line-by-line clamping is that the restoring time constant determined by the impedances Z_1, Z_2, the diode path and the capacitor C_1 shall be small compared with the duration of the pulse.

To ensure this form of operation, it is necessary to estimate the maximum discharge current in the capacitor C_1, and to design the diode bridge to carry more than this current during conduction.

To achieve good clamping in practice :

(1) C_1 is made small, say 0·001–0·01 μF.
(2) The impedances Z_1, Z_2 are made as small as possible.
(3) Clamp pulses are made large, at least twice and sometimes several times the d.a.p. signal excursion on C_1.
(4) Clamp pulses are made equal in amplitude and shape.

This form of clamp has the additional advantage that the reference potential may be varied rapidly, and the level of the signal will follow. This form is therefore used where feedback correction is applied to stabilize black level.[42]

Two-diode Clamp

This is illustrated in Fig. 60, and is commonly used in cases where a fixed reference potential is required and where it is permissible or desirable to take several lines to correct an error in datum level. In the simplest form of application in which noise pulses can occur in the clamp-pulse system this is a desirable feature, and is commonly employed.

When this form of clamp is used, the requirements of low impedance in all current paths are reduced to the level of the reduced performance to be expected.

Suitable values are shown in Fig. 60.

PREDISTORTION AND WAVEFORM CORRECTION

Owing to the non-linear I_A/V_g characteristics of the grid-modulated valve and any subsequent amplifiers, together with regulation of driving stages and supplies, an incoming waveform as Fig. 61 (*a*) will be distorted in the transmitter chain and emerge in the uncorrected state as Fig. 61 (*b*) with the ratio $\frac{x}{y} > \frac{7}{3}$.

The distortion occurs in three distinct forms, which sometimes merge into one another, but for simplicity it is assumed that correction is

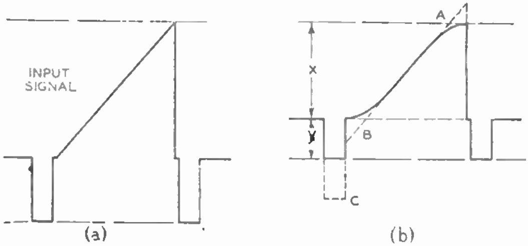


FIG. 61.—(*a*) UNDISTORTED LINE SAW-TOOTH WAVEFORM; (*b*) WAVEFORM DISTORTED DUE TO TRANSMITTER VALVE CHARACTERISTICS.

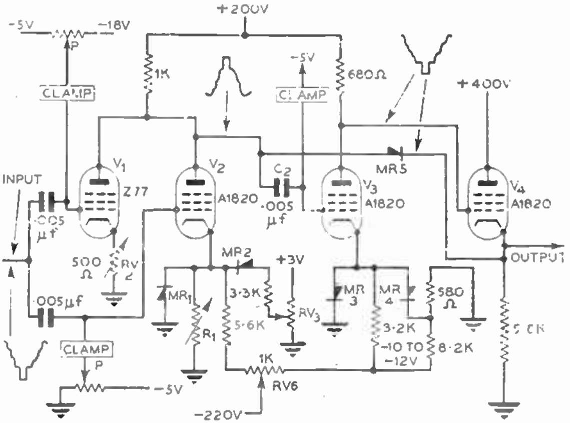


Fig. 62.—Television Waveform Correction Unit.

required (*a*) towards white Fig. 61 (*b*)—A; (*b*) towards black Fig. 61 (*b*)—B; (*c*) to lengthen the synchronizing pulse Fig. 61 (*b*)—C.

A system incorporating these three forms of correction is shown in Fig. 62.

Non-linear Amplification towards White

The valve V_1 is driven with the full input signal, but the clamp reference potential P is returned to a variable bias, which is adjusted to allow V_1 to conduct at the point in the voltage excursion towards " white ", where non-linear amplification is needed. The law of the correcting signal is adjusted by suitable choice of the feedback resistor RV_2. The anodes of valves V_1 and V_2 work into a common load, so that the combined output is the normally amplified signal plus the correction signal.

Synchronizing Pulse Stretching and Stabilizing Circuits

During the period of picture signals the rectifier MR_1 is cut off and the gain of V_2 is largely determined by the cathode feedback resistor R_1. In the period of synchronizing pulse signal the cathode of V_2 is negative to earth and MR_1 conducts, and the gain of V_2 is greatly increased, so stretching the synchronizing pulse. The signal amplitude on the grid of V_2 is sufficiently great to ensure adequate stretching even in the presence of reduced incoming synchronizing pulses. A further degree of synchronizing pulse stretching is provided by V_3 and the similar action of MR_3 and MR_4.

The grid of the output cathode-follower is connected directly to the anode of V_3, and its cathode is connected by the crystal diode to the input of the grid-coupling capacitor C_2. The tips of the synchronizing pulses cause MR_5 to conduct, lowering the effective anode load of V_1 and V_2 and limiting the pulse amplitude. The variable resistance RV_6

acts differentially on the currents in V_2 and V_3, and varies the amplitude of the output synchronizing pulse.

In this system the outgoing pulse amplitude is independent of incoming synchronizing pulse amplitude within the range of picture signal to synchronizing pulse signal ratio 50/50 to 90/10.

Non-linear Amplification towards Black

In the circuit of Fig. 62 RV_3 sets the level at which non-linear amplification sets in, by adjustment of the bias on crystal MR_2. Since V_2 cathode is at earth potential when the signal is at black level, the crystal will conduct when the bias potential provided by RV_3 is positive, and when conduction takes place the gain of V_2 is increased, giving the necessary pre-distortion.

This signal correction system is in current use, and is due to J. E. Nixon.

CLAMP-PULSE GENERATORS

These fall into two broad categories :

(1) Generators in which a pulse-forming oscillator is locked to the incoming waveform.

(2) Generators in which the pulses are derived directly from the incoming waveform.

Each type has its own particular advantages and finds application depending upon the conditions to be met. Thus in systems which employ overall feedback to determine the clamp reference voltage, the

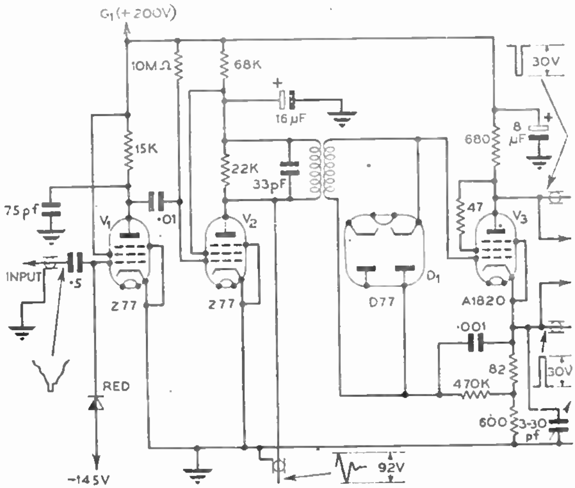


FIG. 63.—RINGING TYPE OF CLAMP-PULSE GENERATOR.

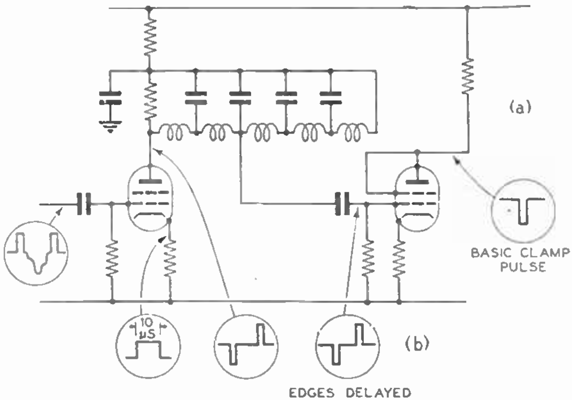


FIG. 64.—BASIC DELAY LINE TYPE CLAMP-PULSE GENERATOR.

locked variety is frequently used in order to keep the system under control when the incoming signal fails.

In systems not employing feedback determined reference potentials, it is often preferable to use the waveform-derived clamp pulses.

Waveform-derived Clamp Pulses

(a) *Ringing Type.* One such arrangement is illustrated in Fig. 63.

The first valve V_1 amplifies the incoming signal with a gain of, say, 50–100, giving 10–20 volts of synchronizing pulse output. The poor frequency response of such a high-gain amplifier is of little consequence, and helps in reducing the effects of short-duration interference pulses in the incoming signal.

A D.C. restored coupling to the next valve V_2 enables V_2 to separate the synchronizing pulses from the composite waveform. In this form of separator a reduced anode voltage on V_2 is advantageous.

The anode current of V_2 flows in a transformer which produces ringing after each transition edge of the pulse. The diode D_1 damps out all except the wanted pulse.

The transformer is tuned to about 170 kc/s and gives a pulse about 3 microseconds in width.

The precise timing of the pulse is unimportant as long as it occurs during the 5 microseconds of " black level ", so that errors in timing due to the simplified separator, etc., which might affect high-precision synchronization are unimportant. The derived pulse is fed to V_3, which provides outputs of both polarities, which may be used to operate on the diode clamps.

The equality of the pulses in amplitude and shape is of some importance, and a trimmer is provided to assist the balance and minimize break-through of the clamp pulses on to the clamped grid.

(b) *Delay Line Type.* A second type of waveform-derived clamp-pulse generator is illustrated in Fig. 64.

First of all the synchronizing pulses are separated and the anode load

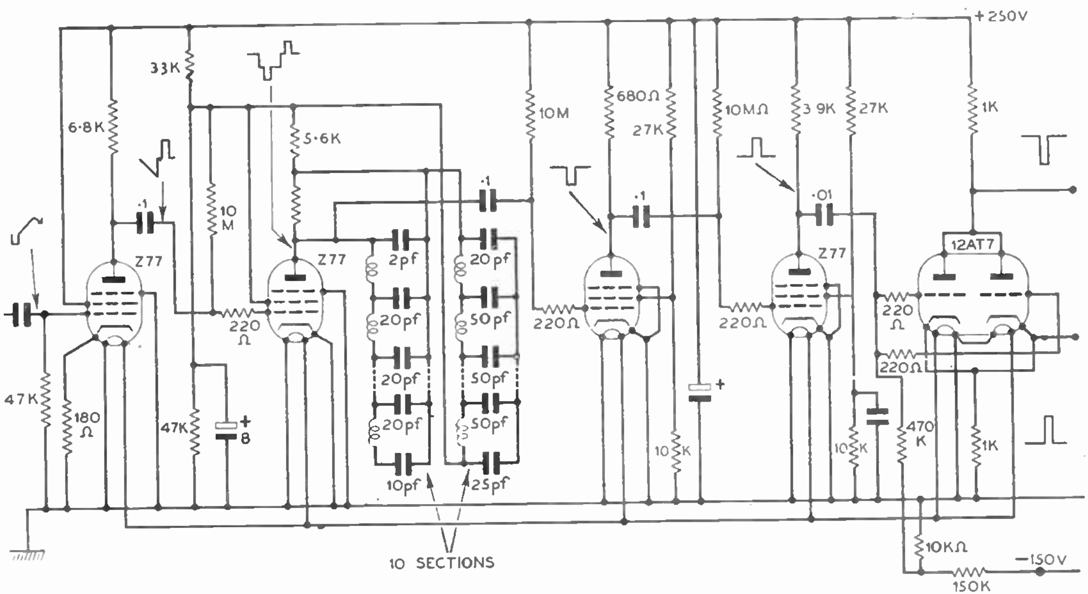


Fig. 65.—Complete Delay Line Clamp-pulse Generator.

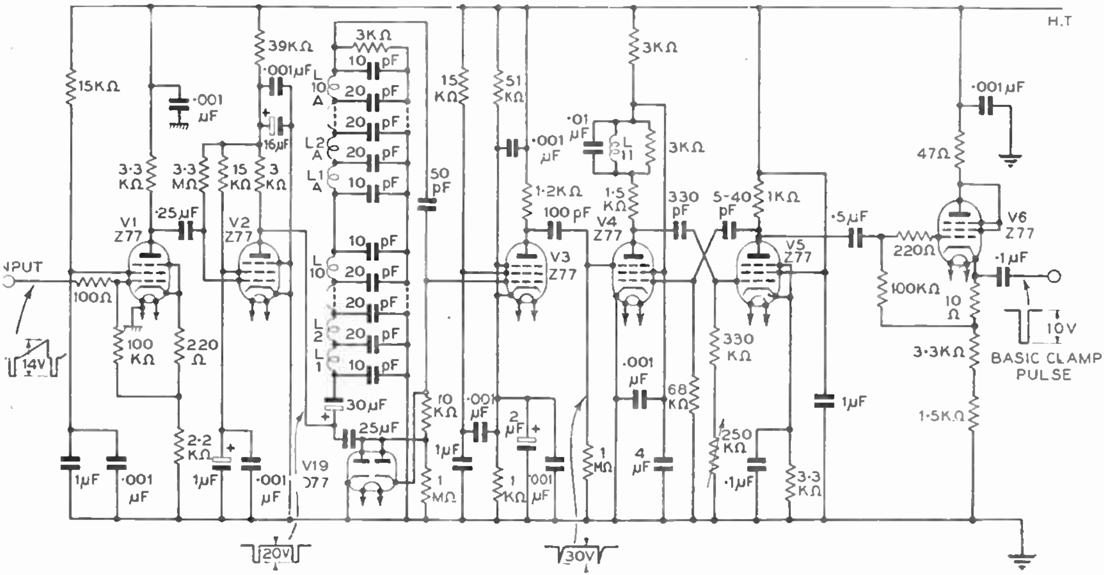


Fig. 66.—Complete Locked Type Clamp-pulse Generator with Interference Immunity.

of the separator consists of a short-circuited delay line D_1 with transmission time of, say, 3 microseconds.

When the separated synchronizing pulse appears at the anode of V_1, it travels along the delay line and is reflected at the shorted end, and arrives back 6 microseconds later at the anode with polarity reversed.

At the anode therefore the separated synchronizing pulses are partially cancelled by the reflected synchronizing pulses, and the result is as shown in Fig. 64.

It will be observed that the second pulse in the resulting output occurs during the black-level period of the original waveform, and may therefore be used directly as a clamp pulse.

In order to move the effective clamp pulse by a safe time margin from the trailing edge of the synchronizing pulse, a further small delay is introduced. This may conveniently be obtained by taking the output from a tapping on the delay line as shown in Fig. 64.

As before the basic clamp pulse is then amplified and/or produced in both polarities for use by the diode clamps.

A complete pulse generator is shown in Fig. 65, which uses a slightly different arrangement of delay lines to achieve the same result.

Waveform-locked Clamp-pulse Generators

These may take the form of a free-running oscillator-type pulse generator such as a multi-vibrator, which is locked to the incoming waveform. One such system is illustrated in Fig. 66.

The separated synchronizing pulses are delayed in the line network and differentiated by the 50 pF–10 k Ω capacitor-resistor combination in the coupling to V_3 to provide an effective locking pulse for the multivibrator V_4, V_5. The tuned circuit in the anode circuit of the multivibrator valve V_4 maintains the line-by-line locking during the frame-suppression period when half-line triggering is present.

Output from the multi-vibrator is taken to valve V_6, the output of which is then amplified, limited and distributed with both polarities.

The circuit including the diode V_{19} is a gating circuit which gives considerable immunity from interference.

This form of clamp-pulse generator is currently used as part of a high-speed line-by-line control of black level, and is due to N. N. Parker-Smith.[42]

POWER SUPPLIES FOR TELEVISION TRANSMITTERS

All power supplies for television transmitters that have to deal with loads which fluctuate at video frequency must either provide a constant output impedance over the video-frequency range, or the output impedance must be so low at all frequencies that it is insignificant.

At high power level, for radio-frequency stages, it is usual to employ the constant-resistance network filter to meet this requirement.

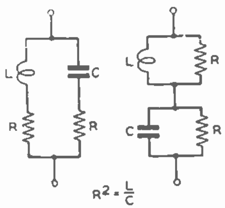


FIG. 67.—TYPICAL ARRANGEMENTS FOR CONSTANT IMPEDANCE NETWORKS. EACH NETWORK IS R AT ALL FREQUENCIES

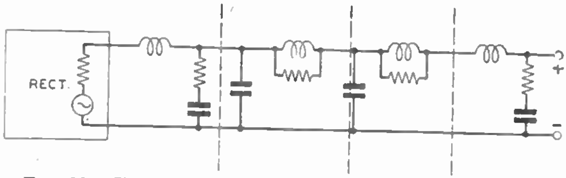


FIG. 68.—FORMS OF CONSTANT-IMPEDANCE NETWORKS AS SUPPLY FILTER.

When more than one modulated radio-frequency stage is supplied from a single source by this method, the constant-resistance filter represents a common coupling impedance, and in some practical cases precludes this form of economy.

In the video-frequency section of the transmitter, common impedance between stages due to supplies is far too troublesome to be included in a design, and the usual procedure is to stabilize the supplies electronically.

Constant-impedance Supplies

For this system use is made of constant-resistance networks as filters between the rectifiers and loads.

Typical arrangements are shown in Fig. 67, and the fundamental relation for all the networks is that if $R = \sqrt{L/C}$, then the impedance

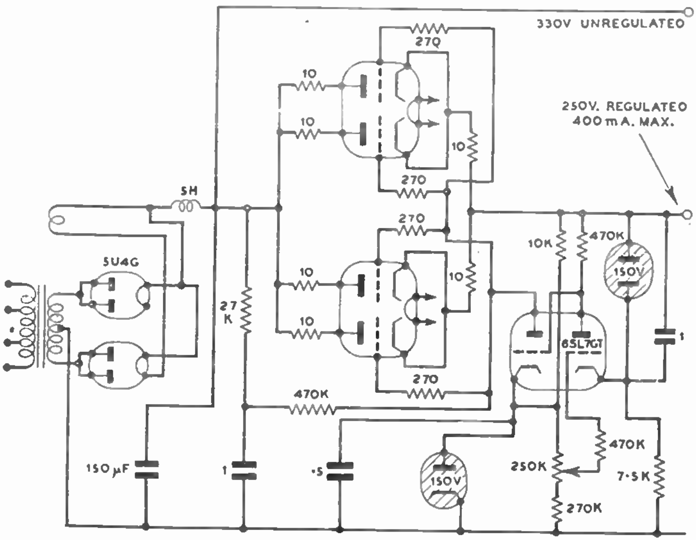


FIG. 69.—SERIES REGULATED POWER SUPPLY.

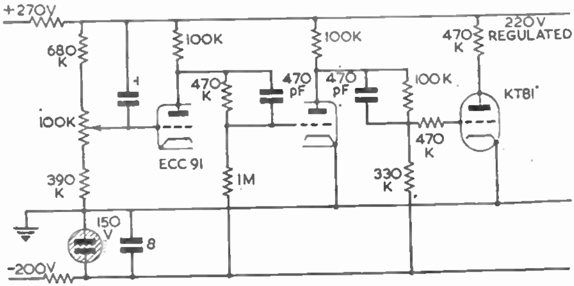


FIG. 70.—SHUNT-REGULATED POWER SUPPLY.

of the rectifier and filter as seen by the load is purely resistive over the video-frequency band.

In practice, owing to the distributed capacitance of the inductors and the inductance in capacitors, several sections are required—tapered in value so that each section deals only with components of current change within the frequency range in which it is substantially a constant resistance. A typical practical filter is shown in Fig. 68.

Stabilizers

These fall into two categories:

(1) series stabilizers;
(2) shunt stabilizers.

Further forms which are a combination of both are occasionally used.

Series Stabilizers

In this form a series valve carries the main load current, and the constancy of the output voltage despite load-current changes is achieved by monitoring the output voltage and applying an amplified control voltage on to the grid of the series valve.

For certain applications in which low-frequency current excursions of the load are high, the series stabilizer becomes uneconomic, as all the current fluctuations flow through the rectifier filter. If the filter is of conventional design, there will be large transient voltages appearing at the anode of the series regulator valves, and considerable margin of control becomes necessary to enable the valves to work satisfactorily under these conditions. In difficult cases constant-resistance filters preceding the stabilizer series valves become necessary.

A typical series-regulated supply in common use for television apparatus is illustrated in Fig. 69.

Series stabilizers are most effective in dealing with supply-voltage variations at reasonably constant loads.

Shunt Stabilizers

In this form of stabilizer the load is shunted by the main stabilizer valves, which are controlled by amplifiers, so that the load on the

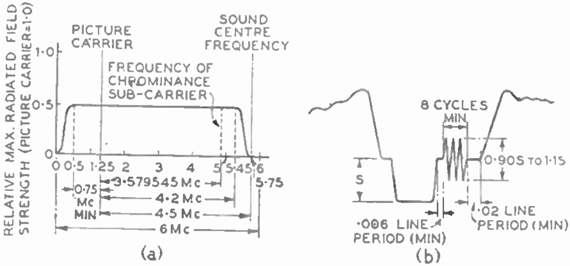


Fig. 71.—(*a*) N.T.S.C. Idealized Picture Transmission Amplitude Characteristic; (*b*) N.T.S.C. Line-synchronization Pulse.

rectifier remains substantially constant. Rectifier supplies of any impedance and with conventional filters are thus quite suitable, and this makes shunt stabilization particularly attractive at high power level, where load-current fluctuations are high.

A typical practical shunt stabilizer is illustrated in Fig. 70, and is suitable for 40 mA load current variation at 220 volts.

Generally speaking, the shunt stabilizer is most suitable in dealing with the load-current fluctuations when the input voltage is relatively constant.

Series-shunt Stabilizers

Where large variation of input-voltage and large variation of load current have to be handled at the same time, it sometimes becomes necessary to use combined series and shunt regulator types.

These are most easily designed by arranging that all load-current fluctuations are made up by the shunt system working at constant input voltage and by arranging that the series system stabilizes the voltage against input voltage fluctuations, at constant current.

COLOUR TRANSMITTERS

The recent adoption of N.T.S.C. colour television standards in the U.S.A. introduces an additional problem into the design of television transmitting equipment. While this system has not so far been adopted in this country, there is every reason to assume that it represents one satisfactory way of achieving receiver compatibility between colour and monochrome systems, and the implications on design are worth noting.

Basically there are two features of the N.T.S.C. type transmitted waveform, illustrated in Fig. 71, which require special consideration :

(1) the fully modulated sub-carrier at a frequency within approximately 0·5 Mc/s of the maximum video frequency to be transmitted;

(2) the existence of the colour synchronizing "burst" in the black-level period.

Sub-carrier Problems

In monochrome transmission it is safe to assume that full-amplitude signals do not occur at the maximum video frequencies. It is assumed

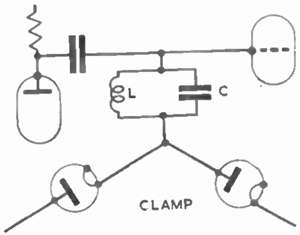


FIG. 72.—CLAMPING CIRCUIT FOR N.T.S.C. CIRCUITS.

in testing television transmitters that for sinusoidal test signals within the range of picture signal amplitude, full amplitude is required only up to 1·6 Mc/s, and that amplitudes then decrease *pro rata* with frequency, i.e., half-amplitude signal at 3·2 Mc/s (for 405-line standards).

In designing for monochrome transmission it is therefore possible to economize in the reactive current availability of the system, provided that the transient response requirement is met.

With a fully modulated sub-carrier at high video frequency, this economy is not possible, and greater reactive current availability must be made. If this current availability is too small, not only will the degree of colour saturation be limited, but a further risk of transmitted sub-carrier phase shift will occur, due to the phase shift often attendant upon valve sources which cut-off during a voltage excursion.

If a sub-carrier phase shift of more than a few degrees occurs, not only is saturation reduced but false colours will also be produced.

The degree to which this affects design is not yet fully known, but it would seem that colour is likely to need at least 30 per cent more reactive current availability than equivalent monochrome transmitters.

" Burst " Problems

The existence of the synchronizing burst has a direct effect on all black-level clamps, since it is during the clamping period that the burst is present.

Fortunately a simple remedy exists. By inserting a parallel-tuned circuit, tuned in to the sub-carrier, in the connection between the clamping diode and the clamped capacitor, the effect of the burst on clamping is removed.

This is illustrated in Fig. 72, and is often achieved by using a self-resonant " choke " as the tuned circuit.

V. J. C.

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APPENDICES

I. Double shunt-regulated Amplifier (Fig. 42)

Output impedance:

$$Z_0 = \frac{(r_{a2} + Z_2)(r_{a1} + Z_1)(Z_5 + Z_6) + (r_{a2} + Z_2)r_{a1}Z_1 - \mu_1\mu_2Z_1Z_2Z_6}{(r_{a2} + Z_2)(Z_5 + Z_6) + (r_{a2} + Z_2)Z_1 + \mu_2Z_1Z_6 + (r_{a1} + Z_1)(Z_5 + Z_6) + Z_1r_{a1} + \mu_1Z_2(Z_5 + Z_6) + \mu_1Z_1Z_2}$$

Gain:

$$A = \frac{V_0}{V_i} = \frac{Z_L}{Z_L + Z_0}\left[\frac{Z_1\{r_{a2} + Z_2 - \mu_2(r_{a1} + \mu_1Z_2)\} - \mu_2Z_6(r_{a1} + Z_1 + \mu_1Z_2)}{(Z_5 + Z_6)\{r_{a1} + Z_1 + r_{a2} + Z_2(\mu_1 + 1)\} + Z_1(r_{a1} + r_{a2} + Z_2 + \mu_1Z_2 + \mu_2Z_6)}\right]$$

Input impedance:

$$Z_i = \frac{Z_L + Z_0}{Z_L}\left[\frac{(r_{a2} + Z_2)(Z_5 + Z_6) + (r_{a2} + Z_2)Z_1 + (r_{a1} + Z_1)(Z_5 + Z_6) + r_{a1}Z_1 + \mu_1Z_2(Z_5 + Z_6) + \mu_2Z_1Z_6 + \mu_1Z_1Z_2}{r_{a1} + r_{a2} + (\mu_2 + 1)Z_1 + (\mu_1 + 1)Z_2 + \{(r_{a1} + Z_1)(r_{a2} + Z_2) - \mu_1\mu_2Z_1Z_2\}/Z_L}\right] - Z_0$$

II. Double-shunt-regulated Cathode Follower (Fig. 43)

Output impedance:

$$Z_0 = \frac{(r_{a1} + Z_1)(r_{a2} + Z_2)(Z_3 + Z_4) + (r_{a1} + Z_1)(r_{a2}Z_3) - \mu_1\mu_2Z_1Z_3Z_4}{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu_1 + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_3 + \mu_2Z_1(Z_3 + Z_4)\} - \mu_1Z_4(r_{a2} + \mu_2Z_1)}$$

and this is zero when

$$(r_{a1} + Z_1)(r_{a2} + Z_2)(Z_3 + Z_4) + (r_{a1} + Z_1)r_{a2}Z_3 = \mu_1\mu_2Z_2Z_3Z_4.$$

Gain:

$$A = \frac{(r_{a1} + Z_1)r_{a2} + \mu_1[(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_3 + \mu_2Z_1(Z_3 + Z_4)] - \mu_1Z_4(\mu_2Z_1 + r_{a2} + Z_2)}{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu_1 + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_3 + \mu_2(Z_3 + Z_4)Z_1\} - \mu_1Z_4(r_{a2} + \mu_2Z_1)} \times \frac{Z_L}{Z_L + Z_0}$$

Input impedance:

$$Z'_i = \frac{Z_L + Z_0}{Z_L}\left[\frac{(r_{a1} + Z_1)(Z_3 + Z_4 + r_{a2}) + (\mu_1 + 1)\{(r_{a2} + Z_2)(Z_3 + Z_4) + r_{a2}Z_3 + \mu_2Z_1(Z_3 + Z_4)\} - \mu_1Z_4(\mu_2Z_1 + r_{a2})}{(r_{a1} + Z_1) + (r_{a2} + Z_2) + \mu_1Z_2 + \mu_2Z_1 + \{(r_{a1} + Z_1)(r_{a2} + Z_2) - \mu_1\mu_2Z_1Z_2\}/Z_L}\right]$$

11. TRANSMITTING AERIALS

11. TRANSMITTING AERIALS

The transmission of intelligence by means of electromagnetic waves has developed primarily because the medium is all-pervading and essentially free from transmission losses. The waves are set up in the medium by an exciter acting as a transformer to convert the radio-frequency power of the transmitter into electromagnetic wave energy as efficiently as possible. The transmitting aerial is the exciter, and its form must ensure that the maximum amount of energy is directed towards the receiver and that the intelligence carried by the energy is not distorted.

Essentially, the best periodic frequency for the waves is settled by the distance between the transmitter and the receiver and by the type of intelligence which is to be sent, but a number of other factors influences the choice: other users, atmospheric noise, man-made noise and various natural phenomena have to be considered.

The design of the aerial is determined by:

(1) The operating frequency (wavelength) selected.
(2) The location of the receiver relative to the transmitter.
(3) The complexity of the intelligence, i.e., the band-width necessary.
(4) The field strength required at the receiver to give acceptable reception of the intelligence.
(5) The location of the transmitting site.
(6) The materials available for use.

The design also involves a knowledge of the nature of electromagnetic waves and of the natural phenomena which influence their progression through regions other than free space.

That part of the spectrum of electromagnetic waves which is used for communication purposes has been classified into decades. Three of these are considered here:

Medium frequency (M.F.).	300–3,000 kilocycles per second 1,000–100 metres wavelength
High frequency (H.F.)	3,000–30,000 kilocycles per second 100–10 metres wavelength
Very high frequency (V.H.F.)	30–300 megacycles per second 10–1 metres wavelength

Radiation

Energy is conveyed away from a conductor carrying a periodically varying current by the electromagnetic field which is set up in the surrounding space. The field has two components, an electric wave with vectors in the same plane as the conductor and a magnetic wave with vectors at right angles to the electric wave. Both sets of vectors are perpendicular to the direction in which the energy travels.

When two conductors are close together and are carrying equal "go and return" currents—as in Fig. 1 (*a*)—the electric and magnetic

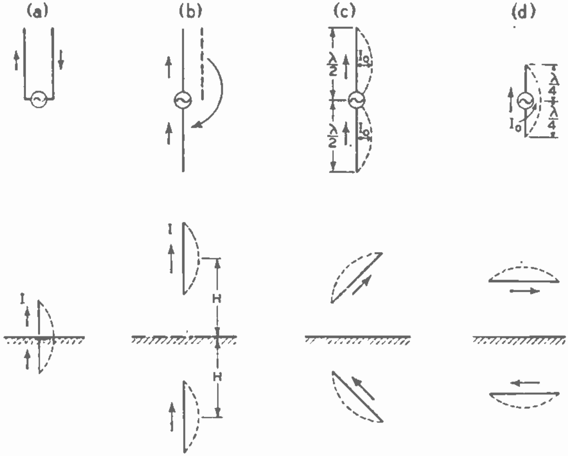


FIG. 1.—AERIALS AND IMAGES.

(*a*) Go and return currents in parallel conductors; (*b*) conductors in line; (*c*) parallel resonant conductors; (*d*) series resonant conductors.

fields of one are exactly cancelled at all distant points by the fields due to the reverse current in the other conductor, and no energy is conveyed away or "radiated" from the pair: it all remains in the electric and magnetic fields bound in the space between the conductors. Now if they are moved apart and placed end to end with the source of energy between them as in Fig. 1 (*b*), the fields no longer cancel at distant points in space, and energy is conveyed out of the immediate vicinity of the conductors by the electric and magnetic fields—together forming the electromagnetic field—to all space. When the conductors are in series or parallel resonance at the generator frequency, that is of a length one-half c/f (where c is the velocity of propagation of electromagnetic waves, and f is the generator frequency ($= \omega/2\pi$) and c/f is the wavelength), a standing wave of current is set up on them, the input impedance is purely resistive and all the input power, apart from losses, is radiated. In the series resonant condition the magnitude of the resistance is usually denoted by R_r and called the "radiation resistance referred to the loop current I_0," so that

$$R_r = \frac{P}{I^2_0}$$

The term I_0 is the maximum value of the standing wave of current on the conductors for the series and the parallel resonant lengths, and R_r is assumed to be located at the point where I_0 is flowing. Two conductors in series resonance as Fig. 1 (*d*) form what is usually called a "half-wave dipole" or more simply a "dipole".

The magnetic field H set up by a current I in a wire of length L at a point at a distance D in the plane which is perpendicular to the wire is given by the expression :

$$H = I.L.\left[\frac{\omega \sin \omega\left(t - \frac{D}{c}\right)}{c.D.} - \frac{\cos \omega\left(t - \frac{D}{c}\right)}{D^2}\right] \text{e.m.u.} \quad (1)$$

The first term represents the radiation field, the second term is the induction field. In the full analysis it can be shown that the energy in the radiation field is lost from the source, whereas that in the induction field returns to the source over the complete time period $1/\omega$, and therefore represents no loss of power. At great distances the field strength will be mainly due to the radiated energy, since this is proportional to the inverse of the distance and the induction component to the inverse of the distance squared.

Polarization

The polarization of an electromagnetic wave is defined in relation to the electric vector, so that in a vertically polarized wave the electric vector is vertical and the magnetic horizontal. It follows that a vertically polarized wave is set up along the surface of the earth when a suitable generator is connected between a vertical wire and ground. The earth can be considered as a reflecting surface, and the field strength of the wave is due to the combined effects of the wire and its image in the surface. Typical radiators and their images are illustrated in Fig. 1. The surface, or Ground Wave, is particularly important on medium frequencies, as it travels for considerable distances in contact with the ground. Energy which is radiated at an angle to the horizontal forms a wave travelling upwards—the Sky Wave—and does not give rise to a field at ground level unless reflected from a conducting layer above the earth, as shown diagrammatically in Fig. 2.

Propagation by Ground Wave

Since the earth is not a perfect conductor, part of the energy of the ground wave is absorbed, and the field strength decreases more rapidly than the inverse of the distance travelled. This effect has been studied in great detail, and groups of curves relating field strength with

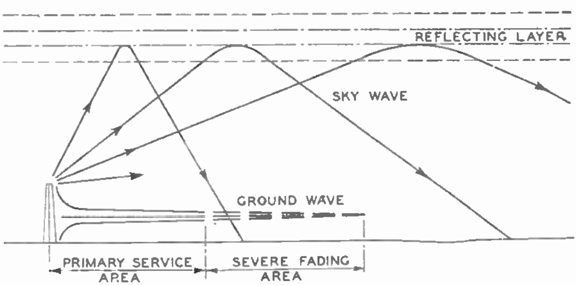


FIG. 2.—GROUND WAVE AND SKY WAVE.

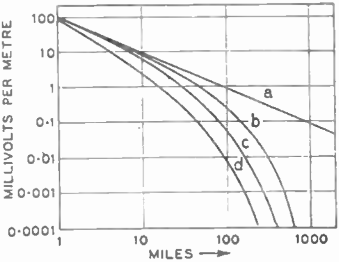


FIG. 3.—GROUND WAVE FIELD STRENGTH OVER A GOOD PATH.

Curve (*a*)—Inverse distance
(*b*)—at 500 kc/s
(*c*)—at 1,000 kc/s
(*d*)—at 2,000 kc/s

Conductivity = 10^{-13} e.m.u.
Permittivity = 15 e.s.u.

distance and frequency for ground paths having specified conductivities and permittivities are available. Fig. 3 shows a typical group plotted for an aerial producing a vertical field strength of 100 mV/metre at a distance of 1 mile. When the field strength at 1 mile has been measured or calculated for a given aerial, the curves give the field strength at other distances when ground attenuation is taken into account and also allow for the curvature of the earth's surface, an effect which becomes important for distances greater than some 35 miles. Ground losses increase as the frequency is increased and, above 1,500 kc/s, so rapidly attenuate the ground wave that the service area is reduced to a few miles. Broadcasting on these frequencies is confined to low-power transmitters designed for a local service. Over the sea, however, the attenuation is very much less, and marine communications can be readily maintained up to 100–150 miles. It will be evident from Fig. 2 that no horizontally polarized ground wave is set up, since horizontal aerials and their images produce equal and opposite fields along the surface of the earth.

Propagation by Sky Wave

That part of the wave which leaves the transmitting aerial at an angle above horizontal—the "sky wave"—varies as the inverse of the distance travelled until it enters the lower regions of the ionosphere. As it penetrates farther into these regions, the wave is also refracted and, under certain conditions, is bent sufficiently to return to the earth's surface at a distance related to its angle of departure and the height of the ionized region concerned. If this distance is within ground-wave range of the transmitter (Fig. 2), the field there will have two components which may differ in amplitude and phase. Slight variations in the height of the ionized layer will alter the sky-path length, and therefore the relative phase of the sky-wave component, causing equivalent changes in the magnitude of the resultant signal. This is the well-known effect of "fading". Fading of this form can put a limit to the service of medium-frequency broadcasting stations, and every effort is made in the aerial design to reduce the strength of the sky wave. On the other hand, it is the return of high-frequency sky waves to earth far beyond the range of the ground wave that makes possible communication to all parts of the globe. The study of the propagation of radio waves over long distances has shown that radiation from the sun acting on atmospheric gases of varying density gives rise to ionized layers at heights between 50 and 500 km. above the earth. This region is called the "Ionosphere"

The Ionosphere

Each one of the rather ill-defined ionized layers has certain dominant characteristics in reflecting or refracting radio waves, which are exploited for long-distance communication. In the lowest region, at a height of between 75 and 95 km., the atmospheric gases are relatively dense, and are ionized where the sun's radiation strikes them, but de-ionize rapidly after sunset. It is sometimes called the " D " layer, and exists during daylight hours only, travelling round the globe with the sun. It is mainly an attenuating region in which the degree of attenuation at frequencies below 2,000 Mc/s is nearly constant, and sufficient to absorb most of the skyward radiation from medium-frequency aerials. Hence there is little fading on medium frequencies during daylight hours in the summer. At a height of about 110 km. the conditions of gas density and ionization change to form the " E " layer, which absorbs little energy from an incident wave. The wave penetrates the layer by an amount dependent on frequency, higher frequencies penetrating farthest, and is progressively refracted until it emerges. The terms " refraction " and " reflection " are practically interchangeable in this connection, and the " effective height " of the layer is that of the equivalent reflecting surface. Thus, at normal incidence, a wave penetrates to the effective height and is then reflected back along the same path. The highest frequency which is reflected at normal incidence is termed the " Critical Frequency ". The " E " layer is also produced directly by the sun's radiation and tends to disappear during night-time. The critical frequency rises with the ionization intensity towards local noon, reaching a maximum of about 3·7 Mc/s during the summer, but falls rapidly after sunset. Medium-frequency fading occurs during the hours of darkness, since the sky wave is no longer absorbed by the day-time " D " layer, but is reflected from the weak residual ionization of the " E " layer and from higher layers.

Another region of ionization occurs between 200 and 400 km. above earth having a critical frequency between 3 and 12 Mc/s and generally known as the " F " layer. During the day it splits into two, a lower " F_1 " and an upper " F_2 " layer. " F_2 " has the higher critical frequency, and is the controlling layer for long-distance communication with its critical frequency (f_c) determining the maximum usable frequency for the path. An incident wave striking the layer at an angle Δ to the normal is reflected if its frequency is less than $f_c . \sec \Delta$. Since

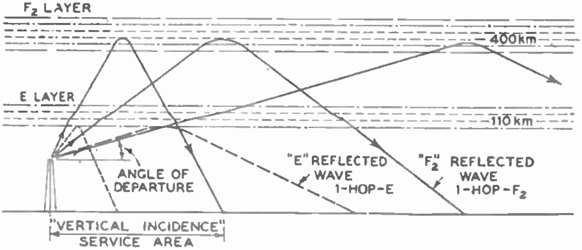


FIG. 4.—HIGH-FREQUENCY SKY WAVES.

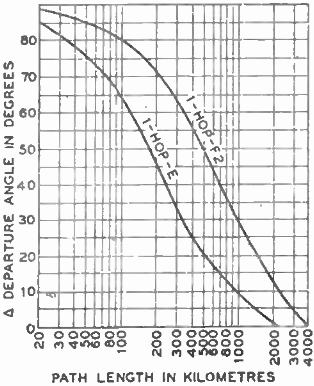


FIG. 5.—OPTIMUM DEPARTURE ANGLES FOR VARIOUS PATH LENGTHS.

the attenuation of the wave varies inversely with frequency, the highest frequency reflected will tend to be used so that the Maximum Usable Frequency (M.U.F.) $= f_c \cdot \sec \Delta$, but in order to take into account unpredictable variations in the F_2 layer a somewhat lower frequency is usually selected, known as the Optimum Working Frequency (O.W.F.).

Choice of Frequency

Sky-wave propagation through the ionosphere is studied all over the world, and records of layer heights, critical frequencies and their periodic variations are plotted at a number of stations. By this means, accurate forecasts of the O.W.F. for any path at any time of the day can be made: see Section 44.

In the design of aerials where it is desired to make best use of the O.W.F., the angle of departure is most important. Fig. 4 shows how this angle determines the distance at which the wave will return to earth for the first time. Over even a short period, however, owing to irregularities in the surface and height of the reflecting layer, the wave actually follows a number of slightly different paths, with the result that the " best angle " becomes a small range of angles. The best mean angle for various distances up to 4,000 km. is shown in Fig. 5. Over longer paths, the mode of transmission is by no means certain, and results sometimes indicate that a series of geometric hops is followed, at others that the energy travels between two layers.

It is, however, well established that the transmitted energy should be concentrated in the low angles of elevation, although the vertical radiation pattern of the array should not be too sharp: $\pm 5°$ to the half-power points is suitable. For similar reasons, the maximum usable frequency is selected from a consideration of only two reflection points for long paths; one at 2,000 km. from the transmitter, and the other at 2,000 km. from the receiver, both points being along the great-circle path between the transmitter and the receiver. For consistent propagation the M.U.F. rarely exceeds 26 Mc/s, though the London television signals on 45 Mc/s were regularly received in South Africa during the summer of 1950—a period of intense sunspot activity. In the very-high-frequency range of 30–300 Mc/s reliable propagation approximates to that of the direct space wave.

Propagation by Very High Frequencies

Aerials designed to work in this range of frequencies can, because of their relatively small dimensions, be placed many wavelengths above the ground, and wave propagation is along the line between transmitter

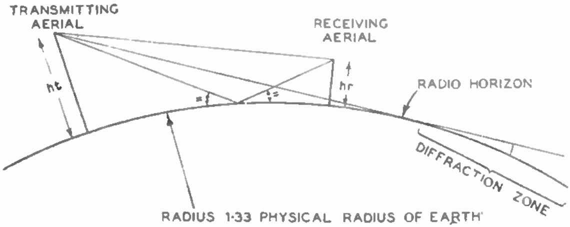


FIG. 6.—VERY-HIGH-FREQUENCY PROPAGATION.

and receiver, partly direct and partly from a ground reflection, as will be seen from Fig. 6. Beyond the direct "line of sight", reception is still possible because :

(*a*) the wave is bent in passing through the atmosphere by virtue of a change of refractive index with height; and

(*b*) diffraction occurs at the surface of the earth after the manner of light waves passing over an edge.

Effect (*a*) can be allowed for by assuming the radius of the earth to be greater, by a factor 1·33, than the physical radius and (*b*) by the use of optical diffraction formulæ. Diffraction patterns can also be set up by obstacles such as ranges of hills, and wide variations of field strength in the neighbourhood of these are to be expected. At points within the radio horizon of a transmitting aerial (that is, the horizon when the earth's radius is 1·33 times the physical radius) the field strength given by a single dipole is

$$E = 2{\cdot}85\sqrt{P}.\ \frac{h_t h_r}{\lambda d^2}\ \text{mV/metre} \quad . \quad . \quad . \quad (2)$$

where P = radiated power in kilowatts;
h_t, h_r = height of transmitting and receiving aerials in metres;
d = the distance from the transmitting aerial in kilometres;
λ = the wavelength in metres.

This is approximately correct for both horizontally and vertically polarized waves, and shows the importance of aerial height on the field at the receiver. The relationship between the received field strength and the height of either aerial is often called the "height gain factor" at the site, and may vary from place to place. It is sometimes negative for sites in hilly country due to abnormal ground reflections.

MEDIUM-FREQUENCY AERIALS

The ground wave set up by a vertical aerial follows the earth's curvature in travelling outwards from the transmitter, and, on medium frequencies, the losses along the path are sufficiently small for a useful amount of energy to be conveyed up to about 100 miles and frequently beyond this distance. This relatively low path attenuation, and the simplicity of the transmitting apparatus, led to the adoption of medium frequencies for the first broadcasting services. The aerials were also simple in form, but, with the growth of broadcasting, more and more attention was given to their design in order to concentrate the energy into the horizontal plane. This was necessary, since above 3° the radiation travels wholly skywards and is lost. Other services adopted medium frequencies for similar reasons: e.g., marine communications where the transmitter is moving and the bearing of the fixed receiver may vary over 360° in azimuth, navigational aids and low-power mobile services generally. In certain special circumstances the sky wave is used, but the great majority of aerial-design problems are concerned with the production of a vertically polarized ground wave.

The Field Strength from a Vertical Aerial

At a distant point the ground-wave field strength is related to

(*a*) the current flowing in each point of the aerial;
(*b*) the height of the current above the ground plane;
(*c*) the physical constants of the ground between the transmitter and receiver.

The aerial designer can determine (*a*) and (*b*), but obviously has little control of (*c*).

At a distance of D km from a vertical aerial of electrical height H radians above a perfect earth, the field strength E is given by

$$E = \frac{60 I_0}{D}(1 - \cos H) \text{ mV/metre} \tag{3}$$

when the current is assumed to be sinuosoidally distributed over the length of the aerial and has the value I_0 amperes at a point which is $\pi/2$ radians from the open end remote from the feed point (see Fig. 8). The feed current I_f is then, $I_0 \cos(H - \pi/2)$, i.e., $I_0 \sin H$.

The aerial will have a resistance R_r to account for the power P

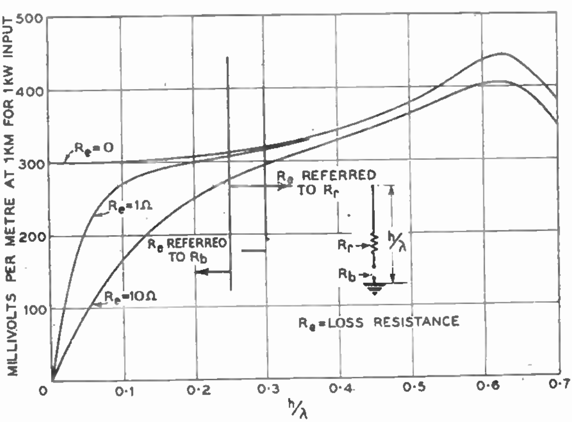


FIG. 7.—FIELD STRENGTH OF VERTICAL AERIALS.

radiated such that $R_r = P/I_0^2$ ohms, where P is measured in watts. If losses are sufficiently small to be neglected, a power of 1 kW will give rise to a field strength

$$E = \frac{1900}{\sqrt{R_r D}} . (1 - \cos H) \text{ mV/metre} \qquad . \qquad . \qquad (4)$$

a value which is very nearly constant for all values of H up to $\pi/2$ and equal to 315 mV/metre at a distance of 1 km for 1 kW input power.

Measurement of Ground-wave Field Strength

The ground-wave field strength of a transmitting aerial is measured at a distance of between five and ten wavelengths in a number of directions and the mean value calculated for the standard distance of 1 km by multiplying each reading by the factor obtained by dividing the actual distance by the standard distances. This makes the justifiable assumption that field strength varies inversely with distance, since, for the short distance considered, the attenuation along the ground can be neglected. The ratio of the square of this mean value to the square of the field strength, given by equation (4) for the aerial is the radiation efficiency.

To take account of losses (power absorbed but not radiated) in the aerial system, equation (4) is modified to :

$$E = \frac{1900\sqrt{P}}{\sqrt{(R_r + R_e)D}} \times (1 - \cos H) \text{ mV/metre} \qquad (5)$$

R_e being the equivalent loss resistance at the point where I_0 is flowing and P the input power in kW. Fig. 7 shows how seriously losses reduce the efficiency of a short vertical aerial, but, as is shown later, it can be improved by increasing H, without actually increasing the height in feet, by the addition of a non-radiating capacitance top.

Radiation Resistance and Feed Point Impedance

As for all circuits whose dimensions are comparable with the wavelength of the exciting voltage, an aerial has distributed inductance, capacitance and resistance, but when the inductance is very much greater than the resistance per unit length, the current along the radiating portion varies sinuosoidally and the radiation resistance is assumed to be lumped at the position of maximum current. When the aerial is less than a quarter of a wavelength long, the maximum value of the sine wave distribution of current does not actually occur on the aerial, and it is more realistic to consider the radiation resistance as being located at the feed point where the current $I_f = \sin H \times I_0$. Calling the resistance R_f, we have

$$R_f = \frac{I_0^2}{I_f^2} . R_r = \frac{r_r}{(\sin H)^2} \qquad . \qquad . \qquad . \qquad (6)$$

and for very short aerials

$$R_f = 40 \tan^2 \frac{H}{2} = 10H^2 \qquad . \qquad . \qquad . \qquad (7)$$

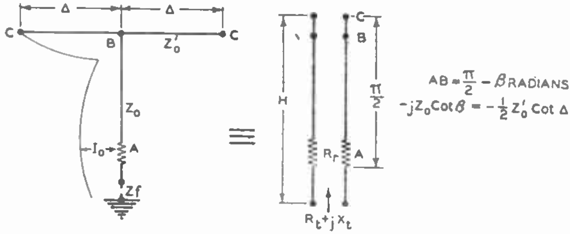


FIG. 8.—TOP-LOADED AERIAL AND EQUIVALENT TRANSMISSION LINE.

For longer aerials, the calculation of radiation resistance becomes somewhat lengthy, but it has been carried out by a number of workers. The curves shown in Fig. 9 have been plotted using values of radiation resistance quoted by W. L. McPherson in *Electrical Communication*, 1938. When these are used to find the field strength from equation (4) it must be remembered that the electrical length $H = 2\pi(h + b)/\lambda$, h being the height of the vertical portion of the aerial, and b the equivalent length of the top capacitance as given by equation (10).

In approximate calculations a vertical aerial can be regarded as a lossless transmission line with a length of H radians and characteristic impedance $Z_0 = \sqrt{(L/C)}$, on which is a standing wave of current $\frac{I_{max.}}{I_{min.}} = \frac{Z_0}{R_r}$ and the feed point or input impedance is then

$$Z_f = R_f - jZ_0 \cot H \qquad (8)$$

The inductance L and capacitance C per unit length of the aerial can be found from published formulæ (for instance, *Bureau of Standards Circular* 74, pp. 237–243) and the characteristic impedance derived as

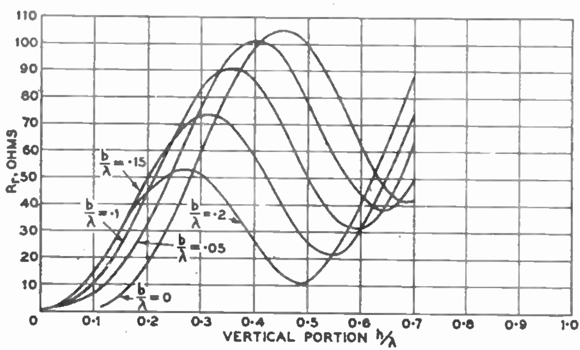


FIG. 9.—LOAD RADIATION RESISTANCE OF VERTICAL AERIAL.
h is the height of vertical portion; b is the equivalent length of top.

above, but the following list will be found of general application. From these values and the value of R_r taken from Fig. 9

$$Z_f = Z_0 \frac{a + j \tan H}{1 + ja \tan H} \quad . \quad . \quad . \quad (9)$$

$$a = \frac{Z_0}{R_r}$$

The characteristic impedance of some simple forms of aerials is given below :

Form of Aerial	*Value of Z_0 ohms*
Single wire 0·25 in. diameter . .	550–600
Two wires spaced 2 per cent of length .	420
Two wires spaced 5 per cent of length .	370
100-ft. stayed mast, 6–9 in. diameter .	320–310
Mast radiator of constant cross-section (average)	245
Masts generally	$Z_0 = 60\left(\log_e \frac{\text{Height}}{\text{Radius}} - 1\right)$

Top Loaded Aerials

Fig. 7 shows that the radiation efficiency of short aerials is greatly reduced by the unavoidable resistance, R_e, of the earth connection, coupling circuits, etc., becoming comparable with the radiation resistance R_r. Capacitance loading can be added to the top of an aerial in the form of horizontal wires, to increase the total electrical length H, and thus the value of R_f, without the need of higher masts. The top wires must be arranged so that the radiation from them is small by making them in the form of a T or like the spokes of a wheel. In Fig. 8 the arms of a T aerial are shown Δ radians long and have a characteristic impedance Z_0' and the aerial is electrically lengthened by β radians where

$$Z_0 \cot \beta = 0{\cdot}5\, Z_0' \cot \Delta \quad . \quad . \quad . \quad (10)$$

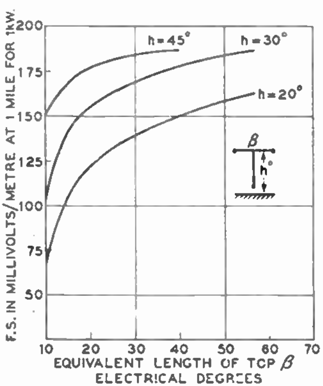


Fig. 10.—Showing the Improvement Effected by Top Capacitance Loading of a Low Aerial with a Poor Earth System.

Z_0 being the characteristic impedance of the vertical conductors. The improvement in the efficiency of an aerial 0·125 wavelengths high as the top capacitance is increased is shown by Fig. 10. The voltage at the ends of a top-loaded aerial is higher, for the same input power, than on the equivalent straight vertical aerial, for by reference to Fig. 8

$$\text{Voltage at } A = \sqrt{P \,.\, R_r} = V_a$$

$$\text{Voltage at } B = V_a\left[\sin\beta - j\frac{Z_0}{R_r}\cos\beta\right]$$

$$= \sqrt{P}\,.\sqrt{\left[R_r\sin^2\beta + \frac{Z_0{}^2}{R_r}\cos^2\beta\right]}$$

At the ends C the voltage is that at the end of an open-circuited line of length Δ (not β) for an input voltage of V_b,

$$\text{Voltage at } C = V_b \sec\Delta$$

It is usual to limit the maximum voltage on a low aerial to 20 kV r.m.s. (10-kV r.m.s. carrier when amplitude modulated) in order to ease the problem of insulation. It will be seen that some control of the voltage is obtained by the choice of Z_0 and Z_0'.

The band-width of short aerials with large-capacitance tops must be investigated by calculating the field strength at the frequencies of the outer sidebands of a modulated carrier. Compared with the value at the carrier frequency it should not change by more than 1 db.

Vertical Radiation Patterns

The general expression for the field strength in the vertical plane and in a direction making an angle θ to the aerial is

$$E = \frac{1900}{D}\sqrt{\frac{P}{R_r + R_e}}\left[\frac{\cos(H\cos\theta) - \cos H}{\sin\theta}\right] \quad . \quad (10)$$

which reduces to equation (5) when $\theta = 90°$.

Examples of the vertical patterns given by aerials of various heights are shown in Fig. 11, from which it will be noted that high-angle radiation is reduced relatively to the ground wave ($\theta = 90°$) as the aerial height increases. This effect is important in the design of anti-fading aerials.

Anti-fading Aerials

As will be seen from Fig. 11, aerials over 0·5 wavelengths in height radiate less sky wave than shorter ones, and the power is concentrated nearer the horizontal plane. Interference at distances where the ground wave is attenuated becomes less, and such aerials have "anti-fading" properties. The angle at which the sky wave is a minimum, θ_0 can be chosen to make the service area of the ground wave a maximum by controlling the height of the aerial; the optimum angle will depend on the conductivity of the ground, the frequency and the height of the layer which reflects the sky wave. An anti-fading aerial is therefore designed for a particular set of conditions, and one design will not necessarily apply to other sites. The optimum value for θ_0 when the ground conductivity is of average value is shown in Fig. 12, and will be

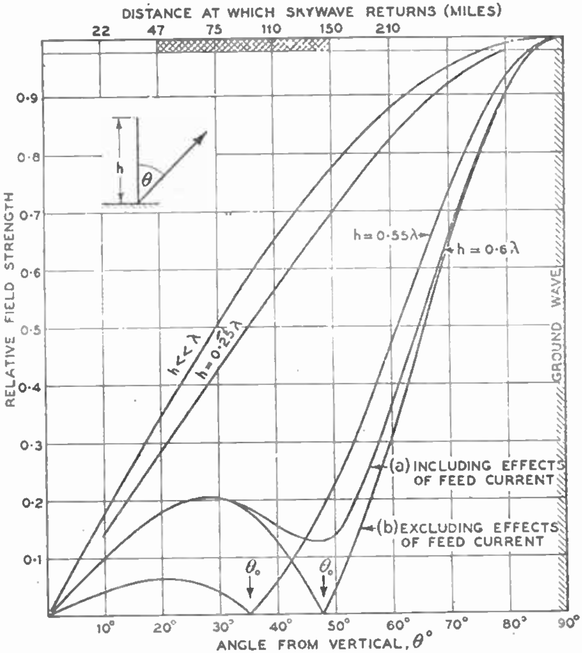


FIG. 11.—RADIATION FROM VERTICAL AERIALS.

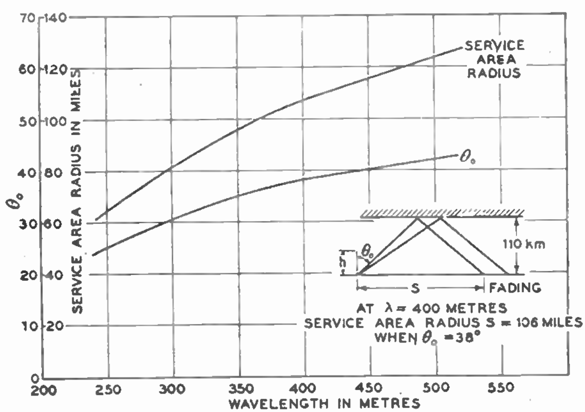


FIG. 12.—OPTIMUM VALUE FOR θ_0 FOR GROUND PATH CONDUCTIVITY = 10^{-13} e.m.u.

(*B.B.C. Quarterly, Vol.* 11, *No.* 3)

seen to lie between 25° and 42° to the vertical for the common broadcasting wavelengths. Interference from adjacent-channel transmissions and man-made static usually reduces the ground-wave service area to the 5 mV/metre contour, so that transmitter powers of over 25 kW are necessary to serve over 50 miles, and anti-fading aerials can do little to enlarge the service area of low-power installations.

The sharp minima in the vertical radiation patterns shown in Fig. 11 are not realized in practice for a number of reasons, of which the most important is the effect of non-sinuosoidal current distribution along the aerial when its diameter is not negligible. A closer approximation to the actual distribution is obtained by considering the current to be

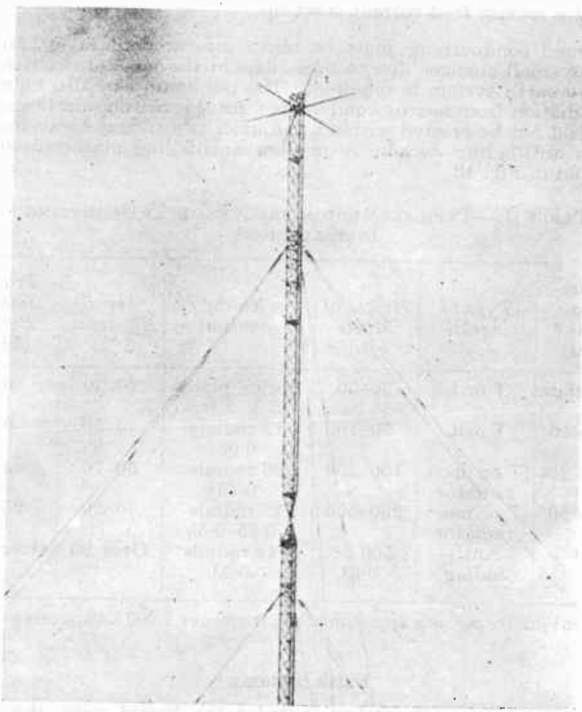


FIG. 13.—A MODERN ANTI-FADING MAST RADIATOR.

The lattice-steel mast radiator for the B.B.C. Third Programme station at Daventry. Height 725 ft. Held vertical by three sets of stays attached at heights of 210 ft. (not shown), 420 and 640 ft. The mast is poised on an insulator capable of withstanding 100,000 volts. 460 ft. above the base the mast is divided by another similar insulator. The mast is fed at this point through a twelve-wire transmission line inside the structure.

(*B.B.C.*)

the sum of two components, one sinuosoidal, the primary current, and the other a linearly distributed feed current maintaining the primary and in quadrature with it. The feed current contributes only a small fraction of the total field in the horizontal direction, but at the angle for which the primary current field is zero (θ_0) the feed-current field still exists, and so tends to " fill in " the minimum as shown in Fig. 11, curve (*a*). This effect can be reduced by :

(*a*) using a very thin wire, i.e., a high value for Z_0, as the radiator to reduce the magnitude of the feed current;
(*b*) feeding the aerial at the current loop;
(*c*) feeding at the current loop and at the base in such a manner that a reverse feed current is set up.

The ground conductivity must be taken into account in (*c*), and to eliminate small changes due to variations in the moisture content an extensive earth system is required. The minimum can also be spoilt by re-radiation from nearby conductors: for this reason, masts and the like should not be erected within a radius of two to three wavelengths from an anti-fading aerial. A modern anti-fading mast radiator is illustrated in Fig. 13.

TABLE 1.—TYPICAL MEDIUM-FREQUENCY TRANSMITTING INSTALLATIONS

*Service Area Radius * (miles)*	*Type of Aerial*	*Height of Masts (ft.)*	*Earth System*	*Aerial Efficiency (%)*	*Transmitter Power (kW)*
Less than 5	T or L	30–50	Buried plates 2 × 2 ft.	20–30	0·1
Up to 10	T or L	50–100	12 radials 0·05λ	40–50	0·5
Up to 25	T or mast radiator	100–200	36 radials 0·25λ	50–70	5·0
Up to 50	T or mast radiator	250–500	72 radials 0·25–0·5λ	70–90	25
100	Anti-fading	500 or 0·6λ	144 radials 0·5λ	Over 90	Over 150

* To 5 mV/metre contour approximately, frequency 1,000 kc/s, average soil.

Earth Systems

The cost of an earth system includes the buried wire, the land, fencing and upkeep. Economics demands that the additional cost of an extensive system be considered in relation to the improvement in service area it is expected to give. Generally speaking, extensive systems can be justified only for high-power transmitters using anti-fading aerials in order to achieve the greatest possible sky-wave reduction and high radiation efficiencies. The radial wires should then extend at least 0·5 wavelengths from the base of the aerial, but for low-power installations a radius of 0·05 wavelengths is sufficient if land is expensive. Table 1 provides a rough guide for the selection of an earth system.

Soft copper wires, not smaller than 0·08 in. dia., can be laid 2 ft. deep by means of a mole plough if the soil is reasonably free from large stones, so allowing the site to be cultivated for most crops. The radial wires should be brazed to a sheet or strip of copper laid near to the base of the aerial and to which all earth connections from the coupling circuits should be made. At high-power installations this central connection is usually a mesh of wires some 20 ft. square immediately under the aerial and, in the case of a mast radiator, is taken over the concrete foundation block and connected to the foundation bolts and other "dead" metalwork. Enclosures housing coupling circuits should be completely screened with wire netting bonded to the mesh at frequent intervals; it is often convenient to provide such buildings with copper damp courses to act as the common link between internal earths, screens and the main buried system. When a restricted earth system is used, it is essential either to connect all nearby metalwork to it or to ensure that this is well clear of wires carrying radio-frequency currents. High losses can occur in such objects as fencing wires if they run close to a low aerial.

Construction

A range of tubular steel and lattice masts, 50–150 ft. high and capable of taking horizontal head loads up to 1,000 lb. is available for supporting T and L aerials. Guys should be sectionalized into lengths not exceeding 0·1 wavelengths with "egg" insulators, one insulator being close to the mast and another close to the bottom of the guy, with the metalwork below the latter connected to the earth system. It is usually best to have all steelwork hot-dip galvanized to reduce maintenance costs, screw-threaded parts being thoroughly greased and protected from the weather. The steel-wire rope halyards supporting the aerial must be anchored at a distance from the base sufficient to prevent them touching the mast when whipping in the wind, and one should be connected to a counterbalance weight some 10 per cent less than the maximum permissible mast-head tension. Sufficient travel should be allowed to relieve the tension when the aerial becomes heavily coated with ice.

Stranded cadmium–copper conductors—7/0·064 in. is suitable—spaced 3–4 ft. apart and insulated at the ends by two porcelain rod insulators are adequate for simple aerials, provided the maximum voltage set up at the ends of the aerial is kept below 20 kV r.m.s.

Mast radiators of constant cross-section are now tending to replace self-supporting and guyed cigar-shaped structures, particularly when anti-fading characteristics are required. When power is fed in at the current loop an insulator is inserted at that point and a transmission line is run through the lower portion of the mast with the outer conductors bonded at regular intervals to the steelwork, and the inner conductors taken to the upper portion of the mast immediately above the insulator. This, and the base insulator, may need to be fitted with corona rings and rainshields to prevent discharges during periods of heavy rain. A triangular section is usually chosen for mast radiators, as this requires less steel and fewer guys and guy insulators, with a consequent saving in the total cost.

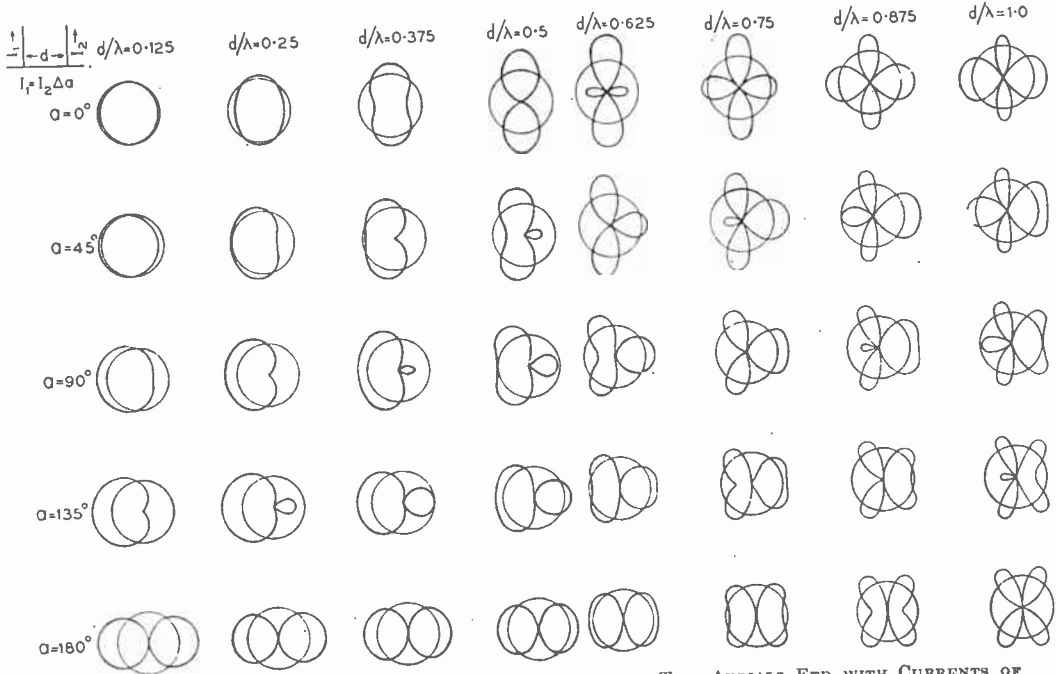


FIG. 14.—HORIZONTAL RADIATION PATTERNS FOR AN ARRAY OF TWO AERIALS FED WITH CURRENTS OF EQUAL MAGNITUDES.

(*G. H. Brown, Proc. I.R.E., January* 1937.)

Directional Aerials

When the service areas of two transmitters operating on adjacent frequency channels overlap, mutual interference can be reduced by suitably shaping the horizontal radiation patterns of their aerial systems. Two or more vertical radiators can be spaced apart and fed with currents of different magnitudes and phases to produce a variety of non-uniform diagrams, some of which are shown in Fig. 14. A cardioid, or heart-shaped, pattern is easily obtained, and is especially useful for preventing radiation in a direction 180° from that required. The principles outlined for the calculation of the diagrams of high-frequency directional aerials can be applied to the medium-frequency case. It is possible to use vertical aerials shorter than 0·25 wavelengths in directional systems, but the resulting pattern is liable to some fluctuation caused by changes in the individual currents, particularly of their relative phase, arising from the effect of climatic conditions on insulators and the ground conductivity.

Medium-frequency Transmission Lines

Co-axial cables, either buried directly or protected by earthenware pipes, are generally the most economical form of transmission line for low-power medium-frequency installations and, coiled up, are a convenient means of adjusting the current phase in directional aerial systems. Rigid copper-tube lines may occasionally be justified, even though they are considerably more expensive than the unbalanced open-wire form, which is often chosen for powers in excess of 50 kW. An open-wire line can be designed with low insulator and copper losses and, provided the outer conductor wires form an efficient shield round the inner conductor, the induced earth currents, and hence the total attenuation, will be small. Four outer wires symmetrically disposed on an 18-in.-diameter circle around one or two inner wires, and supported in spans of about 100 ft. by 12-ft.-high light-steel frames provide a cheap and reasonably efficient means of conveying powers up to 100 kW. A more efficient line has been built to feed the mast radiator, Fig. 13, which is sited 7,500 ft. from its transmitter, consisting of No. 6 gauge copper conductors arranged as shown in Fig. 15 and supported 20 ft. above ground on steel beam sections. This has an efficiency of 93 per cent at 1,000 kc/s, and will carry 250 kW. Table 2 gives the characteristics of a number of commonly used transmission lines.

$$Z_0 = 276\log_{10}\frac{d}{\sqrt{ab}} - 138\,\frac{\left[\log_{10}\frac{d}{b}\right]^2}{\log_{10}\frac{2h}{b}}$$

$$a = r_1\left[\frac{n\rho_1}{r_1}\right]^{\frac{1}{n}}$$

$$b = r_2\left[\frac{m\rho_2}{r_2}\right]^{\frac{1}{m}}$$

$$d = r_2$$

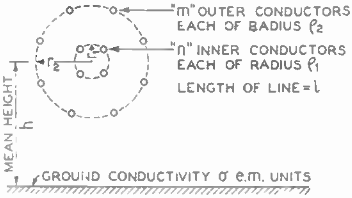


FIG. 15.—MULTI-WIRE UNBALANCED TRANSMISSION LINE.

T

TABLE 2.—TRANSMISSION LINES FOR MEDIUM-FREQUENCY USE

Form of Line	*Characteristic Impedance* (*ohms*)	*Approx. Maximum Power, kW* (1,000 *kc/s*)	*Approximate Loss, db/100 ft.* (1,000 *kc/s*)	*Length in Feet for 90 per cent Efficiency* (1,000 *kc/s*)
Cable—solid polythene dielectric, 1 in. O.D.	50–80	16–24	0·08	550
Cable—semi air-spaced, 1½ in. O.D.	75	40	0·023	2,000
Copper tube, 5 in. O.D.	78	150	0·01	4,500
Open wire :				
5 wires, No. 6 gauge	310	100	0·02	2,250
6 wires, No. 6 gauge	230	150	0·015	3,000
12 wires, No. 6 gauge	220	Over 250	0·005	9,000

The mechanical construction of medium-frequency transmission lines of the open-wire type closely follows power-frequency practice for secondary distribution, except that :

(*a*) conductor wires being closer together, shorter spans are necessary to preserve the spacing ;
(*b*) low-capacitance rod insulators are used ;
(*c*) jumpers must preserve the characteristic impedance of the line ;
(*d*) particular care is taken to earth all outer conductors by a low-impedance connection at each support.

Coupling into Vertical Aerials

Power may be fed into a vertical aerial by any of the following methods :

(1) At the base (this is the simplest method).
(2) At the top, i.e., between the capacitance loading and the vertical portion, in order to achieve a slight gain in forward field on short aerials. Due to mechanical complications it is rarely adopted.
(3) By a shunt connection 0·05–0·1 wavelengths from the base. No base insulation is necessary—a great advantage in regions subject to severe electrical storms. The feed connection should be at an angle not less than 30° to the radiator. The method is not recommended for anti-fading radiators.
(4) At the centre of a half-wave radiator. This reduces the effect of the feed current on the vertical radiation pattern.
(5) At the centre and base. This is a special method used for anti-fading radiators.

The transmitter–aerial system forms a " generator and load " arrangement requiring the usual equivalence of impedances for maximum power transfer, and it follows that the transmission line must be impedance matched to both ends by impedance transforming networks. When unbalanced lines are used to feed vertical aerials unbalanced balanced transformers are unnecessary and coupling circuits simplified. In many cases an inductance in series and a parallel capacitor forming an L section is sufficient, the values being chosen so that by the addition of one element the aerial impedance has either a series or parallel resistive term equal to the characteristic impedance of the transmission line; the second element is then added to cancel the reactive term. The conversion is readily computed from the equivalent expressions, series and parallel, for an impedance,

$$R_s + jX_s = R_p, X_p. \quad R_p = R_s(1 + Q^2) \qquad . \qquad . \quad (11)$$

$$X_p = X_s\left(1 + \frac{1}{Q^2}\right) \qquad . \qquad . \quad (12)$$

$$Q = \frac{X_s}{R_s}$$

Whenever possible the capacitance is put in parallel with the aerial to provide a low-impedance path to earth for currents at harmonic frequencies, but more complicated band-pass circuits may be necessary in some cases.

Common Aerial Operation

Transmitters on a number of different frequencies can be fed into a common aerial by the use of rejectors, consisting of parallel-tuned circuits, to prevent interaction in the output circuits. Each rejector has to pass a current at the frequency of the transmitter to which it is connected in addition to a circulating current of the frequency to which it is tuned. When the difference between the passed and rejected frequencies is small, the total current in the elements of the rejector circuit is many times greater than the passed current, for if

$$\frac{\text{Pass frequency}}{\text{Rejector frequency}} = n = \frac{F_p}{F_r}$$ and the Q of the rejector is greater than 50.

$$\text{Current in inductive element on frequency } F_p = \frac{I_p}{n - \frac{1}{n}} \times \frac{1}{n} \qquad . \qquad . \qquad . \quad (13)$$

$$\text{Current in capacitative element on frequency } F_p = \frac{I_p}{n - \frac{1}{n}} \times n \qquad . \qquad . \qquad . \quad (14)$$

All circuits should be well screened to prevent coupling between them and the aerial or transmitter; a distance of at least one diameter should be allowed between inductance coils and the screen. The inductance of all connections should be negligible in relation to the circuit values. There should always be a D.C. path between the aerial and earth so that static charges cannot build up. A quarter-wave-length of No. 18 gauge wire wound on to a 2–3-in.-diameter former and connected directly between aerial terminal and ground is quite suitable if no other path exists. A horn or ball gap should be fitted between the lowest part of the aerial and ground and adjusted so that it will flash over at a voltage which is about 20–50 per cent greater than the maximum working peak voltage. The setting can usually be obtained with sufficient accuracy by adjusting the gap to flash over at 100 per cent modulation and then increasing the spacing by 50 per cent.

Siting Medium-frequency Aerials

The most important features of a good site are :

(*a*) good ground conductivity up to 5–10 wavelengths of the aerial ;

(*b*) an elevation which is approximately equal to the average elevation of the service area and free from pronounced undulations ;

(*c*) proximity to the centre of the service area ;

(*d*) freedom from tall buildings and overhead power lines ;

(*e*) good accessibility to all services, power, telephone, water, roads, etc.

The relative importance of these features will depend on the power to be radiated. For low-power installations (*e*) may be rated before (*a*), while (*c*) is essential. As the power to be radiated increases, the ground constants become more important, and anti-fading mast radiators should be sited so that (*a*) and (*b*) are achieved.

HIGH-FREQUENCY AERIALS

The high-frequency range extends from 3,000 to 30,000 kc/s, and its main use is for transmission to territories 500–10,000 miles distant by employing the sky wave, bent or reflected by the ionosphere. In zones of high noise level which occur throughout the tropics, frequencies between 2,500 and 7,000 kc/s are used for local broadcasting, the sky wave being radiated at high angles of elevation so that it is reflected back to ground over an area surrounding the transmitting site.

Obviously, the angle at which the wave strikes the reflecting layer determines at what distance it will again strike the ground, and this incident angle will be largely determined by the angle of departure from the aerial. The vertical radiation pattern of the aerial is therefore of first importance, and the maximum amount of energy must be concentrated about the best angle for the distance over which transmission is required. Further, the distant reception area will subtend a relatively small angle in azimuth at the transmitter, so that the radiation can also be confined in the horizontal direction. The effective power in the direction required can therefore be much greater, for the same power input to the aerial, than in the case of a medium-frequency transmitter radiating equally in all horizontal directions. The ratio

$$\frac{\text{Power per unit solid angle in favoured direction}}{\text{Average power per unit solid angle}}$$

is an important parameter of the aerial, and is known as its *directivity*, or *power gain over an isotropic source.*

The field strength produced by an aerial at a distant point is proportional to the square root of the power fed into it, and inversely proportional to the distance, so that the factor $E \cdot D/\sqrt{}$ (kW) will serve to compare aerials; in the direction of maximum-field strength it is related to the directivity by :

$$d = \text{Directivity} = (E_m \cdot D)^2 \times \frac{1}{3 \times 10^4}$$

E_m being the maximum field strength in millivolts/metre and D the distance in kilometres.

The gain in the direction of maximum field strength is defined as $10 \log_{10} d$ db above that of a source radiating equally in all directions through a sphere (isotropic source). The field strength produced by a single half-wavelength dipole in free space, i.e., when far above ground, is not, of course, uniform through the sphere surrounding it, but is maximum in the plane passing through the centre of the dipole at right angles, and is given by

$$E_m = 60 I_0/D \text{ mV/metre}$$

I_0 being the current at the centre in amperes and D the distance in kilometres. An array of many dipoles will, in some direction, produce a maximum field strength, say K times that of a single dipole carrying the same current at its centre as each dipole in the array, and the radiation resistance of the array will be :

$$R_r = \frac{\text{Total input power}}{(\text{Dipole current})^2}$$

then $$E_m = \frac{60K}{D}\sqrt{\frac{1{,}000}{R_r}} \text{ for a radiated power of one kilowatt.}$$

$$\therefore\ E_m \cdot D/\sqrt{\text{kW}} = \frac{1{,}900K}{\sqrt{R_r}}$$

The calculation of K and R_r involves considerable labour for complicated arrays, K being derived from vector addition of the individual field strengths, and R_r from the self and mutual impedances of the dipoles.

The incident field strength at the receiver depends upon the directivity and power fed into the aerial, and the length of the path traversed between ground and ionosphere. In addition, the following factors must be taken into account :

(1) *Absorption in Ionosphere.*—Derived from charts prepared from world-wide observations (see Circular No. 462 of National Bureau of Standards, U.S.A.).

(2) *Interference between waves which have traversed slightly different and variable paths.*—Reduces median field strength by a factor 0·83 (−1·6 db). (The median field strength is that value which is exceeded for 50 per cent of the time.)

(3) *Random changes in polarization of the wave.*—Reduces median field strength by a factor 0·7 (−3 db).

(4) *A Ground-reflection loss when path is by more than one hop.*—A factor of 0·63 allowed for each reflection from the ground.

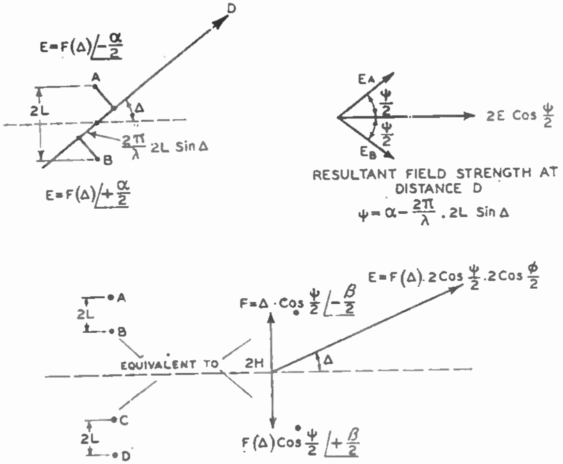


Fig. 16.—Derivation of Vertical Radiation Patterns.

The Calculation of Radiation Diagrams

The array factor K is derived in the following manner. Fig. 16 shows two radiating sources A and B, spaced a distance $2L$ apart, each having a radiation pattern in the plane of the paper which is given by $E = F(\Delta)$. The sources carry equal currents differing in phase by α°. At a distance D very much greater than L the instantaneous field strength, at a point bearing Δ° to the perpendicular to AB, due to source A is

$$F(\Delta) \,.\, \cos\left[\left(\omega t - \frac{\alpha}{2}\right) - \frac{2\pi}{\lambda}(D - L\sin\Delta)\right] \qquad . \quad (19)$$

and due to B

$$F(\Delta) \,.\, \cos\left[\left(\omega t + \frac{\alpha}{2}\right) - \frac{2\pi}{\lambda}(D + L\sin\Delta)\right] . \qquad . \quad (20)$$

the resultant field having a magnitude

$$F(\Delta) \,.\, 2\cos\psi/2 \quad . \qquad . \qquad . \quad (21)$$

where
$$\psi = \alpha - \frac{2\pi}{\lambda} \,.\, 2L\sin\Delta.$$

Thus A and B can be replaced by a source located centrally between them and having a radiation diagram $2F(\Delta)\cos(\psi/2)$. Two such sources spaced $2H$ apart and carrying currents differing in phase by β will similarly be equivalent to one with a radiation diagram, $F(\Delta) \,.\, 2\cos(\psi/2) \times 2\cos(\phi/2)$ where $\phi = \beta - 2\pi/\lambda \times 2H\sin\Delta = F(\Delta) \,.\, K$

when the primary sources are half-wave dipoles for which, as shown above, $F(\Delta) = \text{constant} = \frac{1{,}900}{\sqrt{R_r}}$ mV/metre at unit distance and 1 kW input.

The radiation diagram of any number of sources can be derived in this manner: i.e.. by multiplying the diagram of one by an "array factor". Similarly, the effect of the ground can be taken into account by considering the image of the array as carrying an equal current differing in phase by π radians. If A and B are the aerial elements carrying equal co-phased currents and C and D are their images, $\alpha = 0$, $\beta = \pi$, and the diagram of the array AB, when placed with its centre at a height H above ground, will be

$$E = F(\Delta) \,.\, 2 \cos\left(\frac{2\pi}{\lambda} L \sin \Delta\right) \times 2 \sin\left(\frac{2\pi}{\lambda} H \sin \Delta\right). \quad (22)$$

The second factor is called the Ground Reflection Factor R_h when, as is the case for horizontal polarization, the ground can be assumed to have a reflection coefficient ρ equal to 1 and a phase-change angle α equal to π. In other cases.

$$R = \sqrt{(1 - \rho)^2 + 4\rho \cos^2\left(\frac{\alpha}{2} - \frac{2\pi}{\lambda} H \sin \Delta\right)} \quad (23)$$

Now consider four co-phased radiators spaced half a wavelength apart in free space,

$$2L = \frac{\lambda}{2} \quad \text{and} \quad 2H = \lambda$$

$$E = F(\Delta) \,.\, 4\left[\cos\left(\frac{\pi}{2}\sin\Delta\right) \times \cos\left(\pi \sin \Delta\right)\right]$$

$$= F(\Delta) \,.\, \frac{\sin\left(4\,\frac{\pi}{2}\sin\Delta\right)}{\sin\left(\frac{\pi}{2}\sin\Delta\right)} \quad . \quad . \quad . \quad . \quad (24)$$

The general expression for M such radiators placed with their centre point at a height H above a perfect earth can be derived in a similar manner and will be found to be

$$E = F(\Delta)\left[\frac{\sin\left(M\,\frac{\pi}{2}\sin\Delta\right)}{\sin\left(\frac{\pi}{2}\sin\Delta\right)}\right] . \, 2\sin\left(\frac{2\pi}{\lambda} H \sin\Delta\right) \quad . \quad (25)$$

$$= F(\Delta) \times K$$

In the case of a horizontal dipole $F(\Delta)$ is constant in the plane normal to the dipole and passing through its centre point. In planes containing the dipole the field strength is a function of the angle θ between the dipole and the direction considered given by

$$F(\theta) = \frac{\cos\left(\frac{\pi}{2}\sin\theta\right)}{\cos\theta} \quad . \quad . \quad . \quad (26)$$

The product of (25) and (26) then gives the field strength in any direction defined by Δ and θ from the array.

The total radiation resistance of an array, such as Fig. 17, of M horizontal dipoles carrying co-phased equal currents is the sum of their individual resistances, as modified by the presence of the other dipoles. If mutual resistance is denoted by terms of the form R_{nm} the radiation resistance of No. 1 dipole in the presence of the others and the images in the ground will be $R_r + R_{12} + R_{13} + \ldots \quad \ldots R_{1\,2M}$ and of No. 2 dipole $R_r + R_{21} + R_{23} + R_{24} + \ldots \quad \ldots R_{2\,2M}$

A negative value must be given to the mutual-resistance terms due to the images, for they carry reversed currents. Adding the individual radiation resistances so derived gives that of the whole array,

$$R_r = R_{(\text{No. }1)} + R_{(\text{No. }2)} + \ldots \quad \ldots R_{(\text{No. }M)}$$

Mutual resistance values are tabulated in Table 3, and a full range will be found in the original paper by Pistolkors (*Proc. I.R.E.*, Vol. 17, March 1929). The dipole resistances are clearly not equal but, by end

TABLE 3.—MUTUAL RESISTANCE

Spacing in Line of Elements, d (*Wavelengths*)	*Spacing Perpendicular to Elements*, h (*Wavelengths*)					
	0	0·5	1·0	1·5	2·0	2·5
0	+73·3	−12·3	+4·1	−1·8	+1·2	−0·75
0·5	+26·4	−11·1	+8·9	−5·7	+3·8	−2·8
1·0	− 4·1	− 0·8	+3·6	−6·3	+6·0	−5·7
1·5	+ 1·8	+ 0·8	−2·9	+2·0	+0·2	−2·4
2·0	− 1·0	− 1·0	+1·1	+0·6	−2·6	+2·7
2·5	+ 0·6	+ 0·4	−0·4	−1·0	+1·6	−0·3

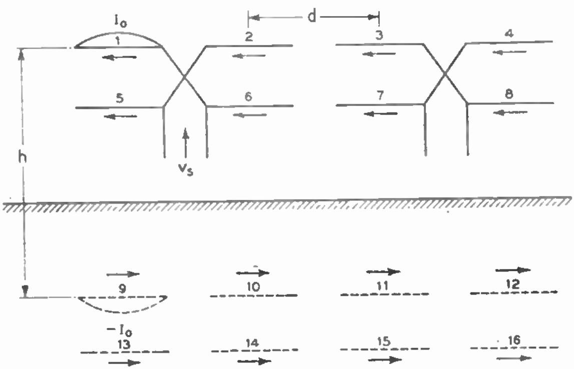


FIG. 17.—RADIATION RESISTANCE OF DIPOLE ARRAYS (*h* AND *d* AS IN TABLE 3).

feeding, as shown in Fig. 17, the currents will be equal, provided the characteristic impedance all dipoles are of similar construction. However, the radiation pattern is changed very little, even when the currents differ considerably, particularly for large arrays.

Practical Aerial Arrays

The following forms of aerial are in general use on high-frequency services :

(a) *Local Broadcasting, 0–200 miles, in zones of high noise level.*—Broadside arrays of dipoles in the horizontal plane 0·15–0·3 wavelengths above ground, radiating upwards between 90° and 60° to the horizontal.

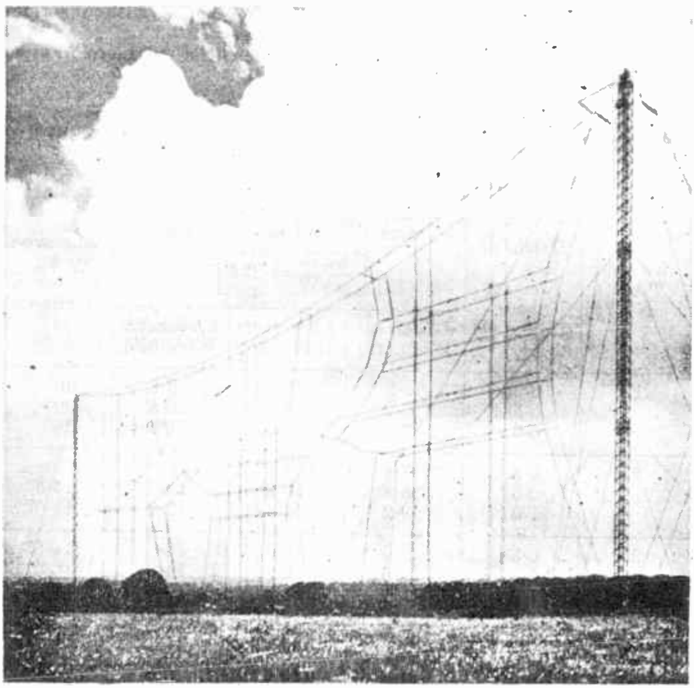


FIG. 18.—B.B.C. AERIAL ARRAY, DAVENTRY.

Three aerials, each with a reflector, suspended between two masts. Each aerial is of the form H.R. 4/4/1·0.

(*I.E.E. Journal*)

(b) *For Broadcasting and General Communications up to 500 miles.*—Broadside arrays of two tiers of half-wave elements in the vertical plane, the lowest being 0·3–0·5 wavelengths above ground, see Table 4.

(c) *For Broadcasting and Point-to-point Services over 500 miles.*—Broadside arrays of eight to sixteen dipoles arranged in tiers of four with reflector curtain, height above ground of lowest dipole 0·5–0·8 wavelength. End-fire arrays of half-wave dipoles spaced 0·25 wavelengths. Rhombic aerials. Stacked rhombics. Vertically polarized " Franklin " arrays.

Of the above types, the broadside array as shown in Fig. 18 in its various forms is the most generally useful but requires higher masts than do rhombics for the same elevation angle of the main lobe. The main beam of radiation can be slewed in azimuth up to 16° and its direction reversed through 180°. Rhombic aerials have the advantage that they will operate over a wider range of frequencies without serious loss of efficiency and can be reversed but not slewed.

Table 4 summarizes the gain and angle of elevation of the main lobes of a number of arrays of the broadside type.

TABLE 4.—GAIN AND ANGLE OF ELEVATION OF THE MAIN LOBES OF BROADSIDE ARRAYS

No. of Tiers Vertically	*Height of Lowest Element above Ground (λ)*	*Gain with Reference to Half-wave Element in Free Space* $\left[\frac{E_m D}{\sqrt{kW}} = 222\right]$			*Elevation of E_m Degrees to Horizontal*
		1 Element Wide (db)	*2 Elements Wide (db)*	*4 Elements Wide (db)*	
1	0·25 0·5 0·6	5·2 6·2 7·0	6·8 8·2 9·0	9·8 10·8 11·6	90° 30° 25°
2	0·3 0·5 1·0	7·8 8·8 9·2	9·8 10·8 11·4	12·4 13·4 13·8	22° 17·5° 11·0°
3	0·3 0·5 1·0	9·0 10·0 11·0	10·8 12 12·8	13·6 14·6 15·6	14·5° 12·5° 8·5°
4	0·3 0·5 0·8 1·0	10·4 11·0 11·8 12	12·4 13 13·8 14	15 15·6 16·4 16·6	10·5° 9·5° 8·0° 7·5°

A reflector adds approximately 3 db to the forward gain given in columns 3, 4 and 5.

Horizontal Dipole Arrays

The characteristic impedance of a twin-wire dipole element consisting of two 12 S.W.G. wires spaced 6 in. is about 390 ohms, giving an end-feed point impedance for a co-linear pair equal to 3,600 ohms. The effect of a reflector is to double the radiation resistance of each element, thus halving the feed-point resistance. Each half of a sixteen-element array fitted with a similar reflector curtain has, therefore, an impedance of 1,900/4 = 475 ohms. The vertical feeder usually consists of two stranded cadmium–copper wires of 7/0·064 in. construction with a characteristic impedance of 600 ohms. To match the whole array to a transmission line the two halves are connected together by equal lengths of feeder, and the impedance is measured at the junction. Assuming symmetry, the impedance of each half of the array at the junction is twice the measured value, and the standing wave ratio (S.W.R.) on the vertical feeders can be derived. Suitable correcting lines are added to the vertical feeders so that each is matched, and the junction point impedance is then close to 300 ohms. This method takes the mutual impedance between the halves of the array into account. Matching is referred to under " Transmission Lines ".

Reflectors

A reflector increases the forward field strength by nearly 3 db and, of course, reduces the field in the reverse direction to a low value, so minimizing interference with other services operating on the same frequency. It is common practice to build the radiator and reflector curtains in the same form so that the direction of radiation can be reversed, either curtain being tuned as a reflector for the other by short-circuiting the vertical feeders at a point which is an odd number of quarter-wavelengths from the lowest element. The optimum position can be found with the aid of a small pick-up loop placed some 5–10 wavelengths either in front or behind the array, forward-to-backward field strength ratios of 10 : 1 being readily obtainable. The efficiency of a reflector increases with the number of elements of which it is composed, and though the spacing from the radiator is not critical, 0·25 wavelengths is usually adopted. 0·15 wavelength spacing is about the minimum, but this reduces the band-width of the whole array and makes the impedance very sensitive to deflections produced by wind acting on the elements of the array.

Slewing

The main lobe of radiation can be deflected, or slewed in azimuth, by advancing the phase of one half of the array relative to the other, the maximum slew boing 16°. Further phase advance produces serious distortion of the lobe with a reduction of field strength in the required direction.

Mechanical Design

Horizontal dipole arrays consisting of two curtains, radiator and reflector, each of sixteen elements impose a vertical load of about 1,500 lb. and require masts up to 350 ft. in height, two masts usually supporting three arrays for three different frequencies. The dipoles of adjacent arrays should be separated by at least one-half of the longer wavelength and a similar separation used between the dipoles and the masts. Horizontal rigging wires within half a wavelength from a dipole should be sectionalized with light insulators spaced 5 ft. and vertical wires sectionalized into 10-ft. lengths. Porcelain rod insulators 8–12 in. long are suitable for dipoles carrying up to 25 kW each. Neither the masts nor their guys need be insulated, but the main supporting ropes between the mast heads should be split into 20-ft. lengths.

The design of the rigging which preserves the form of the array is best done by graphical analysis, and involves a certain amount of "trial and error", since the only fixed points are the mast heads, and the loads on the main support ropes do not act along parallel lines.

Vertical Incidence Arrays

For general purposes a single half-wave dipole supported 0·25 wavelengths above ground is suitable, as it projects a broad lobe of radiation vertically upwards with half-power points at 30° to the horizontal in the plane perpendicular to the element and at 53° in line with the element. The area included in the 70 per cent field-strength contour is 700 × 300 km when the E layer is operative and 900 × 2,000 km by reflection from the F_2 layer at night. Higher gain arrays composed of parallel dipoles spaced 0·5 wavelength in a horizontal plane 0·15–0·3 wavelength above ground can be designed for particular purposes. The radiation pattern will be the product of the array factor and the ground-reflection factor as determined from equations (25) and (26), remembering that the angle for the array factor will be measured from the vertical and must be replaced by $\pi/2 - \Delta$ in the ground-reflection factor.

The Rhombic Aerial

Broadside arrays are composed of resonant elements spaced by a distance determined by the operating wavelength; thus the range of frequencies over which one array will operate satisfactorily is obviously limited by the change of its dimensions in terms of the wavelength. The rhombic aerial is composed of non-resonant conductors which are terminated by a resistance equal to their characteristic impedance to give an input impedance which varies very little as the applied frequency

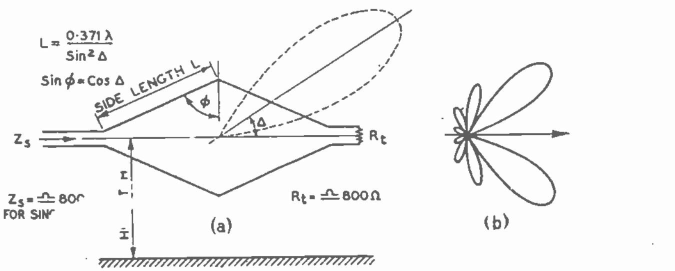


FIG. 19.—THE RHOMBIC AERIAL.

(a) The aerial, (b) radiation pattern of each side (when $L = 2\lambda$) of terminated rhombic.

is changed. A long wire parallel to the ground and terminated in a suitable resistance carries a travelling wave of current and radiates power in the general direction of the wave motion with a maximum inclined to the wire. This angle varies with the length of the wire in terms of the wavelength, becoming smaller as the length increases. Four similar wires can be arranged so that the radiation from each is directed along the same bearing, as shown in Fig. 19, and terminated in a resistance to absorb non-radiated power, which would otherwise be reflected back towards the input and give rise to a lobe of radiation directed 180° from the primary lobe.

The directivity of a rhombic aerial is governed by the length of the wires forming the sides and the height above the ground plane, being a maximum for an angle $\Delta°$, when $H = \lambda/4 \sin \Delta$, and sides of length $0{\cdot}5\lambda/\sin^2 \Delta$. Δ is not the angle at which the radiation pattern is a maximum but this can be aligned on $\Delta°$ by making the length of each side $0{\cdot}371\lambda/\sin^2 \Delta$. Charts for determining the radiation patterns of rhombics will be found in the Department of Scientific and Industrial Research, *Special Report No.* 16 (Radio Research).

When a rhombic aerial is used for transmission, the terminating resistance must be capable of dissipating about one-third of the total power fed into the aerial, and it is usual to employ a dissipative transmission line having conductors of iron or resistance alloy to give an attenuation of 10 db. The radiation efficiency of a terminated rhombic is given by :

$$= (1 - e^{-R_r/Z_0}) \times 100 \text{ per cent} = 70 \text{ per cent approx.} \quad (27)$$

where R_r = the radiation resistance = 700 ohms approx.;
Z_0 = the characteristic impedance of the conductors as a transmission line
= approx. twice the value of the terminating resistance.

Construction of Rhombics

The outstanding advantages of this aerial over the broadside array are its broad frequency response and the simple mechanical construction which can be used.

Either steel or wooden poles are suitable for supporting rhombic aerials, and unbroken guys have little deleterious effect on the radiation pattern. The aerial can be reversed by reversing the positions of feeder and termination resistance and, since no high standing waves occur on the conductors, the voltages are generally lower than for the equivalent broadside array. Conductors formed from two wires in parallel further improve the frequency characteristic, and the aerial can be matched to its transmission line by an exponential transforming section, 1 wavelength long at the lowest operating frequency, in place of frequency selective stubs.

Sites

Installations which carry a world-wide service require up to 400 acres of land to allow the arrays to be spaced apart sufficiently to prevent interference between them. Ideally, sites should be level, free from trees and similar obstructions, and far from ranges of hills. Ground sloping downwards in front of an array tilts the beam downwards by an amount equal to the angle of slope, and similarly upwards when the ground slope is upward. The height of the array is measured perpendicularly to the ground, and is therefore reduced by the ground slope a small amount. The general effect of undulations is to distort the vertical pattern at a distance by non-uniform ground reflections, particularly for high-gain arrays radiating at low angles. It is preferable to avoid placing an array closer than some 1,000 ft. to another in the direction of the main beam. The ground itself should have good conductivity.

Transmission Lines

The open-wire balanced form of transmission line is well suited to use in the high-frequency range, being easily erected and repaired, and having a sufficiently low attenuation for route lengths of up to about 1 km. The open construction allows simple matching devices to be fitted at any point, and the measurement of the current along the line can be made with an ammeter and pick-up loop. Carrier powers of more than 100 kW can be handled by a line constructed throughout from readily obtainable materials and erected by linesmen with no specialized knowledge of high radio frequencies.

A characteristic impedance of between 300 and 600 ohms is most suitable, the lower value being obtained by the use of four wires spaced on the corners of a rectangle and the voltage applied across the horizontal side. A single pair of wires gives the higher value of Z_0 derived from the expression

$$Z_0 = 276 \log_{10} \left[\frac{2S}{d}\right] \quad . \quad . \quad . \quad (28)$$

which for the four-wire construction becomes

$$Z_0 = 138 \log_{10} \left[\frac{2S}{d} \sqrt{(m^2 + 1)}\right] \quad . \quad . \quad . \quad (29)$$

where S = horizontal spacing between conductors;
d = diameter of conductors;
$m = \dfrac{\text{horizontal spacing}}{\text{vertical spacing}}$.

The attenuation at a frequency F (Mc/s) when a standing wave of ratio a, ($= I_{max.}/I_{min.}$) exists on the line is

$$K \cdot \left(a + \frac{1}{a}\right) \sqrt{F} \text{ db/km}$$

K being 0·23 for two-wire and 0·16 for four-wire lines. Attenuation can be neglected in calculating the input impedance, Z_s, and deriving the matching required on an open-wire line feeding power to an array then

$$Z_s = Z_0 \; \frac{a + j \tan 2\pi L/\lambda}{1 + ja \tan 2\pi L/\lambda} \quad . \quad . \quad . \quad (30)$$

where L = the length of line from input to a position of $I_{min.}$

Or in terms of admittance

$$Y_s = Y_0 \cdot \frac{\frac{1}{a} + j \tan 2\pi L/\lambda}{1 + j\frac{1}{a} \tan 2\pi L/\lambda} \qquad (31)$$

At some point on the line the admittance looking towards the load will have a real component equal to Y_0 and a susceptance which can be cancelled by a reactance stub connected in parallel with the line at that point. Then between that point and the input the line will be terminated by Y_0 (or Z_0) and is matched.

Most of the problems associated with the matching of transmission lines are simplified by the use of a calculator which relates impedances, lengths and standing-wave ratios by a family of orthogonal circles and provides a rapid means of solving equations (30) and (31). A polar form of such a calculator has been devised by P. H. Smith (*Electronics*, 1939, Vol. 12, p. 29, and Vol. 19, p. 130) which is applicable to lines of any characteristic impedance through the use of "normalized impedances". An impedance is normalized by dividing it by the characteristic impedance of the transmission line associated with it. Thus equations (30) and (31) are normalized by dividing through by Z_0 in one case and Y_0 in the other.

VERY-HIGH-FREQUENCY AERIALS

Frequencies between 30 and 300 Mc/s comprise the "Very-high-frequency" (V.H.F.) range and are suitable for communication up to about 50 miles, although at the lower end reception at greater distances is frequently possible.

Broadcasting, television, point-to-point communications, amateurs and navigational aids all occupy this range, each service being allocated various bands of frequencies.

V.H.F. aerials fall into three broad categories having either:

(a) a nearly uniform radiation pattern in the horizontal plane, but high directivity in the vertical plane for broadcasting;

(b) a highly directional pattern in both planes for point-to-point communications;

(c) specially shaped radiation patterns for navigational aids, particularly aircraft-landing equipment.

Vertical and horizontal polarization is used in all three categories; directivities greater than in the medium- and high-frequency ranges are usual, since the resonant elements are physically much shorter. The majority of aerials consist of a number of resonant elements, either slots, wires or loops grouped to give the required radiation diagram, but non-resonant aerials, such as the rhombic and the helix, have important applications.

The directivity and power gain of any aerial system over which the current is sensibly uniform are directly proportional to its vertical extension. The mean power gain G relative to a half-wave dipole in free space is then approximately equal to 1·22 times the vertical height in wavelengths occupied by the aerial elements, that is the number of

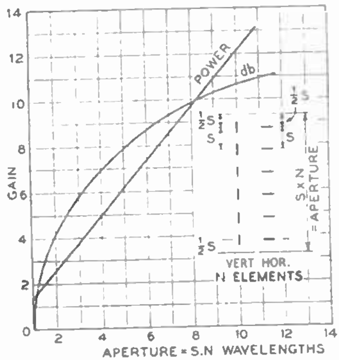


FIG. 20.—GAIN OF STACKED RADIATORS.

elements multiplied by the spacing in wavelengths between elements as shown in Fig. 20. The width to the half-power points of the main lobe of radiation in the vertical plane is approximately $61°/G$.

Aerials for Television Broadcasting

The design must achieve:

(a) the required power gain;
(b) an input impedance characteristic which does not vary by an amount sufficient to cause the radiation of delayed images;
(c) a specified horizontal radiation pattern, usually circular.

The power gain will be related to the service area required and the power output of the transmitter. The effective radiated power (E.R.P.) of the installation is the product of the power output of the transmitter and the gain of the aerial after all losses in the transmission lines and the like have been deducted. Transmitter power, aerial gain and aerial height above ground must be chosen as an economical combination, since deficiencies in one can be compensated by adjustments to the other two, but each can be adjusted only by discrete amounts. Thus it may be cheaper to double the transmitter power instead of the aerial gain. The effective height of the transmitting aerial is not always equal to its height above the site itself, but above some mean datum applying to the service area as a whole, so the field strength over the area will not be directly proportional to the height of the mast. These, and other site considerations, enter into the figure decided upon for the aerial gain.

Impedance Matching

It is a relatively simple matter to match the aerial impedance very closely to the characteristic impedance of the transmission line feeding it at the carrier frequency, but if this matched condition does not hold over the range of sideband frequencies some of the sideband power will be reflected back to the transmitter, and by a second reflection return to the aerial. The time taken for this double traversal of the transmission line will be an appreciable fraction of the duration of one line of the transmitted picture (approximately 100 microseconds) if the transmission line is over 250 ft. long, so that the delayed radiation of a portion of the power will produce a second and distorted picture at the receiver. This is detectable if the input impedance differs from the characteristic impedance by more than ±2 per cent at the carrier and ±17 per cent at the outer sideband frequencies. An increase in diameter reduces the variation of susceptance with frequency of a centre-fed cylindrical radiator, and additional compensation, in the form of a susceptance varying in the reverse manner, can be added in parallel at the feed point. A short-circuited line has this property,

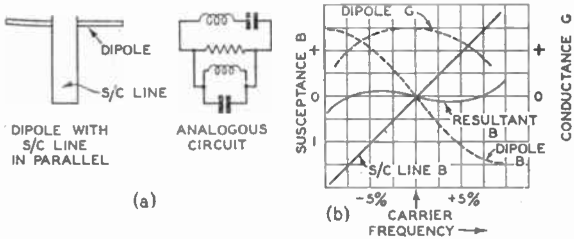


Fig. 21.—Band-width of Dipole with Short-circuited Line in Parallel.

and when its length and characteristic impedance are correctly chosen the band-width of the combination can be two to three times that of the element alone.

Fig. 21 (*a*) shows the circuit analogous to a dipole with a short-circuited line in parallel, the former being series resonant with resistance and the latter parallel resonant. The broken curves of Fig. 21 (*b*) indicate how the conductance and susceptance of the dipole change over a small frequency range about the resonant frequency and the full curve the corresponding susceptance of the short-circuited line. When the two susceptances are added, their resultant is very nearly zero, over a range of ± 5 per cent of the carrier frequency, and the dipole admittance is practically equal to the conductance term alone. This form of compensation is inherent in the folded dipole, since the two sides of the fold, Fig. 22 (*a*), act as parallel short-circuited lines across the feed point, equal and opposite currents flowing internally as shown. The radiation current flows on the outer surface, Fig. 22 (*b*), and divides equally between the two limbs of the fold. The input impedance is therefore four times that of an open dipole of similar construction.

A low value of characteristic impedance—about 25 ohms—gives the best compensation of the dipole susceptance variations, and the element is therefore made from a wide strip, closely folded. The practical form of a high-power television aerial using folded dipole elements can be seen in Fig. 23. The compensating line can be used to support an open dipole element and, by moving the element towards the short-circuited end, the slope of the compensating susceptance can be varied between wide limits without changing the characteristic impedance of the line. The slope is proportional to $1/\sin^2(2\pi x/\lambda)$, x being the distance from the short-circuited end.

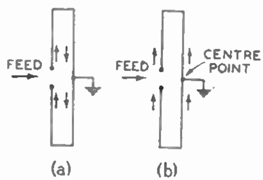


Fig 22.—The Folded Dipole.

Band-width of Tiered Arrays

When tiers of dipoles are built up to form a high-gain array, the band-width of the whole assembly is influenced by the following additional factors, which also have a material effect on the radiation patterns :

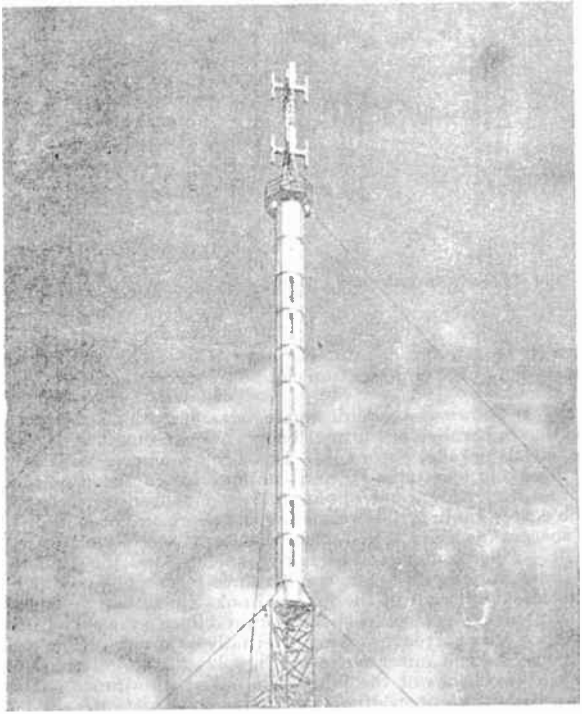


FIG. 23.—HIGH-POWER TELEVISION TRANSMITTING AERIAL USING FOLDED DIPOLES.

The main mast of the B.B.C. Holme Moss station is 750 ft. high and weighs over 90 tons. The close folded dipoles formed from wide strip at the top are for Television broadcasting while the slot aerials in the tubular section are for V.H.F. sound broadcasting.

(*B.B.C.*)

(*a*) the mutual impedance between one element and all others in the array;

(*b*) the variation of the electrical length with frequency of inter-connecting feeders and impedance transformers;

(*c*) the presence of the supporting mast.

The calculation of the effects is laborious, and the solution is usually derived from experiments on small-scale models of the array: in fact reliable plots of the radiation patterns can only be made on such models when placed well away from reflecting surfaces.

Arrays of Half-wave Dipole Elements

A single vertical dipole supported well clear of other conductors has a circular horizontal radiation pattern (H.R.P.), and a number of dipoles placed in line. 1 wavelength apart between centres, could be used to give power gain in the horizontal plane without distorting the pattern. Mechanically such an array is difficult to build as a self-supporting structure, as insulators capable of withstanding the considerable bending moment, and of sufficiently small dimensions, are not available. The elements have to be supported by a mast running parallel to them. This distorts the radiation pattern and affects the band-width (i.e., the frequency range over which the impedance is approximately constant). Other groupings must be adopted, e.g., four co-phased dipoles arranged in a ring 0·4 wavelength in diameter round a mast which is less than 0·1 wavelength in diameter or across the diagonal, give a horizontal radiation pattern uniform to within ± 1 db. This variation can be reduced by decreasing the ring diameter and by increasing the number of dipoles in the ring, but difficulties with the band-width and feeding are introduced, so that four is about the optimum number. One ring has a gain slightly greater than a single dipole element and, by adding further rings each spaced 0·9 wavelengths from the centre of its neighbour, the gain is increased by approximately 3 db every time the number of rings is doubled. The relation between the total length of an array and its gain is shown by Fig. 20.

The presence of guy wires near an array of vertical dipoles would distort the horizontal radiation pattern, so the supporting mast must stand as a cantilever extension to the structure below it and, to withstand the bending moment due to a number of rings, cross-sections exceeding 0·1 wavelength become necessary. Then more pronounced maxima and minima occur in the horizontal-radiation pattern, but by an adjustment of the ring diameter the variation can be limited to ± 2 db for eight tiers of rings on a square mast of 0·36 wavelength side. Eight elements on a ring diameter of 1·28 wavelength, with an inner ring of reflectors 0·78 wavelength in diameter, were used at Alexandra Palace to achieve a near-circular diagram on a mast 0·65 wavelength across.

Intentionally non-uniform diagrams can be obtained by using one, two or three elements only in the ring, leaving one or more faces of the mast open.

In the horizontal plane the field strength of one horizontal dipole varies approximately as the cosine of the angle between the radial and the perpendicular through the centre of the dipole. When two are arranged at right angles to each other on a common centre point and fed with equal currents differing in phase by 90° the resultant pattern is sensibly circular. Vertically mounted tiers of such pairs spaced a wavelength apart form the Turnstile array. Since half the input power is fed into each element of the pair, the gain of a single pair of crossed dipoles is half that of the standard half-wave dipole, and two pairs spaced 0·7 wavelength vertically are practically equivalent to a single vertical dipole. At this spacing the mutual resistance is at its maximum negative value and the horizontal gain is a maximum, but as the number of tiers is increased the optimum spacing approaches 1 wavelength.

Horizontal dipoles can be arranged in a ring round a central mast, four being the usual number, the ring diameter being carefully chosen to compensate for the effect of currents induced in the mast. Even when this is done appreciable variations in the horizontal radiation pattern can occur, for instance, the maximum/minimum ratio is 3 db for a square mast with 0·27-wavelength sides, 4 db for 0·37 sides, and then falls to about 3·6 db when the sides are 0·5 wavelength. More nearly uniform patterns are, of course, obtained when more dipoles are used in the ring, but again the multiplicity of feed points introduces its own problems of impedance transformation and matching.

Alternative modes of phasing the currents in a ring of four vertical or horizontal dipoles are possible :

(*a*) all are driven in the same phase, i.e., the currents have the same direction of rotation round the ring when horizontal (Fig. 24 (*b*));

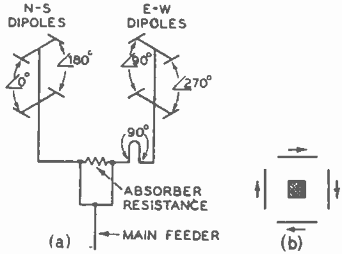
(*b*) the phase is advanced 90° progressively round the ring.

The first gives a more uniform horizontal radiation pattern, and the second less impedance variation, i.e., a greater band-width for the whole array. In (*b*) diametrically opposite dipoles are fed from a common feeder, and similarly orientated pairs in all the other tiers are fed in phase so that the array has then two feed points. These are joined together by feeders which differ by 90 electrical degrees in length, that is one feeder is 0·25 wavelength longer than the other, and delays the phase of the currents in those pairs by 90°, with the result that the relative phases round the ring are 0°, 90°, 180°, 270°. Since the two halves of the array are identical, any deviation of impedance with frequency occurs equally on each, but at the common input connection the reflections set up by such deviations will be 180° out of phase, one having, in effect, twice traversed the added quarter-wavelength of feeder. The reflections cancel, and the input impedance of the whole array remains unchanged. This holds until the frequency band is so wide that the added length of feeder differs significantly from 90°, so that the cancellation of reflections is not complete. Advancing phase, or as it is more commonly known, quadrature feeding, is used in the Super-turnstile array. It should be noted that although the main transmission line to the array input is correctly terminated over the frequency band unequal powers are fed into the two halves, because the impedance of one half is as much greater than the mean as the other is less. The reflected power is not cancelled or absorbed at the junction : it travels on to the other half of the aerial and distorts the horizontal radiation pattern. If the interconnecting feeders are many wavelengths long the delay will be apparent at the receiver through the appearance of a ghost image of the prime picture, but this can be prevented if the reflected power is absorbed by the circuit shown in Fig. 24 (*a*).

It will be apparent that impedance transformation occurs many times in a multi-element array, since each individual feeder must be correctly terminated in its characteristic impedance. Further, the transformers used must be insensitive to changes of frequency within the working range. When the impedance ratio required is small a single quarter-wavelength of a line with a characteristic impedance equal to $\sqrt{r_1 \times r_2}$ can be used (r_1, r_2 being the resistances to be matched), but for larger ratios a double quarter-wave transformer is preferred, the first section

FIG. 21.—ARRAYS OF HALF-WAVE DIPOLES.

(a) Quadrature feeding with power equalization.
(b) Horizontal dipoles in phase feeding.



having a characteristic impedance of $\sqrt{[r_1\sqrt{(r_1 + r_2)}]}$ and the second of $\sqrt{[r_2\sqrt{(r_1 \times r_2)}]}$. Any further correction of susceptance variations is done at selected points on the feeders by introducing, at those points, lengths of line having equal and opposite variations after the manner adopted for the compensation of dipoles.

No insulators are necessary for the mounting of dipole elements on a mast, as the attachment can be made to a point which is electrically equidistant from the ends. For folded dipoles this is the centre of the continuous side of the fold; for open dipoles it is the short-circuit on the supporting stubline compensator. There is no difficulty, therefore, in feeding a balanced element by an unbalanced line, such as co-axial cable, since this has merely to follow the metallic path from mast to feed point, with the outer conductor always in contact and the inner connected to the opposite side of the feed point. The outer conductor, in some designs actually carries the radiating current on its outer surface, the penetration of the current into the conductor being so small in depth that internal and external surface currents cannot interact.

Vertical Slot Aerials

A slot about 0·05 wavelength wide and 0·5 long cut in a large conducting sheet is equivalent, when fed across the centres of the long sides, to a half-wave dipole with the magnetic and electric fields interchanged. The electric field of a vertical slot is horizontal, and the radiation pattern in the horizontal plane is a circle when the sheet is infinitely large. A finite sheet reduces the pattern to a figure of eight with minima along the surface of the sheet. If one considers the slot in relation to a strip dipole which would just fill the slot, the following relationship exists between their centre-point impedances :

$$Z_{\text{slot}} \times Z_{\text{dipole}} = \frac{(377)^2}{4}$$

Since Z_{dipole} is approximately 70 ohms, Z_{slot} is equal to 500 ohms and the band-widths are comparable; wide slots having a greater band-width than narrow ones.

A nearly circular horizontal radiation pattern is given by a slot cut in a cylindrical surface having a diameter of about 0·1 wavelength, and by cutting two or more slots in larger cylinders. If a ratio of maximum to minimum field strength of not more than 3 db is required, the cylinder diameter must not exceed :

0·13 wavelength for 1 slot;
0·175 wavelength for 2 slots;
0·4 wavelength for 3 slots;
0·8 wavelength for 4 slots.

When cylinders with circumferences of about 1 wavelength are used, internal resonances tend to occur and fields are set up which interfere with the field round the slot. It thus becomes difficult to excite the slot in the normal manner by feeding across the centre. The resonance effect is greatly reduced by fitting internal screens behind each slot, so that each is, in effect, a single slot in a small-diameter cylinder, though the screens can follow the chords of the circular section. The resonant length of a slot is increased to about 0·8 wavelength by the addition of screens. As in the case of dipoles, slots can be arranged in vertical tiers to increase gain in the horizontal plane; the same approximate relationship between total aperture and gain holds good (see Fig. 20). The distance between the centres of the slots is usually of the order of 1 wavelength.

It is generally more convenient to feed by an unbalanced line rather than a balanced pair of conductors, as the former can be made to follow the internal surface of the cylinder and has no stray field to interact with the field in the cavity behind the slot. The feeder will not be matched owing to the high impedance of the slot, and the band-width will be impaired, but methods equivalent to the folding of a dipole can be applied which result in the feed-point impedance being divided by a factor of four.

Slotted cylinder aerials have the very important mechanical advantage over dipoles in that they present a smooth surface to winds, so that the total load on the supporting mast is often less than for the equivalent dipole aerial. The slots themselves can be masked by suitable sheets of insulating material such as " Perspex ". Distribution feeders can safely be run in the space between the screens of a slotted cylinder, but care is taken to ensure that they are bonded to the screens at frequent intervals and, as far as possible, to place the feeders in the angle between adjacent screens.

The Super-turnstile Array

A form of radiator intermediate between the horizontal dipole and the vertical slot is employed in the super-turnstile array which was developed in the following manner. Horizontal currents flowing in the sheet surrounding a slot decrease rapidly with the distance from the slot edges, and become negligibly small at more than 0·25 wavelength, so the removal of those portions of the sheet beyond this distance will have little effect. The sheet can also be skeletonized, leaving only seven horizontal bars to carry current, and the impedance/frequency characteristic then becomes similar to that for a single horizontal dipole, most of the current being concentrated in the centre bars. On shortening these, as shown in Fig. 25, the end impedance is raised and more of the current is forced to flow in the outer bars, and the vertical-plane radiatior pattern now approximates to that of two horizontal radiators spaced ½ vavelength apart, and the band-width is appreciably broadened. One element of this form can be adjusted so that its impedance changes by less than 10 per cent over a band of frequencies equal to 25 per cent of the mid-frequency. When a number of the elements is arranged in two vertical, and perpendicular, planes about a central support mast and fed in advancing phase, a high-gain aerial with a reasonably uniform H.R.P. results. The central mast, usually a cylindrical tube, has little effect on the performance when its diameter is less than 0·1 wavelength, but nevertheless the maximum gain is limited by the need to keep the diameter within this figure.

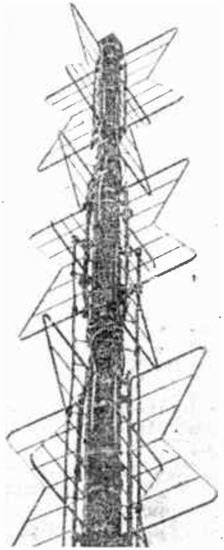


FIG. 25.—THREE-STACK SUPER-TURNSTILE TELEVISION AERIAL.

(*Marconi's Wireless Telegraph Co. Ltd.*)

The elements are spaced vertically 1 wavelength apart between centre points, and each is 0·7 wavelength long and 0·5 wavelength wide at the ends. They are fed by co-axial cables cleated to the central pole, and then to the outer edge of the slot and terminating across the centre of the slot and the pole. The other half of the same element is fed by a cable connected in the reverse manner, i.e., outer conductor to pole and inner to slot edge, and each cable is correctly terminated when the impedance of each " half slot " is made 70 ohms.

Distribution feeders and impedance-matching transformers are all accommodated on the pole, and require careful placing to avoid the introduction of stray coupling between the two perpendicular planes of elements. Adjustments can be done at ground level, the complete aerial being hauled into position in one piece.

Loop Aerials

A circular loop of wire carrying a uniform current gives rise to a circular radiation pattern in the plane of the loop with the electric field in that plane, so that tiers of horizontal loops can be used to concentrate radiation into the horizontal plane equally in all directions. The difficulty is to produce a uniform current round the loop, since, if current is fed in at one point, the periphery must be small in terms of the wave-length, and the radiation resistance will be small also. The difficulty can be largely overcome by dividing the loop into resonant sections on which the current distribution will be approximately sinusoidal, but the combined effect will be similar to a uniform current round the loop. Each section will require its own feed point, and the loop becomes a series of horizontal dipoles with the currents all flowing in the same direction round the ring.

The Kandoian array is a high-gain aerial of this form, and uses triangular loops with half-wave resonant sides very similar in form to a folded dipole. These are supported at the electrical centre by a co-axial line which feeds round the fold to the central feed point so that the outer surface of the line does actually form the radiating surface. The three co-axial lines are then paralleled at the centre of the triangular loop to give an impedance of about 50 ohms. A maximum number of sixteen vertically tiered loops gives an array operating in the 200-Mc/s range with a gain of seventeen times that of a half-wave dipole. The loops are co-phased, and by the use of a special balun transformer they are paired in series so that the common input impedance is kept at a workable value. If N loop feeders were joined directly in parallel the impedance would be $50/N$, by the series–parallel method it is $2 \times 50/\frac{1}{2}N$, or 12·5 ohms for the sixteen-loop array. Mutual coupling between pairs of loops narrows the band-width of the array, but the effect is reduced by fitting non-resonant screens between every other loop. All the feeder cables and transformers are housed inside the triangular mast, together with the cables supplying power to de-icing heaters in each element and aircraft warning lanterns.

The Helical Aerial

As in the case of the rhombic array, a wire many wavelengths long radiates by virtue of the travelling wave of current on it. If the wire is coiled into a helix and complete turns are an integral number of wave-lengths long, all points lying on a line parallel to the axis of the helix will carry currents which differ in phase by integral multiples of 2π radians. In other words, the fields due to these currents will be additive along all radials in the plane perpendicular to the axis of the helix, and a beam of radiation will be produced. The vertical pitch will give rise to a small vertically polarized component, but if right- and left-handed spirals starting from a central point are used, this component can be largely cancelled. In its practical form the aerial is composed of turns 2 wavelengths in circumference (these are wavelengths in the wire taking the velocity of propagation into account) five wound right-handed, and five left with a pitch which makes the total axial length 5 " free space" wavelengths. This forms one unit and is fed at the common centre point, further units being added to give the required total gain. The current along a single helix decreases progressively from input onwards, and the beam produced in the vertical plane would be asymmetrical; it will be evident that this asymmetry is reduced by twinning two helices into a unit as described. A tubular mast lying along the axis acts as the support for the turns, and also as the outer of the co-axial line feeding the units, the inner conductor being brought out through bushing insulators. Stalk insulators of low permittivity and loss hold the turns in position on the mast, and the ends of each unit are connected to the mast, but this does not set up any appreciable reflection owing to the high attentuation along the helix due to radiation. The beam of radiation can be tilted downwards by simply rotating one or more units of the array round the mast so that a progressive phase advance of the currents is achieved.

Gap Filling and Beam Tilting

The field strength in the vertical plane of M, current co-phased elements spaced 2π radians (1 wavelength) apart has been show (equation 25) to vary as the factor:

$$\frac{\sin\left(\frac{M \times 2\pi}{2} \times \sin \Delta\right)}{\sin\left(\frac{2\pi}{2} \times \sin \Delta\right)}$$

which is zero when $\Delta = \sin^{-1} \frac{n}{M}$. $n = 1, 2, 3 \ldots M - 1$

If the array is H ft. above the surrounding district, zones of low field strength corresponding to these zeros will occur at $H\sqrt{\left[\left(\frac{M}{n}\right) + 1\right]}$ ft. from the array, the farthest being, for example, about $1\frac{1}{8}$ miles when $H = 1{,}000$ ft. and $M = 8$. Although this low field strength is usually adequate, there is a possibility of distorted television reception due to the movement of the pattern along the ground with modulation, and large relative changes may occur in the zones. This can be overcome by de-phasing one tier of elements near the centre of the array or by feeding more power to the upper half—a power ratio of 70 : 30 is often used—so tilting the beam downwards. A tilt can also be produced by progressively advancing the phase of the tiers from the bottom upwards. By beam tilting the field strength can be made nearly independent of distance up to 20 miles from the aerial.

F. D. B.

GREAT-CIRCLE BEARINGS AND DISTANCES

Great-circle Routes

The effective use of radiated energy in H.F. and V.H.F. point-to-point services is profoundly influenced by the accuracy with which the bearings of the stations and the distances between them are computed. Errors of a few minutes of arc in the orientation of highly directional arrays may seriously reduce the directional gain, the signal-to-interference ratio and the duration of communication periods. The greater the distance and the sharper the energy concentration, the greater is the need for accuracy in calculation and applying the results to setting out the arrays.

Electromagnetic waves travel by great-circle routes over the earth's surface, since an arc of a great circle is always the shortest route between any two points on the terrestrial globe (see Fig. 26). A great circle is defined as any one of an infinite number of circles formed by a plane passing through the centre of the earth, where it cuts the surface.

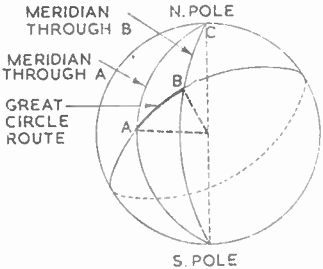


Fig. 26.—Great-circle Routes.

The arc of a great circle is always the shortest route between any two points on the terrestrial globe.

Great-circle Bearings

The great-circle bearings of two stations A and B (Fig. 27) are found by solving the spherical triangle ACB formed by the great circle passing through A and B and the meridians or great circles passing through A and B and the poles. They are calculated from the formulæ:

$$\tan \{(b - a)/2\} = \frac{\sin \{(l_b - l_a)/2\}}{\cos \{(l_b + l_a)/2\}} \cot c/2 \quad . \quad . \quad (1)$$

$$\tan \{(b + a)/2\} = \frac{\cos \{(l_b - l_a)/2\}}{\sin \{(l_b + l_a)/2\}} \cot c/2 \quad . \quad . \quad (2)$$

where a = great-circle bearing of B from A, degrees;
b = great-circle bearing of A from B, degrees;
c = difference of longitude between A and B, degrees (the included angle at the pole);
l_a = latitude of A, degrees;
l_b = latitude of B, degrees.

To simplify calculation, take B as the place of greatest latitude. When dealing with distances of some thousands of miles, the angular quantities $(b - a)/2$ and $(b + a)/2$ must be evaluated with five- or six-figure trigonometrical tables to obtain the necessary degree of accuracy. For distances of a few hundred miles, three- or four-figure calculations will suffice.

Latitudes north of the equator are written positive.
Latitudes south of the equator are written negative.

If $l_b + l_a$ is negative, calculation is simplified by regarding this quantity as positive, giving bearings West of South, and calling the result East of North.

Great-circle Distances

The great-circle distance between A and B is calculated from the equation :

$$\tan d/2 = \frac{\sin \{(b + a)/2\}}{\sin \{(b - a)/2\}} \tan \{(l_b - l_a)/2\} \quad . \quad . \quad . \quad (3)$$

where d = great-circle distance, degrees.

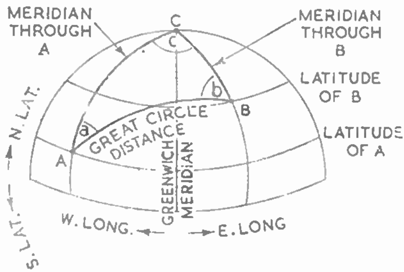


FIG. 27.—GREAT-CIRCLE BEARINGS AND DISTANCE OF TWO STATIONS.

In order to convert the result to linear measure, minutes and seconds must first be expressed in decimals of a degree to the nearest third decimal place before multiplying by one of the following factors :

d, statute miles $= 69{\cdot}057\, d^\circ$
d, nautical miles $= 60{\cdot}000\, d^\circ$
d, kilometres $= 111{\cdot}136\, d^\circ$

EXAMPLE 1 : A directional high-frequency service is to be established between two stations situated near London and Singapore. Find the great-circle bearings at each site and the great-circle distance between sites, given the positions.

London : Long. 0° 7′ W. Lat. 51° 30′ N.
Singapore : Long. 103° 58′ E. Lat. 1° 34′ N.

London is the place of greatest latitude.

Hence
$$l_a = 1^\circ\ 34'\ \text{N.}$$
$$l_b = 51^\circ\ 30'\ \text{N.}$$
$$c = 0^\circ\ 7' + 103^\circ\ 58' = 104^\circ\ 05'$$
$$(l_a + l_b)/2 = 26^\circ\ 32'$$
$$(l_a - l_b)/2 = 24^\circ\ 58'$$
$$c/2 = 52^\circ\ 02'\ 30''$$

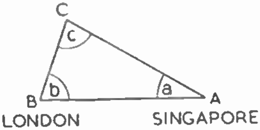


FIG. 28.—EXAMPLE 1.

To evaluate the bearings, take logs of both sides of equations (1) and (2):

$$\tan\{(b - a)/2\} = \frac{\sin\{(l_b - l_a)/2\}}{\cos\{(l_b + l_a)/2\}} \cot c/2$$

$$\log\tan\{(b - a)/2\} = \log\sin\{(l_b - l_a)/2\} - \log\cos\{(l_b + l_a)/2\} + \log\cot c/2$$

	$\log\sin 24° 58' =$	$9{\cdot}625406$
Subtract	$\log\cos 26° 32' =$	$9{\cdot}951665$
		$-\ 0{\cdot}326259$
Add	$\log\cot 52° 02' 30'' =$	$9{\cdot}892159$
	$\log\tan\{(b - a)/2\} =$	$9{\cdot}565900$
	$(b - a)/2 =$	$20° 13'$

$$\tan\{(b + a)/2\} = \frac{\cos\{(l_b - l_a)/2\}}{\sin\{(l_b + l_a)/2\}} \cot c/2$$

$$\log\tan\{(b + a)/2\} = \log\cos\{(l_b - l_a)/2\} - \log\sin\{(l_b + l_a)/2\} + \log\cot c/2$$

	$\log\cos 24° 58' =$	$9{\cdot}957393$
Subtract	$\log\sin 26° 32' =$	$9{\cdot}650034$
		$0{\cdot}307359$
Add	$\log\cot 52° 02' =$	$9{\cdot}892159$
	$\log\tan\{(b + a)/2\} =$	$10{\cdot}199518$
	$(b + a)/2 =$	$57° 44'$

Bearing of Singapore from London

$$b = (b + a)/2 + (b - a)/2$$
$$= 57° 44' + 20° 13' = \underline{77° 57' \text{ E. of N.}}$$

Bearing of London from Singapore

$$a = (b + a)/2 - (b - a)/2$$
$$= 57° 44' - 20° 13' = \underline{37° 31' \text{ W. of N.}}$$

To evaluate the great-circle distance, take logs of both sides of equation (3).

$$\tan d/2 = \frac{\sin\{(b + a)/2\}}{\sin\{(b - a)/2\}} \tan\{(l_b - l_a)/2\}$$

$$\log\tan d/2 = \log\sin\{(b + a)/2\} - \log\sin\{(b - a)/2\} + \log\tan\{(l_b - l_a)/2\}$$

	$\log\sin 57° 44' =$	$9{\cdot}927151$
Subtract	$\log\sin 20° 13' =$	$9{\cdot}538538$
		$0{\cdot}388533$
Add	$\log\tan 24° 58' =$	$9{\cdot}668013$
	$\log\tan(d/2) =$	$10{\cdot}056626$
	$d/2 =$	$48° 43' 30''$

Angular distance = 97° 27′
Linear distance, statute miles = 97·45 × 69·057 = 6,729 statute miles
Linear distance, nautical miles = 97·45 × 60 = 5,847 nautical miles

Determination of Direction of True North

For the purpose of setting out accurately the lines of directional aerials sytems and masts, the direction of True North (the observer's meridian) is established by taking observations with a theodolite of the position of the sun or one of the principal stars, referred to some arbitrary datum line, and noting the time of the observation. The datum line is a line selected between the observer and some outstanding

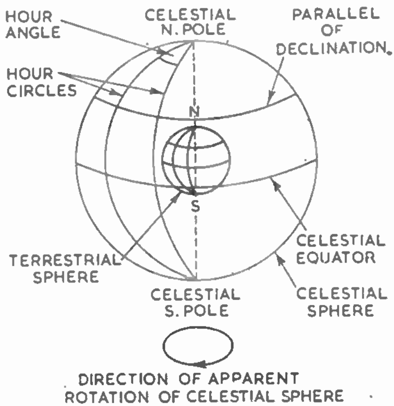


FIG. 29.—DECLINATION AND HOUR ANGLE.

object on the horizon, such as a tree. The bearing of True North is then calculated from the observed declination of the sun, the known latitude of the site and the hour angle by the use of equation (3). It is necessary to define declination and hour angle :

Declination

The latitude of a celestial body on a celestial sphere (Fig. 29). The celestial sphere is an imaginary sphere of infinite radius, analogous to the terrestial sphere and concentric with it, which is considered to rotate about the earth once every 24 hours, the earth being assumed to be relatively stationary.

Hour Angle

The longitude of a celestial body West of the observer, or the number of hours elapsed since the body appeared due South of the observer.

$$\begin{bmatrix}\text{Hour}\\ \text{angle}\end{bmatrix} = \begin{bmatrix}\text{Mean time}\\ \text{of place}\end{bmatrix} \pm \begin{bmatrix}\text{Equation}\\ \text{of time}\end{bmatrix} \pm \begin{bmatrix}12\\ \text{hours}\end{bmatrix}$$
$$= \begin{bmatrix}\text{Local apparent}\\ \text{time}\end{bmatrix} \pm \begin{bmatrix}12\\ \text{hours}\end{bmatrix}$$
$$\begin{bmatrix}\text{Mean time}\\ \text{of place}\end{bmatrix} = \begin{bmatrix}\text{G.M.T.}\end{bmatrix} + \begin{bmatrix}\text{Long. E}\end{bmatrix}$$

The equation of Time is obtained from the Nautical Almanac or other nautical tables.

The hour angle is zero at apparent noon, and is reckoned westward of the site. Since apparent time is reckoned from midnight, 12 hours must be subtracted after midday and added after midnight. If the

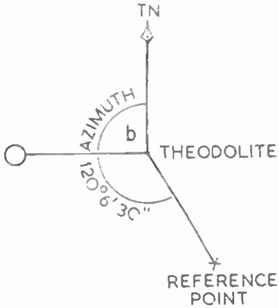


Fig. 30.—Example 2.

hour angle is greater than 12 hours, the sun is East of the site, and the hour angle must be subtracted from 360° to obtain the included angle between the direction of the sun and the direction of the reference line.

It is usual to take three sets of observations and work to the average result.

Example 2: A station is to be sited in a position Long. 9° 09′ W. Lat. 38° 42′ N. The following observations were made on the site with a theodolite of the sun relative to a selected line of reference. Find the direction of the observer's meridian (True North).

Observed horizontal bearing of sun from point of reference	120° 6′ 30″
Times of first and last observations	17 hours 31 minutes 43 seconds B.S.T.
	17 hours 34 minutes 25 seconds B.S.T.
Watch fast by standard time signal	25 seconds

Correction for Time

Times of observations :

1st contact, 17 hours 31 minutes 43 seconds
2nd contact, 17 hours 34 minutes 25 seconds

Mean of times	. . . .	17 hours 33 minutes 4 seconds
Watch fast .	. .	25 seconds
		17 hours 32 minutes 39 seconds
Less summer-time advance	. .	1 hour 0 minutes 0 seconds
		16 hours 32 minutes 39 seconds

Correction for Longitude

Difference for 1°	4 minutes	
Difference for 9° 09′	9·15 minutes × 4 =	36 minutes 36 seconds
Local Mean Time (p.m.)	. .	15 hours 56 minutes 3 seconds

Equation of Time at 16 hours 32 minutes 39 seconds (12/6/1946)

	G.M.T.	*Equation of Time*
From Nautical Almanac {	16 hours	24·0″
	18 hours	22·9″
By interpolation . .	16 hours 32 minutes 39 seconds	23·73″
Local Mean Time . . .	15 hours 56 minutes	3 seconds
Less Equation of Time . .		23·73 seconds
Local Apparent Time . .	15 hours 55 minutes	39·27 seconds

Hour Angle

Local Apparent Time + 12·00 hours	3 hours 55 minutes 39·27 seconds

Declination of Sun at 4 hours 32 minutes 39 seconds G.M.T. (12/6/1946)

	G.M.T.	*Declination*
From Nautical Almanac {	16 hours	23° 08·7′
	18 hours	23° 09·1′
By interpolation . .	16 hours 32 minutes 39 seconds	23° 08·81′

Great-circle Calculation of Azimuth

We now have

Declination, l_a	23° 08·81′ N. = 23° 08′ 49″ N.
Lat. of site, l_b	38° 42′ N.
Hour angle, c	3 hours 55 minutes 39·27 seconds = 58° 54′ 54″
$(l_b + l_a)/2$.	30° 55′ 25″
$(l_b - l_a)/2$.	7° 46′ 36″
$C/2$.	29° 27′ 27″

Applying the great-circle formulæ, equations (1) and (2), and taking logs of both sides, as in the previous example, we have

	log sin 7° 46′ 36″ =	9·131336
Subtract	log cos 30° 55′ 25″ =	9·933413
		−0·802077
Add	log cot 29° 27′ 27″ =	10·248110
	log tan {(b − a)/2} =	9·446033

$$(b - a)/2 = 15^\circ\ 36'\ 15''$$

	$\log\cos\ 7^\circ\ 46'\ 36'' =$	$9{\cdot}995987$
Subtract	$\log\sin\ 30^\circ\ 55'\ 25'' =$	$9{\cdot}710874$
		$0{\cdot}285113$
Add	$\log\cot\ 29^\circ\ 27'\ 27'' =$	$10{\cdot}248110$
	$\log\tan\{(b + a)/2\} =$	$10{\cdot}533223$
	$(b + a)/2 =$	$73^\circ\ 40'\ 20''$

Azimuth $b = (b + a)/2 + (b - a)/2 = \underline{89^\circ\ 16^\circ\ 35''}$

Given the polar diagram of an array, the allowable tolerance in the accuracy of taking observations and staking out the line of the array is best shown by an example.

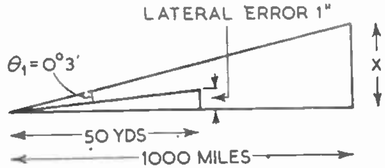


FIG. 31.—EXAMPLE 3.

EXAMPLE 3: The error in computing the bearing of a directional aerial system was 0° 3′ of arc. The lateral error in staking out the line of the aerial amounted to 1 in. in 50 yd. in the same direction. Find: (*a*) the total angular error and (*b*) the distance between the line of maximum propagation and the true direction 1,000 miles distant.

(*a*) Angular error in computation, θ_1 . . . 0° 3′
Angular error in staking out (for very small angles),
$\theta_2 = \text{arc tan } (1/50 \times 36)$
$= \text{arc tan } 0{\cdot}000556$ . . . . 0° 1·9′ approx.

Total angular error, θ_3 . . . . 0° 4·9′

(*b*) Linear error at 1,000 miles distance

$$x = d \tan \theta_3$$
$$= 1{,}000 \times 0{\cdot}00143$$
$$= \underline{1{\cdot}43 \text{ miles}}$$

W. E. P.

CHOICE OF SITE

The choice of a transmitting or receiving site is influenced by the nature of the service, the mode of propagation and the frequency. There are, however, several primary considerations which have a very general application.

Site Characteristics.—In the first place it is an advantage to survey two or more prospective sites and make the final selection by comparing their relative technical merits and the cost of land. The site and surroundings must be flat, free from large wooded areas, well drained and not liable to flooding. Although rocky soil is ideal for mast and building foundations, it makes a poor electrical earth. Clay, loam and limestone, which absorbs moisture readily, have good electrical conductivities and make satisfactory earths; dry sand and gravel are less suitable.

A sufficient area of land must be available for setting out masts, stay anchorages and buildings according to plan, allowing for possible future extensions, as costs may rise or the additional land required may have been occupied later.

Accessibility.—The site must be accessible by a made-up road, capable of carrying the heaviest loads during the constructional period, and within reasonable distance of the nearest town and railway. If the site is remotely situated and no convenient means of transport exists, provision for housing and other staff amenities may have to be included in the estimates.

Regulations and Legal Requirements.—An architect should be consulted on the technical, legal and financial aspects of the project before purchasing land, and plans submitted to the local authorities for approval before contracts are placed. Aeronautical regulations governing the erection of masts and the provision of aviation obstruction lights must be complied with.

Power Supply.—A source of public power supply must be available near the site. If the supply is unreliable and subject to frequent failures, the cost of installing generating plant, either for continuous or emergency operation, will have to be taken into account. If voltage variations at the terminals exceed ± 6 per cent, it may be necessary to install automatic voltage regulators.

Communication Lines.—The cost of renting or erecting and maintaining lines to establish connection with the nearest exchange will need investigation.

Water and Fuel-oil Supply.—A water supply of sufficient capacity for cooling generating plant and supplying the needs of personnel should be available locally. Failing this, expert opinion may be sought on the prospect of obtaining a reliable and constant supply by sinking a bore. Facilities should exist for regular deliveries of fuel oil from the nearest town or railway depot.

Influence of Frequency on Choice of Site

To understand how site features depend on the nature of the service, it will be useful to review the propagation characteristics at different frequencies.

Medium-frequency Transmitters

The useful service area for broadcasting and communications at frequencies below 3 Mc/s is limited to the range of the ground wave.

Although the sky wave is effective at a distance beyond the range of the ground wave, the two combine with a random phase displacement to produce a region of signal instability. The range can be usefully extended only by directing the maximum amount of energy into the ground wave and keeping earth losses in the vicinity of the aerial as low as possible. Good soil conductivity on and around the site for a radius at least ten times the radiator height is an important factor in reducing earth losses. They are further reduced by burying a system of radial earth wires extending up to half a wavelenth from the centre of the radiator Iron railings, wire fences and metallic structures. which also absorb energy from the field, should be situated well out of the vicinity of the radiator, and overhead power and communication lines terminated at the site boundary and connected to the building by buried cables.

The site should be level and at an altitude not less than that of the primary service area. The area is determined mainly by the dimensions of the earth system. A half-wave radiator for a broadcasting service, with its earth system, will occupy a circular area having a diameter equal to the height, but a somewhat greater diameter is usually desirable as a precaution against the possibility of the mast falling on adjoining property. Similarly, in the communications field the area occupied by the earth system for a horizontal tee aerial, and its mast, will be roughly circular about the centre immediately below the point of maximum radiation, and should extend well outside the outer stay anchorages. To these dimensions an allowance must be added for boundary fencing.

High-frequency Transmitters

At frequencies between 3 and 30 Mc/s the ground wave is rapidly attenuated by losses in the soil. Radiation projected at sufficiently high angles to clear the ground contours suffers little attenuation. and is reflected alternately back to earth from the ionosphere and upwards from the ground in a succession of skips. For long-range work, aerial arrays are designed to direct the energy at low angles to the horizon, the aim being to bridge the distance between stations with a minimum number of skips. The site and neighbourhood should, therefore, be devoid of hills and trees which would subtend an angle greater than about 5° at the centre.

The site area is determined by the number and disposition of the aerial arrays. Arrays should be laid out as close as practicable to the station building, and the feeder runs made short and direct to minimize transfer losses. Since attenuation in the feeders increases with frequency, those arrays excited at the highest frequencies should be nearest the building. To avoid distortion of the radiation patterns, the arrays must be disposed in such a way that they do not obstruct each other's line-of-fire. Broadside arrays can adjoin each other, so that a common mast supports adjacent arrays, provided their lines-of-fire do not converge unduly. Rhombic aerials and in-line arrays should have a separation of at least one wavelength.

V.H.F. Stations

Transmission at frequencies above 30 Mc/s is propagated mainly by the direct ray. Normally, the sky wave is not reflected back by the

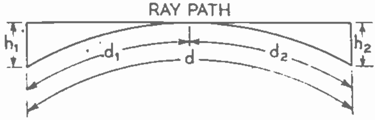


FIG. 32.—VISUAL RANGE OF V.H.F RAY.

ionosphere as it is at lower frequencies. Radiation at angles below the line of sight is attenuated by surface obstructions along the ground and by re-radiation from structures in the path. The closer free-space propagation conditions can be approached, the less will be the attenuation and radiated power required to produce a given signal intensity. For these reasons V.H.F. transmitters and receivers are sited in elevated positions above the surrounding terrain, preferably outside built-up areas, and commanding as nearly as possible a clear optical path. Contour maps of the service area are a valuable aid in selecting a site, which can then be checked by surveying the area and taking field-strength measurements with a mobile transmitter.

V.H.F. aerials are much more compact than high-frequency arrays, and occupy a smaller area of land. The area is determined by the base dimensions of the mast or tower, the station building and the access road. Aerials are erected on light stayed masts or self-supporting towers, usually not more than 100 ft. high. To reduce the length of feeders, towers are commonly built astride the building or closely adjoining it.

V.H.F. Route Planning

The route for a V.H.F. link or chain of stations is planned by preparing a contour elevation similar to that shown in Fig. 33 (*a*) or (*b*) and inserting the ray path. The contours are represented either by reference to the curved surface of the earth drawn to a convenient scale, with the ray path straight, as in (*a*) or to an imaginary flat earth, with the ray path curved to correspond, as in (*b*). An advantage of the flat-earth method of representation is that the contours can be drawn from an ordinary contour map showing heights above sea-level, and the path for the line-of-sight ray and the refracted ray inserted on the same elevation.

The linear distance to the visible horizon varies as $\sqrt{\text{height}}$ of the observer's eye above ground level. If the aerials at either end of the link are raised to heights h_1 and h_2 as in Fig. 32 above ground level, the maximum visual range is given by the sum of the horizon distances d_1 and d_2 from each end, and

$$\begin{aligned} d &= d_1 + d_2 \\ &= c(\sqrt{h_1} + \sqrt{h_2}) \qquad (1) \end{aligned}$$

If the distances are measured in kilometres and the heights in metres, it can be shown from simple plane geometry that the constant $c = 3{\cdot}55$.

Atmospheric refraction, however, causes the ray path to curve in the same direction as the earth's curvature, and extends the range beyond the visible limit. To simplify calculation and representation of the ray path, the path is assumed to remain straight and the earth's radius is imagined to increase by about one-third to correspond. Taking re-

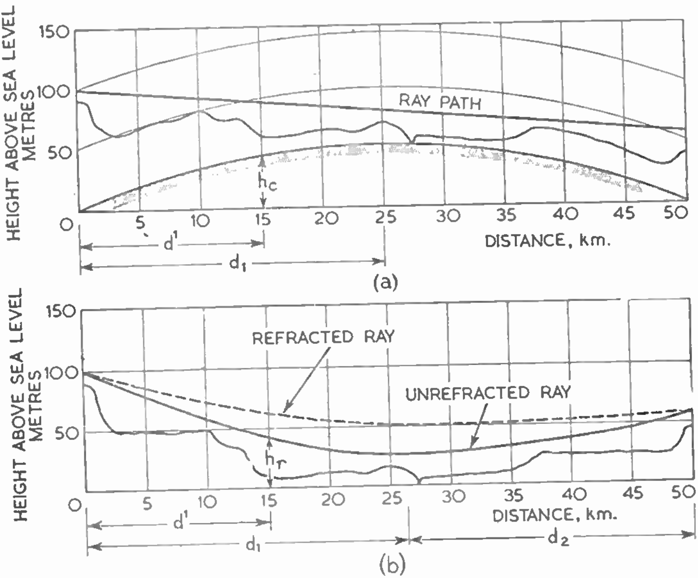


FIG. 33.—CONTOUR ELEVATIONS FOR V.H.F. COMMUNICATION LINKS.
(*a*) Ray path represented to curved earth base. (*b*) Ray path represented to flat earth base. Scales are exaggerated for clarity.

fraction into account, the distance is thus increased about $\sqrt{\frac{4}{3}}$ times, and $c = 4{\cdot}13$ approximately.

On a curved-earth basis, as illustrated in Fig. 33 (*a*), the height of the earth's curvature at any point distant d^1 from one end of the path can be calculated from the expression

$$h_c = d^1(d_1 - d^1)/c^2 \quad . \quad . \quad . \quad . \quad (2)$$

where $c = 3{\cdot}55$.

The curved base line is established by calculating the heights of curvature at several convenient distances, the distances being measured horizontally, not along the curve, and the heights measured vertically, not radially. Since V.H.F. links do not usually exceed 50 km., a suitable horizontal scale would be 1 cm. = 2 km. with a vertical scale 100 or 200 times as large. The terminal elevations and the profile of the terrain are then drawn in, and the ray path is represented as a straight line.

Equation (2) holds for a uniform surface free from irregularities, but a constant clearance height must be added to the calculated results to allow for surface irregularities. When the ray path has been inserted, any obstructions in the path are apparent, and if the route is unsatisfactory, an alternative site is selected and another contour elevation prepared.

If the flat earth base is used, as in Fig. 33 (b), the surface of the earth is represented by a horizontal straight line, with distances marked off to scale. The height h_r of the curved ray path at any point along the earth's surface is calculated for various distances d^1 from the higher end of the path from

$$h_r = (d_1 - d^1)^2/c^2 \quad . \quad . \quad . \quad . \quad (3)$$

or from the lower end $h_r = (d_2 - d^1)^2/c^2$.

The ray path can then be drawn through the points obtained.

With either method the path of the refracted ray can be calculated in the same way from equation (2) or (3) by putting $c = 4{\cdot}13$, and the path added on the same elevation.

Television and Frequency-modulated Broadcast Transmitters

The first need with a V.H.F. sound or V.H.F. television station is an open elevated position. Large cities with concentrated populations are most effectively served from a suburban site; extensive regional areas with distributed populations from a roughly central site. Height is obtained most economically with a hilltop site, but this advantage may be offset by difficulty of access, the provision of power and wator supplies and increased wind and ice loading of the radiator structure. Since the radiator is invariably designed to provide an omnidirectional service, the area in the immediate vicinity must be clear of obstructions in all directions. The presence of hills or high steel-frame buildings near the transmitter or receiver can cause permanent reflections out of phase with the direct transmission. These reflections are translated at the receiver into a phantom image displaced to the right of the true picture, which the directional discrimination of the receiving aerial is not sufficiently sharp to eliminate.

Although calculated estimates of field strength are useful for predicting roughly the performance at a proposed site, they cannot be wholly relied upon, because of the complex effects of varying ground contours. They must be supplemented by a survey of the region and field-strength measurements taken by setting up a mobile test transmitter.

The area of land required is governed chiefly by the space occupied by the mast or tower and the building. For a stayed mast a circular area having a minimum radius from centre to outer stay anchorages of 60–70 per cent of the height is necessary. A self-supporting square tower has a base area with sides 10–15 per cent of the height. In elevated positions, where high winds are encountered, a liberal safety factor must be allowed for these structures. Firm soil and solid foundations are particularly important, and special attention must be given to the protection of towers and buildings against lightning.

Receiving Stations

The site characteristics for satisfactory reception are the same as for transmission. Briefly summarized, these are:

Medium-frequency Reception.—Flat open ground, at least level with the distant terrain. Good soil conductivity. Freedom from screening by local buildings and other structures.

High-frequency Reception.—Flat ground clear of surrounding hills, densely wooded and built-up areas. Good soil conductivity is less important for sky-wave reception.

V.H.F. Reception.—Site area sufficiently elevated above surroundings to give a clear line-of-sight path in the direction of the transmitter with a preference for a downward slope facing the transmitter to divert undesirable ground reflections out of phase with the direct ray.

Television and Frequency-modulated Broadcasting.—A position above surrounding trees, ground contours and buildings, away from busy motoring roads and remote from high hills and metallic structures likely to produce phantom images by reflection.

Of importance in reception at all frequencies is freedom from internal and external electrical interference. Internal disturbances are produced by motors, generators, mercury-arc rectifiers, engine ignition systems and switchgear installed in the station. These are readily eliminated by fitting appropriate suppressors at each source of interference. External interference over the whole radio-frequency spectrum can originate in high-voltage transmission lines, electric railways and trolley-bus systems. Unsuppressed ignition systems in motor vehicles are mainly troublesome on the V.H.F. bands and to some extent on the high-frequency bands. Sites for all types of service should be situated in open country at least 300 yd. from overhead transmission lines, electric railways and primary roads. Power and communication lines entering the site should be buried for a similar distance from the aerials.

W. E. P.

12. RADIO-FREQUENCY TRANSMISSION LINES

12. RADIO-FREQUENCY TRANSMISSION LINES

In order to convey energy in the form of electro-magnetic waves of high frequency from one place to another it is possible to :

(a) Radiate the energy, perhaps directionally.
(b) Guide the wave by using, generally, some metallic structure such as a hollow waveguide, or a surface waveguide, or by guiding the energy as a transverse electromagnetic wave by a pair of metallic conductors. This last technique finds application at all frequencies up to about 10,000 Mc/s, and is the method which will be considered here.

Qualitative Considerations

To appreciate the basic requirements of a transmission line comprising two conductors, consider first the case of energy transfer in the form of D.C. Here, under steady state conditions, it is well known that :

(a) the only characteristic of the transmission line which is of interest is its total resistance, which should be small compared with load resistance;
(b) maximum energy transfer to the load will occur when the load resistance is equal to the internal resistance of the generator (strictly speaking plus the resistance of the line).

Considering now the case where A.C. is flowing in the line, as the frequency is increased it becomes necessary to take into account both the capacitance between the two conductors and also their self-inductance. At sufficiently high frequencies the inductance and capacitance of the line influence the transmission considerably, and under these conditions the input impedance of an infinitely long, uniformly constructed line is not infinity but a pure resistance of numerical value $\sqrt{(L/C)}$, and known as the characteristic impedance, Z_0, of the line, where L = self-inductance of line, henrys/unit length, and C = capacitance of line, farads/unit length.

The Need for Impedance Matching

By analogy with the D.C. case considered above, maximum energy transfer will occur when the generator supplies a load of resistance equal to the generator output resistance. This condition would be fulfilled by, for example, an oscillator of 70 ohms output resistance feeding an infinitely long transmission line of 70 ohms characteristic impedance. Conditions would be unaltered by feeding instead a finite length of line terminated in a 70 ohm resistive load, say a resonant dipole aerial of 70 ohms radiation resistance.

Reflection of Energy in Mismatched Systems

Physically, the effect of a change in impedance in a transmission-line system is to produce partial reflection of the electromagnetic wave, an hence a reduction in the energy absorbed by the load. A perfectly matched system is one in which the transmitted wave leaves the generator and arrives at, and is absorbed by, the load with no *net* reflection. It will be seen later when matching methods are considered that it is possible to convert a mismatched system to a matched one by arranging to add further reflections to cancel those already existing.

PRACTICAL TRANSMISSION LINES

Two general fundamental requirements may be stated as follows :

(1) The transmission line must be a uniform structure with the properties constant along its length. This becomes increasingly important at the higher frequencies and for electrically long lines, particularly in such applications as television transmission, where multiple internal reflections due to changes of characteristic impedance along the length of the line may produce picture distortion or even loss of energy transfer.

(2) The energy absorbed by the line should be small. Some loss is inevitable, due to the resistance of the conductors and the power factor of the dielectric separating them, but the loss may be made small by using highly conducting metals such as copper and aluminium, and low-loss, low-permittivity dielectrics such as polyethylene (polythene), polystyrene, or, better still, air. The line must be of adequate size to carry the requisite power without overheating.

Balanced (Twin) Line

The simplest form of practical line comprises two parallel conducting wires, held at a constant spacing by suitable means. A convenient cable construction of this form comprises two conductors embedded in a flat extruded ribbon of polythene. Such cables are used on systems the terminals of which are balanced with respect to earth (e.g., dipole aerials).

Unbalanced (Co-axial) Lines

If the balanced line is imagined to be rotated about one conductor the result is a co-axial system, which is essentially unbalanced with respect to earth. A simple example would be a single copper wire, covered with an extruded cylindrical coating of polythene and enclosed within a tube of aluminium.

Developments from Simplest Design

From the two basic constructions described have developed a wide range of cables with somewhat modified forms. The changes have been made chiefly to accomplish one or more of the following :

Increased Flexibility.—Mechanical flexibility can be increased by using stranded or braided conductors instead of solid wires or tubes. Alternatively, metal tapes applied helically with a long lay, have been used. In general, thick conductors are unnecessary, since the depth of penetration of the current at high frequencies is very small.

Reduced Losses.—The losses associated with a simple solid-polythene-insulated co-axial cable can be reduced by replacing part of the polythene by air. Various forms of such "semi-air-spaced" cables have been used, one form being made by spiralling a polythene thread around a copper wire, enclosing this structure in an extruded polythene tube and finally applying a braided copper outer conductor and protective jacket of polyvinyl chloride (P.V.C.). Another construction which has been widely used is the air-spaced construction, in which the centre conductor of a co-axial cable is located by discs, of say, polystyrene or polythene, situated at intervals along the cable length. Cellular polythene is also used widely, being more economical than "semi-air-spaced" cables.

High Electrical Uniformity.—For an increasing number of applications, particularly at the higher frequencies, cables with a high standard of electrical uniformity are required. In a non-uniform cable random, and also perhaps periodic, variations of dimensions occur along the length, with the result that the characteristic impedance also varies along the length. This results in a series of impedance mismatches within the cable, with consequent multiple reflection of the electromagnetic wave. The effect is to make the input impedance of the cable vary with frequency, and to cause distortion of a transmitted pulse, since the main signal is followed by a train of echoes. For ordinary purposes the effects of non-uniformities introduced by standard manufacturing technique are not serious, but for special applications high-uniformity cables can be obtained. The problem becomes more serious for the larger, lower-loss cables (such as medium-power television-transmitter cables) in which the unwanted reflected waves are not attenuated so rapidly. Suitable cables with a high degree of uniformity are available, however, such as the helical membrane cable, in which the inner conductor of a co-axial system is located by a helical web of polythene.

CABLE CHARACTERISTICS

Primary Constants

A perfect transmission line would have only two primary constants, inductance, L, and capacitance, C, both expressed per unit length. In addition, a practical line has two further primary constants, namely resistance, R, and leakance, G, per unit length; these are the quantities associated with losses in the conductors and dielectric respectively.

Characteristic Impedance

The characteristic impedance Z_0 is the ratio of voltage to current at any position in a cable sufficiently long for reflections from the far end to be negligible, and also in a short cable terminated by an impedance equal to Z_0. It is substantially a pure resistance at high frequencies, and is given by

$$Z_0 = \sqrt{(L/C)} \text{ ohms} \quad . \quad . \quad . \quad . \quad (1)$$

where L is in henrys/unit length;
C is in farads/unit length.

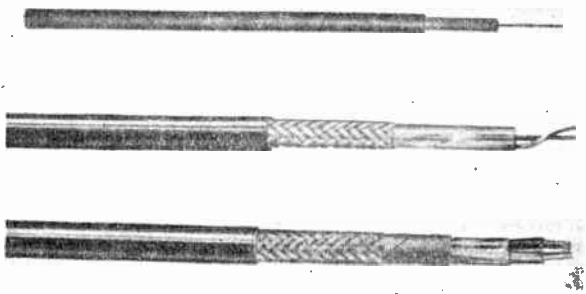


FIG. 1.—GROUP OF SMALL RADIO-FREQUENCY CABLES.
(*The Telegraph Construction and Maintenance Co. Ltd.*)

The more general formula for Z_0, which holds at all frequencies is

$$Z_0 = \sqrt{[(R + j\omega L)/(G + j\omega C)]} \qquad (2)$$

where R is in ohms/unit length;
G is in ohms/unit length (and is normally negligible compared with ωC);
$\omega = 2\pi f$;
f = frequency in c/s;
$j = \sqrt{(-1)}$.

At frequencies where R is not negligible compared with ωL, therefore, Z_0 is greater in magnitude and has a negative reactive component.

For nearly all cases the simple formula is sufficiently accurate above 1 Mc/s.

Attenuation

The attenuation constant is a measure, on a logarithmic scale, of the energy loss in a cable, and may be measured in nepers/unit length or decibels/unit length. In terms of input and output powers P_1 and P_2, or of input and output voltages V_1 and V_2, the attenuation of length, l, of line is

$$\left.\begin{aligned} \alpha l &= \tfrac{1}{2}\log_e\left(\frac{P_1}{P_2}\right) = \log_e\left(\frac{V_1}{V_2}\right) \text{ nepers} \\ \text{or} \quad \alpha l &= 10\log_{10}\left(\frac{P_1}{P_2}\right) = 20\log_{10}\left(\frac{V_1}{V_2}\right) \text{ decibels} \end{aligned}\right\} \quad . \quad (3)$$

$$(1 \text{ neper} = 8{\cdot}686 \text{ db.})$$

The forms of the equations involving voltage ratios are only valid if the input and output impedances are equal.

Two terms contribute to the attenuation, one arising from resistive losses in the conductors, and the other from losses in the dielectric. At high frequencies the attenuation constant is given by

$$4{\cdot}34(R/Z_0 + GZ_0) \text{ db/unit length} \quad . \quad . \quad (4)$$

Phase Constant

The phase constant is a measure of the progressive retardation of phase along a cable. It may be expressed as

$$\beta = 2\pi/\lambda \text{ radians/unit length} \quad . \quad . \quad . \quad (5)$$

where λ is the wavelength in the cable.

Velocity of Propagation

In free space electromagnetic waves are propagated with velocity $C \approx 3 \times 10^8$ metres/sec. In a cable, however, the velocity, v, is less, which means that for a given frequency the wavelength, λ, in the cable is less than the corresponding wavelength, λ_0, in free space. A convenient measure of the velocity of propagation in a cable is the velocity ratio or wavelength ratio,

$$v/c = \lambda/\lambda_0 = 1/\sqrt{\epsilon} \quad . \quad . \quad . \quad (6)$$

where ϵ is the effective permittivity * of the dielectric. The last expression is only valid at high frequencies in normal cables, for which the phase constant, velocity and wavelength depend only on the frequency and dielectric constant, and does not hold for helically wound delay lines.

The velocity of propagation at high frequencies can also be expressed in terms of L and C, the time taken for a wave to travel along the cable being

$$\sqrt{LC} \text{ seconds/unit length} \quad . \quad . \quad . \quad (7)$$

General Expression for the Propagation Constant

The general formula, in vector form, for the propagation constant $\alpha + j\beta$ per unit length, where $j = \sqrt{(-1)}$, is :

$$\alpha + j\beta = \sqrt{[(R + j\omega L).(G + j\omega C)]} \quad (8)$$

As in the case of the general formula for Z_0, this formula must be used at the lower frequencies when R is not negligible compared with ωL. The simple equations (4), (6) and (7) are derived from equation (8) by assuming that R is small compared with ωL and G small compared with ωC.

VARIATION OF CHARACTERISTICS WITH FREQUENCY

Primary Constants

Resistance, R.—High-frequency currents are largely confined to the surface of a conductor, increasingly so as the frequency rises. An increase of resistance with frequency is therefore to be expected, and at high frequencies R rises as the square root of the frequency.

Inductance, L.—If the currents flowing are largely confined to the

* Where permittivity is mentioned, it is the permittivity relative to free space, or dielectric constant, which is meant.

conductor surfaces the inductance is only associated with the dielectric space between the conductors. L is therefore constant at high frequencies. As lower frequencies are approached (say about 1–0·1 Mc/s, depending on the conductivity and diameter of the conductors) the currents begin to penetrate more deeply into the metal, and the inductance rises, reaching a limiting value at zero frequency when the currents are distributed uniformly over the cross-sections of the conductors.

Capitance, C.—This is constant over the entire frequency band, provided, as is the case with polythene, that the dielectric permittivity, ϵ, does not alter. For polythene, the value of ϵ is 2·3.

Leakance, G.—This increases as the product of frequency and power factor of the dielectric. For polythene, the power factor varies only slowly with frequency, and may be taken for most purposes as 3×10^{-4}; for polythene cables, therefore, the leakance increases directly as frequency. The contribution of leakance to the attenuation of a cable is independent of the cable size, and this means that dielectric loss is more important in larger-size cables, where the conductor losses are smaller.

Secondary Constants

Characteristic Impedance.—For a uniform line Z_0 is constant at high frequencies, but at lower frequencies rises due to increase of inductance, and becomes partly reactive.

Attenuation.—That part of the attenuation due to conductor loss increases with the conductor resistance, i.e., at high frequencies as the square root of frequency. The contribution of the dielectric increases as the leakance, and so directly as the frequency.

Phase Constant and Velocity of Propagation.—The phase constant increases uniformly with frequency at high frequencies, when the velocity of propagation is independent of frequency. At lower frequencies the velocity falls and the phase constant varies less rapidly with frequency than at high frequencies.

CALCULATION OF CHARACTERISTICS OF CO-AXIAL AND BALANCED CABLES

Co-axial Cables

For a co-axial cable with solid or tubular non-magnetic conductors put:

d = external diameter of inner conductor, inches;
D = internal diameter of outer conductor, inches;
ρ_1 = resistivity of inner conductor, microhm-cm.;
ρ_0 = resistivity of outer conductor, microhm-cm.;
ϵ = effective dielectric constant of insulator;
p = power factor (strictly loss tangent) of insulator.

Then at high frequencies:

$$L = 14{\cdot}04 \log_{10} (D/d) \quad \mu\text{H}/100 \text{ ft.} \qquad (9)$$

$$C = 7{\cdot}365\epsilon/\log_{10} (D/d) \quad \text{pF/ft.} \qquad (10)$$

$$Z_0 = (138{\cdot}1/\sqrt{\epsilon}) \log_{10} (D/d) \quad \text{ohms} \qquad (11)$$

$$\alpha_c = 0{\cdot}330 \times 10^{-3} \sqrt{f}(\sqrt{\rho_1}/d + \sqrt{\rho_0}/D)/Z_0 \quad \text{db/100 ft.} \qquad (12)$$

where α_c is the part of the attenuation constant due to losses in the conductors, and f is the frequency in c/s.

For a seven-strand inner conductor (D/d) should be replaced by $(D/0{\cdot}939d)$, and the first term $\sqrt{\rho_1}/d$ in α_c should be multiplied by 1·25 to allow for the increased resistance. For a braided copper outer conductor of long lay, the second term $\sqrt{\rho_0}/D$ in α_c should be approximately doubled. The part α_d of the attenuation constant due to losses in the dielectric is given by :

$$\alpha_d = 2{\cdot}77 \times 10^{-6} fp \sqrt{\epsilon} \text{ db/100 ft.} \qquad (13)$$

and is independent of the size and construction of the conductors.

Screened Twin Cables

The following approximate formulæ apply at high frequencies to two small circular conductors symmetrically placed in a circular tubular screen, with :

d = external diameter of conductors, inches ;
D = axial separation of conductors, inches ;
D_s = internal diameter of screen, inches ;
ρ_1 = resistivity of conductors, microhm-cm. ;
ρ_s = resistivity of screen, microhm-cm. ;
$g = d/2D$;
$h = D/D_s$.

$$L = 28{\cdot}07 \log_{10}[(1 - h^2)/g(1 + h^2)] \ \mu\text{H/100 ft.} \qquad (14)$$

$$C = 3{\cdot}683\epsilon/\log_{10}[(1 - h^2)/g(1 + h^2)] \ \text{pF/ft.} \qquad (15)$$

$$Z_0 = (276{\cdot}1/\sqrt{\epsilon}) \log_{10}[(1 - h^2)/g(1 + h^2)] \ \text{ohms} \qquad (16)$$

$$\alpha_c = 0{\cdot}660 \times 10^{-3}\sqrt{f}(\sqrt{\rho_1}/d + 4h^2\sqrt{\rho_s}/(1 - h^4)D_s)/Z_0 \ \text{db/100 ft.} \qquad (17)$$

To obtain the total attenuation constant, the part due to dielectric loss, α_d as given for co-axial cables, has to be added to α_c.

Similar changes as in the previous section should be made to allow for seven-strand conductors, in which case $g = 0{\cdot}47\ d/D$, where d is the overall conductor diameter. For flexible twin cables, however, in which the screen is formed by a helically applied foil or metallized paper supported by a metallic braid, considerable caution is necessary in estimating the loss in the screen and the corresponding contribution to the attenuation.

CHOICE OF CABLE

This section is intended to draw attention to some of the important cable characteristics which should be considered so that the most suitable type can be chosen for a particular purpose.

Mechanical Properties

The most obvious characteristics are the overall diameter and the flexibility, the latter usually being quoted in the form of a minimum bending radius. In some cases the tensile strength may be important,

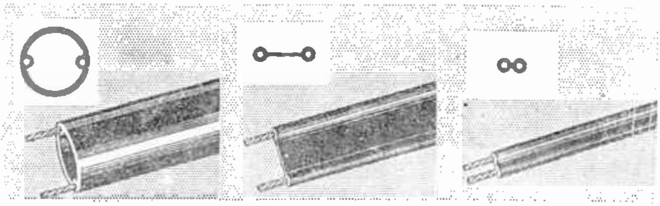


FIG. 2.—150-OHM (*right*) AND TWO 300-OHM BALANCED TWIN FEEDERS.
(*The Telegraph Construction and Maintenance Co. Ltd.*)

particularly of the inner conductor in certain small semi-air-spaced cables. The protective sheath should be chosen according to the risk of mechanical damage or corrosion.

Electrical Properties

The first choice here is that of characteristic impedance, and whether the cable shall be balanced or unbalanced. Preferred standard values of Z_0 are 75 and 50 ohms for the majority of co-axial cables. The choice between balanced and unbalanced may be decided by the terminal equipment, but may be influenced by the degree of screening required. Particularly at the lower frequencies there may be advantages in using a balanced cable to reduce the liability of the cable to pick up external interference.

The maximum permissible attenuation per unit length may set a minimum limit to the cable size, as may the required power (which the cable must carry without overheating) and the maximum voltage on the line (which voltage the cable must withstand without corona or breakdown). It should be noted that if the cable is used mismatched, there will be standing waves on the line, and the maximum voltage on the line may therefore be greater than either the voltage at the input or that at the load.

Finally, for certain uses (e.g., television-transmitter aerial feeders) it is important that the cable should be constructed to particularly high standards of mechanical and electrical uniformity.

Examples of Cables Suitable for Particular Applications

Domestic Television.—For the purpose of connecting television aerials to receivers a range of small cables is available which includes balanced

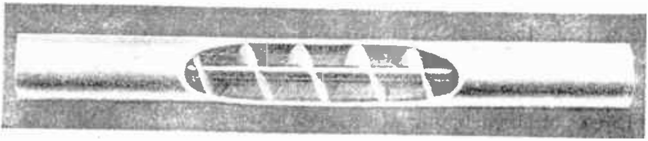


FIG. 3.—CUT-AWAY VIEW OF A HELICAL MEMBRANE CABLE.
The Telegraph Construction and Maintenance Co. Ltd.)

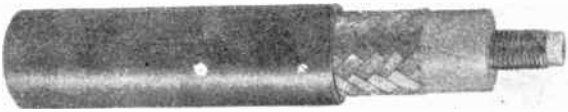


FIG. 4.—A HELICALLY WOUND DELAY CABLE.
(*The Telegraph Construction and Maintenance Co. Ltd.*)

and co-axial types, mostly about 0·25 in. diameter, with 50–80 ohms impedance and attenuations at 100 Mc/s from about 2·5 to 7·5 db/100 ft.

Amateur Radio.—A popular cable for this purpose is the balanced, unscreened, "ribbon" feeder, of 300 ohms impedance, or the equivalent tubular cable which has more stable characteristics under bad-weather conditions. The attenuation at 100 Mc/s is about 1·4 db/100 ft.

Television Relay.—Special screened quad cables have been developed for this application, so that more than one circuit is available on the cable. Aluminium-sheathed co-axial cables have also been used in certain cases, and where a long cable from an aerial to a distribution area is required a high-uniformity, low-attenuation co-axial cable of the helical-membrane type would be a good choice.

Medium-power Television Transmitters.—For this purpose special high-uniformity aluminium-sheathed co-axial cables have been used. A typical cable would have a diameter of about $1\frac{1}{2}$ in., attenuation approximately 0·3 db/100 ft. at 100 Mc/s, and would be capable of carrying at the same frequency a mean power of several kilowatts.

Delay Cables and High-attenuation Cables

A specialized cable type which finds application in, for example, pulse-forming circuits, circuits for colour television, and electronic computors, is the helically wound delay line. Delay cables have very high inductance, produced by winding the inner conductor of a co-axial pair in the form of a cylindrical spiral (see Fig. 4). In terms of the velocity ratio, v/c, the delay of a signal traversing a length of cable is given by

$$0{\cdot}102/(v/c) \text{ microseconds/100 ft.} \qquad (18)$$

At present in this country cables can be obtained with delays of up to 8·5 microseconds/100 ft

For purposes where a standard high attenuation is required, co-axial cables with inner conductors formed from high-resistance alloys are available. Two such cables have attenuation constants at 100 Mc/s of 16 and 200 db/100 ft. respectively.

IMPEDANCE MATCHING

The Effects of Mismatching

As explained in the introductory section, a mismatched system give a smaller energy transfer than a matched system, because part of the energy is lost by reflection. Moreover, if, for example, a television receiving dipole is connected directly to an unbalanced (co-axial) cable, the balance of the dipole is lost and the system becomes more susceptible to outside interference (since "noise" currents induced on the outside of the cable screen are fed into the receiver via the aerial). Examples of impedance transformation from one value to another, or from balanced to unbalanced form, are considered in the following sections.

Impedance Transformers

These may be divided into two groups: group 1 for systems which are either balanced or unbalanced throughout, and group 2 for cases where connection between a balanced and an unbalanced circuit is required.

Group 1 Transformers

The Quarter-wave Transformer.—Consider a hypothetical case where it is required to match a dipole of 60 ohms radiation resistance to a line of characteristic impedance 150 ohms. If the dipole is connected to the main feeder by a short length, l, of line of characteristic impedance Z_0, energy will be reflected both at the junction between the two feeders and at the load. It is possible to make the two reflected waves cancel, giving a system which is effectively matched, by arranging that the two backward waves shall be equal in magnitude and opposite in phase. The two waves will be of equal magnitude if $60/Z_0 = Z_0/150$, and will be of opposite phase if one is delayed by half a wavelength with respect to the other. These conditions are fulfilled by making $Z_0 = 95$ ohms and l equal to one quarter-wavelength.

In general terms, two resistive impedances R_1 and R_2 can be matched by a quarter-wavelength of transmission line of characteristic impedance equal to $\sqrt{R_1R_2}$. When calculating the quarter-wavelength the velocity ratio of the cable must be allowed for, so that

$$\lambda/4 = 2950(v/c)/f \text{ in.} \qquad (19)$$

where f is the frequency in Mc/s.

Stub Matching.—A more general method for introducing a reflected wave to cancel the reflection existing at the load is to introduce near the load a short section ("stub") of line connected across the main feeder. Such a stub would normally be short-circuited at the far end, and the length of the stub and its distance from the load are chosen so that the reflection produced has the required magnitude and phase.

This method is a general one, and can be used when the impedances have reactive components. The method of calculating the lengths and spacing of matching stubs is explained in, for example, reference [1].

Group 2 Transformers (Baluns).

The Triple Co-axial Balun.—The use of a quarter-wavelength of double-screened cable connected as in Fig. 5 (*a*) gives a 1 : 1 transformer between balanced and unbalanced lines. The quarter-wavelength refers to the co-axial formed by the two screens, and the insulation between the screens should therefore be polythene rather than P.V.C., since the permittivity of the latter is liable to vary with the grade used, and furthermore varies with frequency. The presence of the outer co-axial

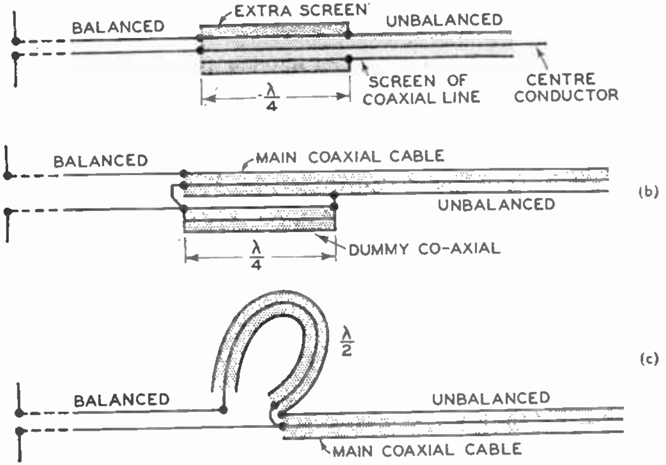


FIG. 5.—BALANCED TO UNBALANCED TRANSFORMERS.

system, short-circuited at a quarter-wavelength from the junction with the balanced circuit, ensures that a very high impedance is presented to currents flowing from the balanced circuit to the outside of the screen of the main co-axial cable.

An Alternative Type of 1 : 1 *Balun.*—Another form of balun working on the same principle uses a balanced short-circuited quarter-wave line. This type comprises an extra quarter-wavelength of the co-axial line arranged and connected as in Fig. 5 (*b*). In this case the quarter-wavelength refers to the balanced line formed by the two screens of the co-axials, the velocity of this line being controlled by the composite dielectric of sheaths and air between the two screens.

A 4 : 1 *Balun.*—This method is useful for connecting a high-impedance balanced load (e.g. folded dipole, open-wire feeder) to a co-axial, Fig. 5 (*c*).

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R. J. S.

13. WAVEGUIDES

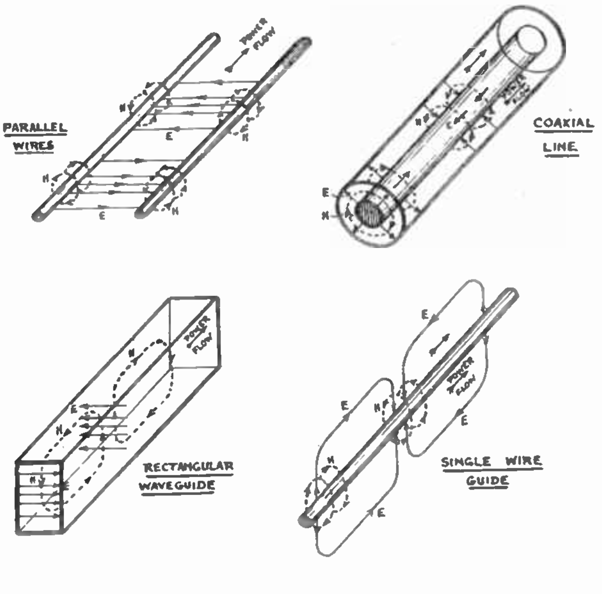
13. WAVEGUIDES

Introduction

An electromagnetic wave is said to be guided from one place to another if, as the result of intermediate arrangements, more power is delivered than would be possible in free space, under otherwise identical conditions.

Thus all transmission lines are, strictly speaking, waveguides, but by common usage the term has come to refer more particularly to some form of hollow metal tube, dielectric rod or even a single wire, acting in support of a wave (Fig. 1).

It is important to bear in mind that the energy transmitted always travels through the associated dielectric medium, and the guide itself, which is generally some form of conductor, then extracts a small proportion of that energy for services rendered *en route*, in directing the flow along the path required. Air at atmospheric temperature and



FIELD GUIDES EMPLOYING CONDUCTORS

FIG. 1.—VARIOUS MEDIUMS ACTING AS WAVEGUIDES.

pressure is capable of carrying as much as 1,000 kW/sq. cm. of cross-sectional area, so that there is no fundamental difficulty in getting a very large power through space if it is guided properly and does not suffer heavy losses in the process.

At low frequencies it is convenient and legitimate to regard the provision of an uninterrupted path through a conductor for the flow of a current as the essential requirement for the transmission of electrical energy. In that case we know that if the conductor is properly insulated the current does not leave it, and there is therefore no difficulty in directing the transfer of energy which resides in the electromagnetic field of the adjacent dielectric medium.

At high frequencies a very different situation arises, because displacement currents become at least equally important and the current streamlines then follow complicated patterns which are not necessarily any indication of the way in which energy travels from one point to another. To get information about the behaviour of a high-frequency guide attention must therefore be concentrated on the distribution of the electromagnetic field in its vicinity.

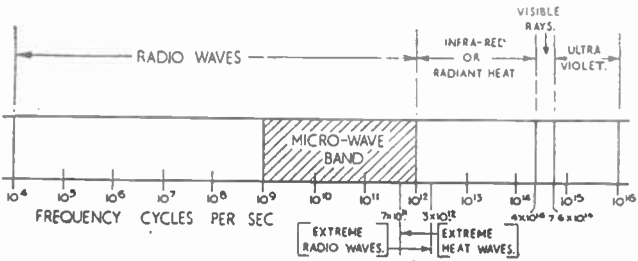


FIG. 2.—ELECTROMAGNETIC SPECTRUM.

Radio waves, as part of the electromagnetic spectrum, include what is called the microwave band, extending from about 1–1,000 kMc/s (see Fig. 2). When the frequency is raised to very high radio values the cross-sectional dimensions of any practical waveguide necessarily become comparable with the wavelength, and this introduces important considerations in providing for the support of the wave.

The discussion which follows deals particularly with the hollow metal tube as a guide for a wave trapped within it and propagated along it without serious attenuation. The parallel-wire line, as a familiar transmission system, will be used to introduce the subject and to show that the tubular form of guide is a logical development at ultra-high frequencies. Other forms of guide will be considered only very briefly.

The Development of a Skeleton Waveguide from Transmission-line Elements

A hollow metal waveguide can be regarded as built up, in the transverse plane, of ordinary transmission-line elements whose length is comparable with the wavelength. In pursuing this approach to the

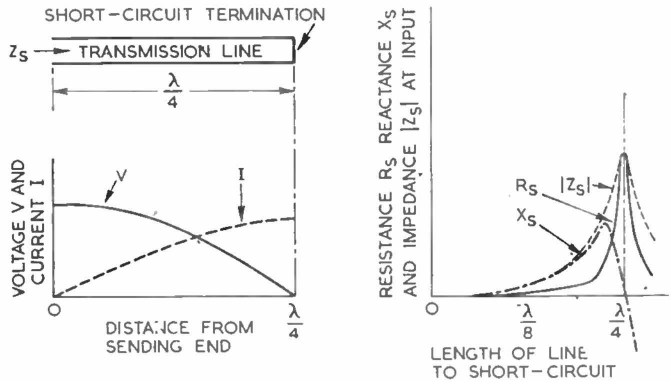


FIG. 3.—QUARTER-WAVELENGTH TRANSMISSION LINE CHARACTERISTICS.

problem we are specially interested in the short-circuited quarter-wavelength parallel-wire line associated with the usual T.E.M. wave, as shown in Fig. 3. It is well known that such a line, supporting a standing wave, has a high input impedance, which is almost purely resistive if the losses are small. The voltage and current, or the equivalent transverse electric and magnetic fields, are so distributed along the line that one is a maximum when the other is a minimum, making their ratio large at the sending end. Thus short-circuited quarter-wavelength elements attached transversely to two parallel bus-bars at intervals along their length (Fig. 4) should cause no serious disturbance to a wave transmitted along the bus-bars. Any number of such quarter-wavelength elements or stubs can be arranged side by side both above and below the bus-bars, and in this way a skeleton waveguide of rectangular cross-section may be built up (Fig. 5). Points P_1 and P_2 (see Fig. 4) on adjacent stubs equi-distant from the bus-bars will be at the

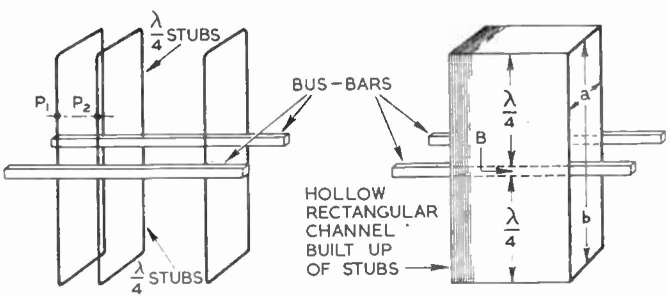


FIG. 4.—RECTANGULAR GUIDE BUILT UP OF TWO PARALLEL BUS-BARS WITH QUARTER-WAVELENGTH STUBS.

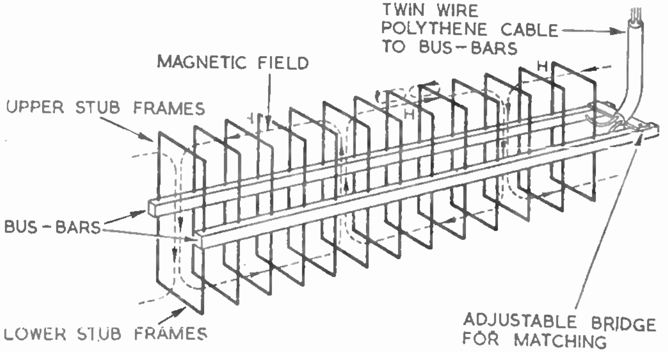


Fig. 5.—Outline of a Skeleton Waveguide Showing an H_0 Travelling Wave.

same potential, and therefore the walls of the guide can be made of continuous metal sheet, forming a closed channel without disturbing the wave propagated along the inside. In the skeleton waveguide. from which this closed channel is derived, the bus-bars must have a finite width B, and hence the transverse dimension of the guide in the direction of the projecting stubs must be rather greater than half the wavelength. It is important to remember that in addition to the longitudinal wave on the bus-bars there is always a transverse standing wave on the stubs, so that a resultant field distribution represents, from that point of view, a composite wave and takes a new form. If the width of the bus-bars were reduced to zero, the wave would be simply reflected backwards and forwards across the guide, without making any longitudinal progress at all. This corresponds to the cut-off condition for which the wavelength is λ_c. It follows that an arrangement of this kind as a waveguide is only of a convenient size for microwaves. The other transverse dimension of the guide is not critical, and provided it is sufficiently large to avoid electrical breakdown in the presence of field strengths arising under working conditions, no restrictions need be

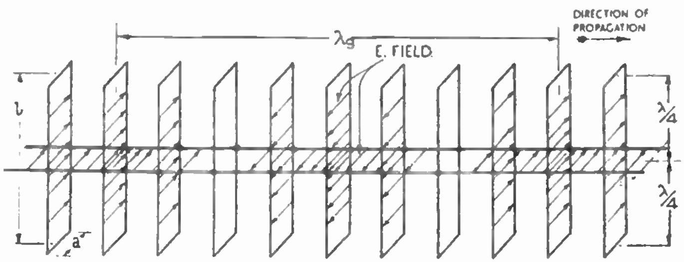


Fig. 6.—Electric Field Distribution along an Open-wire Transmission Line Enclosed by $\lambda/4$ Insulator Frames Above and Below the Line.

imposed. At a frequency of 3,000 Mc/s a guide of internal dimensions 3 in. × 1 in. or 2½ in. × 1 in. is commonly employed, whilst at 10,000 Mc/s a guide of 1 in. × ½ in. is generally regarded as suitable.

The distribution of the electric and magnetic fields associated with wave propagation inside a closed metal channel can be traced without difficulty in relation to the equivalent skeleton waveguide. Thus for the electric-field component we must have the usual sinusoidal spatial distribution along the length of the bus-bar, with lines of force stretching across from one to the other (Fig. 6). In the transverse plane we must necessarily have a standing wave so that the electric field between the two parallel wires forming the sides of any one of the stubs falls to zero at the short-circuited end.

To obtain the magnetic field distribution it is convenient in the first place to consider a standing wave on the bus-bars as well as on the stubs. This results in the two component field patterns being stationary in

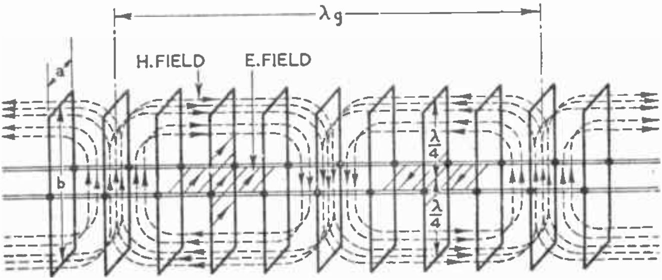
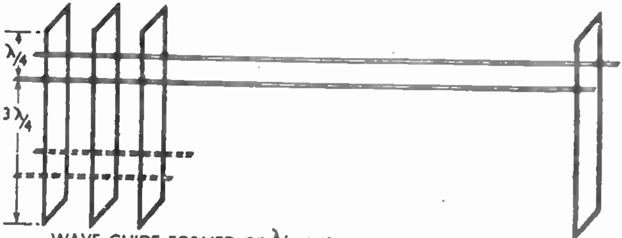


Fig. 7.—Magnetic and Electric Field Distribution along an Open-wire Transmission Line Enclosed by $\lambda/4$ Stubs Above and Below the Line. Standing Wave Condition.

space, and the composite field picture can therefore be more readily delineated. For a pure standing wave the transverse component of the electric field is a maximum at a point where the transverse component of the magnetic field is zero, and vice-versa. This corresponds with a spatial displacement of a quarter wavelength in the direction of propagation between the electric and magnetic field distributions so that there is no power flow. Thus in Fig. 7 we have a strong longitudinal magnetic field at the ends of stubs situated about mid-way between points at which the transverse magnetic field of the standing wave on the bus-bars is a maximum. A little consideration shows that the resultant magnetic field distribution then consists of a series of loops, end to end along the inside of the guide. The magnetic field arising from the stubs is entirely longitudinal, since, between adjacent stubs, the individual fields cancel one another much in the same way as occurs between turns of a closely wound solenoid carrying current (see Fig. 5). In a later section we shall consider the case of a longitudinal travelling wave on the bus-bars and how the transverse components of the electric and magnetic fields are then spatially brought into step with one another, without altering their individual configurations.

It will be observed that the wave which is supported by this hollow



WAVE-GUIDE FORMED OF λ/4 INSULATOR FRAMES ABOVE THE TRANSMISSION LINE AND 3λ/4 INSULATOR FRAMES BELOW IT

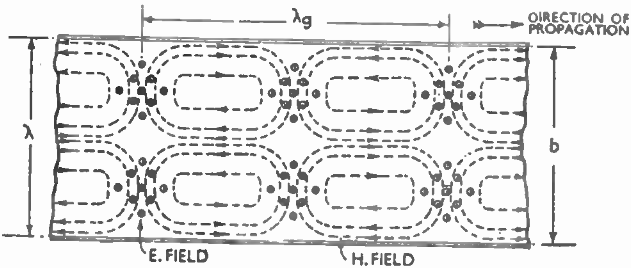


FIG. 8.—ELECTROMAGNETIC WAVE IN A RECTANGULAR GUIDE WHEN $b > \lambda$.

metal guide differs from the familiar T.E.M. wave in that there is a longitudinal component of the magnetic field in addition to a transverse component. All such waveguide modes are characterized by a component *either* of the electric field *or* of the magnetic field in the direction of propagation, and they are therefore described as E and H types of wave respectively.

The argument, which we have already employed, leading to the construction of a skeleton waveguide, can be extended to include arrange-

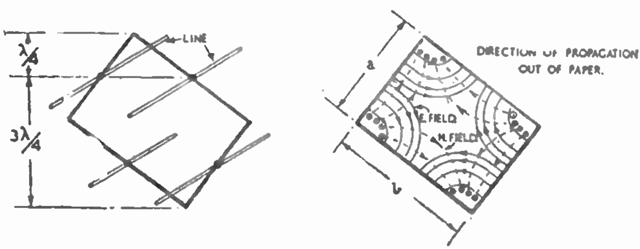


FIG. 9.—DEVELOPMENT OF A WAVE MODE OF HIGHER ORDER IN A RECTANGULAR GUIDE.

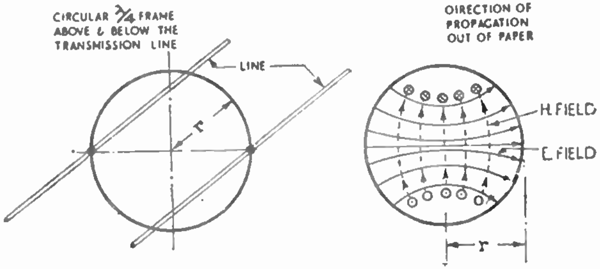


FIG. 10.—DEVELOPMENT OF SIMPLE WAVE MODE FOR CYLINDRICAL GUIDE.

ments suitable for other wave modes inside hollow metal channels. Thus a stub having the required high input impedance can be any odd multiple of a quarter-wavelength long, and with the rectangular stub this leads to higher-order modes of the same general type as previously derived (Figs. 8 and 9). Alternatively, the stubs can be of semi-circular form giving a field distribution suitable for wave propagation along the inside of a cylindrical metal tube (Fig. 10). These are all H-type waves, but we shall see later that E-type waves can also be propagated in certain cases.

Electric and Magnetic Field Distributions in Proximity to a Metal Surface

To get a proper understanding of the properties of waves guided by metal surfaces it is necessary to take into consideration the conditions that have to be satisfied at the boundary with the surrounding dielectric. In the first place let us suppose that the metal is a perfect conductor. The electromagnetic field cannot then penetrate the surface (Fig. 11). Thus the electric field component, which is supported by a charge on the surface, must be entirely *normal* to it. The magnetic field, on the other hand, supported by a current on the surface, must be entirely *tangential* to it. Consequently, if a tangential component of the electric field or a normal component of the magnetic field exists in the dielectric remote from the surface, both of these components must fall to zero as the surface is approached.

In general, at a boundary between any two arbitrary media we have both normal and tangential components of the electric and magnetic fields, designated respectively E_n and E_t or H_n and H_t, with appropriate subscripts to indicate to which side of the boundary reference is made (Fig. 12). If the media have permeability and permittivity values of $\mu_1\kappa_1$ and $\mu_2\kappa_2$, then :

(i) The *normal* electric and magnetic *flux densities* must be continuous across the boundary so that :

$$\kappa_1 E_{n1} = \kappa_2 E_{n2} \quad . \quad . \quad . \quad . \quad (1)$$

and

$$\mu_1 H_{n1} = \mu_2 H_{n2} \quad . \quad . \quad . \quad . \quad (2)$$

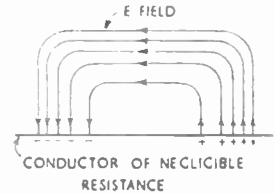
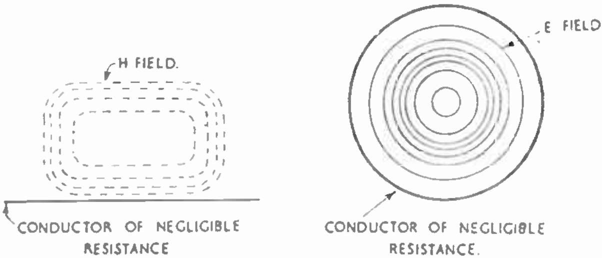


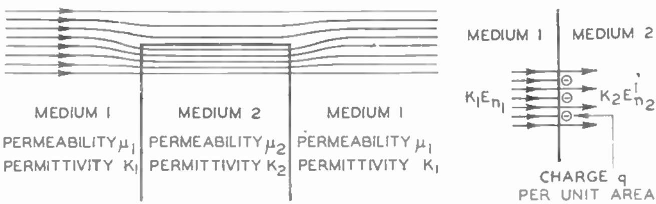
Fig. 11.—Electric and Magnetic Field Distribution in Proximity to a Conductor of Negligible Resistance.



In the case of the normal *electric* field this may be partly supported by charge on the surface, and therefore

$$\kappa_1 E_{n_1} = \kappa_2 E'_{n_2} + q \quad . \quad . \quad . \quad (3)$$

where q = charge per unit area; and
E'_{n_2} = normal component of electric force in the second medium after passing the charge q.



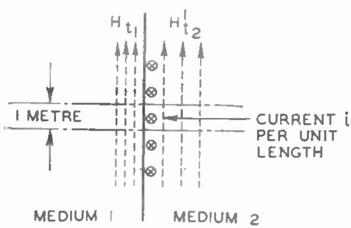


Fig. 12.—Fields at the Boundary of Two Different Media.

(ii) The *tangential* electric and magnetic *forces* must be continuous over the boundary so that :

$$E_{t_1} = E_{t_2} \quad . \quad . \quad . \quad . \quad (4)$$

and

$$H_{t_1} = H_{t_2} \quad . \quad . \quad . \quad . \quad (5)$$

In the case of the tangential magnetic force at the surface this may be partly supported by a current on the surface and therefore

$$H_{t_2} = H'_{t_2} + i \quad . \quad . \quad . \quad . \quad (6)$$

where i = current per unit length measured in the direction of H_t; and

H'_{t_2} = tangential magnetic force in the second medium after passing the current i.

At the boundary of a perfect conductor surrounded by a perfect dielectric we have already seen that there can be no tangential component of E and no normal component of H.

Thus $\qquad E_{t_1} = E_{t_2} = 0 \quad \text{and} \quad H_{n_1} = H_{n_2} = 0$

Moreover, since the electromagnetic field does not penetrate the surface, the normal electric flux must be supported entirely by charge on the surface, whilst the tangential magnetic force must be supported entirely by current on the surface.

Hence we have

$$\kappa_2 E'_{n_2} = 0 \quad \text{and} \quad \kappa_1 E_{n_1} = q$$

with

$$H'_{t_2} = 0 \quad \text{and} \quad H_{t_1} = i$$

The finite conductivity of the metal forming the walls of an actual guide makes little difference to our discussion of the way in which the field is distributed within the guide, since at these ultra-high frequencies the penetration into the metal is, in any circumstances, exceedingly small. For our present purpose we shall therefore neglect the effect of the resistance of the guide. It is important nevertheless to remember that in calculating, for example, the attenuation of a metal guide, the tangential component of electric field with the corresponding normal component of magnetic field at the surface is of vital consequence.

The Synthesis of Various Waveguide Modes and Their Characteristics

We have seen that waves capable of support within a closed metal channel can be regarded as derived from the superposition of T.E.M. components simultaneously propagated transversely and longitudinally on the equivalent skeleton waveguide structure.

In the transverse plane there must always be a standing wave which, being practically pure, consists of equal and oppositely directed travelling waves, whilst in the longitudinal plane there may be either a single outward-bound travelling wave to provide for energy flow in that direction or something in the nature of another standing wave. Taking the simplest case of a single longitudinal travelling wave from left to right and combining this in turn with each of the oppositely directed transverse component waves, we get two T.E.M. waves propagated diagonally across one another. We should clearly expect such a system

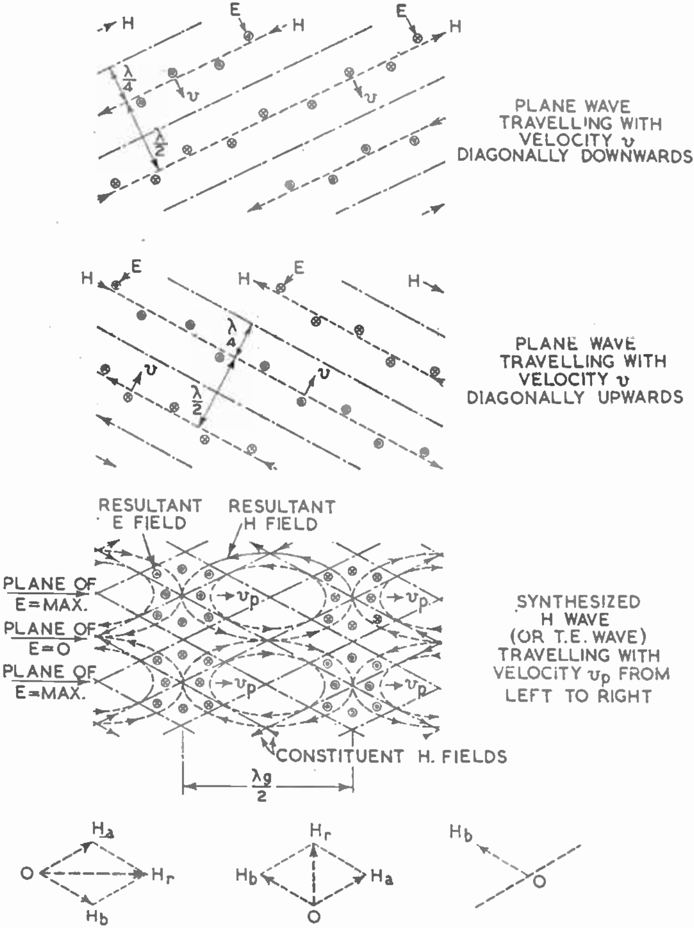


Fig. 13.—Synthesis of H Wave from Two Plane Waves Moving Diagonally Across One Another.

to yield a waveguide mode, and that in fact proves to be the case. Thus consider two free-space plane waves of the same frequency, each moving forward diagonally from left to right with velocity v and having the magnetic field in the plane of the paper. Combining these two waves together by the usual process of plotting the resultant field distribution as shown in Fig. 13, we find magnetic loops in the plane of the papor arranged end to end in layers, whilst there is a concentration of electric field perpendicular to the paper at the ends of those loops. This spatial coincidence of the maximum values of the transverse electric and magnetic field components is in accordance with our assumption of a single longitudinal travelling wave transmitting power horizontally from left to right. The composite wave which arises in the manner described is clearly of the H type and of the same form as that already deduced for support within a closed channel of rectangular cross-section. Since each constituent plane wave is propagated with the same velocity v at right angles with its own wave front, the resultant composite wave pattern moves forward without distortion horizontally from left to right. In general, the wavelength λ_g of the composite wave and its pattern velocity v_p are both greater than the corresponding quantities λ and v respectively of the constituent plane waves.

Remembering that a magnetic field adjacent to a conducting surface must be tangential with it, whilst an electric field approaches the surface at right angles, we can decide very easily where metal plates can be introduced into the field without disturbing it. Thus plates may be interposed perpendicular to the paper in Fig. 14 between any two successive layers of magnetic loops, and the field on one side of such a plate may be removed altogether, leaving the field on the other side to be supported by currents in the surface. In this way an H type of wave can be trapped between two parallel metal sheets. The extent of the field along the other lateral dimension can also be restricted by metal plates placed anywhere in the plane of the paper. We are able therefore to form a closed metal channel of rectangular cross-section containing within its b dimension (see Fig. 14) an integral number n layers of complete magnetic loops, and having an a dimension of any convenient value sufficient to avoid sparkover. In describing this type of wave by the letter H we find it convenient to append subscripts, the first of which represents the number of half-wavelengths along the a dimension and the second the corresponding number applicable to the b dimension. The wave considered is therefore designated an H_{0n} mode, and in the simplest case, for which there is only one layer of magnetic loops, we have an H_{01} mode.

Referring to Fig. 14 the wavefronts OM and ON of the two constituent plane waves move from M to P and from N to P respectively, each with velocity v, in the same time as the composite wave moves from O to P with the velocity v_p.

Hence $$v/v_p = \cos\psi \quad . \quad . \quad . \quad (1)$$

But $$v = f\lambda \quad . \quad . \quad . \quad . \quad (2)$$

and $$v_p = f\lambda_g \quad . \quad . \quad . \quad . \quad (3)$$

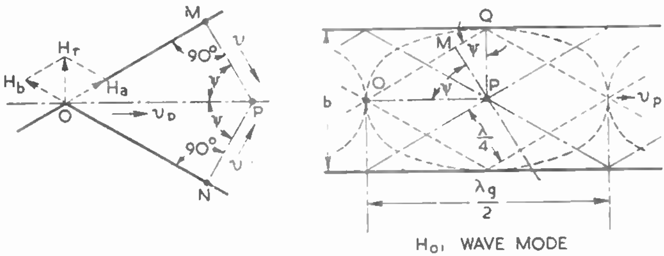
where f = frequency and λ_g = wavelength of the composite wave along the guide

so that $$\lambda/\lambda_g = \cos\psi \quad . \quad . \quad . \quad (4)$$

From the diagram it will also be seen that

$$MP = \lambda/4 \quad OP = \lambda_g/4 \quad \text{and} \quad PQ = b/2$$

with $$\sin\psi = \frac{MP}{PQ} = \lambda/2b$$



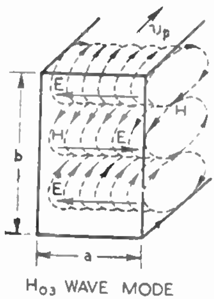


FIG. 14.—H WAVES SUPPORTED BY A HOLLOW METAL CHANNEL.

and since $\qquad \sin^2 \psi + \cos^2 \psi = 1$

we find $\qquad (\lambda/2b)^2 + (\lambda/\lambda_g)^2 = 1$

or $\qquad 1/\lambda_g^2 = 1/\lambda^2 - \dfrac{1}{(2b)^2}$

We have already defined the cut off wavelength λ_c for this H_{01} mode by the relationship $b = \lambda_c/2$.

Thus we ha

$$1/\lambda_g^2 = 1/\lambda^2 - 1/\lambda_c^2 \qquad . \quad . \quad . \quad (5)$$

Although this expression (5) has been derived for a particular waveguide mode, it is in fact of general application, and embraces E-type waves as well. From (1) and (4) it is clear that, since we must always have $\psi > 0$ to provide for the transverse standing wave, therefore $v_p > v$ and $\lambda_g > \lambda$.

For a guide of rectangular cross-section having $b = 3$ in. and $a = 1$ in. supporting an H_{01} wave mode at 3,000 Mc/s we get

$$\lambda_c = 2b = 15{\cdot}2 \text{ cm.}$$

and from (5) $\qquad \lambda_g = 13{\cdot}3$ cm. with $\psi = 41°$

Any attempt to excite such a guide at a wavelength greater than 15·2 cm. results in the setting up of an evanescent field which decays

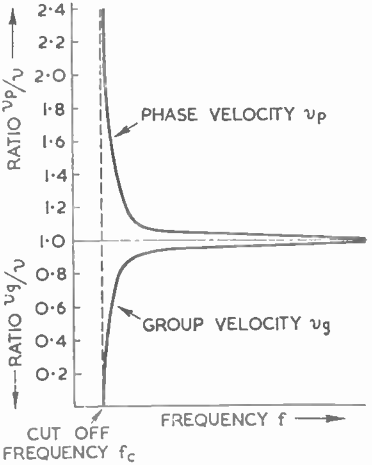


FIG. 15.—PHASE AND GROUP VELOCITIES.

$v_p v_g = v^2$

very rapidly and exhibits no wave distribution along the length of the guide. These evanescent fields find useful application in attenuators.

Referring again to Fig. 14, the constituent plane waves each have a velocity component v_g along the guide such that

$$v_g = v \cos \psi \quad . \quad . \quad . \quad . \quad . \quad (6)$$

This is the velocity with which energy travels along the guide, and is called the group velocity.

Combining (1) and (6), we have

$$v_p v_g = v^2 \quad . \quad . \quad . \quad . \quad . \quad (7)$$

At cut-off $v_p = \infty$, and a long way from cut-off $v_r \longrightarrow v$.

Thus v_g is always less than the free-space velocity v, as indeed it must be (see Fig. 15).

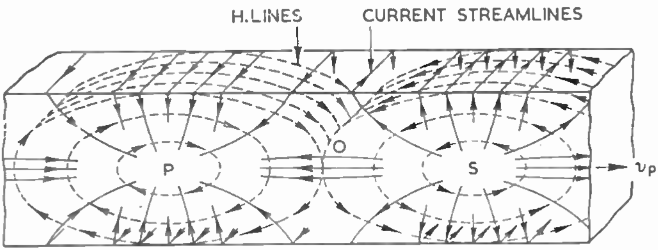


FIG. 16.—CURRENT STREAMLINES FOR H_{01} WAVE IN A RECTANGULAR GUIDE.

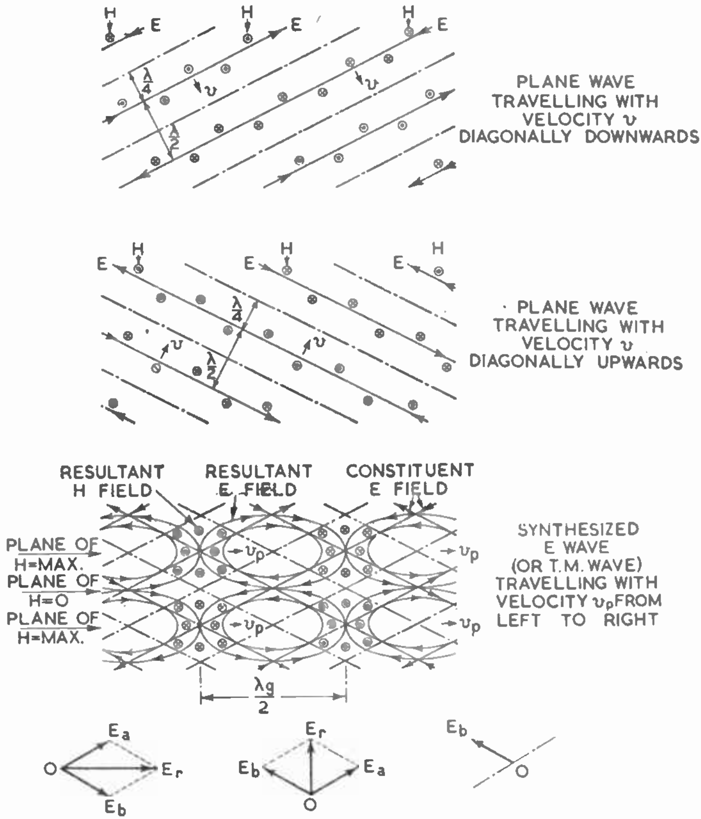


Fig. 17.—Synthesis of E Wave from Two Plane Waves Moving Diagonally Across One Another.

It is instructive to trace the distribution of the current streamlines in the walls of a guide, remembering that the current always flows at right angles with the adjacent tangential magnetic field. Fig. 16 shows the current distribution for an H_{01} mode in a guide of rectangular section. Since the currents diverge from the point S and converge on the corresponding point S′ on the opposite face of the guide, we should expect to accumulate a negative charge at S and a positive charge at S′. If the wave pattern remains stationary in space, as when no energy flows along the guide, the points S and S′ are fixed relative to the guide, and an electric field varying harmonically in time stretches across the inside from S′ to S. Here we have an axial displacement of a quarter

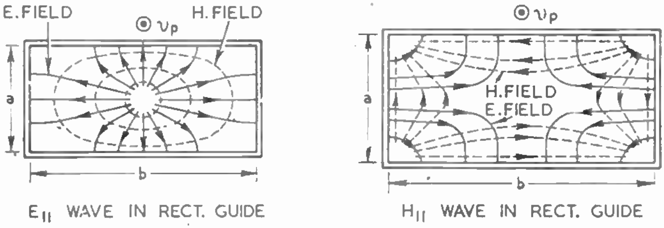


FIG. 18.—E_{11} AND H_{11} WAVES IN A RECTANGULAR GUIDE.

of a guide wavelength between the maxima of the transverse components of the electric and magnetic fields. If, however, the wave pattern is moving forward, we must bear in mind that for sinusoidal time variations of the field the current and charge are a quarter of a period out of step. There is therefore zero charge at S when the current flowing away from S is a maximum, but the full charge appears at this point, regarded as stationary in space, a quarter of a period later. The wave pattern has then travelled forward a distance of $\lambda_g/4$, so that in effect the charge appears at the point O on the wave pattern. Thus

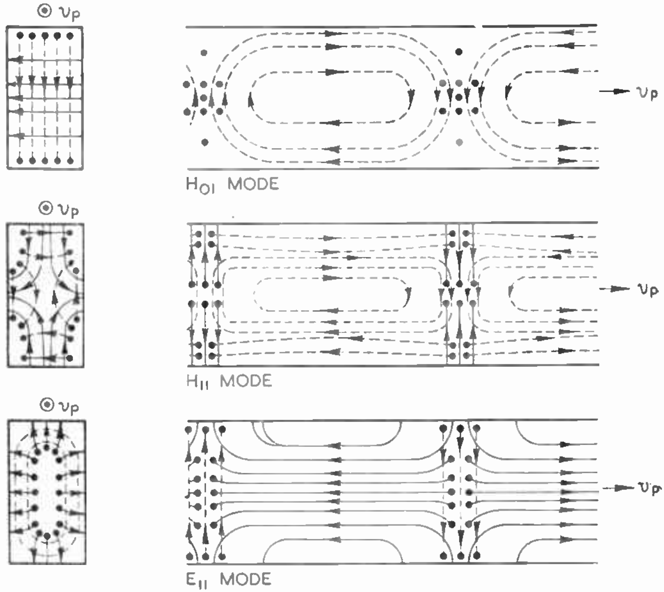


FIG. 19.—WAVE MODES IN A RECTANGULAR GUIDE.

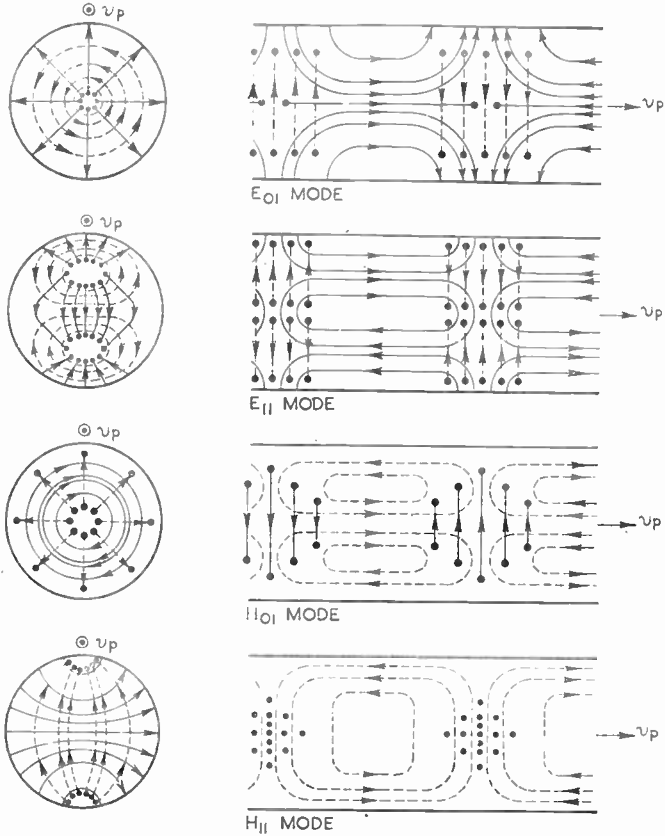


FIG. 20.—WAVE MODES IN A CYLINDRICAL GUIDE.

with a pure travelling wave, for which the guide is delivering power, the electric field stretches across between O′ and O, and the maxima of the transverse components of the electric and magnetic fields coincide in space as required.

Just as H-type waves can be synthesized by the superposition of two plane waves moving diagonally across one another, so E-type waves can be obtained in much the same way. It is clear that if we turn each of the constituent waves of Fig. 13 through a right angle so that the electric field lies in the plane of the paper, an exactly similar composite field diagram is derived but consisting of loops of electric force end to end in successive layers (see Fig. 17).

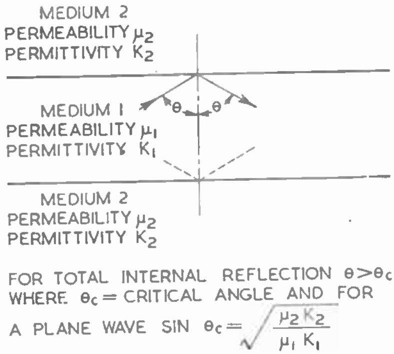


FIG. 21.—DIELECTRIC WAVEGUIDE.

In this case we can insert horizontal metal plates perpendicular to the paper so that they bisect the loops and thus trap a self-supporting E-type wave between them. No plates can be introduced in the plane of the paper, and therefore this form of wave cannot be supported in a guide of rectangular cross-section.

The E_{11} wave mode is the simplest that can be associated with such a guide (see Fig. 18).

It is not difficult to establish according to the principles enunciated a variety of wave modes appropriate to guides of rectangular or circular cross-section as shown in Figs. 19 and 20. All these wave modes clearly satisfy the essential boundary conditions. The co-axial line is capable, when there is sufficient spacing between the inner and outer conductors of supporting waveguide modes as well as the familiar T.E.M. form.

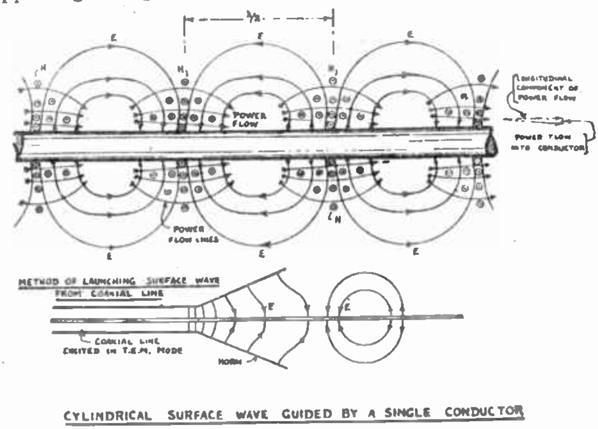


FIG. 22.—SMALL COPPER WIRE ACTING AS A CONDUCTOR FOR AN E TYPE ELECTROMAGNETIC WAVE.

Dielectric Waveguides and Single-wire Transmission Lines

The conception of plane waves reflected between the top and bottom of a hollow rectangular metal guide as the basis of propagation in that case suggests that a rod of solid dielectric in air without any surrounding conductor might, by reason of the phenomenon of total internal reflection, be induced to behave in much the same way (Fig. 21). If a wave travelling in a dense dielectric strikes the boundary of a less-dense dielectric at an angle of incidence greater than a certain critical value θ_c, then ideally all the energy is reflected. As with hollow metal waveguides there is for the solid dielectric rod of given thickness a minimum frequency at which such a condition can exist. The losses in even the best solid dielectrics are considerable, and therefore guides made exclusively of these materials are limited to short lengths only.

Electromagnetic waves of the E type can quite easily be made to travel along the outside of a single conductor, such as a small copper wire (see Fig. 22). If the conductor is straight, no radiation of energy takes place outwards into the surrounding space, so that the conductor acts as an effective waveguide between transmitter and receiver. There is no cut-off frequency, but below the microwave band the field tends to spread a long way from the conductor unless special arrangements are used to avoid this. Although radiation at bends is a serious problem, the device is likely to find some useful applications.

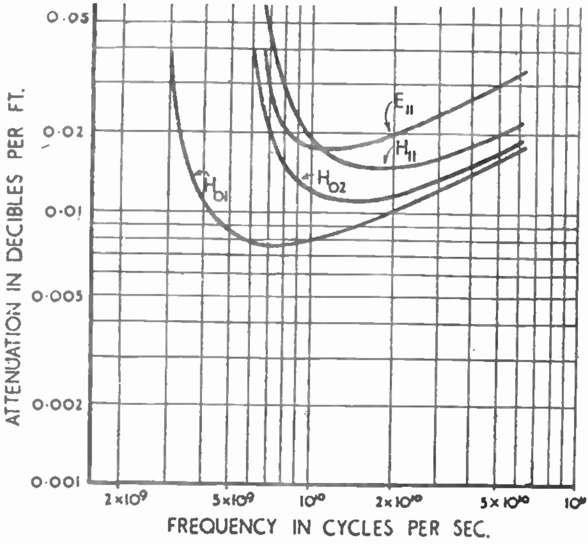


Fig. 23.—Attenuation in an Air-filled Rectangular Copper Guide 2 in. × 1 in.

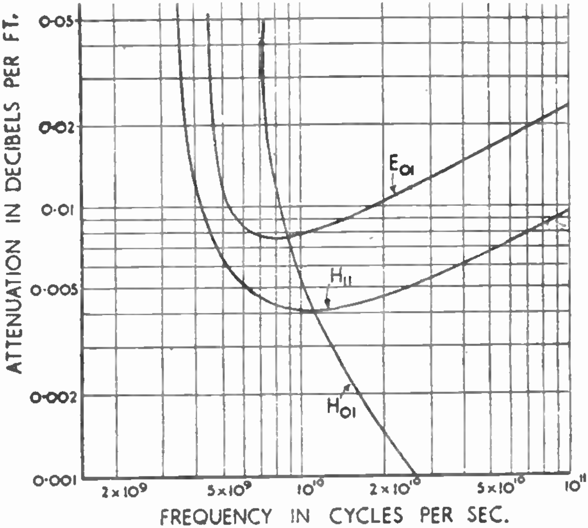


FIG. 24.—ATTENUATION IN AN AIR-FILLED CYLINDRICAL COPPER GUIDE OF 2 IN. DIAMETER.

Attenuation in Waveguides

Figs. 23 and 24 show typical curves of attenuation in guides of rectangular and circular cross-section respectively. It is interesting to observe that the H_{01} mode in a circular-section guide has a unique characteristic showing a progressively decreasing attenuation with increase of frequency. This may lead to the use of that wave mode for trunk communication purposes.

At 3,000 Mc/s a good cable has an attenuation of the order of 0·5 db/metre, whereas a hollow metal waveguide may have an attenuation as small as 0·025 db/metre.

Dielectric guides are little better than high-frequency cables, and much less adaptable.

The single-wire line would normally have an attenuation of the same order as that of the equivalent hollow metal waveguide.

Acknowledgments

Figs. 2, 6, 8, 9, 10, 11, 23 and 24 are reproduced from the writer's *Microwaves and Waveguides*, by kind permission of the publishers (Messrs Constable and Co. Ltd.).

Figs. 1 and 22, reproduced by courtesy of The Television Society, are from the writer's Fleming Memorial Lecture to the Society, and were originally published in the *Journal of the Television Society*, Vol. 6, July–September 1952.

H. M. B.

14. BROADCASTING RECEIVERS

14. BROADCASTING RECEIVERS

This section deals mainly with design trends in sound broadcasting receivers since the end of the Second World War in 1945. A review of the general features of post-war designs, with tabulated information on receiver classification for United Kingdom and export models, is followed by notes on the more important developments in particular components. Modern circuitry is analysed, stage by stage, and complete circuits of a number of typical models are provided. Except where otherwise stated, only British practices are covered.

GENERAL FEATURES OF POST-WAR DESIGN

Reliability

There was been a constant striving to improve reliability on the part of set manufacturers and component manufacturers alike. The improvements have been more marked in the export than the home field, because tropical climates necessitate much greater attention to this subject. Improvements in the reliability of individual components are discussed later. Reliability, however, also depends on careful choice of the most suitable components and materials by the set manufacturer and on the way they are used. The following are examples of this :

(1) Capacitors are liable to breakdown or to cause trouble through developing too low an insulation resistance unless chosen with care. For example, the chassis/earth capacitor in a live-chassis receiver must be capable of continuous operation at the A.C. mains voltage. Similarly, a capacitor coupling the anode of one valve to the grid of the next can cause a damaging positive voltage to appear on the grid if its insulation resistance ever falls too low.

(2) Resistors must be chosen so that their wattage ratings are not exceeded, care being taken to relate these to the ambient temperature. If a resistor's ambient temperature is unusually high, the normal wattage rating will be too high a figure to take : neglect of this precaution may lead to premature failure.

(3) Other components that may suffer through insufficient attention to the effects of heat are electrolytic capacitors, metal rectifiers, mains transformers, and even valves, which although designed to run hot, need adequate ventilation, especially power valves and rectifiers.

(4) The set's performance may suffer due to the effects of heat on the oscillator-circuit components and intermediate-frequency transformers, causing frequency drift and necessitating irksome re-tuning of the receiver at intervals as the set warms up. Apart from attention to ventilation, spacing with respect to very hot components, and careful choice of material for coil formers and switch insulation, etc., drift may be reduced by the use of negative temperature coefficient capacitors.

Ventilation

Adequate ventilation is aided by measures such as the following :

(*a*) Holes in the chassis around components such as valves or transformers, and air ingress to the under-chassis space via slots in the cabinet bottom or the bottom of the back.

(*b*) The production of a " chimney effect " by allowing air ingress only near the bottom and egress only near the top of the cabinet.

(*c*) Shields for preventing the heating by hot bodies of those which must be kept cool.

(*d*) Heat deflectors, e.g., for preventing the cabinet roof from becoming too hot.

(*e*) The provision of cooling fins or other large cooling surfaces in contact with components whose temperature must be kept low.

Climatic Considerations

All components and materials must be chosen so as to withstand the rigours of the climate in the regions where the sets will be used. For export sets this generally means withstanding tropical conditions, although not usually so severe as for Service equipment. Even in Britain there are areas where the relative humidity may be very high for days on end, and the protection needed against moisture ingress from this cause is not far short of that required for tropical areas.

For test purposes, satisfactory figures to take for maximum room temperatures are 30° C. for home sets and 45° C. for tropical export sets. Humidity tests are more complicated, and the reader is referred to Specification R.I.C. 11 or to British Standard Specification B.S. 2011 for complete details, including information about various destructive atmospheres and mould growths.

Economy

There has been a constant striving to bring prices down without appreciably reducing performance standards.

Examples of circuit practices which assist in economy are described later, and include :

(1) Resistance–capacitance smoothing, with or without hum-cancellation circuits, in order to eliminate expensive smoothing chokes. This has been facilitated by new techniques in electrolytic capacitor manufacture enabling much larger capacitances to be contained in existing or even smaller sizes of can.

(2) The omission of automatic bias resistors with their attendant by-pass capacitors from the cathode circuits of frequency-changer valves and amplifier valves.

(3) The omission of the by-pass capacitors from the audio-frequency amplifier and output valve cathode resistors.

(4) The use of a mains auto-transformer. This may be made considerably smaller than a double-wound transformer doing the same job. This is because of the opposition of primary current and secondary H.T. current in the main winding (and secondary heater current if series-heater valves are used). It brings, however, the disadvantage of a " live " chassis, or of a live cathode line if an isolated chassis is still preferred.

The money saved by an auto-transformer is negligible, however, if a mains filter, an earth terminal and pick-up sockets are fitted. This somewhat surprising result is due to the cost of the mains filter, the isolating capacitor for the earth terminal, three isolating capacitors plus a resistor for the pick-up sockets and other extra items needed as a result of live-chassis or live bus-bar technique. These include: a current-limiting resistor for the rectifier valve; a by-pass capacitor for the rectifier valve to prevent modulation hum; and the aerial isolating capacitor.

Apart from economic considerations, the mains auto-transformer is very useful in a small receiver, where the production of heat in voltage-dropping resistors would be troublesome, provided that operation on D.C. mains is not required; for example, in a small mains or mains/battery receiver in a thermoplastic cabinet.

Improved designs have reduced the intrinsic cost of some components. Several also contribute to overall economy because their reduced size results in smaller chassis, and, if desired, smaller cabinets. Examples of these are valves, ganged variable capacitors, intermediate-frequency transformers and electrolytic capacitors.

Size of Chassis and of Receivers

The reduction in the sizes of components has permitted the use of smaller, lighter, chassis. This in turn has meant that very small receivers can be made, and, even where for acoustic or appearance reasons a larger cabinet is desired, the chassis may occupy but a small portion of the total volume of the cabinet.

Small receivers which may be mains operated, battery operated or mains/battery operated have become increasingly important. The smallest set, however, is the " personal portable ", fitted with carrying handle and weighing only four pounds or so including batteries. The components mainly responsible for this are the layer-built H.T. battery and the miniature 1·4-volt filament valve with B7G (seven-pin) base.

A further considerable reduction in the size and weight of the filament battery (or filament portion of a combined battery) has become possible by the use of B7G valves having only half the filament wattage of the earlier types. Transistor receivers are discussed separately.

Safety

There have been determined efforts by manufacturers to improve the safety of their receivers during the last few years, and the British Standard dealing with this subject (B.S. 415) has recently been revised.

The following is a brief review of the principles involved in B.S. 415, together with examples of practical measures which may be employed.

The receiver must be so designed and constructed that there is no risk of serious electric shock or danger of fire either in normal use or under fault conditions, e.g., breakdown of insulation, the accidental touching of two metal parts or their bridging by another metal part.

To ensure this, tests have to be carried out both under normal use (including those conditions of mains voltage, receiver signal conditions, external connections and control settings most likely to cause trouble) and with faults deliberately created, e.g., by short-circuiting insulation, unless this is so good that it can be considered completely free from the possibility of breakdown.

TABLE 1.—TYPICAL RECEIVERS EXCLUDING V.H.F./F.M. MODELS (MARKETS, WAVE-RANGES AND POWER SUPPLIES)

Market	*Wave-ranges*	*Mains*		*Mains/Battery*		*Battery*		
		A.C.	*A.C./D.C.*	*Mains/All-dry*	*Mains/Vibrator*	*All-dry*	*Wet L.T.* Dry H.T.**	*Wet L.T. Vibrator*
Britain and export to Europe	Long + medium	Portable Small table Medium table Console Radiogram	Portable Small table Medium table	Portable		Portable	Small table Medium table	Car-radio
	Long + medium + short	Small table Medium table Big table Console Radiogram	Small table Medium table Radiogram	Portable		Portable	Small table Medium table	Car-radio
	Long + medium + 2–4 short	Medium table Big table Radiogram	Medium table Big table Radiogram					Car-radio
	Long + medium + 1–3 short + 2–8 bandspread	Medium table Big table Radiogram	Medium table Big table Radiogram		Medium table			Medium table Car-radio
Export outside Europe	Medium *or* short	Portable Small table	Portable Small table			Small table		Car-radio
	Medium + short *or* 2 short	Portable Small table	Portable Small table			Small table		Car-radio
	Medium + 2–4 short	Small table Medium table Radiogram	Small table Medium table		Medium table	Small table Medium table		Medium table Car-radio
	Medium + 1–3 short + 2–8 bandspread	Medium table Big table Radiogram	Medium table Big table Radiogram	Portable	Medium table			Medium table Car-radio

* These batteries are now very rarely used.

Practical measures of design and construction include the following examples

(1) The receiver housing should be such that it is impossible to bring a finger close to a live part. (" Live " being either under normal or fault conditions; the revised B.S. 415 describes a standard test " finger " of hinged construction, and its use.)

(2) Adequate clearance and creepage distances should be maintained between conductors which are not at substantially the same potential.

(3) Insulating material used in components and in the receiver generally should be of good quality.

(4) The leakage current which can flow to earth from any terminals accessible to the customer should be so small that no more than small harmless shocks occur if the terminals are touched or wires connected to them are held. The aerial leakage current is a special case, since a man working on a ladder might lose his balance by receiving only a very small shock, and a particularly low limit has to be set here. This limit determines the maximum value of an aerial-isolating capacitor, and the minimum value of a static discharge resistor (see below) if this be connected to a live conductor, either direct or through a sufficiently large capacitance.

(5) The aerial terminal should be provided with a D.C. leakage path in order to prevent breakdown of isolating components by the accumulation of static charges.

(6) It should not be possible to touch a live part with a plug or wire when trying to insert or connect it.

(7) Under normal or fault conditions the temperature of components should remain low enough for flames, sustained arcs or smouldering not to develop. Current and/or temperature fuses may be used to achieve this. It will often be found that more fuses than one are required, and great care is necessary in choosing the types and ratings of fuses and their positions in the circuit. This is because it is easy to choose a current fuse that will blow when a short-circuit develops, but it may be found that this also blows with switching-on surges. Examples of surges when switching on are those occurring in the primary circuit of a mains transformer, in the supply circuit of a rectifier followed by a reservoir capacitor (switching on when either cold or warm with a metal rectifier or when warm with a valve rectifier), and in the supply circuit of a chain of series-run valve heaters (when cold).

Quality of Reproduction

Improvements in quality of reproduction have taken place chiefly in high-fidelity receivers, the markets for which are in general small because of their necessarily high cost and because the receiver and its loudspeaker occupy too much space for the average householder.

High-fidelity reproduction is discussed in Section 37.

In ordinary mass-produced domestic receivers there have been a few improvements, however. These include: loudspeakers; pick-ups and recordings; negative feedback; compensated volume control; and the reduction of cabinet boom.

The low-price mass-produced loudspeaker has improved; how this affects broadcast receivers is described later.

TABLE 2.—TYPICAL RECEIVERS (MISCELLANEOUS FEATURES)

Features	*Portable (i.e., with Handle)*	*Small Table*	*Medium Table*	*Big Table*	*Console*	*Radiogram*		*Car-radio*
						Table	*Floor*	
Min. size approx.[1]	2½ × 6 × 8 in.	4 × 6 × 9 in.	7 × 10 × 15 in.	10 × 16 × 22 in.	—	—	—	—
Max. size approx.[1]	7 × 12 × 17 in.	7 × 10 × 15 in.	10 × 16 × 22 in.	12 × 19 × 25 in.	—	—	—	—
Cabinet material	Plastic, wood, or plastic + wood, or plastic + metal	Mainly plastic, sometimes wood, sometimes metal	Mainly wood, sometimes plastic or wood + plastic	Mainly wood, rarely plastic	Mainly wood, rarely plastic	Mainly wood, sometimes wood + plastic	Wood	Metal
Loudspeaker(s)[2]	One, 3–6¼ in. diam.	One, 5–6½ in. diam.	One, 6½–8 in. diam.	One or two, up to 10 in. diam.	One or two, up to 12 in. diam.	One, 5–8 in. diam.	One or two, up to 12 in. diam.	One or two, 5–6½ in. diam.
Number of valves, excluding rectifier and tuning indicator[3]	4–8	3–5	3–5	5–7	4–7	3–4	3–8	4–7
Self-contained aerial(s):								
Home	Yes	Usually	Usually	Usually	Usually	Usually	Usually	—
Export	Yes[4]	Usually	Sometimes	Sometimes	—	Sometimes	Sometimes	—
Pick-up sockets	No	Sometimes	Usually	Yes	Yes	—	—	No
Pilot light (mains)	Usually	Usually	Yes	Yes	Yes	Yes	Yes	—
Radio-frequency amplifier:								
Mains	No	No	Sometimes	Yes	Sometimes	No	Sometimes	—
Battery or mains/battery	Sometimes	No	Sometimes	—	—	—	—	—
Separate oscillator valve (battery or mains/battery only)	Rarely	Sometimes	Sometimes	—	—	—	—	—
Push-pull output	Sometimes	No	No (except in battery sets)	Usually	Sometimes	No	Sometimes	Sometimes

[1] Approximate sizes are only given as a rough guide where there may be doubt what is meant by the type-names.
[2] Elliptical speakers of equivalent size are increasingly used instead of circular ones.
[3] A tuning indicator may be found on any of the types except the smallest, since a battery-operated one is now available.
[4] For short waves, a telescopic aerial is often used, disappearing into the cabinet when not required.

The cheap mass-produced pick-up has improved considerably, though at some loss in robustness. In practice, this change is probably the most significant improvement. In addition, recording technique has improved and the microgroove record has been introduced: the latter has other virtues besides long duration of play; for example, low surface noise, provided that the record is kept clean.

Negative Feedback

Negative feedback has been increasingly used for (among other things):

(1) reduction of distortion and of hum;
(2) loudspeaker damping, one result of which is a better transient response;
(3) tailoring or levelling audio-frequency responses;
(4) tone control;
(5) compensated volume control, i.e., the frequency response varies with the setting of the volume control to suit the human ear characteristic;
(6) preventing a high-gain radio-frequency amplifying valve from being overloaded by strong signals;
(7) minimizing the effects of variations in supply voltage and of components and valve tolerances on gain.

Compensated Volume Control

Compensated volume control—sometimes called physiological compensation—may be obtained by several methods, e.g., by negative feedback which varies with the volume-control setting, by using a volume

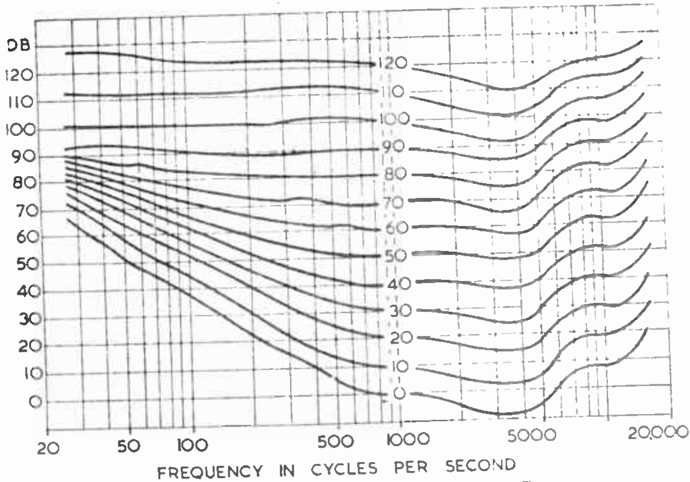


Fig. 1.—Fletcher–Munson Equal Loudness Contours.

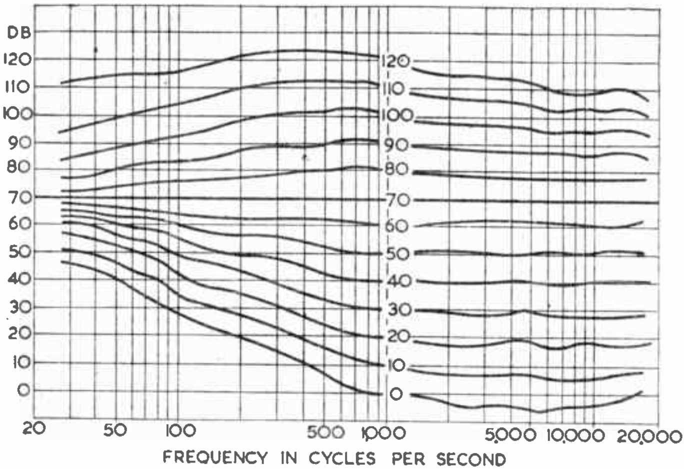


FIG. 2.—DIFFERENCES BETWEEN FLETCHER–MUNSON EQUAL LOUDNESS CONTOURS PLOTTED FROM THE 70-PHON CONTOUR.

control having a tap on it to which a frequency-conscious circuit is connected, or by using a tone control ganged with the volume control.

Compensated volume control is useful because people tend to listen to music at home at lower levels than natural, and it is the aim of the designer to achieve a realistic effect whatever the setting of the volume control. To see how this is attempted it is helpful to study human ear characteristics, such as the equal loudness curves [1] due to Fletcher and Munson, reproduced in Fig. 1, or better, in this case, the redrawn curves of Fig. 2, in which the contours of equal loudness are replaced by curves showing the differences in decibels compared with the 70-phon contour (an arbitrarily chosen level). It will be seen that considerable compensation is needed at the bass end of the spectrum and hardly any at the treble end. It is assumed that the receiver's electro-acoustic response, without compensation, is satisfactory at this level, and we wish to make the response vary with the volume control so as to sound the same at other levels. The compensated volume control shows its worth chiefly in avoiding "thin-ness" in the reproduction at low volume-control levels, and "boom" at high levels. The system is at its best when used in a receiver having very good automatic gain control, so that normal volume is always obtained at approximately the same control setting, irrespective of the strength of the radio-frequency input.

If the automatic gain control is not very good, as is the case with the undelayed or little delayed type usually employed in middle and low-priced receivers, the compensation could be theoretically correct for only one particular radio-frequency input. However, despite this, it has been found that the system greatly improves the performance of such receivers.

For receivers in very small cabinets the compensation may be deliberately over-emphasized so that the extra bass makes up for the poor acoustic bass response resulting from the small baffle area. In the usual low-priced receiver there is no gain to spare for bass lift (which, of course, has to be obtained by reducing the gain at middle and high frequencies) when the volume control is turned up for weak station reception. However, when listening to strong stations there is gain to spare, and by making the relative bass increase occur at low volume-control settings, use of this can be made.

Reduction of Cabinet Boom

Measures for getting rid of cabinet "boom" have been used more frequently, for example:

(1) The use of shallow cabinets, ranging from the cabinet which is little more than a flat baffle to the box-type with relatively small front-to-back dimensions.

(2) The use of sub-baffles thick enough to reduce considerably the vibration of the front panel of the cabinet.

(3) The use of strengthening struts in the cabinet for reducing panel vibration.

The resistance strain-gauge consisting of a piece of very fine resistance wire glued to a cabinet panel, and connected via a bridge circuit to a response-recording apparatus is a very useful instrument for investigating panel resonances. Several gauges can be fixed to different places and their responses recorded.

The strain gauge forms one arm of a D.C.-balanced bridge, the detector terminals being connected to the microphone terminals of a response-recording apparatus of the type normally used for taking loudspeaker responses. Resistan variations unbalance the bridge, causing an alternating voltage to be applied to the apparatus, when the loudspeaker is fed by an audio oscillator. The frequency of this oscillator is varied automatically in sympathy with the motion of a strip of paper on which a curve is drawn with ordinates proportional to the amplitude of the resistance variations, the abscissæ being proportional to the frequency. One form of the apparatus uses an ink

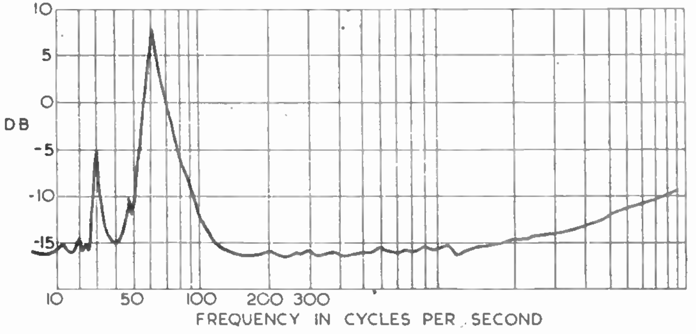


FIG. 3.—ELIMINATION OF CABINET BOOM—I.

Response curve of a cabinet plotted by means of a resistance strain gauge showing severe resonances.

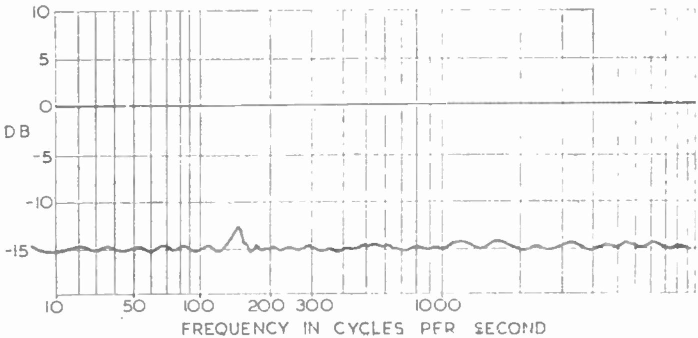


FIG. 4.—ELIMINATION OF CABINET BOOM—II.

Response curve of the same cabinet after the addition of two metal struts of T-section.

recorder; the origins of the curves shown in Figs. 3 and 4 were produced by an instrument employing a beam of light and photo-sensitive per.

Selectivity

Improvements in the selectivity of receivers for medium- and long-wave reception have become necessary because of the increases in power and number of broadcasting stations, particularly in the European area. It is thus of considerable importance to appreciate fully the two main types of interference affecting superheterodyne receivers.

The first type consists of tuneable whistles produced whenever $mf_p \sim nf_h \simeq f_i$, where m and n are integers, f_p is the frequency of undesired station, f_h is the local oscillator frequency and f_i is the intermediate frequency. It can only be cured or minimized by having sufficient discrimination against the undesired signal, before the frequency changer; that is sufficient radio-frequency selectivity, and/or by careful choice of the I.F. in relation to station frequencies.

The second type consists of side-band chatter (a background of the higher audio frequencies from a station whose frequency is very near to that of the desired station), accompanied by a fixed frequency note (heterodyne whistle) if the carrier frequency difference is low enough to come within the ear's range. This is called "adjacent channel interference", and can only be cured or minimized by having sufficiently selective intermediate-frequency amplifier circuits.

Types of Tuneable Whistles

The expression $mf_p \sim nf_h \simeq f_i$ means that tuneable whistles are likely to occur whenever the interfering station or one of its harmonics (produced either by the station or in the frequency changer) differs from the intermediate frequency by a frequency equal to that of the oscillator or one of its harmonics.

This general formula covers nearly all types of tuneable whistles, provided that it is remembered that sometimes the desired signal (let us call it frequency f_s) can create its own interference whistle without the need for a separate f_p.

The most troublesome ones are as follows:

Type (1). *When* $m = 1$ *and* $n = 0$, we have $f_p \simeq f_i$ (the frequency of the interfering station is approximately equal to the intermediate frequency).

This type of interference comes mainly from shipping transmitters, and is therefore noticed chiefly in coastal districts. It may produce whistles accompanying every station heard.

Type (2). *When* $m = 2$, $n = 0$ *and* $f_s = f_p$ (i.e., the station creates its own interference) we have:

$$2f_s \simeq f_i$$
$$f_s \simeq \tfrac{1}{2} f_i$$

(The frequency of the desired station is approximately half the intermediate frequency.).

Type (3). *When* $m = 2$, $n = 1$ *and* $f_s = f_p$ (i.e., the station creates its own interference), we have:

$$2f_s - f_h \simeq f_i$$
$$\text{or } 2f_s - (f_i + f_s) \simeq f_i$$
$$\text{or } \quad f_s \simeq 2f_i$$

(The frequency of the desired station is approximately twice the intermediate frequency.)

Type (4). *When* $m = 3$, $n = 2$ *and* $f_s = f_p$ (i.e., the station creates its own interference) we have:

$$3f_s - 2f_h \simeq f_i$$
$$\therefore \quad 3f_s - 2(f_i + f_s) \simeq f_i$$
$$\therefore \quad 3f_s - 2f_s \simeq 3f_i$$
$$\therefore \quad f_s \simeq 3f_i$$

(The frequency of the desired station is approximately three times the intermediate frequency.)

Type (5). *When* $m = 2$ *and* $n = 1$, we have:

$$2f_p - f_h \simeq f_i$$
$$\therefore \quad 2f_p - (f_i + f_s) \simeq f_i$$
$$\therefore \quad 2f_p - f_s \simeq 2f_i$$

In practice, this is only serious when the frequency of the interfering station is fairly close to that of the desired station.

Then $\quad f_p \simeq f_s$

and $\quad 2f_p - f_p \simeq 2f_i$

$$\therefore \quad f_p \simeq 2f_i$$

i.e., it is only serious when the frequency of the interfering station (and of the desired station) approximates to twice the intermediate frequency.

Type (6). *When* $m = 1$ *and* $n = 1$, we have:

$$f_p - f_h \simeq f_i$$
$$\therefore \quad f_p - (f_i + f_s) \simeq f_i$$
$$\therefore \quad f_p - f_s \simeq 2f_i$$

(The frequency of the undesired station differs from that of the desired one by approximately twice the intermediate frequency.)

This is called second-channel or image interference, the first name distinguishing it from adjacent-channel interference, and the other arising from the similarity to an image in optics, because the interfering station frequency is as far "behind" the oscillator frequency as the desired signal is "in front of" it..

Type (7). *When there are two strong f_p's say f_p' and f_p'' (i.e., two strong or local interfering stations)* and any of the following three relationships occur :

$$\text{(i) } f_p' - f_p'' \simeq f_i$$
$$\text{(ii) } 2f_p' - f_p'' \simeq f_i$$
and
$$\text{(iii) } f_p' - 2f_p'' \simeq f_i$$

Type (8). *When the frequency of the interfering signal is approximately equal to that of the local oscillator.*

The second detector then receives two intermediate-frequency signals differing only slightly in frequency, so that an audible beat is produced between them.

Type (9). *When the receiver is very close to one or more high-power transmitters* (the more transmitters, the worse the interference) there may be a large number of whistles due to such a variety of causes as to make them individually unidentifiable and incalculable.

Prevention or Reduction of Tuneable Whistles

Type (1) can be reduced to negligible proportions by using a wavetrap tuned to the intermediate frequency, often called an intermediate-frequency rejector.

Types (2), (3), (4) *and* (5) can only be cured or minimized by careful choice of the intermediate frequency. No amount of radio-frequency selectivity will eliminate them.

Types (6), (7), (8) *and* (9). Type (6) is common; types (7), (8) are rare. Type (9) is fairly common under present conditions. To help cure these, good radio-frequency selectivity is necessary. If this is not quite adequate, it may be supplemented by one or more wavetraps tuned to the frequencies of interfering stations. If one of these is also on occasion a desired station, the presence of the trap will not generally matter, because the signal will usually be strong enough to give good reception even after attenuation by the wavetrap.

It is also desirable to consider types (6), (7) and (9) when choosing the intermediate frequency, because this may enable cheaper radio-frequency circuits to be used than otherwise would be the case.

Cross Modulation

Through the rectifying action of non-linear portions of radio-frequency amplifier or frequency-changer characteristics, the modulation of a strong undesired signal may be transferred to the carrier of the desired signal, so that both modulations appear in the audio output. Measures for reducing this effect are as follows :

(1) Good radio-frequency selectivity (including the circuit before the radio-frequency amplifier, if any).

(2) Choosing valves with low cross-modulation characteristics (variable-mu valves are the best in this respect).

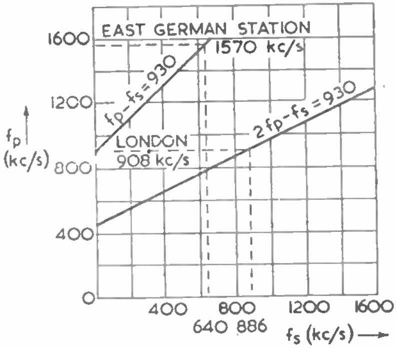


FIG. 5.—WHISTLE INTERFERENCE CHART.

(3) By-passing the cathode resistor(s) to radio but not to audio frequencies, so that the audio-frequency voltages due to cross modulation appear across it, and choosing the value of the cathode resistor(s) so that these voltages cancel the cross modulation of the desired carrier.

(4) By the application of negative feedback, e.g., by using an unbypassed cathode resistor.

Choice of Intermediate Frequency

A chart can be drawn permitting one to see readily whether a given transmitter is likely to interfere with others, for a given intermediate frequency. Fig. 5 shows such a chart, in which f_p is plotted against f_s for different types of whistle and for different intermediate frequencies.

For example, in the chart we have plotted the two straight lines: $2f_p - f_s = 2f_i$ (Type 5) and $f_p - f_s = 2f_i$ (Type 6).

Then taking 465 kc/s as the intermediate frequency, to plot these lines we have for example: two points for type (5), (0,465) and (1,500, 1,215); and two points for type (6) (0,930) and (570, 1,500).

As examples of interference we then have London (908 kc/s) interfering with the Welsh Home Service on 881 kc/s (type 5), and an East German station on 1,570 kc/s interfering with the Third Programme on 647 kc/s (type 6).

A complete chart will have many more lines than are shown in Fig. 5.

Further information on this subject has been published by the European Broadcasting Union.[2]

Pre-selection

Methods of obtaining satisfactory radio-frequency selectivity are described later (Circuitry). In general the best system for medium and long waves is the band-pass input circuit. This, however, requires a triple-gang variable capacitor, and can therefore be used only in the more expensive sets.

Fairly good results can be obtained with care in circuit and coil design using single-tuned input circuits requiring only a two-gang variable capacitor for aerial and oscillator tuning.

The interference on short waves which is within the designer's power to eliminate or minimize is chiefly image or second-channel interference; adequate pre-selection before the frequency changer is here essential for good performance. A tuned radio-frequency amplifying stage is a great help, and good results are possible on band-spread ranges with a two-gang capacitor, if each aerial circuit be pretuned to the centre of the

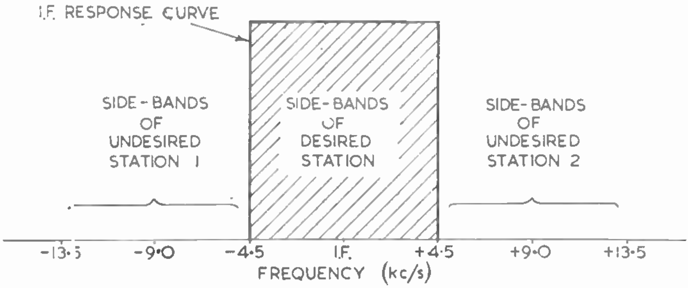


FIG. 6.—IDEAL INTERMEDIATE-FREQUENCY RESPONSE.

band and each anode circuit be gang-tuned, with a further improvement if an image-rejection circuit is employed.[3]

Intermediate-frequency Selectivity

The problem here is to eliminate adjacent-channel interference without cutting the higher audio frequencies. Unfortunately, the problem is insoluble : a compromise is necessary. For a perfect solution—even if there were no stations using frequencies not officially allocated, the standard 9-kc/s separation being maintained—the transmitters would have to restrict their modulation to 4·5 kc/s and the receivers have an intermediate-frequency band-width of 9 kc/s. To reproduce satisfactorily the audio frequencies up to only 4·5 kc/s would mean a vertical-sided, flat-topped, intermediate-frequency amplifier response (Fig. 6).

As this is impossible—and also undesirable because of the restricted modulation frequencies—considerable compromise is necessary. The B.B.C. endeavours to make 16 c/s and 10 kc/s the cut-off frequencies for transmitter modulation, so that listeners with wide-band equipment and favourable reception conditions (local signals strong in comparison with signals on nearby frequencies) may enjoy the best reproduction of which their sets are capable.

It is clear, therefore, that a receiver must have a variable intermediate-frequency band-width if it is to give the best possible performance in all areas. However, the cost of this refinement, and the fact that the number of places where the wide band-width may be used without interference has fallen considerably, have resulted in its being discontinued in all but a few of the most expensive receivers.

The compromise adopted for the normal receiver without variable selectivity and with only two intermediate-frequency transformers results in an intermediate-frequency response curve similar to that shown in Fig. 7. Because it is impossible to produce a curve with vertical sides, considerable sideband cutting is necessary to obtain even the degree of selectivity represented by this curve. Figures often used to express the performance due to such a curve are the band-width 6 db down from the peak, the attenuation in decibels (or as a ratio or percentage) 9 kc/s off tune each side, and the slope of each side in db/kc/s (between 6 and 30 db down).

The fall-off in treble response due to sideband cutting has to be made

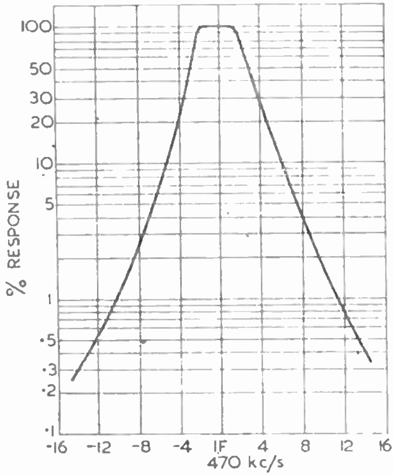


FIG. 7.—TYPICAL PRACTICAL INTERMEDIATE-FREQUENCY RESPONSE CURVE.

good as far as possible by tone-correction (treble rise) in the audio-frequency circuits of the receiver.

Typical figures for the band-width 6 db down are 5–8 kc/s; for the attenuation 9 kc/s off tune, 30–40 db; and for the slope, 4–6 db/kc/s.

To attain such figures, the transformers—in their cans but before being put in the receiver—would have a Q-factor of approximately 120–140, and a kQ of approximately 1·0–1·2 (where k is the coupling coefficient, and Q is the geometric mean of primary and secondary Q-factors, often different due to effects such as the position of the coils with respect to capacitors and to the top of the can). Damping in the receiver would reduce the Q-factor so that in practice the coupling would be critical ($kQ = 1$, giving maximum gain) or a little more or less, depending on what compromise has been chosen.

DEVELOPMENTS IN COMPONENTS

Self-contained Aerials

The " Plate " or Capacitance Aerial

This usually consists of a piece of metal foil fixed inside the cabinet, as far away from the chassis as possible. It is used for all wavebands in mains receivers, but is seldom found in battery sets. When there is no earth connection to the set, as is frequently the case, the sensitivity is quite good, because the capacitance to earth of the " aerial " completes the return path to earth of signals picked up in the mains wiring. The system is, however, naturally susceptible to electrical interference. Adding a good short-lead earth reduces sensitivity, but a long-lead connection (say 20 ft. or more) may increase the signal-to-interference ratio. This is also true of a third-lead or conduit " earth ", because the interference voltage in this is usually less than in the mains wiring itself.

A variant of this aerial is a short length of wire about 6–10 ft. long, sometimes called a " throw-out " aerial, connected to the aerial terminal of the set. It may be used alone, fixed to a wall or allowed to hang and lie on the floor; or it may be joined to a longer indoor or outdoor aerial.

The "Mains" Aerial

A plug is provided for insertion in the receiver-aerial socket: it is connected by a flexible lead either to an insulated wire of the same length as, and contained in, the set mains lead, or via an isolating capacitor of 30 pF or so, to one side of the mains. This has a poor performance from the points of view both of signal collection and of electrical interference. It is not now used much in this country.

The Telescopic Vertical Aerial

This is used for battery or mains/battery portable receivers for short-wave reception. When extended, its height is of the order of 40–50 in. Naturally its performance is poor compared with that of a normal indoor aerial of length about 20–40 ft. It is, however, superior for short waves to any frame aerial of convenient form, and is therefore used for portable sets for want of a better type. Since its characteristics are fixed, the loose coupling used for open aerials of unknown characteristics is not necessary. It is usual therefore to connect it to the "top" of the first tuned circuit, so that its electrical properties form part of that circuit.

The Loop or Frame Aerial

For *medium waves* the loop usually constitutes the tuned-circuit inductance. It is kept as far from metalwork as possible, and is designed for high Q factor and to embrace the largest possible area.

For if E = field strength, V = voltage across tuning capacitor, $\omega/2\pi$ = frequency, n = turns, A = area, L = inductance (fixed by the frequency cover and the maximum of the variable capacitor), we have:

induced voltage $\frac{V}{Q} \propto \omega n E A$ approx.

and $L \propto n^2 A$ approx. so $n \propto \sqrt{\frac{L}{A}}$ approx.

$$\therefore \quad V \propto Q\omega E A\sqrt{\frac{L}{A}} \propto Q\sqrt{A}\,\omega E\,\sqrt{L} \text{ approx.}$$

The directional properties are useful for reducing interference from an undesired station by orienting with the axis of the loop pointing to the station approximately: the theoretical figure-of-eight polar diagram with sharp nulls and broad maxima is, however, modified in practice because of capacitance pick-up, unless the loop be shielded or balanced to earth.

To satisfy the high Q-factor requirement may require the use of Litz wire, say 9/40 S.W.G., but quite acceptable results can be obtained with solid wire of about 26 S.W.G. There are several satisfactory ways of winding the loop. Examples are: single-layer solenoid, wound flat coil and solenoid folded flat. They may be wound in—or bent afterwards into—various shapes to suit the cabinet style, and supported by tying to supports or retained by moulded plastic fitments or in any other suitable manner. It is advantageous to hold the turns together, for example by impregnating, by squeezing between adhesive tapes (not suitable for tropical zones) or by using wire covered with a plastic material which bonds adjacent turns together when heat is applied by the momentary passage of an electric current through the wire. A typical " working " Q-factor would be 90 at 1 Mc/s. A few turns can be added to form a coupling coil for connection of an external aerial.

For *long waves* the loop may constitute the tuned circuit inductance, but this is expensive and awkward to manufacture because of the large number of turns of fragile wire required to obtain the necessary inductance. Frequently, therefore, the medium-wave loop is series-loaded by a small inductor with dust-core adjustment, to bring the inductance up to the value required. This is not an efficient arrangement, as the signal-collecting part of the total inductance is small. A better method is to couple the medium-wave loop to the long-wave tuning inductor by a tight coupling winding, with coupling turns chosen by experiment for maximum sensitivity. The arrangement constitutes a matched auto-transformer, basically similar to the method of reference 4, which gives the theory. The complete long-wave coil may also embody an additional coupling coil for connection of an external aerial.

The same basic principle of a loop (with an inductance much less than that required for tuning) being transformer-matched to the tuned circuit is sometimes carried a stage further by using a *single-turn loop* for both medium and long waves. For the Q-factor of this to be high enough for good results, a large conductor cross-section is essential (of the order of $\frac{1}{16}$ sq. in. minimum). In general, the sensitivity obtained with this system is not as great as that already described. The single-turn loop, however, has the advantage that, by connecting its centre to earth, pick-up of man-made interference can be made very small.

The Ferrite-rod Aeria

This aerial, made possible by the use of ferrite material,[5, 6] is basically a loop aerial using a very small coil wound on a high-permeability rod which concentrates the flux from a large area. The null point is obtained with the loop at right angles to the direction of the transmitter (as with any loop), that is, with the rod pointing towards the transmitter.

A typical design for medium and long waves comprises a rod 8 in. long by $\frac{5}{16}$ in. diameter, with coils wound on thin-walled formers, using 53 turns of 32 S.W.G. enamel spaced 32 turns per inch for medium and 138 turns of 42 S.W.G., D.S.C. for long waves, the initial inductance tolerance being about 1 per cent. The best results are obtained with spaced solenoid coils; this is not possible on long waves unless separate rods are used for each band. Fine adjustments are made later during R.F. alignment of the receiver, by sliding the coils on the rod, afterwards preventing further movement by applying a quick-drying adhesive. The assembly should be kept as far as possible from metal-work, and also from heat, bearing in mind the variation of permeability with temperature.[5]

FIG. 8.—FERRITE ROD AERIAL FOR MEDIUM AND LONG WAVE-BANDS.

To date the performance of aerials of this type has been inferior to that of the best tuned loops, but their long-wave performance is much superior to that of the medium-wave loop, series-loaded for long-wave reception.

Such high Q-factors are obtainable, owing to the low losses and high permeability of the ferrite rod, that on long waves especially it is easily possible to cut the higher audio frequencies due to the poor sideband response of a sharply tuned circuit; this point must be watched in choosing the wire gauges. Sideband cutting may be made very small if a ferrite rod aerial be used as the first inductor in a band-pass circuit, using a three-gang capacitor for this and the oscillator circuit. The small size of the ferrite-rod aerial makes it suitable for very small receivers, and also for mounting in large receivers on a pivot, a knob being fitted to rotate the rod through 180 degrees in a horizontal plane so that it may be oriented for minimum signal from an interfering station.

Radio-frequency Coils

These include aerial, radio-frequency amplifier coils and transformers, and oscillator coils.

There has been an increasing tendency to abandon air-cored coils, usually on formers of synthetic-resin-bonded paper, in favour of iron-dust-cored coils on formers of plastic material, such as moulded phenolic resin or polystyrene. The latter makes a cheaper former, but has the disadvantage that it does not remain rigid at the lowest temperature at which a wax impregnation or dip can be done. Textile covered coils need an impregnation or dip if their Q-factors are not to suffer from moisture ingress: phenolic resin formers are needed for such coils. The greatest advantage of the iron-dust-cored coils is that their inductance can be easily adjusted when they are mounted in the receiver, thus enabling the calibration and tracking to be made correct at the low-frequency end of each wave-range (or at the mid-band frequency in the case of band-spread ranges). Other advantages are: small size, economy in wire and lower self-capacitance.

In the case of short-wave oscillator coils a metal slug is often used instead of an iron-dust core: it is frequently helpful in securing an approximate constancy of oscillator voltage over the wave-range.

More recently ferrite cores have been used instead of iron-dust cores. (See Section 30, "Inductors and Transformers", and reference[5].)

Improved Q-factors necessary for higher-selectivity medium-wave aerial circuits have been obtained by using more expensive "Litz" (e.g., 9/46 S.W.G. or 12/48 S.W.G.) and/or two-pie windings.

Cheaper receivers with limited space have successfully made use of a single former holding both medium- and long-wave windings, each band having an adjusting core at its own end of the former.

Intermediate-frequency Transformers

Iron-dust Cores.—Early iron-dust cores, used for intermediate-frequency transformers, were smooth-walled cylinders cemented to threaded brass rods for screw adjustment. These have been replaced in general by iron-dust cores with moulded screw-threads: this arrangement is cheaper and provides a better Q-factor.

Ferrite Cores.—Recently it has become possible to make intermediate-frequency transformers of the same performance as the best iron-dust-core type, but considerably smaller, by using ferrite cores, and ferrite rods for reducing the effect of the can on the Q-factor. Alternatively, a better performance transformer with no increase in size is possible. The ferrite used is a non-metallic, homogeneous magnetic material, the result of research into mixed crystals, usually manganese-zinc ferrites or nickel-zinc ferrites.

The crystal structure is cubic, which avoids stresses, thus contributing to the high initial permeability and low hysteresis loss. The resistivity is very high, resulting in very small eddy current losses. Since, unlike iron dust, the ferrites do not depend on insulating material between metal particles for their high resistivity, the air-gap effect of such insulation is absent, and the permeability is much higher than that of the best iron-dust material. (See Section 30 and reference[5].)

The ferrites are useful not only as cores: in addition, a small ferrite rod may be used to aid and adjust the magnetic coupling between primary and secondary windings. The added magnetic coupling also enables the coils to be spaced so as to keep stray capacitance coupling very small, reducing asymmetry of response.

A typical Q-value for a ferrite intermediate-frequency transformer in a $1\frac{1}{8}$-in.-diameter can is 140.

For miniature designs, ferrite rods are useful for placing between can and coil to reduce the effect of the can on the Q-factor.

Winding Technique.—A worthwhile improvement of Q-factor is possible by winding each coil in two or three " pies " instead of one, thus reducing the distributed capacitance and bringing at least half of the turns nearer to the core.

Another way of improving the Q-factor, and incidentally obtaining a more compact winding, is by using silkless " Litz " wire, e.g., where a plastic coating on the enamel has been made to flow by momentary heating by an electric current, thus bonding all the strands together, or waxed enamelled " Litz ". These wires, and some new pre-fluxed wires, which do not require stripping before soldering, may be used also for R.F. coils

Sense of Connection of Coils in the Receiver

This should always be such that any stray capacitance coupling assists the magnetic coupling, because the reverse, i.e., capacitance coupling opposing magnetic, would result in too great a variation of total coupling, and thence of band-width, in production receivers.

This means that co-axial designs, with both coils wound in the same direction, should have the start of one coil " earthy ", if the start of the other coil be " hot ", or " hot " if the start of the other coil be " earthy ".

With " binocular " construction, on the other hand, the desired sense is obtained with either both starts " earthy " or both starts " hot ".

Detailed information on intermediate-frequency transformer design is given in references[7,8].

Methods of Reducing Damping

Damping by valve input and anode resistances and/or diode load resistances can be reduced either by reducing the inductance or by coil-tapping.[8] The former method is generally used only for the secondary of the final intermediate-frequency transformer, where the damping is likely to be greatest. In the latter method one or more coils may be tapped, usually with 0·5–0·75 of the turns across the source or load, according to the damping. Both methods reduce the gain, but the gain reduction may be worth-while in order to improve the selectivity. For example, in a typical case of a set using two identical miniature intermediate-frequency transformers originally designed for requirements less stringent than those of the present day, a reduction of the secondary inductance of the second (final) transformer to about one-third of its previous value increased the relative attenuation at 9 kc/s off resonance by 6 db. for a gain reduction of 3 db.

Protection against Moisture (R.F. and I.F. Coils)

The introduction of various textile-free wire coverings suitable for wave-winding has made unnecessary the elaborate sealing methods sometimes used with coils of textile-covered wire.

With textile-covered wires, one system was not to impregnate at all, but to oven-dry the coil assembly so as to drive out all moisture, and then to seal it in a can by spinning the open end over a neoprene-faced, synthetic-resin-bonded paper disc in much the same manner as that employed for electrolytic capacitors. The trimming problem may then be dealt with in three ways :

(i) The intermediate-frequency transformers are pre-trimmed using a dummy receiver or a trimming gear with capacitances to represent the average stray capacitances of a production receiver. There are several difficulties with this method : it also means a different gear or a different capacitance adjustment for every type of set.

(ii) Trimming holes are provided, the design being such that moisture ingress to the coils themselves is impossible.

(iii) Trimming holes are provided and have a means for sealing them against moisture ingress after trimming is complete. It is desirable for such seals to be removable and replaceable by a serviceman, in case re-trimming is necessary due to transit shift (a good design should avoid this except for the very roughest type of journey), or because whistle interference in the customer's area renders a slight change of intermediate frequency necessary, or because of drift over a period of years due to one or more of a variety of causes (this is unlikely with modern components, but has to be considered).

Iron-cored Transformers and Chokes

For many years the practice has been to try to prevent moisture ingress by some form of impregnation, sometimes followed by a dip or dips in a sealing compound. Probably the most economical process for satisfactory tropical protection is a wax impregnation followed by a bitumen dip. This, however, permits operation only up to 90° C., at which temperature the wax will be in a liquid or semi-liquid state inside the bitumen sealing coat. There are varnish impregnants and processes designed for operation at higher temperatures, but generally only the more expensive ones are completely satisfactory, and these tend to be too costly for the normal domestic receiver, due to their long processing time. The best of the known methods—used in the Services—is to seal the component, oil-impregnated, in a can filled with oil. It also is usually too expensive for most domestic receivers, and is better than necessary for the great majority of operating conditions.

Windings Operating at a Positive Potential

When a winding operates at a positive potential with respect to the core and/or to another winding, it is possible for electrolytic action to remove copper from any bare places on the wire, such as "pin holes" in the enamel, provided something in the nature of an electrolyte exists. In other words, corrosion occurs, eventually resulting in a break in the winding. Inspection of such an open-circuited winding reveals a green spot at the point of break, so the phenomenon is often called "green-spot" corrosion. In spite of care being taken to use the purest paper obtainable, cheekless paper-interleaved windings are much more troublesome in this respect than pile-wound bobbins, leading to the conclusion that the paper, when not completely dry, acts as a suitable electrolyte. Experiments have shown that wax impregnation alone is useless for preventing the moisture ingress which causes this corrosion, even in temperate zones, but that wax impregnation followed by a bitumen dip is very good, provided that the bitumen dip is carefully done, avoiding pin-holes. However, a pile-wound moulded bobbin is quite good for domestic-receiver transformers in the tropics if wax impregnated, and for home use if entirely unfinished. Textiles should be avoided as far as possible, although varnished cotton tape between the windings of an unfinished pile-wound bobbin gives very little trouble. For the wax-impregnated bobbin, paper insulation may be used between the windings; this appears not to be at all as troublesome as the paper between all the layers of a paper-interleaved winding, perhaps because in the latter there are many possible paths for leakage *along* the paper to the core.

A better space-factor is obtained with pile-wound bobbins than with bobbin-less paper-interleaved construction, but the latter facilitates the achievement of good inter-winding insulation, because with it there is no tendency for the end turns of one winding to slip down to the winding beneath. If there are more than two windings, or if the windings are sectionalized to reduce leakage inductance, it may be better to employ the paper-interleaved method, using wax impregnation and a bitumen dip-coat, the latter being necessary even if the set is for use only in the temperate zone.

Examples of windings operated at positive potentials with respect to the core and/or other windings are output-transformer primaries, synchronous-vibrator transformer secondaries and smoothing chokes in positive leads. (It is not usually convenient to put a smoothing choke in a negative lead, because this prohibits the use of a common can for the electrolytic capacitors each side of the choke: electrolytic capacitors with a common positive terminal are not usually made because of trouble from leakage between cathodes.)

A recent innovation is the transformer using polyester film insulation and high-temperature synthetic enamel wire covering. Impregnation is not necessary and the operating temperature can be made much higher than the maximum permissible with older materials, resulting in smaller power transformers.

Detailed information on this subject is given in reference [9].

Loudspeakers

Improvements which have taken place in the type of loudspeaker normally used for domestic receivers include the following :

(1) Dust-excluding construction has become almost universal, rendering a bag unnecessary.

(2) The ability to withstand tropical exposure has improved (e.g., cone-paper is treated so that it does not grow mould).

(3) The first result of the use of new magnet alloys was to eliminate the electrically energized loudspeaker. Still further improvements in magnet alloys have considerably reduced the weight of a loudspeaker magnet of normal flux density (6,500–10,000 lines/sq. cm.) and made possible very high flux densities (of the order of 17,000 lines/sq. cm.).

(4) The bass resonance has been lowered, and a wider, flatter response obtained, particular in the smaller sizes.

(5) The inverted (or " wafer ") loudspeaker has been developed. In this the magnet sits within the cone, resulting in a very compact unit.

(6) " Pot " types with very low external flux have been developed.

In addition, a form of electrostatic loudspeaker suitable for domestic receivers has been developed, intended for use only at the higher audio frequencies (above about 5,000 c/s). It requires a voltage of the order of 250 volts for polarization, so the normal receiver H.T. supply may be used for this purpose. As selectivity requirements severely limit the audio-frequency response of amplitude-modulation receivers, its chief use is for frequency-modulation receivers and radio-gramophones.

Increased use has been made of the elliptical loudspeaker : it will often fit into a cabinet-space too narrow for a circular loudspeaker of similar performance. Unfortunately, this frequently means that the major axis of the ellipse is horizontal, so that use is not made of the elliptical loudspeaker's property of spreading the higher frequencies in the plane of the minor axis.

For further information on loudspeakers, see Section 32.

Resistors

Fixed Carbon Types.—The composition type, unprotected save by paint, has largely been superseded by the ceramic- or plastic-cased type, removing the danger of accidental contact with other components and improving stability by reducing moisture ingress. There has been a raising of wattage limits for a given size, and an extension of ceramic-tube construction to higher wattage values, which were till recently still of the painted type.

The high-stability, cracked-carbon type has become much more widely adopted when stability is important. In effect, this means that this type should be always called for if tolerances of 5 per cent or better are required. Its price is now little more than that of 5 per cent composition types, whose stability is still not always sufficient to warrant their use where a close tolerance is needed, although being constantly improved.

The use of silicone varnish for coating these high-stability resistors has given better protection against moisture ingress, thereby still further improving their stability. The silicone coating also withstands heat much better than earlier lacquers. This type is also available in a ceramic case.

Fixed Wire-wound Types.—An economical type has been evolved using a silicone compound as a coating for the wire round its ceramic tube. This silicone compound stands heat without crazing, stands up well to hot, damp conditions and prevents corrosion.

Temperature-sensitive Types.—These and their uses are fully described in Section **3**, under " Semi-conductors ". To date probably their most important use in the domestic radio field has been in improving the scale-illumination brightness in universal mains receivers.

Variable Composition Resistors.—These are mainly used in broadcast receivers as volume and tone controls.

In the past the greatest defects have been change of resistance due to the effect of moisture, and noise.

Tracks have been improved so that moisture no longer has much effect, and although noise troubles have not been eliminated, they have been much reduced, e.g., by the use of carbon brushes both for track and collector ring.

The receiver designer can also reduce noise by arranging that no D.C. flows through the control, e.g., detector diode current or amplifier-valve grid current, the latter being the worse offender.

For further information on resistors, see Section 28.

Valves

The physical size of valves has been greatly reduced, and glass bases have become almost universal. Most of the valves using the International or the British Octal base have been replaced by types using seven-, eight- or nine-pin all-glass bases. In addition, the heater wattages have in general been considerably reduced.

Apart from the obvious advantage of the smaller chassis space conferred by the size reduction, there is the advantage of lower losses at the higher frequencies due to the glass bases and shorter lead lengths.

The reduction in heater wattage means smaller mains transformers in A.C. sets and smaller dropping resistors in A.C./D.C. sets; it also facilitates keeping electrolytic capacitors and cabinet roofs cool enough; and makes it easier to keep frequency drift down to reasonable limits.

The range of battery valves with 1·4-volt filaments has made possible the all-dry battery set, which increases in popularity both as a table and as a portable set. The range has recently been extended by the introduction of new types with only half the filament consumption of the older types.

For further information on valves, see Section 23.

Metal Rectifiers

Convection-cooled metal rectifiers for power supplies have become smaller and cheaper, so that in some cases they can compete with valves for price, because of savings in the mains transformer, the absence of a valve-holder and the fact that no series protection resistor is needed. Their chief disadvantage is that it is sometimes difficult to keep down to their maximum permissible operating temperature, especially in sets designed for the tropics. Their chief use has been in battery/mains receivers, where the provision of power at a suitable voltage and current for the heater of a rectifier valve would be awkward and expensive.

Miniature types—with current ratings of the order of 1 mA—are suitable for bias supplies or battery economy circuits.

A recent development is the contact-cooled rectifier, very much more compact than the convection-cooled type, and dependent for its cooling on contact with the receiver chassis, there being no cooling fins. Whereas the convection-cooled type should be mounted with its fins in a vertical plane, the contact-cooled type is not restricted as to mounting position. The forward resistance has been made so small that a worthwhile reduction (of up to 20 per cent or so) in the A.C. voltage supplied to the rectifier may be achieved, compared with that required for a valve.

For further information on metal rectifiers, see Section 25.

Batteries

A great reduction in the size of H.T. batteries for portable receivers has resulted from the development of the layer-built battery. It is limited to cells of small capacity which are, however, perfectly satisfactory for the H.T. requirements of portable receivers.

The layer-built battery has about two-thirds the weight and half the volume of the cylindrical-cell type for the same voltage and capacity.

" Balanced life " combined H.T. and L.T. batteries have various advantages compared with separate units. The arrangement is more compact, and battery changing simpler. More important, the L.T. and H.T. voltages fall at rates related to give the best performance and the longest life, the valve manufacturers' recommendation being that the L.T. voltage should have dropped to 1·0 volt per cell when the H.T voltage has fallen to 0·75 volt per cell.

Further information on batteries is given in Section 36.

Fuses

Current-operated Fuses.—In broadcast receivers by far the most used current-operated fuse is the cartridge fuse to B.S. 646B, usually 1¼ in. long. The specification says that it must be capable of carrying its rated current for 1,000 hours and should blow within 1 minute after the current reaches a value 1·75 times the rating.

In many applications the B.S. 646B fuse may be unsatisfactory because it blows on a surge. In such cases a special surge-proof fuse may be used, e.g., the Mag-Nickel delay fuse, or the spring-operated type, in which a soft-soldered joint holds against the force of a spring until the excess current has melted the joint.

B.S. 646B lays down a standard colour code : see Table 3.

Heat-operated or Temperature Fuses.—The primary purpose of a temperature fuse is to reduce fire risk by breaking a circuit when the temperature of a component such as a transformer exceeds a certain limit.

Fig. 9 illustrates a typical temperature fuse fitted to a mains transformer. A copper strip B is well insulated and fixed between the H.T.

TABLE 3.—COLOUR CODE FOR CURRENT-OPERATED FUSES

Colour	Rating (A)	Colour	Rating (A)
Green and yellow	0·010	Green	0·750
Red and turquoise	0·015	Blue	1·0
Eau-de-Nil	0·025	Light blue	1·5
Salmon pink	0·050	Purple	2·0
Black	0·060	Yellow and purple	2·5
Grey	0·100	White	3·0
Red	0·150	Black and white	5·0
Brown	0·250	Orange	10·0
Yellow	0·500		

(British Standards Institution)

and heater windings, the order of winding being : primary, screen, H.T., heater and rectifier heater. It projects at both ends of the coil, having a lug at one end to which the start of the primary winding is soldered, and at the other end is jointed to a phosphor-bronze spring A by a fusible alloy of low melting point (95° C.). The other end of the spring is secured to the insulated tag jacket of the transformer by a tag which also serves as connection to one mains lead.

If the temperature of the transformer rises sufficiently to cause the heat conducted along strip B to melt the fusible alloy, the spring A separates from the strip, thus breaking the mains circuit. This arrangement has been arrived at after experiments to ensure that whichever winding or part of a winding be short-circuited, the temperature rise of the hottest part of the transformer cannot exceed 135° C., when measured 2 minutes after the fuse has opened, thus complying with the revised B.S. 415 (fire risk in impregnated interleaved coils).

The fusible alloy is a mixture of bismuth, lead and tin in the proportions which produce a eutectic alloy, that is, one which melts sharply at a single definite temperature. Any other proportions would have an undesirable plastic range.

It is obvious that such a circuit-breaker is proof against surge. As an example of the time taken for the temperature fuse to operate, a sample opened 1¾ minutes after the inner half of the H.T. winding had been short-circuited.

When rejointing an opened temperature fuse of this type, it is not

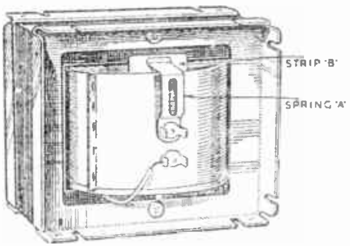


FIG. 9.—TEMPERATURE FUSE FITTED TO A MAINS TRANSFORMER.

normally necessary to add more alloy, because the quantity of alloy used in the original joint is usually sufficient to permit repair by the application of heat alone. Care should be taken to avoid adding any solder other than the special alloy. Any flux used either during manufacture or repair must, of course, be non-corrosive, because the slightest trace of corrosive flux close to fine-wire windings might cause damage.

Capacitors

The following brief notes on capacitors refer mainly to changes that have taken place in their use in domestic broadcast receivers. For further information on the subject reference should be made to Section 29.

Ceramic

Ceramic capacitors have improved recently in several ways: (1) there is a greater variety in size, capacitance, working A.C. and D.C. voltages and in temperature coefficient; (2) for a given capacitance, size has been reduced; (3) temperature stability of high-permittivity types has improved. The most popular form is the tubular contruction, but cup and disc types are also used.

The protective envelope may be a wax coating (several thin dips), an outer ceramic tube with sealed ends or a phenolic coating with wax impregnation. Where tolerances of 10 or 20 per cent are acceptable (e.g., oscillator feed capacitors, intermediate-frequency filter capacitors, and radio-frequency coupling capacitors) their compactness and low price make ceramic capacitors eminently suitable. However, they are not generally used for tuned circuits, unless deliberately selected to cancel a positive temperature coefficient of the remainder of the circuit, to reduce frequency drift.

Drift Correction

The following is an example of the process used for choosing a drift-correction capacitor.

Measure the average intermediate-frequency drift and oscillator drift of at least three prototype sets, starting 3 minutes after switching on.

Let the intermediate-frequency drift be . . . . . i kc/s per ° C.
Let the oscillator drift be . . h kc/s per ° C.
(in practice i and h are usually negative).

Let f_h (kc/s) be the oscillator frequency (assumed higher than the signal frequency).

Let C (pF) be the total circuit tuning capacitance.

For net zero drift the oscillator drift should be the same as the intermediate-frequency drift, i.e., i kc/s per ° C. or $\frac{i}{f_h} \times 10^6$ p.p.m./° C. in frequency or $-\frac{2i}{f_h} \times 10^6$ p.p.m./° C. in LC product (where p.p.m. is parts per million).

But at present the LC product drift is $-\frac{2h}{f_h} \times 10^6$ p.p.m./° C.

Assume that this drift is entirely due to capacitance change, i.e., that the inductance does not drift. This is strictly incorrect, but the error is usually small enough to be ignored for an approximate solution : the final value depends on trial and error.

Let C_n be the part of C which is intended to be replaced by a capacitor with a negative temperature coefficient, of $-n$ p.p.m./° C. (usually $n = 750$).

Then, equating total capacitance change per ° C., we have :

$$-\frac{2i}{f_h}10^6 C = -nC_n - \frac{2h}{f_h} \times 10^6 (C - C_n)$$

(This assumes that $C - C_n$ has the same coefficient as the original C, which although incorrect, is near enough for our purpose.)

$$\therefore \quad C_n\left(\frac{2h\,.\,10^6}{f_h} - n\right) = C\,.\,\frac{2\,.\,10^6}{f_h}(h - i)$$

$$\therefore \quad C_n = \frac{C\,.\,\dfrac{2\,.\,10^6}{f_h}(h - i)}{\dfrac{2h\,.\,10^6}{f_h} - n}$$

$$= C\,.\,\frac{2\,.\,10^6(h - i)}{2h\,.\,10^6 - nf_h}$$

In a typical case,

$h = -0{\cdot}6$ kc/s per ° C., $i = -0{\cdot}1$ kc/s per ° C., $n = 750$ p.p.m./° C., $C = 50$ pF, $f_h = 2{,}000$ kc/s

$$C_n = 50\,\frac{2\,.\,10^6(-0{\cdot}5)}{-750\,.\,2{,}000 - 2(0{\cdot}6)10^6}$$

$$= 50\,\frac{10^6}{(1{\cdot}5)10^6 + (1{\cdot}2)10^6} = 18{\cdot}6 \text{ pF}$$

If it be impossible to replace all the calculated capacitance by a negative temperature-coefficient capacitor, as much as possible should be so replaced, and the negative temperature-coefficient capacitor should be moved to a warmer place, e.g., near to a valve, so that its capacitance change is increased.

Tubular Foil and Paper Capacitors

Capacitors with solid impregnants have been found generally unsuitable where appreciable A.C. stress (over 100 volts) or over 1,000 volts D.C. are involved. For such uses, petroleum jelly, or better, special viscous oils, are advisable, the latter requiring leak-proof containers. Petroleum jelly does not leak from ordinary containers except at unusually high temperatures.

It is often necessary to ensure that capacitors whose insulation resistance may fall to low values are not, for example, used for coupling from anode to grid. The insulation of the simplest types of wax-coated, cardboard-cased capacitors may, for instance, fall as low as 5 MΩ within a year or two, even in a temperate climate. Table 4 indicates the approximate value to which the insulation of various types may ultimately fall; it should be recognized, however, that manufacturers are constantly improving their products.

TABLE 4.—TYPICAL ULTIMATE VALUES OF INSULATION RESISTANCE

Type of Case	*For United Kingdom Use* ($M\Omega$)	*For Export to Tropical Areas* ($M\Omega$)
Simplest waxed-cardboard tube	5	(not suitable)
Phenolic-resin moulding . .	20	5
Hard wax moulding . .	100	10
Better waxed-cardboard tube (spun over deep concave metal end caps)	1,000	100
Metal tube with neoprene seals .	1,000	1,000

Variable Air Dielectric Capacitors

The tendency here has been towards smaller physical sizes with only very small reductions in capacitance swing (due to closer spacing of the smaller plates) and capacitance tolerance, and little or no increase in the effect of microphony. These changes have been accompanied by a reduction in cost because of the simpler construction employed: bent-up frame or U-type instead of bar frame or E-type. In addition, there is a miniature U-type which has a slightly smaller capacitance swing and a larger tolerance. While satisfactory for medium- and long-wave sets, this construction could not always be relied on for short-wave use, because it is prone to microphony troubles. It is, however, very useful for miniature and portable receivers.

The mounting of the variable capacitor must be such that the frame and shaft are not appreciably distorted. Grommets are the usual mounts, in order to reduce microphony as well as to avoid stresses in the frame; but spring mounts have been used. Grommets should be of a soft plastic material, such as P.V.C., or rubber treated so as to prevent perishing. The mechanical arrangements must be such that the grommets are not over-compressed, whatever the tolerance on the components; on the other hand, they must not be so loose that the capacitor rocks from side to side when the receiver is being tuned. It is advisable to have two out of the usual three grommets spaced well apart, in a line at right angles to the spindle.

The tension in the drive cord should be no greater than necessary to secure a smooth, non-slip drive, since too great a tension would result in too much wear on the cord and exert too much side pull. Permanent side pull may be made negligible by arranging that the cord enters and leaves the drive drum by the same tangent, which should coincide as nearly as possible with the line joining two of the mounting grommets. The latter condition assists in minimising side-to-side rock when tuning.

Alternatively, a Bowden-wire arrangement may be used, so that the internal wires exert the desired torque on the drum while the outer sleeves are fixed to both the chassis and the capacitor. There is then negligible side pull on the capacitor during tuning; the wires just slide within the sleeves as they pull the drum round.

Another method of driving the capacitor is to use a slow-motion gear drive so that the cord drum is reduced to about 1½ in. diameter for a scale length of 8 in. This method is useful where a large drum cannot be accommodated but where a fairly long scale is desired. The drive may use two spring-coupled co-axial toothed wheels, both engaged with the driving pinion, so as to avoid back-lash.

One cause of microphony, especially if the drive drum be rather large, is that the drum acts as a diaphragm, driven by the sound waves from the loudspeaker. Plenty of large holes in the drum will usually cure this trouble.

Electrolytic Capacitors

The chief improvements making for increased reliability are: (1) the almost exclusive use of sealed aluminium containers, and (2) the exclusion of all metal other than aluminium from the construction.

The chief enemy of reliability is corrosion: therefore all foreign matter must be rigorously excluded; and this requires a perfect seal from the atmosphere, a design which specifies materials of high purity, and great care and cleanliness in manufacture.

Great reductions in size have been achieved by replacing plain anode foils by either etched foils or by gauze hot-sprayed with aluminium (sometimes called "fabricated plate" construction).

Voltage.—The manufacturer specifies two voltages, one the maximum working voltage and the other the maximum surge voltage.

The working voltage is equal to the sum of the direct voltage and the peak ripple voltage under conditions which make this sum a maximum. That is with maximum marked mains voltage input (plus 6 per cent for Electricity Supply allowed variation) to that mains tapping which produces the largest direct voltage. The voltage may be measured by means of a peak-reading voltmeter, or a close approximation may be made by adding $\sqrt{2}$ times the product of r.m.s. ripple current and reactance to the direct voltage.

The surge voltage is that which occurs every time the set is switched on or under certain accidental conditions such as the open-circuiting of the H.T. line due to a dry joint.

Ripple Current and Ambient Temperature.—In some uses, notably as a reservoir or integrating capacitor, considerable A.C. flows through the capacitor. The high-power factor of an electrolytic type then results in a temperature rise which, unless limited, will cause permanent damage.

It follows that the user must be careful to keep the ripple current below a maximum laid down by the manufacturer, and also to ensure that the ambient temperature does not exceed a stipulated limit.

The calculation of ripple currents is a complicated matter, depending on several factors, including the D.C. supplied, the source impedance, the forward rectifier resistance and the capacitance value. Such calculation is not recommended; it is better to measure the current in a prototype receiver, being careful to do this under conditions giving the highest value likely to occur, and using a thermal meter to avoid wave form error. However, to get a rough idea as to whether a given capacitor will do, the ripple current may be taken as 1·25–1·5 times the D.C. for full-wave circuits and 2–3 times the D.C. for half-wave circuits.

CIRCUITRY

The Aerial Circuit

Good selectivity before the frequency changer is essential for present-day broadcasting conditions, especially in Europe. A useful figure for expressing selectivity of the type required here is the image ratio, in decibels, i.e., the ratio : radio-frequency input of image frequency to radio-frequency input of signal frequency, both for standard output. If only a single tuned circuit be used, the high Q-factor necessary for good image rejection will result in some side-band cutting on long waves and to a certain extent on medium waves, at least at the low-frequency end of the band, resulting in a partial loss of the higher audio frequencies.

Probably the best and most used system for a single tuned circuit is mutual inductance coupling with a high-impedance primary, resonating below the low-frequency end of the band with the combined capacitance of the aerial, the primary self-capacitance, and added shunt capacitance where used (the latter usually on long waves only). Whatever aerial is used (an aerial for medium and long waves may be regarded as mainly capacitative, of value approximately 30–450 pF, depending on length, with 100 pF as the average), to avoid feedback troubles the resonant frequency should be well clear of the intermediate frequency. This results in a primary inductance of about 3,000 μH for medium waves. The coupling factor, a compromise mainly between gain, image ratio and tracking error requirements, is of the order of 15–20 per cent and the Q-factor about 160 at 1 Mc/s, out of the set, falling to about 100 for the "working" value in the set. The 3,000-μH primary may also be used for long waves by placing it between the medium- and long-wave tuned windings on the same former, and switching a capacitance across it to lower the long-wave resonance to a suitable value. This latter depends on whether it is desired to obtain a fairly uniform performance over the whole of the band, in which case the resonance would be at about 0·8 of the lowest tuning frequency with the smallest expected aerial, or to provide maximum gain at the frequency of a particular station such as Droitwich : for this a series capacitor of about 400 pF would be used to limit the effect of aerial change, and the resonance made to occur at 200 kc/s with a 100-pF dummy aerial.

Greater uniformity of aerial gain over the medium waveband may be obtained by adding "top-capacitance" coupling to the magnetic coupling, i.e., connecting a capacitor of value about 3 pF between the aerial terminal and the grid end of the tuned winding.

The relationship between adequate image ratio, Q-factor and loss of higher audio frequencies may be theoretically considered in an approximate way as follows, calculating Q from image ratio as described elsewhere and in reference [10]. For an image ratio of, say, 50 db at 200 kc/s (the aerial coupling improves this to, say, 60 db in practice) the "working" Q would be about 60, resulting in a band-width of $200/Q =$ 3·4 kc/s, i.e., the response is -3 db at about 1·7 kc/s.

To avoid interference from stations operating on or near the intermediate frequency, a rejector circuit should be included. This is usually connected across the primary winding of the aerial coil, and takes the form of a capacitor of about 30 pF in series resonance with an inductor of Q-factor about 150, tuned by an iron-dust core to the intermediate frequency. In a cheap set the rejector may be omitted, but

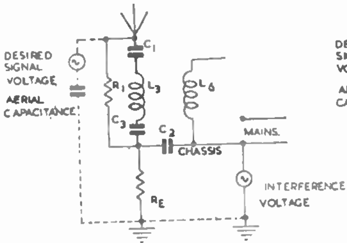


FIG. 10.—AERIAL INPUT CIRCUIT FOR NON-ISOLATED CHASSIS.

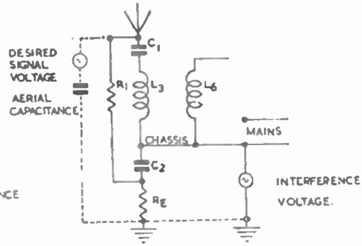


FIG. 11.—ALTERNATIVE AERIAL INPUT CIRCUIT FOR REDUCTION OF MAINS-BORNE INTERFERENCE.

provision should be made for easy fitting of one by a dealer in areas where this type of interference prevails.

Better image rejection than is possible with a single-tuned circuit, and at the same time freedom from serious side-band cutting, can be obtained by using a band-pass input circuit. This method is little used today because of its cost (it requires extra coils and a three-gang capacitor), but details are given in reference [10].

Receivers with Non-isolated Chassis

Fig. 10 shows a medium-wave aerial circuit for a non-isolated chassis, simplified by omitting the wave-change switch; C_1 (500 pF) is the aerial isolating capacitor, which prevents the aerial becoming live in the event of an accidental short-circuit to chassis of any of the aerial-coupling coils or of their associated wiring, e.g., in the wave-change switch. Like service is performed by C_3 (2,000 pF) for the earth lead: C_2 is a chassis/earth capacitor, made as large as possible consistent with avoiding appreciable electric shocks to anyone handling the earth lead (0·01 μF).

The inductor L_3 is the high-impedance coupling coil, and L_6 the tuned winding. It will be noted that R_E is the resistance of the earth connection (no " earth " has an entirely negligible resistance).

Fig. 11 shows a circuit which is often used. It effectively isolates the aerial and earth sockets from the " live " chassis (not quite as well as Fig. 10 though, because there the aerial has C_1, C_2 and C_3 in series to the chassis instead of C_1 alone in Fig. 11). However, Fig. 10 is preferable, in spite of the extra cost of C_3, for the following reason connected with mains-borne interference.

In Fig. 11 the interference voltage between mains and earth produces a voltage across C_2, part of which reaches L_3 via the aerial capacitance, and is passed on to the rest of the set in the same way as the desired signal. In Fig. 10, however, this cannot occur. It is true that the interference voltage across C_2 will cause small circulating currents via the distributed capacitance between L_3 and L_6 or via " top " capacitance coupling if used, but the signal produced at the grid end of L_6 by this means is small compared with that produced there in Fig. 11. There is also an interference voltage across R_E: this is kept small by using a good " earth ".

Filter chokes in the mains leads can be added to reduce the interference voltage across C_2: they are obviously useless unless an earth connection is provided.

Since there is no D.C. connection between the aerial and earth, due to the isolating capacitors, it is necessary to provide a path which will discharge static electricity collected by the aerial, thus avoiding the possibility of insulation breakdown; this is provided by a resistance R_1 (1 MΩ).

Short Waves

(a) *Large Cover Bands.*—For domestic receivers it is usual to regard the aerial as a resistance of 400 ohms, and to choose coupling-coil turns experimentally for a compromise between gain, image ratio and signal-to-noise ratio. In practice, the aerial impedance and phase angle vary widely over the short wavebands, so the 400 ohms is only a very rough average.[11] Loose coupling (of the order of 20 per cent or less) is therefore advisable in order to avoid undesirable effects of aerial resonances on the tuned circuit.

(b) *Bandspread Bands.*—If a radio-frequency amplifying stage is included, it is satisfactory to have one coil for each band and to pre-tune it to the centre of the band. If the aerial circuit precedes the frequency changer directly, probably the best arrangement is to pad the gang capacitor to restrict its cover, and couple the aerial by capacitance. Fig. 12 shows an economical arrangement in which only the coils are switched, and one capacitor (C_1) serves both as added "top" capacitance coupling on the medium-wave and S (3·2–9·3 Mc/s) bands, and as aerial coupling on the bandspread bands.

The Radio-frequency Amplifier

Not many domestic receivers employ a radio-frequency amplifier today, because it adds to the cost, and quite good results can be obtained without one. However, it can improve performance considerably, especially on short waves.

The figures in Table 5 based on the radio frequency amplifier to be described here, illustrate the improvements to be expected.

The 6F1 valve, having a high slope (9 mA/volt) yet moderately low anode current, produces less noise than the variable-mu type, which also has a much lower slope. This results in quite a good signal-to-noise ratio (e.g., 20 db for a signal of 7 μV at 15 Mc/s). The receiver of Fig. 24 incorporates this amplifier.

Aerial Circuit

On medium waves a band-pass aerial circuit is used: its good selectivity reduces the possibility of cross-modulation in the radio-frequency valve. In Fig. 13 mutual

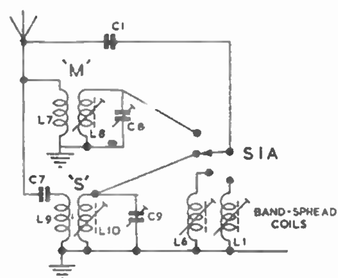


FIG. 12.—ECONOMICAL ARRANGEMENT FOR AERIAL INPUT CIRCUIT WITH BAND-SPREAD SHORT-WAVE BANDS.

TABLE 5.—EFFECTS OF ADDING R.F. AMPLIFIER

Waveband	*Set with Radio-frequency Amplifier and Band-pass Input for Medium Waves*		*Set without Radio-frequency Amplifier but with Good Single-Tuned Aerial Circuit*	
	Gain Aerial to Mixer Grid (db)	*Image Ratio (db)*	*Gain Aerial to Mixer Grid (db)*	*Image Ratio (db)*
Medium . . .	16–21	70–74	9–12	35–55
3–9 Mc/s . .	24–26	30–57	9–10	18–29
Bandspread Bands, 11–49 m. . .	16–42	33–65	6–14	16–18

inductance and shunt capacitance coupling between the two tuned circuits (L_{14}, C_{33} and L_{10}, C_{32}) are provided by L_9 and C_{12} (0·02 μF). This mixed type of coupling reduces band-width variation as the set is tuned through the medium-wave range.

In order to reduce cross-modulation from strong stations, the un-by-passed cathode resistor R_4 (150 ohms) introduces negative current feedback with 6 db gain reduction, which can be increased by 20 db by operating the local/distant switch S_1. This puts R_7 (1,800 ohms) into circuit, but does not alter the bias to the radio-frequency valve, since the grid is D.C.-connected to the junction of R_4 and R_7 via R_6 (10,000 ohms).

To compensate for the reduction of valve-input capacitance, which would cause detuning with increase of feedback, a similar capacitance is coupled into the grid circuit by a much larger capacitance C_{11} (390 pF), transformer coupled by the winding L_{50}. It is done this way because the capacitance required (2 pF) is too small for switching directly across the tuned circuit.

The local station switch has four positions, and can be set to increase the feedback on medium waves or on the 3–9-Mc/s band, either individually or together. There is also a position for maximum gain on both bands.

The 3–9-Mc/s band uses a single gang-tuned aerial circuit.

Each bandspread band has a circuit pretuned to the centre of the band, with tapped capacitance aerial matching.

R_5 (39 ohms) prevents parasitic oscillation due to an unavoidably long lead to the gang capacitor.

Anode Circuit

On medium waves the anode circuit (Fig. 14) is basically a shunt *m*-derived low-pass filter with a cut-off frequency of 1·9 Mc/s. R_9 and R_{11} are the terminating resistances; and C_a and C_b, made up of the valve and circuit strays, are the input and output capacitances. The components L_{15} and C_{17} (3·3 pF) resonate at 2·6 Mc/s, giving the filter a sharp cut-off, which is a maximum at approximately the highest image

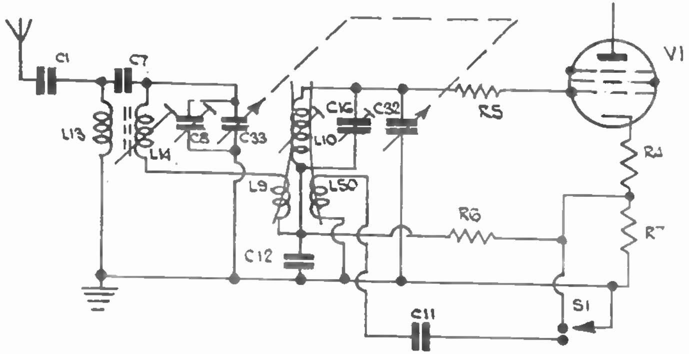


Fig. 13.—Band-pass Tuning Circuit for Radio-frequency Amplifier.

frequency on the medium-wave band. In addition, L_{17} and C_{22} (56 pF) form a series rejector circuit tuned to the intermediate frequency. This not only secures good rejection of signals of intermediate frequency (50 db for ratio of intermediate-frequency signal to desired signal at 550 kc/s for same output) but also improves the signal-to-noise ratio: "noise" here includes all radio frequencies; the rejection of a narrow band centred on the frequency at which the frequency changer is most sensitive (i.e., the intermediate frequency) reduces the noise in the frequency changer output.

For the 3–9-Mc/s band (Fig. 15) the inductance L_{16} is tuned by the gang capacitor C_{33} and the trimmer C_{18}. The primary inductor L_{15} is the same coil that is used in the low-pass filter for medium waves, and is tuned by the valve and stray capacitances C_b to about 1·8 Mc/s.

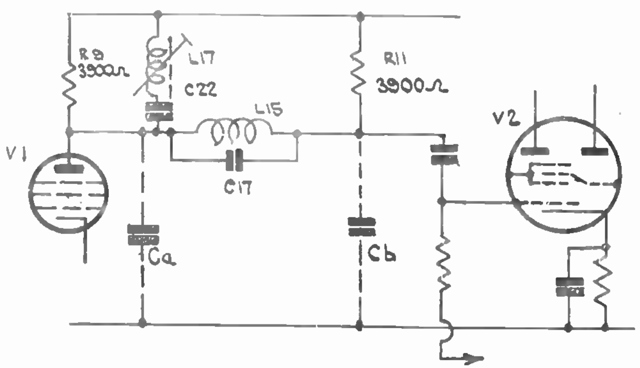


Fig. 14.—Medium-wave Interstage Coupling.

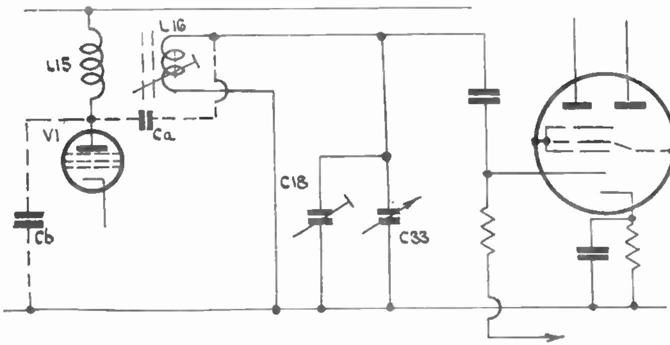


FIG. 15.—INTERSTAGE COUPLING FOR 3–9-Mc/s BAND.

The stray capacitance C_a acts as " top " coupling which combines with the mutual inductance ($k \doteqdot 0{\cdot}1$) of L_{15} and L_{16} to give a gain constant within ± 1 db over the band.

On the bandspread bands the capacitor which couples the tuned circuit to the anode is chosen so as to resonate with the tuning inductor at the image frequency. It is sufficient to use the same coupling capacitor for each pair of adjacent bands. As an example of the *extra* image rejection obtained by this method, this figure is 20 db for the 25-m. band. The method necessitates the oscillator frequency being lower than the signal frequency.[3]

On the four lower-frequency bandspread bands the cathode resistance is switched to a higher value in order to reduce the gain (although not designed as a variable-mu valve, the 6F1 characteristics permit limited control of gain by bias change).

Simplification Using a Two-gang Capacitor

In the above radio-frequency amplifier, one section of the three-gang capacitor is not used on the bandspread bands, and by dispensing with it on the medium waveband and the 3–9-Mc/s band a two-gang capacitor could be used, at some sacrifice of performance on these bands.

On medium waves a high Q, high-impedance primary aerial circuit would replace the band-pass arrangement, the same low-pass filter being used in the anode circuit. Gain and signal-to-noise ratio on medium waves need not suffer from this change. On the

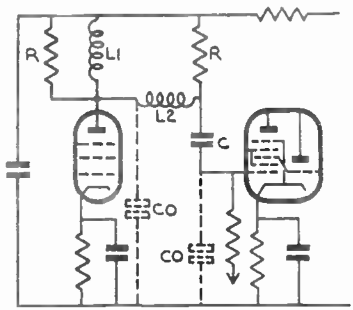


FIG. 16.—ALTERNATIVE ARRANGEMENT FOR 3–9-Mc/s.

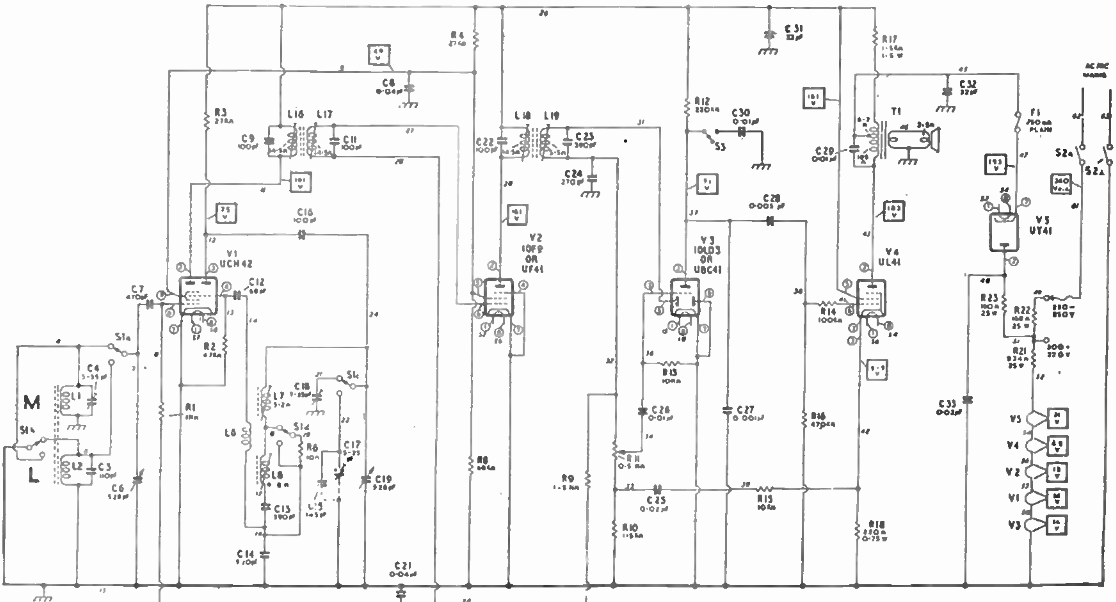


FIG. 17.—TYPICAL SMALL A.C./D.C. RECEIVER WITH FERRITE-ROD AERIAL.

other hand, image ratio, freedom from cross-modulation and from side-band cutting would suffer somewhat, although the loose coupling made possible by having the high gain of the amplifier means that image ratios of the order of 55 (including the effect of the anode circuit) could be obtained.

On the 3–9-Mc/s band the gang-tuned anode circuit could be replaced by a circuit such as that of Fig. 16, in which L_1 resonates with the 6F1 anode "strays" at about 3·75 Mc/s and L_2 with the frequency changer grid "strays" at about 8 Mc/s, the gain peaks being damped by resistors R of about 2,500 ohms each. In this way, gain and signal-to-noise ratio would be affected little, but the image ratio would fall to about 18–29 db.

The Frequency Changer

The design of frequency changers in mains-operated broadcast receivers has not changed appreciably in recent years, the valve being almost always a triode-heptode or triode-hexode and the circuit a standard one with small variations, chiefly in the type of oscillator feedback employed. A typical circuit is shown in Fig. 17, which also illustrates an economy measure now almost universal in low- and moderately-priced receivers: the omission of the cathode biasing resistor and its attendant by-pass capacitor. With this omission the cathode current is kept within its limit by lowering the screen volts. Under no-signal conditions a small negative bias is applied to the frequency changer from the automatic gain control diode. This bias voltage varies according to the type of diode and its life, being of the order of 0·5–1·2 volt. It is the net result of contact potential (positive) and initial velocity of electrons (negative).[12]

For battery receivers, frequency changers, such as the heptodes DK92 (filament current 0·05 A.) and DK96 (filament current 0·025 A.) have been produced. A typical receiver employing the DK92 (or equivalent 1C2) is shown in Fig. 19. The following recommendations, however, apply to both the DK92 and the DK96; see reference [13].

The oscillator circuit should be of the tuned-grid type with an anode reaction winding. This avoids too great an oscillatory voltage on the second grid (virtual oscillator anode), which would reduce the conversion gain. Anode series feed is used for better oscillator drive and better high-frequency performance.

For medium and long waves the reaction winding should be wound in a single layer on top of or underneath the tuned winding. For short waves the reaction winding should be of thin wire (34–38 S.W.G. approximately) interwound with or close to the earthy end of the tuned winding. No booster coil is needed for the DK92, but one should be used with the DK96 for a large-cover short-wave band. For example, for a 6·0–18·7-Mc/s band, a booster circuit (Fig. 18) consisting of an inductor resonating with a capacitor of 68 pF at about 4·8 Mc/s would be connected across the reaction winding. The booster circuit increases the drive at the low-frequency end of the band, where unaided it would be weak.

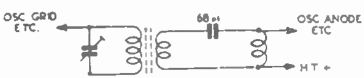


FIG. 18.—OSCILLATOR BOOSTER CIRCUIT.

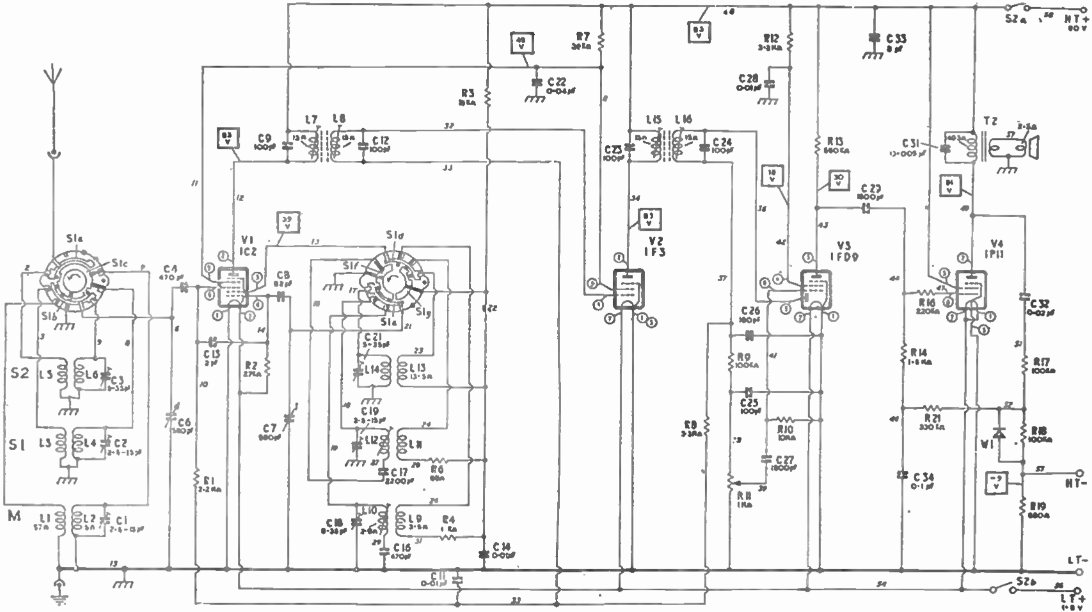


FIG. 19.—FOUR-VALVE, THREE-WAVEBAND RECEIVER WITH BATTERY ECONOMY CIRCUIT.

With both valves, a neutralizing capacitor is necessary to avoid excessive "pulling" on short waves above about 13 Mc/s. It is adjusted for minimum oscillator volts on the signal grid. The value required is about 2 pF, and varies but little with frequency. At one time a variable capacitor was required, and was often constructed by winding a few turns of insulated thin wire round a thick wire, later removing turns to adjust the capacitance. but a close-tolerance fixed capacitor suitable for this work has now become available. For the 6·0–18·7-Mc/s band, neutralization at the upper frequency trimming point (around 17 Mc/s) is satisfactory.

Sometimes it is desired to manufacture mains and battery versions of the same basic receiver: economical production is then assisted by using common components as far as possible, and the designer can help here by starting with the battery set. The same coils can then be used in the mains versions, with in most cases no change other than shunt feed instead of series feed (to avoid re-modulation feedback along the H.T. line due to higher gain in the mains set) and a higher oscillator anode feed resistance to limit oscillation amplitude.

The Intermediate-frequency Amplifier

In recent years there has been very little change in the design of intermediate-frequency amplifiers, apart from the improvement in transformers already described. Variable selectivity (usually done by switching into circuit a tertiary winding closely coupled to one coil and in series with the other) is now seldom found, partly because there are few places where a wide band-width can be used.[14]

As in the case of the frequency changer, the cathode bias resistor and by-pass capacitor have been increasingly omitted from the intermediate-frequency amplifier valve in mains-operated receivers. A defect is that the diode bias may be so small that intermediate-frequency valve grid current flows and damps the intermediate-frequency transformer, causing a drop in sensitivity with inputs small enough to generate little or no automatic gain control voltage (A.G.C.). With this system it is therefore inadvisable to carry out production testing with a small audio output such as the 50 mW frequently employed in the past, but to use instead a larger output such as 500 mW, so that sufficient automatic-gain-control volts are generated to bias the valve beyond grid current cut-off. Otherwise the "spread" of sensitivity in production sets will be very large and will make more difficult the detection of faults in the receiver.

Only rarely, in the most expensive receivers, is more than one valve used in the intermediate-frequency amplifier of a mains receiver, but one extra valve is sometimes added to a battery receiver using 1·4-volt valves, in order to increase gain, and this may use a resistive load for economy, giving a gain of about 15–20 db.

The Detector

By this is meant the rectifier which produces an audio voltage from the modulated intermediate-frequency voltage. It is sometimes called the "second detector" to distinguish it from the mixer or first detector. Design has changed little in the past decade, the diode being almost universally employed. The negative direct voltage developed by this diode is frequently used to provide automatic-gain-control bias for

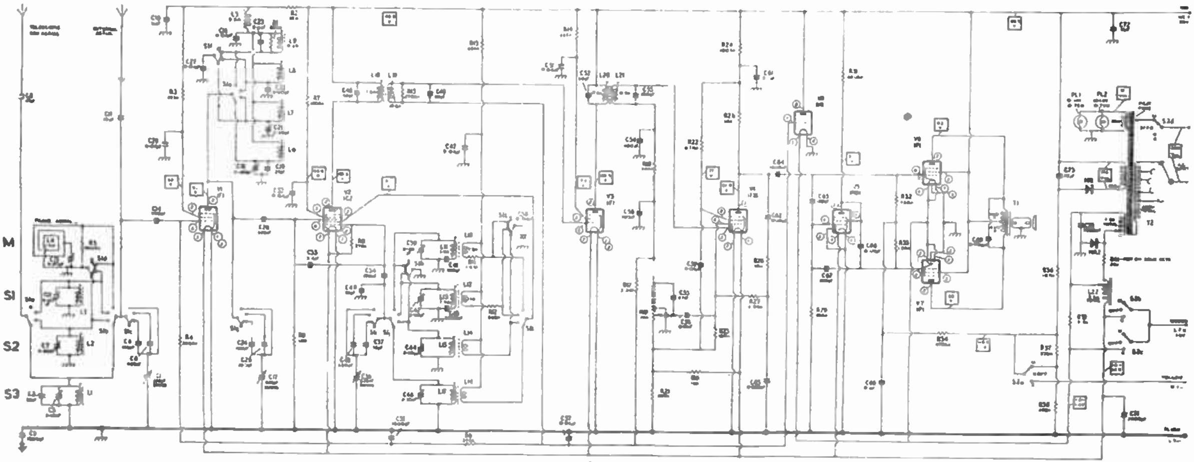


FIG. 20.—PORTABLE A.C. MAINS BATTERY RECEIVER FOR MEDIUM AND SHORT WAVES, WITH R.F. AMPLIFIER.

some or all of the preceding valves, and to operate a cathode-ray tuning indicator.

Medium-wave broadcasting stations have tended to increase their average depth of modulation, especially during the winter, in order to combat more effectively interference from foreign stations. This increase of average modulation depth has made it more than ever necessary for receiver designers to provide a high value of A.C./D.C. load ratio for the diode detector, for it can be shown [15] that the A.C./D.C. load ratio is approximately equal to the value of modulation depth at which distortion begins to be severe. The A.C./D.C. load ratio can be increased in a few ways. For example, the diode load can be tapped, so that the volume control only shunts its lower portion at A.C., or the volume control can *be* the diode load, and grid leak bias with a high value of grid leak (about 20 MΩ) can be used for the following audio-frequency amplifier. The latter method results in a flow of D.C. through the track, which can increase the possibility of noise, although the blocking capacitor necessary with this form of bias prevents grid current from flowing via the slider, thus reducing noise from this cause.

In calculating the A.C./D.C. load ratio, *no* A.C. shunting must be overlooked : in addition to the audio take-off there may be automatic gain control and tuning indicator loads. The automatic-gain-control filter resistor should be made as high as possible consistent with not exceeding the valve-makers' figures for maximum grid-circuit resistances of the automatic-gain-control-fed valves, and the tuning indicator should if possible be fed from the automatic-gain-control line instead of direct from the signal-diode load. Often automatic-gain-control volts may be obtained from a tap on the diode load.

If an extra valve can be afforded, audio-circuit A.C. shunting can be almost entirely removed by inserting a cathode follower or grounded anode amplifier between the detector and the first audio-frequency amplifier, because heavy negative feedback gives the cathode follower a very high input impedance.

Further information on diode detectors and the reduction of distortion therefrom has been given elsewhere.[16]

Automatic Gain Control

For the majority of receivers, the trend has been away from delayed automatic gain control, partly on cost grounds and also because the

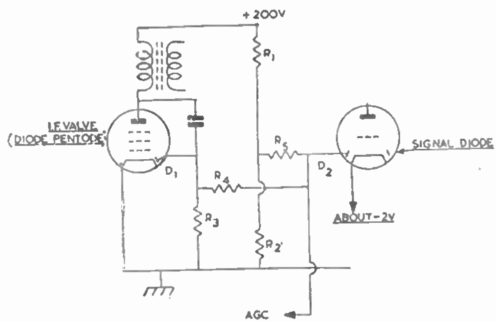


FIG. 21.—AUTOMATIC-GAIN CONTROL CIRCUIT WITH NEGLIGIBLE DIFFERENTIAL LOADING.

simple system usually employed (with delay from bias on the automatic-gain-control diode) causes distortion over a small range of signal levels, due to the diode input impedance changing as the diode conduction starts and stops with varying carrier amplitude. The effect is reduced by a high ratio of diode A.C. load resistance to tuned circuit impedance.

A better method is to use an extra "side-chain" intermediate-frequency amplifier valve for automatic gain control alone, or to employ two diodes as well as the signal diode (Fig. 21): D_2 conducts due to the positive potential applied from the divider R_1R_2 via R_5, until the signal on D_1 produces a sufficient negative voltage (applied to D_2 via R_4) to neutralize the positive potential and stop D_2 conducting. If R_4 is made at least twice R_3, the loading of D_1 on the intermediate-frequency transformer will be constant enough for the distortion due to varying load to be negligible. Until D_2 stops conducting, its anode will have a small positive potential with respect to its cathode; hence the return of the cathode to a negative point to supply a negative voltage to the grids of the controlled valves under weak signal conditions.

Sometimes a variant of this circuit is used in which the intermediate frequency valve suppressor and cathode are used instead of D_2; this method has, however, given trouble due to large variations in "diode" resistance from valve to valve.

Audio-frequency Amplifier

The design of audio-frequency amplifiers for normal domestic receivers has changed little in the past decade, the only significant trends being the increased use of negative feedback in association with the output stage and of grid leak bias.

The latter arrangement is economical because it renders unnecessary the electrolytic by-pass capacitor used to prevent loss of gain by negative feedback when a cathode self-bias resistor is used, and it assists in obtaining a good A.C./D.C. load ratio when the amplifier valve is directly capacitance-coupled to a volume control used as (or D.C.-connected to) the diode load.

The method makes use of the grid current which flows through a high grid leak (of the order of 5–20 MΩ) as a result of several causes (chiefly contact potential, initial electron velocity and gas current combined), to produce negative grid bias. The amount of static bias is determined by the intersection of the grid-leak load line and the total grid current curve, and is of the order of 0·5–1·2 volts for the usual indirectly heated valve. Distortion in the form of peak-clipping occurs due to partial rectification, but for normal outputs the overall distortion is hardly any greater than with cathode bias, provided that the source impedance is kept small enough. For example, with an EBC41 triode using a 20-MΩ leak, a source impedance not greater than 0·2 MΩ, an anode load of 0·22 MΩ, fed from a 150-volt supply, and followed by a grid resistor of 0·68 MΩ for the following valve, the total distortion is 2·5 per cent for an output of 8 volts r.m.s., the input being 0·17 volt r.m.s. For 5 per cent total distortion the output and input become 14 volts and 0·29 volt compared with 16·5 volts output and 0·38 volt input, if the optimum cathode bias be used instead of grid-leak bias.

The Output Stage

Here again the c[illegible]ef trend lies in the increased use of negative feedback (discussed lat[illegible]

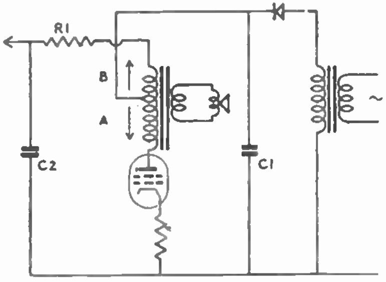


FIG. 22.—HUM CANCELLATION WINDING ON OUTPUT TRANSFORMER.

Another very popular trend for low- and medium-priced receivers is the use of a hum-cancellation winding on the output transformer. So few extra turns are needed that they can usually be added to an existing design with little or no alteration to the number of primary turns. The output transformer can then be connected direct to the reservoir capacitor (32 or 50 μF) instead of via a smoothing choke or resistor, thus saving this component and requiring less output from the mains transformer, in some cases permitting electrolytic capacitors of lower working voltage to be used (Fig. 22).

The operation is as follows: the A.C. due to the hum voltage present at the tap on the primary winding branches into the two parts A and B. By suitably choosing the tap ratio in relation to the smoothing resistor R_1 and the internal resistance (r_a say) of the valve, the magnetic fields of A and B approximately cancel, so that no appreciable current of hum frequency flows in the speech coil. Mathematically, if I_a be the A.C. through A and r_a, I_b that through B and R_1, while N_a and N_b are the turns, then $I_aN_a = I_bN_b$ (for equal ampere-turns) and $I_ar_a = I_bR_1$ (for equal potentials at ends of primary, i.e., zero potential difference across primary; we neglect the voltage across C_2)

$$\therefore \quad \frac{N_a}{r_a} = \frac{N_b}{R_1}$$

There will be some residual hum because r_a is a non-linear resistance and because of hum entering the valve via the control and screen grids.

Push–Pull Stages

Many systems of phase-reversing for push–pull operation have been used; one is the self-balancing phase reverser (Fig. 23).

The voltage feeding one output valve is stepped down to the grid of V_6 by a voltage divider R,R. V_6 produces the desired phase reversal, its anode being used to feed the other output valve. The self-balancing feaure is achieved by heavy local negative feedback via xR, so that with carefully calculated resistance values and tolerances, the output voltage of the phase reverser is substantially equal to its input, whatever the gain of V_6 itself (within wide limits).

A formula for resistor relationships may be derived as follows:

Let V be the alternating input voltage; $-V$ is then the output.

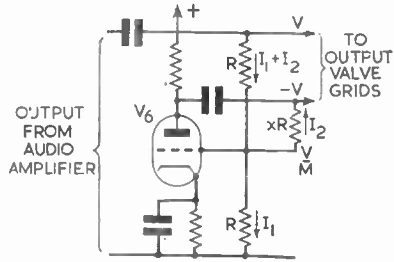


FIG. 23.—SELF-BALANCING PHASE REVERSER FOR USE WITH PUSH-PULL OUTPUT STAGES.

Taking the other currents and voltages as in Fig. 23, where M is the gain of the phase-reversing valve without feedback, we have :

$$-V = \frac{V}{M} - xRI_2 \quad . \quad . \quad . \quad . \quad (1)$$

and

$$V = RI_1 + R(I_1 + I_2)$$

$$= \frac{2V}{M} + RI_2 \quad . \quad . \quad . \quad . \quad (2)$$

since

$$\frac{V}{M} = I_1R_1$$

Eliminating $\frac{RI_2}{V}$ from (1) and (2), we have

$$x = \frac{M+1}{M-2} \quad . \quad . \quad . \quad . \quad (3)$$

It will be noticed that the higher M is, the less dependent on M is x. For other phase inverters, see references [17, 18, 19].

The push–pull output valves themselves are used most often in Class AB_1 (for classification of amplifiers, see Section 23), in order to achieve economy of power-supply components.

When still greater economy is desired, as in battery receivers, Class B_1 or "Q.P.P." is frequently used. This is really a "push–push" system, since each valve "pushes" while the other does nothing. Although fixed bias gives the best results with Class B_1, bias from the voltage drop across a resistor in the H.T. negative feed is frequently used because it avoids the need for separate bias batteries or H.T. batteries with special tap sockets, into which, it was found, users tended to put the wrong plugs.

Further information on push–pull circuits is given in references [17, 18, 19, 20].

Battery Economy Circuits

In addition to the well-known Class B_1, or "Q.P.P." circuit, a lesser known circuit for a single-ended output stage has been used : see Fig. 19. The standing bias voltage produced across R_{19} by the total H.T. current flowing through it is sufficient to bias V_4 so that it draws very little current. When an audio-frequency voltage is applied to V_4, it is necessary to reduce the bias to normal to avoid distortion and loss in output. This is done by rectifying a portion of the audio anode voltage (fed via C_{32}, stepped down by R_{17}, R_{18}) using a small metal rectifier connected so as to produce D.C. of opposite polarity to the bias across R_{19}. The resultant voltage is fed to V_4 grid via the filter circuit R_{21} and C_{34}, the grid leak R_{14} and the grid stopper R_{16}.

Power Supply

Mains Receivers

In power-supply circuits for ordinary mains broadcast receivers the only significant changes in the years under review have been as follows :

(i) The availability of small electrolytic capacitors of large capacitance has enabled the smoothing choke (which had already replaced the speaker-field winding owing to the development of the high-flux permanent-magnet speaker) to be dispensed with in most cases. Nowadays one electrolytic capacitor unit suffices for the smoothed H.T. supply of a low-priced mains receiver. The unit contains two capacitors, usually of 32 or 50 μF each, for use as reservoir and filter capacitor, with resistance smoothing often aided by an output transformer with hum cancellation. Sometimes a further stage of smoothing, using a resistor and tubular capacitor of 0·1 or 0·25 μF, is added for the first audio-amplifier valve.

(ii) The development of smaller and cheaper convection-cooled metal rectifiers has resulted in their use instead of valves in some cases. This tendency is likely to increase as more types of contact-cooled rectifier become available at competitive prices.

(iii) The development of A.C.-mains-type rectifier valves with better heater-cathode insulation has resulted in many receivers having the rectifier heater wired in parallel with the receiving-valve heaters, thus eliminating the rectifier-heater winding from the power transformer. Some valve manufacturers do not, however, approve of this technique, holding that it lowers reliability.

In a small A.C./D.C./battery portable set for 200/250-volt mains, the problem of dispersing the heat developed in dropping resistors, without unduly raising the temperature of other components and the cabinet, is a difficult one, solvable, however, by measures such as those already described under "Ventilation" (page 14–3). If the same set were altered to A.C. operation with a mains transformer, the consumption in watts would fall in a typical case from 19 to 10·5 watts, with filaments still series-fed through a resistor from the 90-volt H.T. supply, or to 7 watts, with filaments parallel-fed from a separately rectified low-voltage supply. Thus A.C. transformer operation eases or removes the heat problem.

A.C./D.C./Battery Receivers

The usual arrangement is for one rectifier to supply both H.T. and filaments, the rectified voltage being dropped to a low value for a series filament chain by a resistor and to 90 volts for the H.T. by another resistor. The filaments need individual shunt resistors in order to keep their voltages within prescribed limits (allowing for the different cathode currents), and to ensure that the limit voltage(s) of the filament-smoothing electrolytic capacitor(s) is/are not exceeded. A resistor in series with the rectifier has its value chosen so that substantially equal output voltages are obtained on D.C. and A.C. mains.

Particular care is necessary in choosing mains tappings and resistor tolerances in order to keep within the valve manufacturer's limits for filament voltage. The problem is eased if a lamp is used as a barretter, and it may be used also for scale illumination.

Detailed information on this type of power supply is given in reference [21].

A.C./Battery Receivers

There are several systems in use. In one, for example, a transformer produces a 90-volt D.C. supply, and the voltage is dropped for a series filament chain by a resistor. In another, a transformer has two secondary windings, each of which feeds a rectifier, one producing H.T. at 90 volts and the other L.T. at 1·3 volts after smoothing. The latter arrangement is preferred by valve manufacturers, because the valves are designed primarily for constant-voltage operation. An important point concerns the choice of reservoir capacitor for the filament supply, because the normal tolerances of this, $-20 + 50$ per cent or even $-20 + 100$ per cent, can cause large variations in filament voltage unless the nominal capacitance is chosen high enough. A plotted curve of reservoir capacitance against output voltage is the best way to decide. In a typical case (with a smoothed supply of 225 mA) the curve flattened out at about 3,000 μF, which meant that the reservoir's nominal value would have to be at least this, unless a lower value with special close tolerances were used.

Some degree of L.T. voltage stabilization may be achieved by using a metal rectifier connected in shunt.

Battery Receivers

The set with a 2-volt accumulator and a dry H.T. battery has nearly disappeared. This results from the increasing electrification of homes and the convenience of all-dry batteries.

The table set worked from a car-type accumulator with a vibrator and transformer—at one time in great demand for export markets—is also tending to be replaced by one operating solely from dry batteries, presumably because of the inconvenience of transporting and re-charging a car-type battery. The larger "all-dry" sets usually have push–pull output.

Negative Feedback

Seven main applications of negative feedback, numbered (1)–(7), have been listed earlier (page 14–8).

In broadcast receivers (excluding high-fidelity equipment) it is not usual to employ feedback for reduction of distortion and of hum *alone* (1); the feedback is usually made frequency selective for tailoring or levelling responses (3), perhaps in conjunction with tone control, (4) and/or compensated volume control (5).

An example of partially compensated volume control by negative feedback appears in the small A.C./D.C. mains receiver of Fig. 17, where R_{16}, C_{25} and R_{10} result in a diminution of middle and high frequencies as the volume control is turned down, giving an effective bass

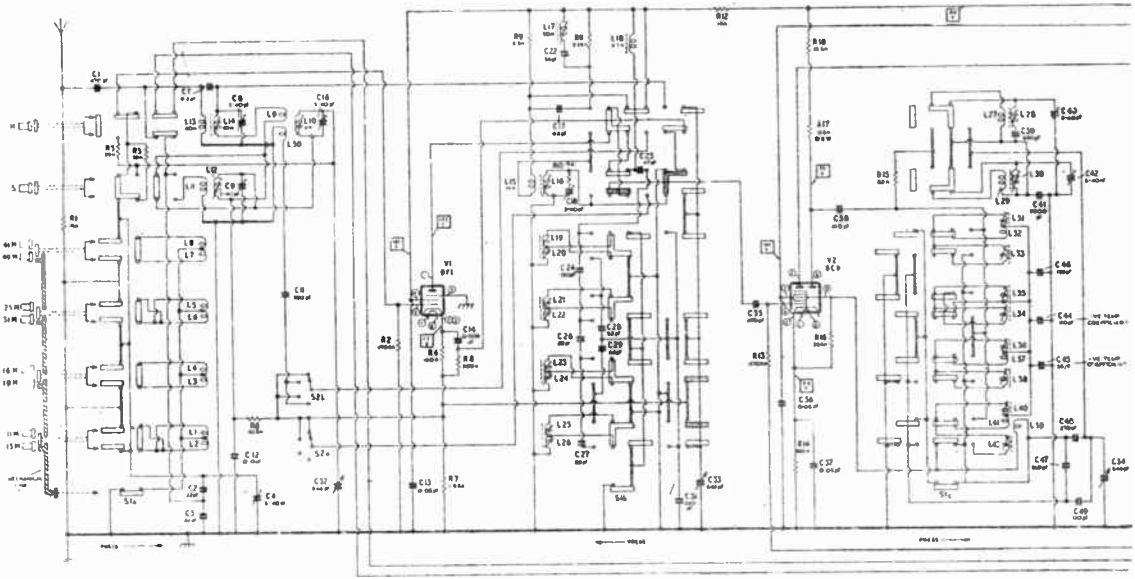


FIG. 24 (a).—EXPORT RADIOGRAMOPHONE WITH RADIO-FREQUENCY AMPLIFIER, PUSH-BUTTON WAVE-BAND SELECTION, PUSH-PULL OUTPUT STAGE AND TUNING INDICATOR.

Two pick-up heads are used with switched compensation for standard and long-playing records.

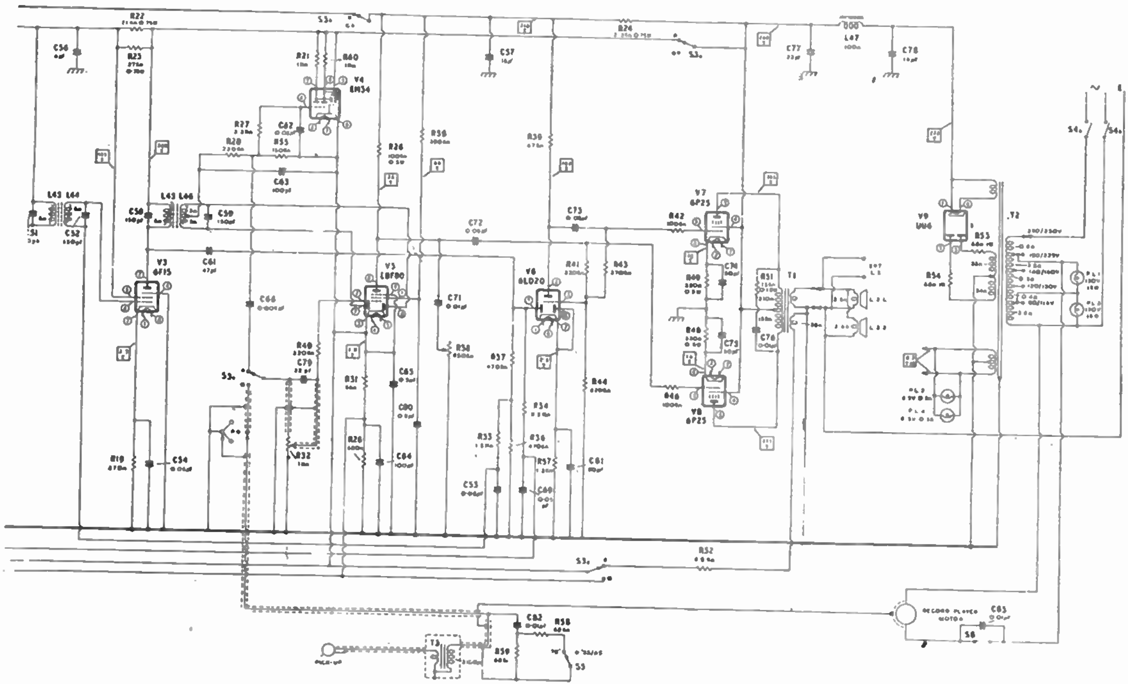


FIG. 24 (*b*).—INTERMEDIATE- AND AUDIO-FREQUENCY STAGES OF EXPORT RADIOGRAMOPHONE.

lift. The feedback almost disappears when the volume control is at maximum, so the loss of sensitivity is negligible.

An example of (6), preventing a high gain radio-frequency amplifying valve from being overloaded by strong signals, is given in Fig. 13.

Negative feedback needs care in use to avoid phase changes great enough to produce positive feedback, and hence, if the loop gain be enough, oscillation. Mathematically, if the loop gain and phase-angle shift be plotted on polar co-ordinates, the amplifier will be stable if the locus does not include the point 1,0. An approximate rule, covering most cases and avoiding the necessity for plotting the locus, is: if the loop gain be less than unity when the phase angle is +180° and when it is −180°, the amplifier is stable. See also pages 1–51, 10–26 and references [22] and [23]. Component tolerances, warming-up conditions, mains-voltage changes and speaker-load variations must all be taken into account. Oscilloscope checks are advisable for showing supersonic oscillation, sometimes only over one narrow portion of the signal cycle, over a narrow band of audio frequencies.

Tone Controls

There are four basic types of tone control, each of which may be used alone or in combination with one or more of the others:

(1) adjustable treble rise (or high frequency intensification);
(2) adjustable treble cut (or high frequency attenuation);
(3) adjustable bass rise (or low frequency intensification);
(4) adjustable bass cut (or low frequency attenuation).

Treble rise may be obtained without further loss of gain, if negative feedback is in use, by reducing the feedback at high frequencies. The gain loss can be made virtually negligible by arranging for the feedback (and with it the treble rise) to diminish as the volume control is turned up. The treble rise may be adjusted by using a switch or variable resistance to restore the high-frequency feedback in varying degrees.

Treble cut may be obtained without the use of feedback by shunting a capacitor across a portion of the audio circuit. It may be adjusted (1) by switching different capacitors, (2) by using a variable capacitor or (3) by using a variable resistance in series with one capacitor. For interference reduction the variable-resistance method is the least suitable, since—for any resistance value other than zero—a horizontal "shelf" is formed in the response curve; this is because the resistance sets a limit to the attenuation at frequencies above those at which the capacitor's reactance is negligible compared with the resistance.

Treble cut may be obtained with negative feedback by reducing the feedback at middle and low frequencies, using a switch or variable capacitor or resistor to vary the effect.

Bass rise may be obtained without further loss of gain, if negative feedback is in use, by reducing the feedback at low frequencies. The loss can be made virtually negligible by arranging for the feedback (and with it the bass rise) to diminish as the volume control is turned up. The bass rise may be adjusted by using a switch or variable resistance to restore the low-frequency feedback in varying degrees.

Bass cut may be obtained without the use of feedback by reducing the size of the coupling capacitor between two valves. It may be adjusted by switching different capacitors, or by shunting a variable

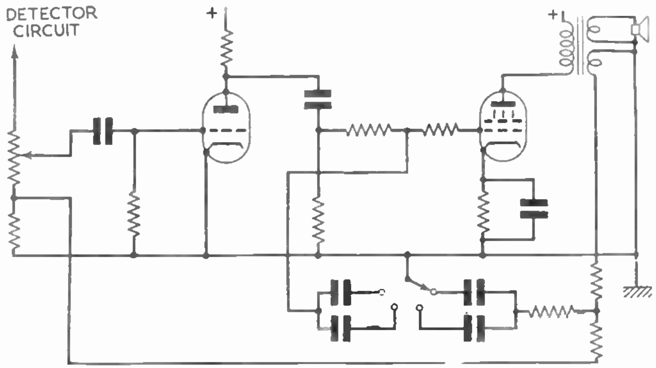


Fig. 25.—Tone Control Circuit Providing Two Positions of Treble Rise and Two of Treble Cut Using a Four-way One-pole Switch (Irrelevant Parts Omitted).

resistor across the capacitor. The shunted variable-resistor method has the disadvantage that the lowest frequencies are not attenuated as much as may be desired at intermediate control settings, whereas low-to-middle frequencies may be attenuated too much. Here again, this is because—for any resistance value other than zero—a horizontal " shelf " is formed in the response curve.

Bass cut may be obtained with negative feedback by reducing it at middle and high frequencies, using a switch or a variable resistor to vary the effect.

Figs. 25 and 26 give two examples of tone-control circuitry. Fig. 25 uses negative feedback to give two positions of treble rise, and capacitance shunting to give two positions of treble cut. The effectiveness of the treble cut diminishes as the volume control is turned down, because here the feedback is at its maximum and tends to level the frequency response at high frequencies. If this effect is not desired the capacitance shunting may be transferred to a pre-volume control position, care being taken to reduce worsening of the diode A.C./D.C. load ratio by the use of a series-isolating resistor.

In Fig. 26 the potentiometer gives treble rise in one direction (by reducing the feedback at high frequencies), and treble cut in the other (by increasing the effect of a capacitance shunt). The rheostat adjusts the amount of bass rise.

An example of very wide range bass and treble controls, both providing rise and cut, and employing an extra valve section to make up for the resulting loss of gain, is shown in Fig. 39. The connection of C64 to the tap on the treble control reduces considerably the " shelf " effect mentioned above and increases the ability of the instrument to reduce interference such as needle scratch without too much sacrifice of the desired high frequencies.

Pick-up Connections and Radiogramophones

Switching.—It is usual to provide pick-up sockets on all mains receivers except the smallest. Usually they are switched across the

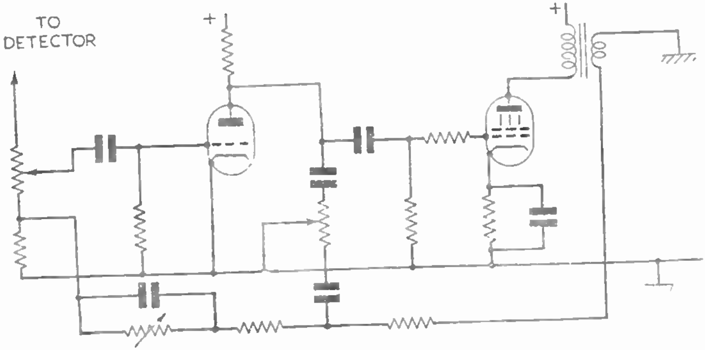


Fig. 26.—Tone Control Circuit Providing Continuously Variable Bass and Treble by Adjustable Feedback Networks.

volume control by a radio gramophone switch or by contacts on the wave-change switch, but switching is sometimes omitted in the cheaper receivers: in this case the user has to remove the " live " pick-up plug when wishing to listen to the radio, and to tune away from stations when the gramophone side is desired. In addition, the diode which remains connected may cause distortion.

Even when a changeover switch disconnects the detector audio output from the volume control for gramophone use, strong stations may break through unless further switching is carried out, such as disconnecting the H.T. supply to the frequency changer.

Audio Gain.—The normal receiver with a high-gain triode as the only audio amplifier preceding the output valve has insufficient gain for some types of pick-up if the correct compensation for recording characteristics—including long-playing types—is to be provided. Gain is usually insufficient with moving-coil pick-ups and some moving-iron types. Crystal types, on the other hand, provide sufficient output.

To cover the less-sensitive pick-ups, extra gain can be obtained in several ways. For example, a pentode can replace the triode, approximately doubling the gain, or a pre-amplifier valve can be used. This may be either an extra valve, or an existing valve in the set may be given a second function when the gramophone switch is operated. The triode section of the frequency-changer, or the intermediate-frequency amplifier valve, is occasionally used for this purpose.

Non-isolated Chassis.—Three pick-up sockets, each with an isolating capacitor, are necessary if hum is to be minimized. If only two sockets—as in Fig. 27—were used, part of the hum voltage between the cathode line (at mains potential) and earth would appear across one isolating capacitor in series with the grid-cathode circuit of the valve. The lower isolating capacitor cannot be made sufficiently large to reduce this hum without exceeding the limit prescribed under the revised Safety Specification B.S. 415. If, however, three sockets are used—Fig. 28—it will be seen that no part of the hum voltage appears in the grid-cathode circuit. In practice, there will still be a little hum because

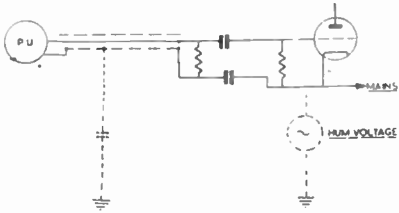


FIG. 27 (*left*).—TWO-SOCKET PICK-UP CONNECTION FOR NON-ISOLATED CHASSIS—UNSUITABLE OWING TO HUM VOLTAGE.

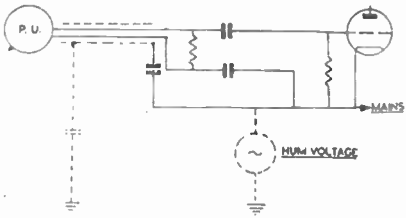


FIG. 28 (*right*).—THREE-SOCKET PICK-UP CONNECTION —HUM VOLTAGE MUCH REDUCED.

the simplified circuit ignores several stray capacitances, such as those between the conductors inside the screened lead and the screen.

The resistor across the pick-up sockets is necessary for crystal pick-ups, since a high electrostatic potential difference can be built up across such a pick-up, causing distortion unless its insulation resistance be fairly low or a conducting path be provided.

Compensation for Recording Characteristics

The necessary compensating components vary considerably according to the type of pick-up and the receiver audio response. The problem also arises as to how different compensating networks can be connected when a standard head (for records with the present British 78-r.p.m. recording characteristic) is changed for a long-playing head.

Where cost is a secondary consideration a switch can be mechanically coupled to the speed-change lever on the record player or changer and this switch can, of course, contain sufficient contacts to perform all network changes necessary for near-perfect compensation for both heads.

A cheaper method, when the now common three-pin plug-in heads are used, is to utilize a cross-connection in one head and if necessary a different cross-connection in the other, to modify a compensation network connected at the other end of the three-way lead. Alternatively, or in conjunction with this method, a small resistor may be fixed in the coupling adaptor between one pick-up and arm, with either no resistor or a different value in the adaptor for the other pick-up.

Progress in crystal pick-up design has resulted in turn-over models with electro-mechanical compensation such that satisfactory results are now possible both on standard and long-playing records without any network *switching*, and with either no electrical compensation at all or a very simple network.

Further information on compensation is given in Section 34.

V.H.F. BROADCASTING RECEIVERS

A recent development in Britain has been the commencement of a regular V.H.F. broadcasting service to overcome the deficiencies in the normal service, caused by interference from foreign stations in the medium and long wavebands. After prolonged trials by the B.B.C. based on experimental transmissions from Wrotham using both amplitude modulation and frequency modulation the Television Advisory Committee recommended frequency modulation. The latter system has been in use in the U.S.A. for a considerable time and in Germany and a number of other countries during the post-war period.

The advantages of a V.H.F. service compared with the present medium- and long-wave service are :

(i) Reception is less troubled by interference from undesired stations.

(ii) The spacing of V.H.F. stations can be such that the adjacent-channel selectivity requirements of the receiver are met without the side-band cutting now severely experienced in medium- and long-wave receivers coping with the standard 9-kc/s spacing. This means that a good high-frequency audio response is possible.

The main reasons why the Television Advisory Committee recommended frequency modulation for the V.H.F. service are :

(i) The transmitter cost is much less than for amplitude modulation.

(ii) The difference in receiver costs appears to be extremely small.

(iii) The problem of accommodating all the stations required for a national coverage of three programmes in the available 88–100 Mc/s band appears to be more easily solved by a frequency modulation service.

Frequency-modulation Receivers

Since the combined amplitude-modulation/frequency-modulation receiver is undoubtedly the most popular type, the following brief description applies to a typical receiver of this category.

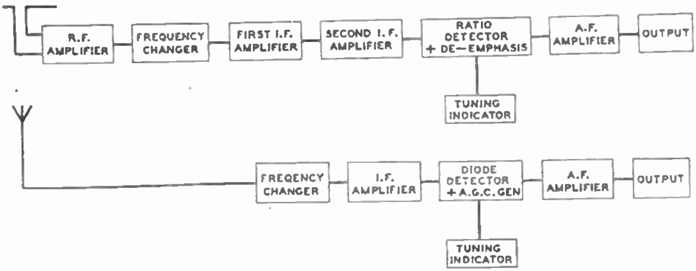


FIG. 29.—SCHEMATIC CIRCUIT DIAGRAM OF F.M./A.M. RECEIVER.

In practice, most valves will be used for both F.M. and A.M. reception. Intermediate frequency F.M. $\simeq$ 10·7 Mc/s, A.M. $\simeq$ 470 kc/s.

Fig. 29 is a schematic diagram of the frequency-modulation and amplitude-modulation circuits.

Within this framework, several different valve arrangements are possible :

Example (1)

FREQUENCY-MODULATION RECEPTION

(i) The radio-frequency amplifier is a separate pentode or grounded-grid triode (not used on amplitude modulation).

(ii) The frequency changer is a separate triode, as self-oscillating additive mixer, with the signal voltage fed to a bridge-balanced null point on the oscillator coupling coil, in order to reduce oscillator radiation, and positive feedback to remove the damping of the triode on the intermediate-frequency transformer (not used on amplitude modulation).

(iii) The first intermediate-frequency amplifier valve is the heptode section of a triode-heptode, the triode section not being used on frequency modulation (both sections used on amplitude modulation).

(iv) The second intermediate frequency amplifier valve is a variable-mu pentode which may have a higher slope than the type hitherto used for amplitude-modulation sets (also used on amplitude modulation).

(v) The ratio detector valve is one bulb containing two well-matched low-resistance diodes (not used on amplitude modulation). It is followed by a de-emphasis circuit (frequency modulation only).

(vi) The tuning indicator is a normal electron-beam type (also used on amplitude modulation).

(vii) and (viii) The audio-frequency amplifier and output valves follow normal amplitude modulation practice (both also used on amplitude modulation as also are detector and automatic-gain-control diodes in the same bulb as the amplifying triode).

AMPLITUDE-MODULATION RECEPTION

This follows normal practice.

Example (2)

The radio-frequency amplifier and frequency changer are triodes as in Example (1), but are contained within the same bulb, which incorporates a screen between the two sections.

Example (3)

The ratio-detector diodes and one diode for amplitude-modulation detection and automatic gain control are contained within the same bulb as the triode audio-frequency amplifier, making the valve a triple-diode-triode.

DESIGN OF COMBINED V.H.F./A.M. RECEIVERS

The remainder of this sub-section is devoted to the design of combined amplitude-modulation/frequency-modulation receivers and of adaptors for use with existing amplitude-modulation receivers.

The amplitude-modulation portion will not be described further except where it links with the frequency-modulation part. Throughout, the term "frequency-modulation" (abbreviated F.M.) will be taken to mean frequency-modulation in the very high frequency band, and "amplitude modulation" (abbreviated A.M.) to mean amplitude modulation in the low-, medium- and high-frequency bands.

The Aerial

A suitable type of aerial is the half-wave dipole described in Section 21. For connection to the receiver, 70–80-ohm balanced twin feeder is suitable; normally this will be unscreened, but in areas where electrical interference (usually ignition) is very bad, markedly greater freedom from such interference will be obtained by using screened balanced twin feeder.

In areas of signal strength greater than about 1 mV/metre it will often be found that a good internal cabinet aerial will suffice. It is convenient to make such an aerial of two strips of paper-backed foil about 2 in. wide, glued to the cabinet roof and sides, or to the cabinet back. Unless the cabinet is very large, the 62 in. or so needed to make a half-wave dipole cannot be accommodated, and it will be necessary to use a loading coil at the centre to tune the aerial to the centre of the frequency band. It should be noted that Band II will be ultimately 87·5–100 Mc/s; the B.B.C. are using only 88·1–94·5 Mc/s for as long as the 95–100-Mc/s portion is occupied by other services.

The loading coil can conveniently be made from 18 S.W.G. tinned copper wire, a typical coil having six turns of internal diameter $\frac{5}{16}$ in. It may be wound on a former or be self-supporting, depending on the vulnerability of its mounting position.

The inductance should be adjusted until checking with a grid-dip meter (see Section 38) indicates resonance at the band-centre frequency. A radio-frequency bridge is used to measure the impedance at the ends of the coil in modulus and phase angle; if the phase angle is not quite zero an adjustment to the inductance should be made. A half-wave-length piece of cable is a useful device for linking the coil and the bridge without changing the impedance.

Tapping points (equidistant from the centre and 70–80 ohms apart in impedance) are then calculated approximately, by assuming that the turns ratio is equal to the square root of the impedance ratio. The taps are checked by the radio-frequency bridge, being shifted if necessary. If the tapping points occur at geometrically inconvenient positions on the coil, it may be necessary to wind a new coil of the same inductance as the first, but having a different diameter and, therefore, a different number of turns.

Such a foil aerial for a medium-sized table receiver (see Table II, page 14–7) may be expected to deliver into the aerial-coupling coil of a receiver, a signal voltage approximately 50 per cent of that delivered by an unloaded, full-sized dipole situated in the same field.

Sensitivity Requirements

To estimate the voltage available at the receiver terminals in order to get some idea of the required sensitivity of the receiver, use must be made of the data published by the B.B.C. on the various transmitters. Field-strength maps show an area surrounding the transmitter which lies within a contour of average field strength 1 mV/metre; this is referred to as the *first-class service area*. Between this contour and another of average field strength 250 μV/metre is the *second-class service area* in which some trouble with ignition interference may be experienced. It is important to note that the published contours are of mean field strengths; there will be local pockets of considerably smaller or larger field strength scattered about the area.

In addition, the contours apply only to a receiving aerial at a height of 30 ft., the field strength falling at lower heights approximately as the height ratio. Buildings can further attenuate the signal, even by 30 db in the worst cases. However, by using either an outside aerial or an upstairs aerial or a roof aerial, it can be assumed that a field strength of at least 25 μV/metre is obtained, at the edge of the second-class service area. Neglecting feeder loss, this means that the voltage at the aerial terminals of the set will be $\frac{25\lambda}{2\pi}$ μV (see Section 21, page 19) or approximately 12·5 μV. This is equivalent to a generator e.m.f. of 25 μV in series with the feeder's characteristic impedance.

It is usual to express receiver sensitivity as the "open-circuit voltage" or signal-generator e.m.f. required to produce standard output (usually 50 mW). The actual receiver terminal voltage, if the input impedance matches the feeder correctly, is half of this e.m.f.; this latter figure is not generally used. Signal-generator output calibrations usually refer to e.m.f. or open-circuit voltage, but the practice is not universal, so if in doubt the maker's instructions should be consulted. For further information see reference [29].

To ascertain what sensitivity is required, consider a typical receiver with a maximum useful output of 2·7 watts. The open-circuit voltage of 25 μV must produce this output when the signal is modulated, say, 90 per cent. Therefore, output for 30 per cent modulation = $\frac{2\cdot 7 \times 1000}{(90/30)^2}$ = 300 mW, and "open-circuit voltage" input for 50 mW output = $\frac{25}{\sqrt{300/6}}$ = 10 μV approx.

This gives an idea of the *minimum* sensitivity required.* Both absolute sensitivity and noise-limited sensitivity should be of this order or better, and preferably the receiver should also be providing a useful amount of A.M. rejection at this level, in order to minimize ignition interference and also to make negligible the distortion which occurs sometimes with F.M. due to multipath reception causing virtual A.M. This subject is referred to later in connection with the detector stage, and see also reference [29], which describes a suggested test procedure, for definitions of the measurement terms used above.

The Radio-frequency Amplifier

It is possible to design and make a receiver which gives quite a good performance, from the point of view of its user, without a radio-frequency amplifier. It is very doubtful, however, whether the oscillator radiation from such a receiver could be made sufficiently low to avoid interference with other V.H.F. receivers (fundamental frequency) and with Band III television receivers (second harmonic). To minimize the latter it is usual to employ an oscillator frequency lower than the signal frequency.

* To meet the popular demand for satisfactory operation on internal cabinet aerials at appreciable distances from the transmitter, a sensitivity of the order of 1 μVm. has become desirable.

Interference limits which have been suggested and which at present constitute the targets aimed at by many designers, are: radiation field 3 m. away, fundamental 300 μV/m. and second harmonic 500 μV/m.; terminal voltage on 75 ohms, fundamental 750 μV and second harmonic 1,750 μV; voltage between either main and earth, 500 μV. (If the second harmonic falls inside Band III, the limits become 200 μV/m. and 1,000 μV.)

To keep oscillator radiation down to reasonable limits then, it is usual to preface the frequency changer with a radio-frequency amplifier, and, in addition, so to couple the two stages that oscillator volts fed back to the aerial circuit are largely cancelled out. A radio-frequency amplifier is also desirable because it assists in obtaining a good signal-to-noise ratio, and gives some protection against image interference and intermediate-frequency breakthrough.

Choice of Valve

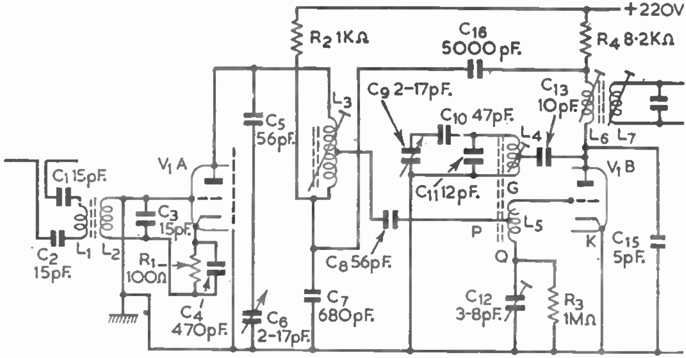
Examples of suitable radio-frequency amplifier valves are: pentode, grounded grid triode and cascode (see Section 15, page 17). The latter consists of a grounded-grid triode preceded by a grounded-cathode triode, and is probably the best arrangement if cost has not to be considered, giving the high gain of a pentode without its partition noise. However, the most popular arrangement is the grounded-grid triode, mainly because by using a double triode for radio-frequency amplifier and self-oscillating additive mixer, a very economical circuit results; in addition, a better signal-to-noise ratio is possible with a triode than with a pentode.

In the grounded-grid triode amplifier the grid acts as a screen between the output electrode (anode) and the input electrode (cathode), so that neutralization is unnecessary. The low input impedance (approximately $\frac{1}{g_m} \cdot \left(1 + \frac{R_L}{r_a}\right)$ where g_m is the mutual conductance, R_L is the anode load resistance and r_a is the anode A.C. resistance) of about 150 ohms is no great disadvantage when fed from a 70–80-ohm feeder by means of a suitable matching transformer, the secondary of which is fix-tuned to the centre of the band. Fig. 30 shows a typical radio-frequency amplifier and mixer circuit.

Input Transformer

The matching can either be for minimum noise or for maximum power transfer, or for a compromise between the two. In practice, it will usually be found that the signal-to-noise ratio is good enough, if a satisfactory power match is achieved. The coupling should be made as tight as possible; in Fig. 30, capacitors C1 and C2 tune out the residual leakage inductance in order to ensure a resistive input impedance for matching the feeder. They should thus be adjusted for minimum voltage across the aerial terminals, while the coupling turns are adjusted so that the voltage across the terminals is the same as that across a 70–80-ohm resistor temporarily replacing the receiver. This gives maximum power transfer and terminates the feeder correctly.

A gain about twice that obtainable with an orthodox grounded-grid amplifier is possible if the tuned secondary be grounded at a tap instead



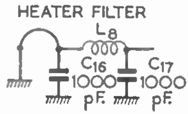


FIG. 30 (*above*).—V.H.F. R.F. AMPLIFIER AND MIXER, USING DOUBLE TRIODE WITH GANGED CAPACITOR TUNING.

L8 is self-resonant at 104 Mc/s.

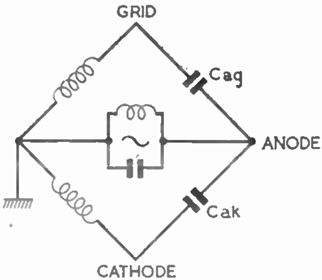


FIG. 31 (*right*).—SIMPLIFIED BRIDGE CIRCUIT TO EXPLAIN NEUTRALIZATION OF TRIODE R.F. AMPLIFIER.

The two coils shown together constitute the secondary of the aerial coil, which is tapped to chassis.

of at the grid. In this arrangement the valve is working partly as a conventional grid-driven amplifier. Neutralization is then necessary to avoid instability. However, if the tap is placed at the correct point, neutralization is achieved, as shown by the simplified bridge of Fig. 31; when the bridge is balanced none of the anode/ground voltage appears between grid and cathode.

A centre-tap on the primary winding may be connected to the A.M. aerial terminal so that the F.M. aerial (internal or external) will function also on A.M. reception.

Since a high Q is not necessary for the fix-tuned matching transformer, this can be wound of fairly thin wire, e.g., 24 S.W.G., on a small, wire-ended, iron-dust core.

Permeability and Capacitor Tuners

It is possible to use either permeability or ganged-capacitor tuning for the radio-frequency amplifier and mixer. The latter system is used in the receiver of Fig. 39; it is convenient for a combined A.M./F.M. set because ganged capacitors are available having suitable V.H.F. sections (10–17 pF cover) in addition to the usual sections (of the order of 500 pF swing) for short, medium and long waves. The vanes of a good V.H.F. section should be very well spaced and preferably thicker than those used in variable capacitors for domestic A.M. sets.

The permeability tuner has the advantage of compactness and also of freedom from the unwanted coupling, which can take place in the spindle of a ganged capacitor unless the more expensive split-spindle type be employed. Both these advantages make it easier to keep oscillator radiation within safe limits with permeability tuning.

The circuit of Fig. 30 uses a ganged-capacitor tuner; Fig. 32 shows a modified circuit using a permeability tuner.

Amplifier Anode Circuit

The tuned-anode circuit is usually tapped down to the oscillator circuit in order to reduce the loading of the oscillator on the tuned circuit. The best tap position can be found by experiment; as a guide 0·6 from the bottom end is a typical value.

The Mixer

The Tuned Circuit

In Fig. 30 the oscillator tuned circuit is connected to the anode by C13, which is also part of the tuning capacitance of the primary of the first 10·7-Mc/s intermediate-frequency transformer. It should be noted that the impedance of the oscillator circuit, in series with C13 for intermediate-frequency tuning, is very low at 10·7 Mc/s. The anode coil may be tapped down to the anode-coupling capacitor in order to reduce the damping of the valve on the oscillator-tuned circuit, and to assist in keeping down oscillator radiation. A tap about 25 per cent from the top end is a suitable point.

Reduction of Oscillator Radiation

The coupling coil in the grid circuit of Fig. 30 is tapped at its electrical centre to receive the signal voltage; this centre-tap, together with C12 and $C_{g\text{-}k}$, results in a considerable reduction of oscillator radiation voltage. This is apparent when the bridge of Fig. 33 is considered. In practice, the balance is imperfect because the simplified circuit omits stray couplings and resistances, e.g., the valve-input resistance. C12 is made variable so that the bridge may be balanced for minimum oscillator voltage at the aerial terminals.

Other measures which assist in reducing the unwanted oscillator voltage are: the tapping of the oscillator anode into the tuned circuit; the capacitance potential divider C13 : (C15 + $C_{a\text{-}k}$) (these two reduce the voltage at the mixer anode, and hence that at the amplifier anode, due to the capacitance between these electrodes); careful screening between signal and mixer circuits and of the whole mixer; avoidance of chassis currents at oscillator frequency; reduction of oscillator volts fed back from the amplifier anode to its cathode (the neutralizing bridge of Fig. 31 can assist in this); and the provision of a filter in the heater circuit of V1 (C16, L8 and C17 in Fig. 30).

Chassis currents can be detected by using a small search coil at the end of the measuring set feeder, or by using the measuring set to

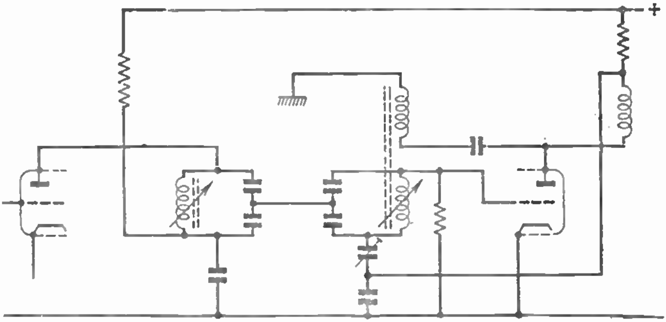


FIG. 32 (*above*).—R.F. AMPLIFIER AND MIXER, USING DOUBLE TRIODE AND PERMEABILITY TUNER.

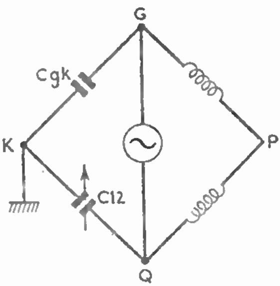


FIG. 33 (*right*).—SIMPLIFIED BRIDGE CIRCUIT TO EXPLAIN REDUCTION OF OSCILLATOR VOLTAGE AT POINT P (V1B OF FIG. 30).

investigate the radio-frequency potential difference between points on the chassis. One example of the causes of unwanted chassis currents could be : C15 earthed too far from L4 and the cathode of V1B (C15 and L4 should be earthed at the same point, which should be close to the C9 earthing brush and also to the earthing of V1B cathode). Another example might be the cathode of V1A earthed in the path of the chassis current between points in the oscillator circuit, e.g., between L4 and V1B cathode.

Improving Intermediate-frequency Gain and Anti-squegg Precautions

The grid leak R3 of the oscillator may be made high (e.g., 1 MΩ), as this assists in the reduction of the noise level. Such a high value would normally result in squegging, but this can be avoided by feedback as described below.

The reduction in sensitivity which would normally result because of feedback by $C_{g\text{-}a}$ at intermediate-frequency is neutralized by the application of positive feedback via C16, across C7. The effective internal resistance of the triode at intermediate-frequency is thereby raised from about 9,000 ohms to about 60,000 ohms. Care must be taken not to make C7 too small, or instability will result. Although the C16–C7 circuit produces positive feedback at intermediate-frequency, it produces negative feedback at the comparatively low frequency at which squegging would normally occur; this is because at such low frequencies the reactance of the intermediate-frequency transformer primary is negligible, so that the anode load is mainly resistive.

The conversion gain from the aerial terminals to the secondary of the first 10·7-Mc/s intermediate-frequency transformer which can be achieved with the semi-grounded-grid circuit is of the order of 500. Thus with 0·5 μV input across the terminals, the intermediate-frequency voltage at the heptode grid is about 250 μV.

Coil Construction

The amplifier tuned-anode coil and the oscillator coil are conveniently wound on low-loss formers (e.g., polystyrene) with 6-mm. iron-dust cores, fine-threaded for ease of adjustment. The tuned windings may be 18 or 19 S.W.G. copper wire, or copper foil about $\frac{3}{32}$ in. wide and 0·005 in. thick. The best coating is silver, lacquered against corrosion in impure atmospheres; but quite good results have been obtained with coatings of enamel or tin. The oscillator-coupling winding, centre-tapped, can be of fairly thin enamelled copper wire, e.g., 30 S.W.G. The tuned windings should have about $\frac{1}{8}$ in. spacing between turns; the coupling coil is close-wound, preferably using bifilar winding.

Oscillator Drift

It is very important to keep oscillator drift down to a very small amount, 20 kc/s being about the maximum permissible. This can be achieved by, first, arranging the layout for the least possible heating of oscillator circuit components, and secondly, by making the fixed capacitor in the oscillator-tuned circuit (C11 in Fig. 30) a type with a suitable negative temperature coefficient. For further information on this subject see pages 14–27, 15–14, 29–12.

Layout and Wiring Precautions

Very short radio-frequency leads and very careful positioning of components are necessary in amplifier and mixer circuits operating at 100 Mc/s, partly in order to secure low oscillator radiation as mentioned earlier, but also to avoid undesirable feedback. It should be remembered that a straight 1-in. length of 23 S.W.G. wire has a reactance of about 16 ohms at 100 Mc/s.

The Intermediate-frequency Amplifier

Before designing the intermediate-frequency amplifier, it is essential to have some idea of the gain required. The ratio discriminator chosen for the F.M. detector requires for limiting at least 1 volt at its transformer secondary, and this value also produces about 50 mW output when used with a typical triode audio-frequency amplifier and a pentode output valve, with a signal modulated 30 per cent. Working from a signal of about 250 μV at the heptode grid, a gain of about 4000 is required between the heptode grid and the ratio detector transformer secondary.

This can just about be achieved by a two-valve intermediate-frequency amplifier, consisting of the heptode portion of the triode-heptode used for mixing in the A.M. portion of the set, and the variable slope intermediate-frequency amplifier valve used for the normal intermediate-frequency amplifier, the latter usually being one of the more recent types having a maximum slope of about 4 or 6 mA/volt. For economy, such a circuit is often used, and it will be described first, followed by a description of how it may be improved, especially from the points of view of ignition-interference suppression and avoidance of distortion due to multipath reception, by the addition of an extra valve.

Economical Amplifier

To obtain sufficient gain with the economical two-valve arrangement, rather small tuning capacitors have to be used in the intermediate-frequency transformers, of the order of 15 pF or less.

It is not necessary to use A.G.C. on F.M., although the ratio detector does supply a voltage which may be used for this purpose. A.G.C. can be troublesome because variations in the valve-input capacitance with a change in bias can cause detuning, resulting in distortion and reduced noise rejection. Although this effect can be reduced to negligible proportions by negative feedback from an un-by-passed cathode resistor, this method causes appreciable gain reduction and is difficult to apply to the heptode valve. To date the advantages and disadvantages of A.G.C. are so evenly balanced that current practice has not yet moved decisively in favour of or against it.

Simple series connection of the F.M. (10·7 Mc/s) and A.M. (470 kc/s) intermediate-frequency transformers is usually satisfactory : the exception is the primaries immediately following the heptode. To avoid undesirable effects due, for example, to an oscillator harmonic at 10·7 Mc/s being transmitted along the intermediate-frequency chain on short or medium waves, it is advisable to switch the heptode anode to the appropriate primary. Sometimes "shorting-out" of the 10·7-Mc/s intermediate-frequency windings is adopted to avoid such troubles; and also "shorting-out" of the 470-kc/s windings, because the presence of these circuits in series with the 10·7-Mc/s circuits may upset F.M. performance.

Neutralization.—Feedback via the anode/grid capacitance of the pentode intermediate-frequency amplifier can sometimes cause distortion of the response curve. Neutralization is then advisable. A typical circuit is shown in Fig. 34 (*a*), with the equivalent bridge circuit depicted in Fig. 34 (*b*). It will be noticed that the anode H.T. decoupling capacitor is taken to screen instead of to chassis. Incidentally, this circuit also illustrates an alternative to simple series connection of the two sets of intermediate-frequency transformer primaries.

However, if the response-curve distortion is slight, it may be remedied by wobbulator alignment adjustments, without recourse to special circuits.

Band-width.—The band-width of the intermediate-frequency amplifier should be as narrow as possible consistent with a reasonable allowance for drift, and an adequate band-width for the wanted signal : this is because the peak value of the noise signal passed by the amplifier is proportional to the square root of the band-width. Band-width limitation may also prove necessary later in order to obtain sufficient adjacent-channel selectivity, but this is unlikely in view of the way in which the transmitters are spaced in frequency and in geographical location. The band-width for points 3 db down should thus be of the order of 200 kc/s plus a suitable drift allowance, resulting in a total of about 230 kc/s.

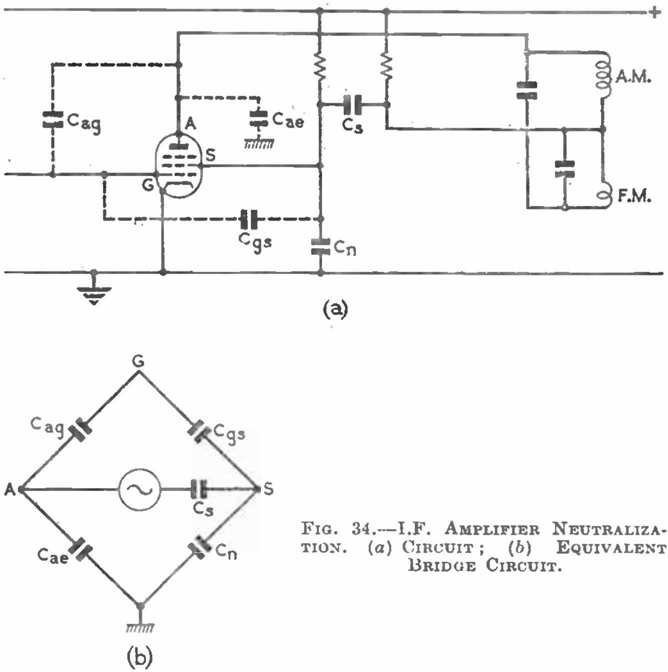


FIG. 34.—I.F. AMPLIFIER NEUTRALIZATION. (a) CIRCUIT; (b) EQUIVALENT BRIDGE CIRCUIT.

(The maximum deviation of the B.B.C. transmission is 75 kc/s; for the relationship between the band-width required, the deviation and the maximum modulation frequency, see Section 6, page 30).

Coupling Factor.—For maximum gain and for the most linear phase characteristic (necessary for minimum audio-frequency distortion), critical coupling (kQ equal to 1) is desirable, but it is satisfactory to make the kQ factor about 1·2 (i.e., to over-couple slightly) to obtain the desired flat response in the pass-band.

Coil Construction.—Solenoid windings on moulded formers, fine-threaded internally for 6- or 8-mm. dust-cores, are usually employed, the windings being of enamelled copper wire, of approximately 30 S.W.G.

The 470-kc/s and 10·7-Mc/s coils may be in separate screening cans or in the same can. In the latter case a tall " chimney " type of can may be used, or a rectangular can with the coils mounted side by side. Side-by-side mounting is probably the best arrangement, both for economy and good mechanical design, and is at least as good electrically as other types, although it requires more chassis area.

Drift.—Drift will generally be low enough, if polystyrene formers and silvered-mica or plastic-film tuning capacitors are used.

Improved Amplifier Using Extra Valve

By using an extra pentode in front of the ratio discriminator, several advantages accrue. The gain of each stage does not need to be so high, so that larger capacitors (of the order of 30–40 pF) can be used for tuning the primary and secondary inductors of the intermediate-frequency transformers. There is then unlikely to be any trouble, such as distortion of the response curve due to feedback, or shift of tuning due to the change of valve-input capacitance with grid-current bias under strong signals or with A.G.C. if used.

Limiting.—In addition, the extra valve can be made to act as a limiter for stronger signals, while still behaving as a high-gain amplifier for weaker signals. The amplitude-rejection properties of a ratio discriminator are not perfect—as will be shown later—so that pre-limiting is an advantage.

For this system of partial limiting the pentode, a high-variable-slope type, is not run with low anode and screen voltages as in the saturated-amplifier type of limiter frequently employed before the Foster–Seeley phase discriminator (see Section 8), but differs from an ordinary amplifier merely in having a resistor shunted by a capacitor connected in series in the grid circuit. Typical values for the capacitor and resistor are 0·1 MΩ and 100 pF; these providing a time constant of 10 μS. Too small a time constant would not allow the A.M. voltages to develop enough bias to cut down the gain of the valve sufficiently, while too large a time constant would not cope with the short rapid pulses of ignition interference.

The F.M. Detector

Reasons for Choosing the Ratio Detector

The demodulator almost universally employed in domestic A.M./F.M. receivers is the ratio detector. Before describing this circuit it is useful briefly to examine the reasons for its popularity. There is not space to consider all known types, but three will be mentioned here: the Foster–Seeley phase detector (see Section 8, page 23), the gated-beam valve, and the nonode. The Foster–Seeley detector provides no self-limiting, and therefore requires an extra valve for limiting, whereas the ratio detector, the gated-beam valve and the nonode are self-limiting devices. The Foster–Seeley detector is therefore rarely used in domestic radio receivers, although its use may be justified in the more expensive sets because it produces less distortion (up to about 1 per cent, compared with about 3 per cent for the ratio detector, both figures being at 75 kc/s deviation).

The *gated-beam valve* [30] has two grids operating in a quadrature phase relationship at centre frequency, and the phase difference varies about this 90° value as the frequency varies. The quadrature voltage is developed by space-charge coupling to a special grid to which an extra parallel-tuned circuit is connected. Limiting is by control of the electron-beam current by a slot aperture. The sensitivity and A.M. rejection properties of this valve are of the same order as those obtained with a ratio detector plus a normal audio-frequency triode. However, the ratio detector provides a voltage which may be used for A.G.C. and/or a tuning indicator; and since, in a F.M./A.M. set, a triode is almost always needed for audio-frequency amplification (and usually

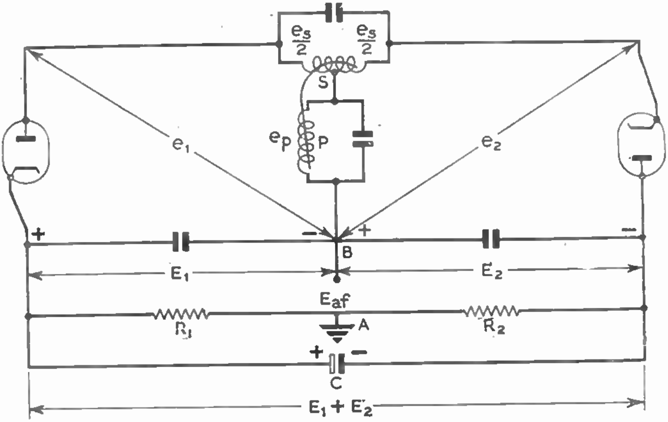


FIG. 35.—BASIC CIRCUIT OF RATIO DETECTOR.

e_s = secondary voltage; e_p = primary voltage; e_1, e_2 = frequency-dependent voltages; E_1, E_2 = rectified voltages produced by e_1, e_2; E_{af} = audio output voltage between points A and B.

also for gramophone facilities) the ratio detector again scores. In addition, the gated-beam valve is expensive to make.

The *nonode* (Mullard EQ80,[31]) has seven grids, of which the second, fourth and sixth are screen grids and the seventh a suppressor grid. To each of the control grids (g_3 and g_5) an intermediate-frequency voltage of at least 8 volts r.m.s. is applied, each derived from a separate tuned intermediate-frequency transformer winding. The grids are so arranged that anode current flows only during those time intervals when g_3 and g_5 are simultaneously positive, and the current then has a value dependent only on the potentials of g_1 and g_2, which are kept constant. The mean value of the rectangular anode current pulses depends on the phase difference between the g_3 and g_5 voltages, which in turn depends on the frequency deviation. It does not depend on their amplitude; hence the limiting action. The audio-frequency output is more than enough to feed an output valve, being of the order of 22 volts for 100 per cent modulation, but the nonode has the major disadvantage compared with the ratio detector of requiring a minimum input of 8 volts r.m.s. as opposed to about 1 volt r.m.s. for a ratio detector. Points in its favour are that the first grid may be used in a circuit for inter-station noise suppression, and that the A.M. rejection properties are rather superior to those of the ratio detector. However, for the same reasons as those given for the gated-beam valve, it does not provide such an economical arrangement as the ratio detector when used in a combined A.M./F.M. receiver.

The Ratio Detector

Only a comparatively brief description of this ingenious device can be given here. A fuller account of the theory of operation is given in reference [28]. Like the Foster–Seeley phase discriminator, the ratio detector depends for its operation on the phase difference which exists between the primary and secondary voltages of a loosely-coupled intermediate-frequency transformer. At the centre-frequency, i.e., the frequency of zero deviation to which both primary and secondary circuits are tuned, there is a 90° phase difference. For negative deviations the phase difference is greater than 90°, and for positive deviations less than 90° (or vice-versa with reversed connections).

Also, as in the Foster–Seeley detector, two diodes are required, but one is reversed, and the load resistors R_1, R_2 are shunted by an electrolytic capacitor C, the combination having a time-constant of the order of 0·2 second.

It will be seen from Fig. 35, which is a circuit simplified to explain the basic theory, that the voltages E_1 and E_2 are obtained by rectifying e_1 and e_2, each of which is the vector sum of the primary voltage e_p and a half-secondary voltage $e_s/2$. The vector diagrams of Fig. 36 show e_1 and e_2 for zero, negative and positive deviations.

Suppose that e_p is amplitude modulated at an audio frequency to a percentage 100 M, i.e., e_p becomes $e_p\ (1 + M)$. Then e_s, and with it e_1, e_2 and E_1, E_2 will also be modulated to the same percentage (or so we will assume for the moment; we shall find later that the action of the ratio detector actually prevents this).

Let $E_1 = nE_2$

Since $\dfrac{E_1\,(1 + M)}{E_2\,(1 + M)} = \dfrac{E_1}{E_2} = n$, we see that this ratio is independent of the amplitude modulation.

Now $E_1 + E_2$ is constant for a given mean carrier level, despite the amplitude modulation, because the large time constant $C(R_1 + R_2)$ ensures this. Let $E_1 + E_2 = V$, and let $R_1 = R_2$

Then $V = E_1 + E_2 = E_2\,(n + 1)$

$\therefore \quad E_2 = \dfrac{V}{n+1}$ (which is therefore not amplitude modulated)

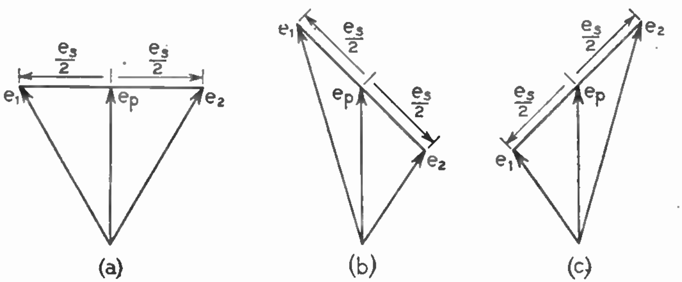


FIG. 36.—VECTOR DIAGRAMS FOR RATIO DETECTOR.

(a) Zero deviation (on resonance); (b) positive deviation (above resonance); (c) negative deviation (below resonance).

The voltage between the points A and B

$$= V_A - V_B$$
$$= \tfrac{1}{2}V - E_2$$
$$= V\left(\tfrac{1}{2} - \frac{1}{n+1}\right)$$

It depends therefore only on the *ratio* of the voltages E_1 and E_2 and is independent of their common amplitude modulation, as long as the period of this is small compared with the time constant $C(R_1 + R_2)$. Hence the name " ratio detector ".

Thus A and B are the points from which the audio-frequency output voltage E_{af} is taken. V, and with it E_{af}, will, of course, vary with *slow* variations in e_p. E_{af} also $= \tfrac{1}{2}(E_1 + E_2) - E_2 = \tfrac{1}{2}(E_1 - E_2)$. (This compares with $E_1 - E_2$ for the Foster–Seeley detector.)

The graph of E_{af} against frequency deviation is similar to that given in Fig. 31 of Section 8 (page 8–24). To ensure linearity over a wide enough frequency band, the peak separation is usually made about 350 kc/s. The great ingenuity of the device lies in its ability to hold $(E_1 + E_2)$ constant while permitting $(E_1 - E_2)$ to vary in accordance with the frequency modulation.

In practice, measures which are not apparent from the above simplified theory have to be taken to achieve good results. In particular, use has to be made of the following: imagine a sudden increase of carrier amplitude; the resultant increase of diode current will result in a virtual increase of damping because $E_1 + E_2$ is held constant, and the effective load resistance (voltage/current) therefore falls. This increased loading tends to counteract the increase in voltage by increasing the damping on the tuned circuits. For a sudden *decrease* in carrier amplitude, the corresponding reduction of damping is still more important, because were it not so the diodes would cease to conduct, being subject to a net negative bias (the charge voltage of the capacitor C minus the new applied voltage), thus causing distortion. It is therefore important that the circuits are normally heavily damped by the diode loading, so that when the applied voltage momentarily becomes less than that to which C is charged, the increase in diode resistance as the bottom bends of the characteristics are reached will reduce the damping and permit the applied voltage to rise again. Thus the diodes will not be completely cut off.

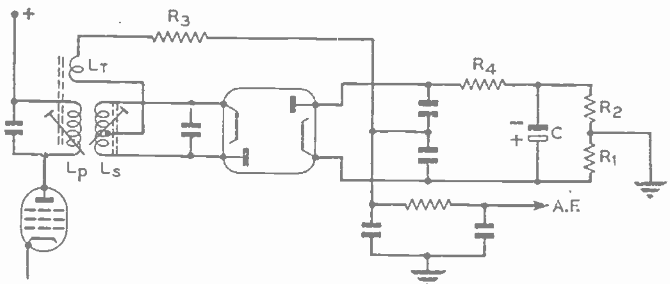


Fig. 37.—Practical Ratio Detector. Balanced Circuit.

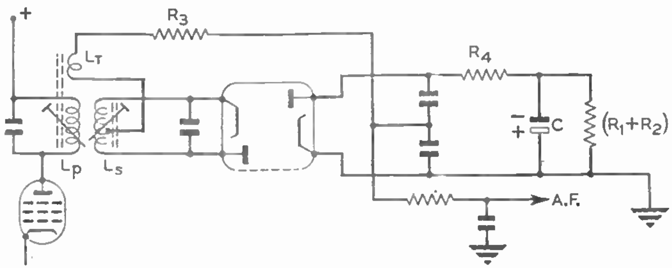


FIG. 38.—PRACTICAL RATIO DETECTOR, UNBALANCED CIRCUIT.

Again, results in practice are not so good as simple theory indicates, and it is found that the maximum amount of downward amplitude modulation that can be handled by a ratio detector without diode cut-off causing distortion is of the order of 75 per cent. As multipath reception (see reference [26], page 275) can result in much greater percentages of downward amplitude modulation, additional limiting before the ratio detector is an advantage.

Low-resistance, well-matched diodes are essential. Typical values are about 200 ohms each, with a maximum ratio of 1·5 : 1.

Ratio Detector : Practical Circuits

Fig. 37 shows a practical circuit of a ratio detector. It will be seen that a tertiary winding L_T is used to provide a virtual primary coil dissociated from H.T., and with the turns chosen for impedance matching (the working Q of the secondary is low because of the necessary heavy damping). This tertiary winding is closely coupled to the primary. The unloaded Q of the secondary is as high as possible, so that when the damping is reduced during downward A.M. the applied voltage has ample opportunity to rise. It is advisable to make the ratio (unloaded Q : loaded Q) at least 4 : 1. R_3 and R_4 assist in balancing out unwanted A.M. ; the best results will be obtained if at least one of them is a pre-set variable, because only thus can valve and coil tolerances be allowed for, The absence of a corresponding resistor to R_4 in the "lower limb" is due to the asymmetry brought about by stray capacitances.

L_t is bifilarly wound so that its two halves are of equal inductance and coupled equally to the primary, whatever the position of its inductance-adjusting iron-dust core. A grooved former is advisable so that the bifilar winding may be accurately spaced and stable ; this assists in keeping low the drift with temperature change. 6-mm. dust cores are suitable for primary and secondary inductance adjustment.

Typical circuit values [28] are :

Primary, unloaded, Q = 70 ; primary, with diode loading, Q = 40.

Secondary, unloaded, Q = 89 ; secondary, with diode loading Q = 21.

Secondary/primary coupling = 50 per cent of critical.

Tertiary turns = approximately one-quarter of total secondary turns, wound on top of "cold" end of primary for tight coupling.

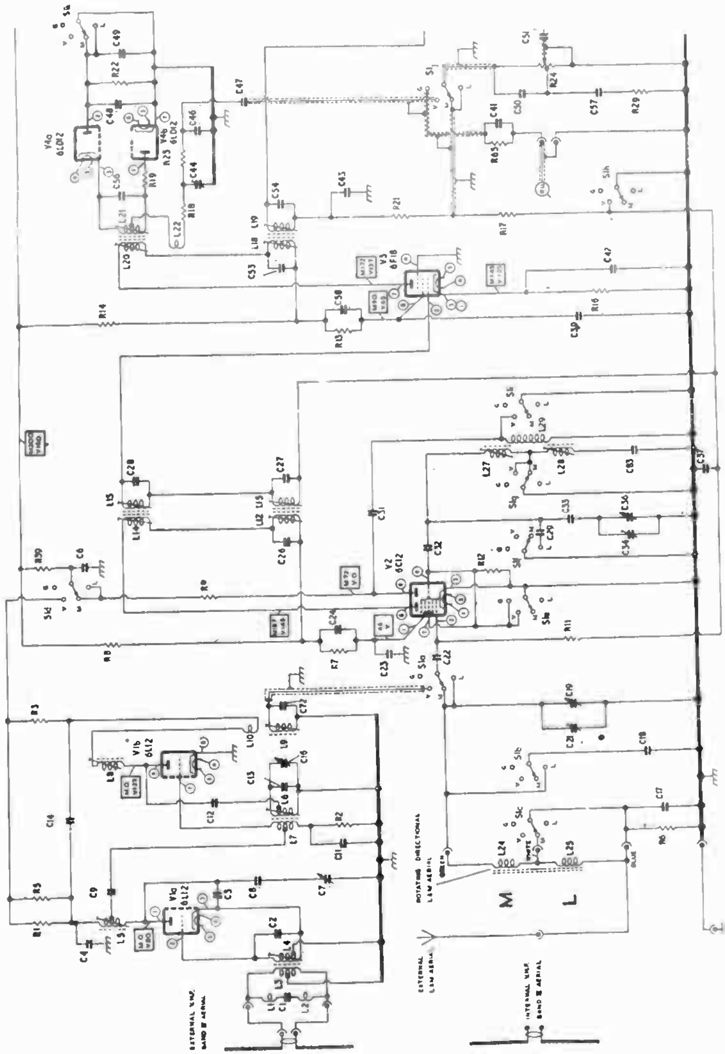
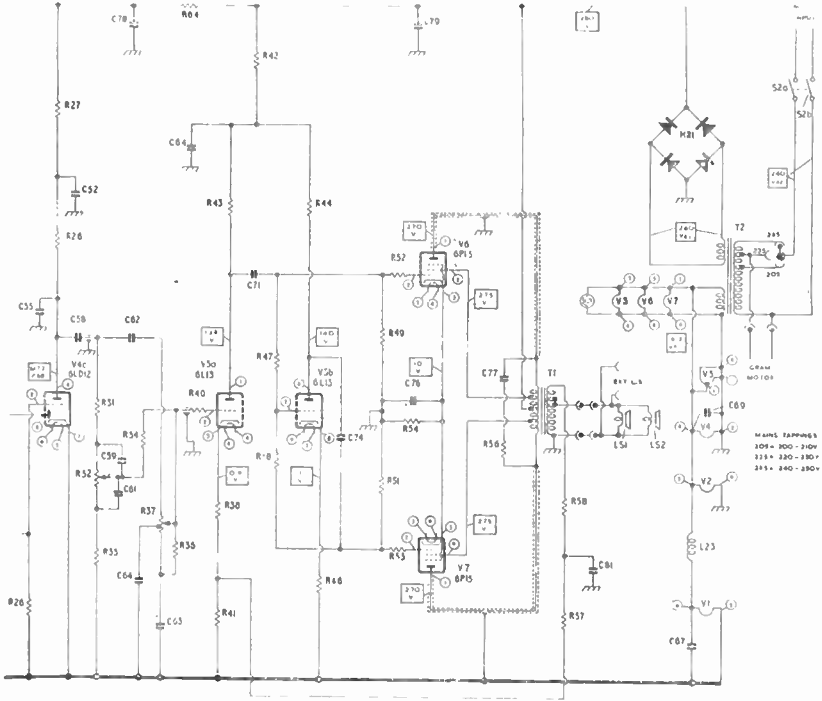

V4a 6LD12
V4b 6LD12
V3 6F18
V2 6C12
V1a 6L12
V1b 6L12
M
L
ROTATING DIRECTIONAL L&M AERIAL
EXTERNAL L&M AERIAL
EXTERNAL V.H.F. BAND II AERIAL
INTERNAL V.H.F. BAND II AERIAL

FIG. 39. — REPRESENTATIVE MODERN V.H.F. F.M./A.M. RADIOGRAMOPHONE.

Resistors

R1	18 kΩ	R48	1 MΩ
R2	220 kΩ	R49	680 kΩ
R3	4·7 kΩ	R51	680 kΩ
R5	22 kΩ	R52	10 kΩ
R6	22 kΩ	R53	10 kΩ
R7	47 kΩ	R54	150 Ω
R8	2·2 kΩ	R56	15 kΩ
R9	33 kΩ	R57	3·3 kΩ
R11	470 kΩ	R58	3·3 kΩ
R12	47 kΩ	R63	1·5 MΩ
R13	39 kΩ	R64	4·7 kΩ
R14	2·2 kΩ	C1	22 pF
R16	150 Ω	C2	10 pF
R17	2·2 MΩ	C3	2·7 pF
R18	47 Ω	C4	470 pF
R19	39 Ω	C6	4700 pF
R21	100 kΩ	C7	10·9 pF sw...
R22	33 kΩ	C8	68 pF
R23	39 kΩ	C9	82 pF
R24	500 kΩ	C11	10 pF
R26	10 MΩ	C12	15 pF
R27	100 kΩ	C13	15 pF
R28	68 kΩ	C14	4700 pF
R29	15 kΩ	C16	10·9 pF swing
R31	68 kΩ	C17	3900 pF
R32	250 kΩ	C18	15 pF
R33	6·8 kΩ	C19	510 pF swing
R34	39 kΩ	C21	3–30 pF
R36	47 kΩ	C22	4700 pF
R37	250 kΩ	C23	0·005
R38	1·8 kΩ	C24	4700 pF
R39	1 kΩ	C26	220 pF
R40	100 kΩ	C27	220 pF
R41	220 Ω	C28	15 pF
R42	47 kΩ	C29	47 pF
R43	220 kΩ	C31	1800 pF
R44	220 kΩ	C32	100 pF
R46	2·2 kΩ	C33	425 pF
R47	1 MΩ	C34	4–40 pF

(continued on page 72)



C36	510 pF swing	C48	4700 pF	C57	0·04	C71	0·01 μF
C37	0·04	C49	4 μF	C58	0·01	C72	10 pF
C38	0·01	C50	100 pF	C59	0·002 μF	C74	0·01 μF
C39	0·005	C51	0·01	C61	0·02 μF	C76	50 μF
C41	47 pF	C52	0·1	C62	330 pF	C77	0·005 μF
C42	0·04	C53	220 pF	C63	0·005 μF	C78	50 μF
C43	100 pF	C54	390 pF	C64	0·002 μF	C81	0·02 μF
C44	50–250 pF	C55	220 pF	C67	0·01 μF	C83	230 pF
C46	0·001	C56	56 pF	C69	0·01 μF	C84	16 μF
C47	0·01						

Fig. 38 shows an alternative practical circuit, suitable for use with a triple-diode-triode having one "hot" cathode for connection to L_s, and one earthy cathode common to the triode, the A.M. diode and one F.M. diode. (Fig. 37 is not suited to this, and would use either a double diode or two germanium diodes.) It is only at very low-frequency unwanted A.M. that Fig. 37 shows its superiority over Fig. 38; at such frequencies, where some variation of voltage across C may occur, the balanced circuit keeps such variations from the audio-frequency output, but the unbalanced circuit permits them to appear there.

There are many minor variations on these circuits.

The circuit diagram of a complete receiver, Fig. 39, shows one such variation and provides typical component values.

Precautions.—To prevent drift with changing temperature, stable capacitors are necessary as well as stable coils. These may well be of the silvered-mica or plastic-film type.

The ninth harmonic of the intermediate frequency may be troublesome unless the ratio-detector components are well screened and/or lead lengths are kept very short; a heater-filter capacitor of about 1,800 pF is usually necessary for the detector valve.

Tuning Indicator.—The D.C. voltage developed across the stabilizing capacitor C is sometimes used for A.G.C. and may also be used for operating a tuning indicator. A better tuning indicator would be one worked from the D.C. at the audio-frequency take-off point (Fig. 37 only), since this is zero for the true on-tune condition.

De-emphasis.—In both Fig. 37 and Fig. 38, it will be noticed that the audio-frequency output passes through an R–C filter. This is a de-emphasis network inserted to restore the audio-frequency balance, because the upper audio-frequencies are emphasized by an R–C network of time constant 50 μS at the transmitter. The overall effect is to improve the signal-to-noise ratio. It is not necessary to make the de-emphasis time constant exactly 50 μS; the value may be chosen to give the desired overall response when taken in connection with the rest of the audio-frequency circuits and the loudspeaker. The pre-emphasis provides, in fact, a useful reserve of upper audio-frequencies which may to a certain extent be used to compensate for unavoidable losses in the last stages of the receiver.

Audio-frequency Amplifier, Output Stage and Loudspeaker(s)

In the less expensive sets, these follow A.M. receiver practice, but in higher-priced receivers it is usual to make improvements so as to take advantage of the better high-frequency response available with F.M. due to the avoidance of side-band cutting which takes place in the necessarily highly selective A.M. receivers. These improvements include the use of loudspeakers of smoother and extended response, or combinations of loudspeakers with frequency-dividing networks (see page 32–12), better output transformers, and measures to reduce distortion (e.g., higher power-output valves, negative feedback).

Switching from A.M. to F.M. Reception

A study of Fig. 39 will show how this is done in a typical receiver. Points to note are :

(i) There is no switching at V.H.F., the heptode grid being switched to the A.M. signal circuits or the F.M. first intermediate-frequency transformer.

(ii) If the set had a short-wave band, the first 10·7-Mc/s I.F. transformer primary would have to be switched out or short-circuited on A.M. to avoid passing oscillator harmonics along the I.F. amplifier and the generation of A.G.C. voltages by these. This effect is much smaller at medium wavelengths and is made negligible in the receiver of Fig. 39 by the short-circuiting of C49.

(iii) The short-circuiting of V2 triode grid to cathode on F.M. to avoid gain reduction due to grid current bias, since this grid is connected to the third grid of the heptode.

Adaptors

The design of an F.M. adaptor to feed into the pick-up sockets of an A.M. receiver can follow on the lines of the complete receiver already described, with the following exceptions :

(i) There is no A.M./F.M. switching.

(ii) Almost certainly a permeability tuner is a better choice than a ganged capacitor tuner, because the advantages of a combined capacitor are lost (see page 14–60).

(iii) The ratio detector valve can be a double diode or a pair of germanium diodes; an audio-frequency amplifier is unnecessary, since receivers with pick-up sockets always incorporate one.

Test Procedure

A useful suggested procedure is given in detail in reference [29].

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The author's thanks are due to Messrs. Murphy Radio Ltd. for permission to publish many of their diagrams; and to several colleagues for their kind assistance.

L. D.

TRANSISTOR PORTABLE RECEIVERS

Transistorized portable receivers and record players appeared on the U.K. market in 1956, and since then have formed an increasing percentage of battery-operated equipment. The advantages of low power consumption, low-voltage operation, small size, long life and robust construction are particularly attractive for small portable or "personal" receivers. The reduction in battery consumption can be substantial: a typical transistorized portable receiver may have a standing current of about 8 mA from a 9-volt battery, rising to 55 mA at 300 mW output and about 25 mA average for music. This compares with the 125 mA at 1·4 volts and 10 mA at 90 volts for the average four-valve portable receiver using low-consumption valves. The running cost of a transistorized receiver with a Class B output stage is of the order of 0·15 of a penny per hour. At present, however, the initial cost of a transistorized receiver is appreciably above that of an equivalent valve receiver.

The equipment so far marketed falls into four main groups:

(1) Portable receivers having a maximum output of about 200–500 mW and with a sensitivity and performance comparable with the standard four-valve battery receiver. Such a receiver will usually have six transistors plus a germanium crystal diode detector, and the output stage will consist of two transistors operating in Class B, in either a symmetrical push–pull circuit or in the so-called "single-ended" push–pull arrangement.

(2) The midget "personal" receiver using four or more transistors plus one or more crystal diodes, with a Class A single-output transistor providing a maximum output of the order of 20 mW, and sometimes including a headphone jack for use in noisy environments. Several models of this type have used reflex circuits permitting a single transistor to be used as both an I.F. amplifier and an A.F. driver stage.

(3) Small "hybrid" portable receivers using valves in the early stages with a push–pull transistor output stage. Such receivers require an H.T. supply, which can be derived from a battery or from a ringing-choke D.C. transistor converter, but overall consumption is considerably less than for conventional valve receivers, in which the output stage accounts for 40 per cent of L.T. and over 50 per cent of H.T. consumption.

(4) Small portable battery-operated record reproducers, comprising basically a 2–300-mW amplifier using three or four transistors and a low-consumption battery-operated gramophone motor; consumption of the motor is usually about 80 mA at 5 volts.

Transistor-circuit practice can be best illustrated by consideration of a receiver of type (1). The complete receiver commonly has an additive "self-oscillating" frequency changer; two stages of I.F. amplification; a crystal diode detector which also provides an A.G.C. potential; an A.F. driver; and a push–pull Class B output stage.

Input Circuits

Ferrite-rod aerials with a matching winding provide a convenient means of obtaining an impedance transformation to match the low input resistance of the frequency changer. Since the only connection from the rod aerial at high impedance is that to the wave-change switch or, in the case of single-band models, to the gang tuning capacitor, the effective stray capacitances across the tuned circuit can be kept lower than with valve circuits; because of this the capacitance tuning swing of the tuning capacitor and associated trimmer can be much lower: the tuning swing may be as low as 110 pF (compared with about 520 pF for valve circuits), though values between 170 and 350 pF are more common with two-waveband models.

The Frequency Changer

Although some models have used separate transistors for mixing and as a local oscillator, the more popular arrangement is a single transistor as a self-oscillating additive mixer, which is more economical and generally has better frequency stability. Fig. 40 shows typical circuits,

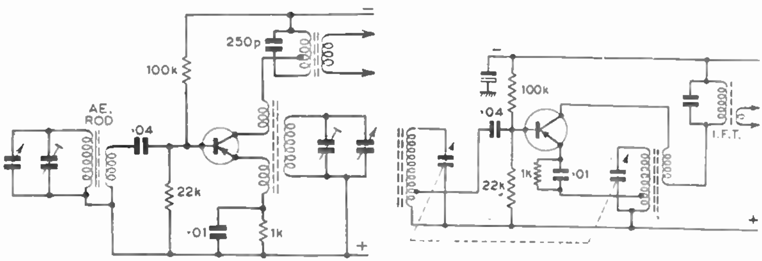


FIG. 40.—TYPICAL TRANSISTOR FREQUENCY-CHANGER CIRCUITS.

simplified by the omission of wave-change details. This comprises a common-emitter mixer combined with a common-base oscillator in which positive feedback is obtained from the collector circuit; either a low-impedance tap or coupling winding is connected in the emitter circuit. The tuned-circuit inductance should be of high Q construction to allow for transistor damping and to maintain oscillation with the low voltages likely to be encountered towards the end of the useful life of the battery. To increase Q a ferrite-pot core may be used.

To provide satisfactory tracking without padding capacitors, the oscillator section of the tuning gang may have specially shaped vanes. Typical values are aerial tuning 175 pF swing; and oscillator tuning 120 pF swing.

To provide ready oscillation when the receiver is first switched on, the oscillator transistor may be initially biased for Class A operation; subsequently, rectification of the oscillator voltage at the emitter provides additional bias, driving the transistor into Class B operation. This stabilizes the amplitude of oscillation and also tends to reduce the

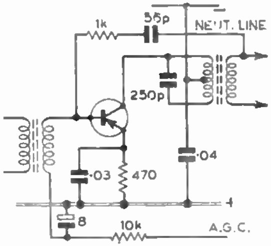


FIG. 41.—TYPICAL I.F. STAGE.

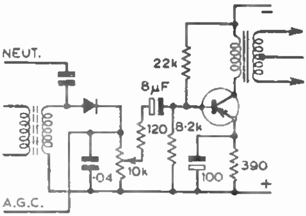


FIG. 42.—TYPICAL DETECTOR AND A.F. DRIVER.

falling off in the performance of the stage towards the higher limit of the frequency range of the transistor.

Coupling between the aerial and oscillator circuits must be kept to a low value to prevent spurious oscillation; for this reason a screening shield may be required between the two sections of the tuning gang.

Intermediate-Frequency Stages

The gain which may be achieved per stage, without risk of instability, with current types of transistors is of the order of 30 dB, and it is customary to include two I.F. stages; an alternative arrangement found in some smaller "personal" receivers is one I.F. stage with controlled regeneration.

Because of base-collector capacitance, which corresponds roughly to the grid–anode capacitance of a triode valve, it is necessary to include components to neutralize the effect of positive feedback via this capacitance. Neutralization can be achieved by balancing out this feedback by introducing external feedback paths of opposite effect. Where this path is purely capacitive, it may subsequently be necessary to change the value of the neutralizing capacitor when the transistor is replaced. However, it is common practice to introduce resistance into the feedback circuit to give an unilateralized stage, and by careful design of the coupling circuits the spread of base-collector capacitances can be accommodated, making it unnecessary to adjust values should it be necessary to change transistors.

Detector and A.G.C.

A germanium diode is almost invariably used for detection, with the D.C. voltage developed across the load resistance fed back to the first I.F. stage to provide A.G.C. For *pnp* transistors, the A.G.C. line will be positive in respect of chassis (the greater the positive voltage the less will be the stage gain). Apart from controlling the gain, the A.G.C. line may be used to provide a small forward bias for the detector diode in order that the quiescent operating point is located near the "bend" of its characteristic; operation in this condition will improve efficiency on weak signals and also helps keep the input resistance of the detector

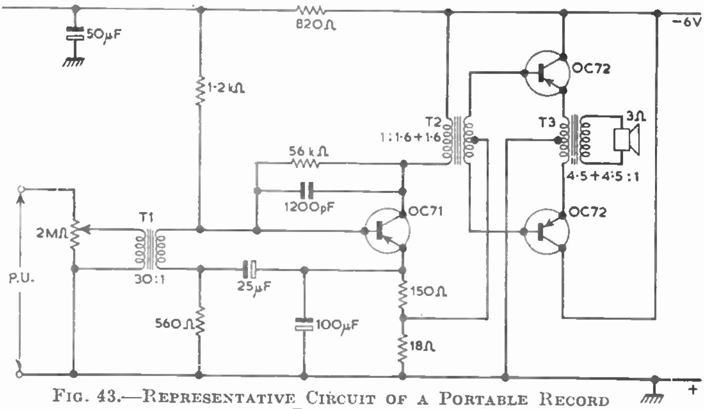


FIG. 43.—REPRESENTATIVE CIRCUIT OF A PORTABLE RECORD REPRODUCER.

reasonably constant for varying signals, so that it presents a fixed loading to the second I.F. stage.

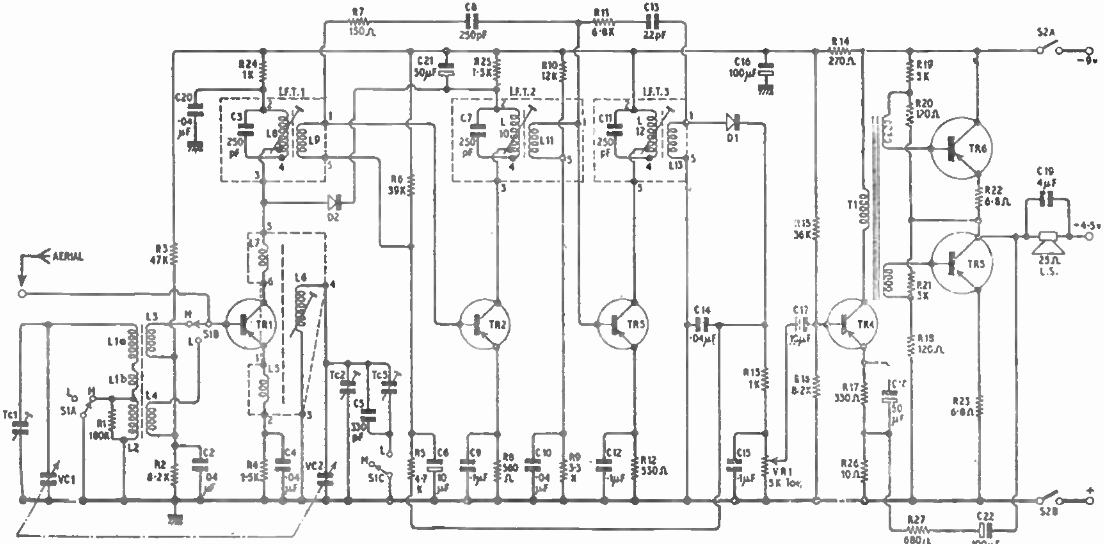
Since a relatively high circulating current may exist in the diode circuit, the position of the diode by-pass capacitor and the associated wiring needs careful consideration in order to minimize radiation of the harmonic frequencies present in this circuit; excessive harmonic radiation is likely to produce spurious whistles and instability, particularly at the second harmonic (about $2 \times 470 = 940$ kc/s).

The simplest A.G.C. systems may provide insufficient range of control, and various circuits incorporating additional diodes have been used to increase the range. One method is to use a second germanium diode to damp the first I.F. transformer: see Fig. 44.

Owing to the dual detector/A.G.C. function, it is important, when connecting or replacing this diode, to observe polarity.

A.F. Driver

For the "full specification" receiver, an A.F. driver stage is invariably included; in smaller receivers a reflex circuit, which enables an I.F. stage to be used also for A.F. amplification, may often be found. The A.F. driver is generally a common-emitter amplifier operating in Class A with bias stabilization. A high value miniature electrolytic capacitor and resistor is used to decouple the "H.T. line" so as to prevent distortion as the internal resistance of the battery increases with age. Negative feedback, derived from the output stage, is commonly applied via the emitter circuit of the driver stage. In order to stop thermal run-away, a resistor, with associated by-pass capacitor, is usually included in the emitter circuit; it should be noted that although in a circuit diagram these components appear to resemble the cathode bias components of valve circuits, their purpose differs. Five per cent tolerance bias stabilization resistors are usually fitted.



(*Ultra Electric Ltd.*)

Fig. 44.—Representative Circuit of Transistor Portable Receiver TR1 XA102; TR2 XA101; TR3 XA101; TR4 XB103; TR5, 6 XC101 D1 CG12E; D2 CG6E.

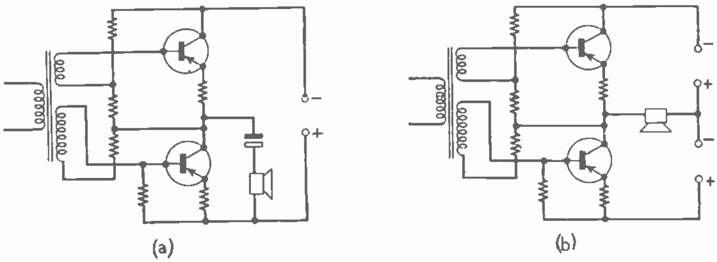


FIG. 45.—" SINGLE-ENDED " PUSH-PULL OUTPUT CIRCUITS.

Output Stages

While the early stages of most receivers so far marketed in the United Kingdom tend to follow roughly similar lines, there has been a more noticeable variation in the output stage.

One popular arrangement, which corresponds to conventional valve practice, is to use two matched transistors in a Class B push–pull common-emitter stage, usually with impedance-matching transformers in the input and output circuits: see Fig. 50. The output transformer may be omitted if the impedance of the loudspeaker speech coil is about 120 ohms with a centre-tap. Also widely employed have been two variations of the so-called " single-ended " push–pull circuit, see Fig. 45.

The single-ended circuit makes it easier to eliminate the output transformer (output impedance of the order of 30–80 ohms); and, since in small portable receivers this component may have an efficiency only of the order of 50 per cent, the omission of the transformer can contribute an appreciable battery saving by providing the same acoustic output for lower electrical output. The single-ended circuit is also easier to stabilize thermally, and the lower losses in the stabilizing components tend to compensate for its theoretical lower sensitivity compared with the symmetrical push–pull circuit. One disadvantage of the single-ended arrangement is the difficulty of avoiding distortion with large signals.

Another arrangement occasionally used is to have two transistors in a common-base configuration with input and output transformers; this arrangement provides less gain than the common-emitter circuit, but at less distortion.

Phase inversion is normally provided by the driver transformer, which may have either a centre-tapped low-impedance secondary or two separate balanced secondary windings to permit independent biasing of the output transistors. While a useful increase in gain can be obtained from this transformer, in some designs a deliberate mismatch may be introduced, since a signal-source impedance several times that of the transistor input resistance helps to decrease distortion. The D.C. resistance of the transformer windings should be kept to a minimum, and it may be advantageous for the two halves of the secondary to be bifilar wound.

The bias to be applied to the output transistors needs careful determination if battery economy is to be combined with a minimum of "cross-over" distortion, throughout the useful life of the battery. Cross-over distortion, which can occur also in Class B valve stages, arises from the fact that each transistor conducts and amplifies during alternate half-cycles of the audio input. If the bias conditions are such that any discontinuity occurs at the change-over point, the two half-cycles of the amplified waveform will not fit exactly together, the resultant distortion being highly objectionable (see Section 26). It should be noted that with transistors, bias is a *current* which flows through the emitter-base circuit: unlike valves, increasing the bias will cause an increase in current; while "cut-off" is a condition of minimum, but not zero, collector current.

Since the battery voltage will gradually diminish, it is difficult to avoid some degree of cross-over distortion towards the end of the useful life of the battery. Negative feedback can help, and in practice there will always be standing current above that of the "cut-off" condition; it is also usual to fit 5 per cent tolerance bias stabilization resistors. A common emitter resistance can help to compensate for slight differences in the characteristics of the two transistors and, as with the driver stage, is a guard against thermal runaway. Receivers of this class are normally designed for operation within the temperature range 15–45° C. Thermistors are frequently used as part of the bias network to reduce sensitivity to temperature variations.

Further information on transistor circuits is given in Section 26 Servicing of transistor receivers is dealt with in Section 39.

CAR RADIO RECEIVERS

The design of a car-radio set and the installation arrangements for the various sections into which it must be divided for convenience requires different consideration from the design of a radio for domestic purposes.

The set itself must be capable of producing a useful output from a very small signal input, which varies through a far wider range than the signal received by a domestic set. In general, it should also provide a greater output, be of smaller physical dimensions, and place a limited load on the vehicle electrical supply system. The introduction of transistors has made it possible to reduce appreciably the power drawn from the car battery.

The construction of the set and accessories must be such as to withstand the intense vibration which could be experienced, particularly overseas, under some conditions of travel, and to tolerate wider variations of temperature and humidity than are experienced in domestic radio.

Interference suppression is discussed in Section 33.

Car Aerials

Car aerials, of whatever construction, are rather poor collectors of signal due both to their positioning and their low effective height. Consequently, a car radio receiver must have a greater sensitivity and better signal-to-noise characteristic than an equivalent domestic receiver. To satisfy this condition a tuned R.F. amplifier stage is often employed; this also helps to improve the A.G.C. characteristic, an important factor in a car radio, which must cope with wide variations of signal.

Electrically, the aerial is capacitive; as this aerial and its screened cable connection to the set are fixed once the set has been installed, the aerial can be coupled more closely into the first R.F. circuit than is possible in a domestic radio. Variations in capacitance are taken up by an aerial trimmer, which should always be adjusted after a set and aerial have been installed in a car. Usually this trimmer is adjusted for either maximum signal or maximum noise near a tracking point. Manufacturers' instructions should be followed, as it may be necessary to trim on each waveband or on one waveband only.

The higher the aerial, the better. Mostly wing- or roof-mounted aerials are used nowadays; the under-car aerial, due to its poor performance, has fallen into disuse. The choice between a wing or roof aerial is usually governed by mechanical considerations; i.e., ease of installation and accessibility. The roof aerial has better " signal collecting qualities ", but is usually short to prevent the aerial fouling garage lintels. The wing-mounted aerial can be longer; although there may be some screening effect, due to the body of the car, its higher capacitance to " earth " allows closer coupling into the aerial tuned circuit. One advantage of the roof-mounted aerial is that it is less susceptible to various forms of static interference.

There is a current fashion for mounting aerials on the rear wing. The only difference with this arrangement is the extra length and consequent greater capacitance of the screened cable. Depending on the input circuit, no signal loss may necessarily be incurred, but care must be taken to ensure that when the aerial trimmer is adjusted a true maximum is reached. If this cannot be done, either a very low-capacitance cable should be used or the input circuit may require modification, for example, any fixed capacitance across the input may have to be removed or reduced.

Wing-mounted aerials must be of sound construction, and particularly well sealed against ingress of water. Owing to the speed of cars, the water pressures under a wing are very high; and an aerial base mounted on the wing is subjected to this pressure combined with a bombardment of dirt. When a car radio develops poor performance, the earth leakage of the aerial should be checked, especially where the fault develops after a spell of wet weather.

Aerial Input Coupling

Since the aerial is capacitive, the closest coupling to the aerial-tuned circuit will be possible when a permeability tuner is used. This and other factors make for almost general use of permeability tuners on car radio receivers. Tracking on such a tuner is effected by the shaping of the tuning coils or by mechanical linkages, and the only adjustment necessary is of external trimmers. Unless in possession of detailed alignment procedure, it is highly inadvisable when servicing the set to change the relative positions of the movable dust cores in the tuner.

Power Supplies

Car radio receivers have to operate from the vehicle's 6-, 12-, occasionally, 24-volt battery.

Vibrator Supplies

Until recently, the usual method of obtaining the required high-tension was by means of a vibrator convertor. Normally, rotary convertors were used only in communication equipment, where greater power is needed.

As the battery voltage cannot be directly transformed, it is necessary first to convert it to alternating current, transform it up to required voltage and then rectify it. The convertion to A.C. is done by a vibrator; and, in the case of a synchronous vibrator, the rectification as well. The vibrator is basically a self-excited vibrating reed, with one or more sets of contacts.

The exciter coil may be connected either in series, when an auxiliary contact is required for its operation, or in shunt, when no extra contacts are required. The operation of a vibrator convertor can be understood with the help of Fig. 46.

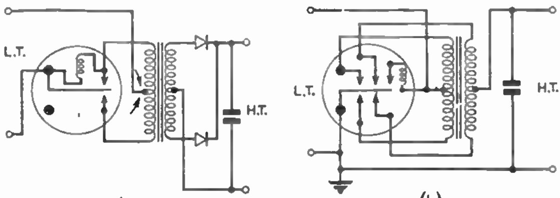


FIG. 46.—VIBRATOR POWER UNITS.
(*a*) Interrupter vibrator with shunt drive; (*b*) synchronous vibrator with series drive.

The current from the battery flows in turn through each half of the vibrator transformer; the voltage induced in the secondary due to change of flux in the core is a series of positive and negative square waves. When this voltage is rectified and smoothed, the required H.T. voltage is obtained.

The exciter coil as shown is energized on application of battery voltage and short-circuited when the vibrator makes contact on one side. In a series-excited vibrator the exciter current flows through the coil via an auxiliary contact on the reed. One of the disadvantages of a series-excited vibrator is that after prolonged storage the auxiliary contact may slightly oxidize and the vibrator may be difficult to start: on a shunt vibrator the oxide film is mechanically broken down on starting.

To prevent excessive sparking and consequent contact wear, the vibrator transformer is tuned by the buffer capacitor. The action can be more easily understood by reference to Fig. 47. After the current through the transformer is interrupted by the reed leaving one contact, the circuit formed by the inductance of the primary of the vibrator transformer and the reflected value of the buffer capacitor is shock excited into oscillations. The frequency of this oscillation is adjusted in such a way that by the time the reed approaches the opposite contact the voltage on this contect is almost equal to the voltage on the reed. It is therefore important to treat as one unit the vibrator and the vibrator transformer with its associated buffer capacitor. Every component, on failure, must be replaced by its exact equivalent. If this is not done, the life of the vibrator will be much reduced; furthermore, excessive sparking is likely to cause serious interference

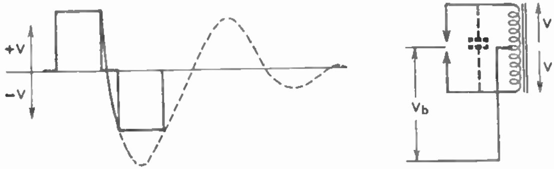


FIG. 47.—PRINCIPLES OF VIBRATOR ACTION.

The operation of a self-rectifying vibrator is similar to the interrupter type except for the addition of an extra set of contacts which are used for reversing the voltage from the secondary of the transformer in synchronism with the reversal of the primary current. Consequently, rectification occurs and no separate rectifier is required. However, as the structure of the synchronous vibrator is more complex, this type tends to be less reliable than the interrupter one.

In this type of vibrator the polarity of the H.T. circuit follows the polarity of the L.T. circuit. It is therefore necessary, if the earth polarity of the vehicle is changed, to interchange the connections either to the primary or to the secondary of the vibrator transformer. On some types of vibrators reversal of polarity is facilitated by making the base symmetrical; the vibrator can then be plugged in its socket in two ways, spaced at 180°. One position is connected for a negative earth system, the other for a positive earth. Thus, when changing from one system to the other, the vibrator may be withdrawn, turned through 180° and replaced.

Generally, a different vibrator converter is required for either 6-, 12- or 24-volt supply. Sometimes vibrators for both 6- and 12-volt supplies are used, either with the exciter coil tapped for the appropriate voltage, or a series resistor connected externally in series with the exciter coil. The windings of the transformer must also be changed when changing from one voltage to another; and series–parallel switching of windings is incorporated on such a set.

Vibrator units for 24-volt s .pplies may be more complicated than 6- or 12-volt ones; this is because there is a danger when switching on of an arc being maintained across the contacts, since ionization of the air between contacts takes place at the higher voltages. To prevent this, a limiting resistor must be connected into the supply on first switching on. After the vibrator has started, this resistor is short-circuited either automatically by a relay, or with a three-position "off-start-on" switch. On low-power 24-volt vibrator convertors the limiting resistor may be left in circuit permanently, where the consequent lowering of efficiency is acceptable.

Transistor Convertors

Transistor convertors may now be used in place of vibrator convertors. Basically, a transistor instead of a vibrating reed is used as a switch to interrupt current; the transformation and rectification being similar to the vibrator system. Further details are given in Section 36.

The transistor or transistors are generally connected in a type of relaxation oscillator circuit; the frequency of oscillation being determined chiefly by the parameters of the transformer.

Certain types of single-ended circuits must not be operated without load, as the output voltage will rise considerably beyond the rated value of the components. A typical circuit is shown in Fig. 48.

The efficiency of transistor convertors can be made higher than that of either vibrator or rotary convertors; also due to higher operating frequencies possible with a transistor convertor, transformer size can be reduced and the overall weight and bulk of such convertors can be very small.

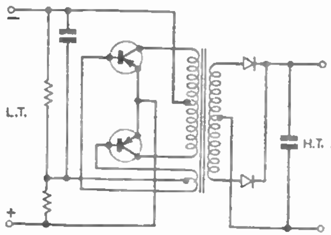


FIG. 48.—TYPICAL PUSH-PULL TRANSFORMER-COUPLED TRANSISTOR H.T. UNIT.

Some car radio sets use very small transistor convertors for supplying the H.T. required for operation of the R.F. and I.F. valves only; the audio and output stages use transistors and operate directly from the battery voltage.

Hybrid Receivers

A recent development has made the use of either vibrator or transistor convertors unnecessary. This type of receiver is termed "hybrid", because it makes use of both transistors and valves. The valves used in the early stages are operated directly from either a 6- or a 12-volt "H.T." line. Useful voltage and conversion gains are obtained under these conditions, but power amplification is not practicable. The audio output power is obtained from a power transistor, which will also operate directly from the low voltages available, does not require heaters and is more efficient in its operation than a valve.

The valve stages of such a set follow fairly conventional lines. A tuned R.F. stage is followed by a frequency changer, feeding a single I.F. amplifier. Detection and A.G.C. are then either accomplished by thermionic or crystal diodes, and there are A.F. amplifier and driver stages.

Component values differ slightly from conventional circuits, not only in the value of various resistors, but also because a small positive standing bias is needed, in addition to grid-leak biasing of the frequency changer and I.F. stages.

An R.F. amplifier stage is invariably employed as, due to lower amplification of valves operating on 12 volts, optimum gain is required. Additional reason for the use of an R.F. stage is the added selectivity which this stage provides in front of the frequency changer, so helping to reduce spurious responses, which may arise from the inferior linearity of 12-volt valves, compared with conventionally operated valves.

The A.F. stage consists of a voltage amplifier valve, which is in turn followed by a driver stage. The driver is in effect a power amplifier capable of delivering a few milliwatts of A.F. output, which is transformer coupled to the transistor output stage. Some power from the valve is required, as transistors are true power amplifiers. Owing to their low input impedance, power—and not only voltage—is required to drive them.

Transistor Output Stages

Power transistors differ from small amplifying transistors mainly by their size, and by the amount of current they will handle. Because of the high powers dissipated in these transistors, additional means of disposing of the resultant heat is required, in order that the temperature of the germanium junction is kept at a safe level. Consequently, power transistors are designed to be clamped to a relatively large mass and area of metal. This is termed the " heat sink ".

This heat sink is designed to help to get rid of the heat both by convection and radiation. It may take various forms, such as a finned plate or a box; in some instances the case of the set itself may be used. It is always important that the heat sink is not covered when installed, and also that some freedom of air circulation is maintained. Also it is inadvisable to mount the transistor unit with its heat sink directly into the air stream of a car heater.

Power transistors are designed for good heat transference from the transistor element to the metal casing; to help matters, one of the transistor electrodes, the collector, is connected directly to the metal casing of the transistor. As it is sometimes impossible to have the collector connected electrically to chassis, it may be necessary to have either the transistor heat sink insulated from the chassis, or the transistor body insulated from the heat sink. The latter is usually done by interposing, between the transistor and its heat sink, a thin mica sheet only a few thousandths of an inch thick. Mica is particularly suitable because of its small heat resistance combined with good electrical insulating properties. To increase heat transfer from the transistor body to the heat sink, all the touching surfaces, including the mica washer, are usually coated with silicon grease.

When a power transistor is replaced, it is important to maintain good electrical insulation, while making sure that the heat transfer is not impeded. While power transistors are semi-conductor triodes, having only three electrodes, namely emitter, base and collector, their characteristics in many ways resemble those of pentode valves. To date, in the United Kingdom, only *p–n–p* types of germanium power transistor have been used; with these the "live" supply voltage is always negative. This must be borne in mind very carefully, as the reversal of supplies may cause serious damage to the transistor.

The usual circuit configuration for transistor amplifiers is the common-emitter circuit. This roughly corresponds to the grounded-cathode circuit of a valve amplifier; the emitter corresponds to the cathode, the base to the grid and the collector to the anode. The common emitter is the basic configuration without any inherent feedback, consequently giving highest gain. Another possible circuit is the common-base arrangement, which may be likened to a grounded-grid valve amplifier: gain of such a stage is small, as there is 100 per cent current feedback. Finally, there is the common-collector amplifier corresponding to a cathode follower: again gain is lower than in a common-emitter circuit due to 100 per cent voltage feedback. In practice, all three arrangements may be found in transistor amplifier circuits, though the common-emitter circuit predominates.

The amplifier stage may use one or more transistors. The working of a single-transistor Class A amplifier (Fig. 49) is analogous to a valve

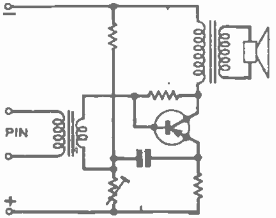


FIG. 49.—SINGLE TRANSISTOR CLASS A OUTPUT STAGE.

Class A amplifier, in that a steady collector (anode) current is flowing, which is adjusted by the value of fixed bias. The signal, which is superimposed on the bias voltage and supplied in series with it, controls the flow of the collector current. This collector current in turn flows through the load formed by the output transformer and its reflected load.

The difference from a valve amplifier lies mainly in the input current being amplified, and the signal as well as the bias being supplied as current. Owing to the high spread of the current amplification factor and the input resistance of the transistors, the value of the bias current is usually adjusted on test, by means of a potentiometer incorporated in the bias network. This current must be readjusted to its correct value when the transistor is replaced or circuit changes are made.

The bias potentiometer forms a ready-made means of checking the correct functioning of the transistor, as the collector current must show change when the bias potentiometer is moved. Care must be taken that a safe value of collector current is not exceeded.

A drawback of the Class A amplifier is the high collector dissipation of the transistor in comparison with the A.F. power output, which will always be less than 50 per cent of the dissipated power. The power taken from the battery will be constant and independent of signal.

If two transistors are used, much more efficient working can be obtained by Class B operation. A Class B amplifier will be more efficient not only in terms of conversion of battery supply to audio power (efficiencies of 70 per cent can be reached) but also more power can be obtained from a given size and dissipation rating of the transistor. This is because the current drawn by the transistors will be proportional to signal. When there is no signal, only very small current flows; when signal increases, the current taken also increases. As the average value of music or speech is considerably lower than the peak value, the average power dissipation in the transistors is lower than in Class A, and consequently the heat sink can be much smaller or omitted altogether; conversely, from a transistor of a given size and rating, much greater output can be obtained in Class B. As Class B circuits have only a very small standing current, there is usually no need, and consequently no provision, for individual bias adjustment.

A typical circuit is shown in Fig. 50.

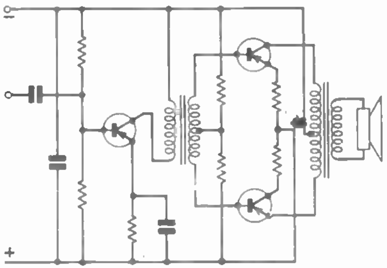


Fig. 50.—A.F. Driver and Class B Output Stages.

In order to obtain somewhat similar characteristics to Class B operation without the need for two power transistors, a "sliding bias" circuit is used in some sets. This is basically a Class A circuit, using one power transistor. The bias, however, is not fixed, but made proportional to the average value of audio signal handled. Consequently, the average power taken from the battery supply and the transistor dissipation can be made lower for a given peak power output, than in a conventional Class A amplifier.

The variable bias current is usually derived from the output stage itself; a part of the audio output being rectified and superimposed on the small standing bias. Care must be taken that the charging-time constant is short and the discharge-time constant is long as compared with the audio-frequencies. In order to achieve this, a directly coupled pre-amplifier transistor stage is sometimes added, the rectified part of the output being applied to this stage together with the fixed bias. With this arrangement the pre-amplifier-stage power gain is added to the "feedback" loop, with consequent lowering of power drain for bias requirements. Thus, the discharge-time constant can be kept fairly long. A disadvantage of "sliding bias" is that a certain amount of harmonic distortion, particularly transient distortion, is inevitably introduced.

With any type of power-transistor stage, it is important always to have a loudspeaker or output meter connected in circuit. Should the set be switched on without a load being connected across the output transformer secondary, the high voltage developed across the inductance of the output transformer may irreparably damage the transistor. This effect is somewhat similar to "flash-over", which can occur in a valve amplifier in similar circumstances.

Sets using the "hybrid" principle can be made for reception on all normal wave-bands, including Band II for V.H.F./F.M. reception. While the reduction of the drain on the vehicle battery is less than is possible with a fully transistorized receiver (largely owing to the heater power required for the valves), an appreciable power economy is possible compared with a conventional valve receiver.

F. J. G.

15. TELEVISION RECEIVER DESIGN

15. TELEVISION RECEIVER DESIGN

TELEVISION RECEIVER PRACTICE

The pattern of television receiver design has tended to become rather more standardized in recent times compared with the decade following the re-opening of the television service in 1946; design efforts have been directed mainly in the direction of better engineering and value for money.

Since the introduction of commercial television some form of multi-channel tuner is always used, either using selected separate coils for each channel, as in turret tuners, or switched coils, as in the incremental tuner or a combination of both. Permeability tuners are occasionally found.

Now that there are a number of frequency modulation (F.M.) radio broadcasting stations operating in Band II, many television receivers also include facilities for receiving these programmes, the purely television section of the receivers being switched out during the reception of F.M. radio.

The 17-in. tube is currently by far the most popular size of cathode-ray tube, 14-in. tubes being used only in the lowest-priced sets and some portables. Many tubes are now electrostatically focused (see Section 24). With the advent of the 110-degree scanning-angle tube, however, even 21-in. tubes can be accommodated in quite compact cabinets; while 17-in. portables are quite small and light, one particular model being only 15½ in. high, 18¾ in. wide and 13¾ in. deep.

Most receivers include a number of features now considered more or less essential, e.g., sound and vision automatic gain control, sound and vision interference limiting and, at least, in most fringe receivers, some form of flywheel control on the line time-base.

As brightness levels of receivers have increased, and because in most receivers the D.C. component of the received signal is partly lost before the signal is applied to the cathode-ray tube, it is possible under some conditions for the lines occurring during the frame flyback period to become visible as widely spaced sloping lines. Most receivers now use a method of blanking out these lines; usually by applying to the cathode-ray tube a suitable pulse from the frame time-base.

The H.T. supply for most receivers usually consists of a simple half-wave rectifier to which the mains supplies are applied via a resistor tapped to accommodate different input voltages. It is somewhat cheaper and generally lighter to have large capacitances and small inductors to provide a given degree of H.T. smoothing; this also allows the voltage drop across the choke to be kept small. With very large electrolytic capacitors, 100–200 μF, used for H.T. smoothing, electrolytic charging currents have become very high, placing considerable strain on valve rectifiers, so that metal rectifiers are often used. The voltage drop across these rectifiers on low mains supplies is often appreciable, and many receivers have facilities for shorting-out the rectifiers when low-voltage D.C. mains are used.

Silicon power rectifiers may well supersede selenium types, as they have negligible voltage drop, less than 1 volt compared to 12–20 volts

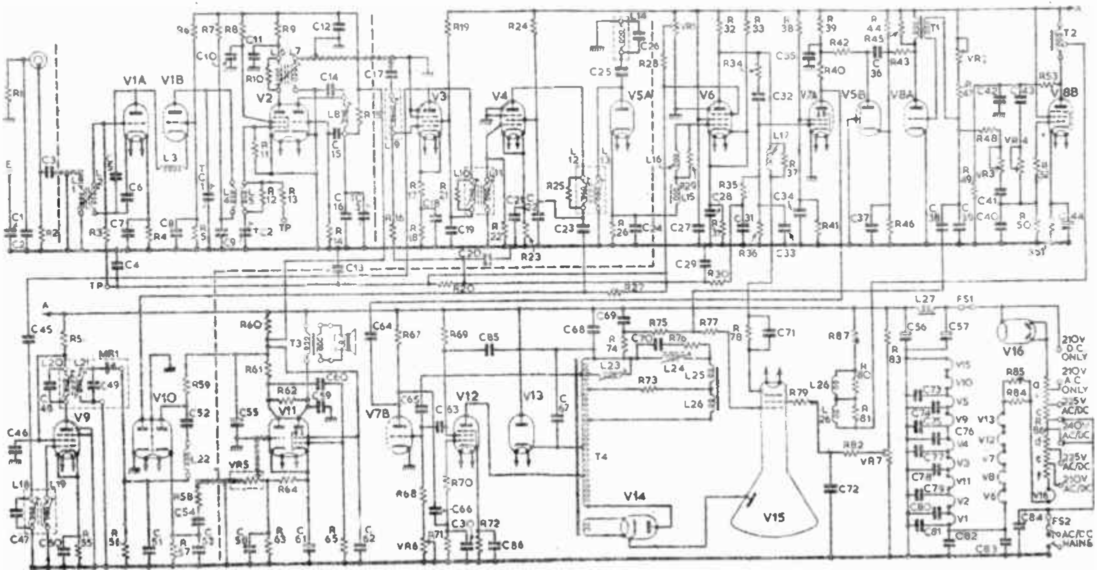


Fig. 1.—Representative Circuit of a Band I/III 17-in. Receiver.

for selenium, are very small and can operate safely at temperatures in excess of 100° C.

The introduction of printed-circuit technique to television receivers has drastically changed the form of many receivers. Because a design has to be completely developed before mass-production printing—and because of the difficulty or impossibility of modifying a design once printed—manufacturers have tended to use several small printed circuits clipped or bolted on to a main framework. Such panels are easier to print, wire up and test, and, if modifications have to be made to them, the minimum waste occurs. In addition, many receivers have detachable or replaceable panels to ease the service problem.

The circuit diagram of a typical 17-in. receiver with a 90° deflection picture tube is shown in Fig. 1. The valve types and functions are as follows: V1 30L1 (PCC84, 7AN7) cascode R.F. amplifier; V2 30C1 (PCF80, 9A8) frequency changer; V3 EF85 (6BY7) common vision and sound I.F. amplifier; V4 30F5 vision I.F. amplifier; V5 6D2 (EB91, 6AL5) vision demodulator (A) and frame sync. pulse shaper (B); V6 30F5 video amplifier; V7 30FL1 sync. separator (tetrode) and line oscillator (triode); V8 PCL82 (16A8) frame blocking oscillator (triode) and frame output (pentode); V9 30F5 sound I.F. amplifier; V10 6D2 sound interference limiter (A) and A.G.C. clamp (B); V11 30PL1 audio amplifier (triode) and audio output (tetrode); V12 30P4 (PL36) line output; V13 U191 efficiency diode; V14 U26 E.H.T. rectifier; V15 CRM173 picture tube; V16 PY32 H.T. rectifier; MR1 CG6E sound detector.

RADIO-FREQUENCY AND INTERMEDIATE-FREQUENCY AMPLIFICATION

As in modern radio-receiver practice, the majority of the amplification in a television receiver takes place before the sound and vision detectors. For a considerable period British designs assumed the reception of one channel only, with the result that many were of the T.R.F. type, and could be retuned to another channel only if major alterations were undertaken.

Latterly this concept has disappeared; all receivers can be tuned from channel to channel readily, and for this reason the superheterodyne principle is used exclusively.

Basic Requirements

In a superheterodyne receiver the intermediate-frequency amplifier is responsible for most of the gain, and also for determining the precise shape of the response curve. Due to the fairly rigid conditions imposed by vestigial sideband transmission, the response must be held to within fairly close limits if various distortions are not to appear on the picture; and with this in mind the circuits, operating frequency and valve types used must be chosen with some care. In any case a commercial receiver has always to be designed with a view to reasonable production economies, and many ingenious ideas have been introduced into this part of the set in order to achieve the required results with the minimum number of components.

The tuner must select the wanted channel and convert the vision and sound signals to the intermediate frequency. It therefore contains, basically, a mixer and local oscillator for the purpose of frequency conversion. Additionally a signal-frequency amplifier stage is nearly always included. Selectivity against unwanted signals is of the utmost importance, and to provide this prior to the converter with tuned circuits alone is most difficult. Without adequate facilities, many signals at various frequencies could beat with the local oscillator to produce spurious interference with the wanted signal at the intermediate frequency.

As was pointed out earlier, a modern receiver of high sensitivity will provide a good signal-to-noise ratio when operating at full gain. Achieving this result depends upon careful design in the tuner section. The mixer inherently produces much random noise, and the effect of this must be overcome by means of a radio-frequency amplifier giving a low noise contribution.

Much trouble may be caused by radiation from the local oscillator of a television receiver, especially if it occurs at a frequency liable to cause beat interference with a nearby receiver, perhaps operating on another channel. Apart from careful screening, the radio-frequency stage will provide a large part of the attenuation needed to prevent the local oscillator output from reaching the aerial.

Adequate arrangements must also be made to separate fully the sound and vision intermediate-frequency signals within the receiver; the possibility of the visi n signal components appearing in the sound channel and vice-versa must be entirely eliminated, except, perhaps, under the most unlikely cor ditions (e.g., a signal strength several times higher than the overload level, or complete mistuning). To this end various trap circuits are included in the vision intermediate-frequency channel to attenuate the sound signal as much as several hundred times, and care must be taken to prevent other forms of coupling between the sound and vision amplifiers.

THE RADIO-FREQUENCY AMPLIFIER

Noise Considerations

As has already been stated, the first purpose of the signal-frequency amplifier in a superheterodyne receiver is to provide gain prior of frequency conversion sufficient to render the large noise contribution to the mixer stage unimportant. The random noise appearing with weak signals is then due only to that received with the signal, the noise generated in the aerial radiation resistance and the noise contributed by the first circuit and valve—i.e., the radio-frequency amplifier. If large, this noise voltage can severely limit the maximum usable sensitivity of the receiver, and in good designs every effort is made to reduce it to an extent where the aerial and received-noise predominate.

At frequencies below 50–60 Mc/s, using modern valves, the noise in the first stage will be largely due to shot and, in the case of a pentode, partition noise, which, for purposes of calculation, is normally assumed to occur in an imaginary resistance R_{eq} in series with the grid of the valve. With R_{eq} taken as 1,000 ohms it is just possible, under these conditions, to design, for a receiver having the desired bandwidth for the British television system, a radio-frequency stage which has a noise factor (N) of 5 db (noise factor being defined as the ratio of the total noise at the output from the source of signal and the receiver divided by the noise due to the signal source alone). If the noise factor is 5 db, then to a large extent the receiver will not unduly limit the range of reception, and for a signal-to-noise ratio of about 12 db—which experimentally has been found to be the limit of useful reception—the minimum usable signal is of the order of 10 μV.

When it is desired to achieve similar results in Band III (174–216 Mc/s), the question of transit-time damping has to be taken into account. In the first place the type of valve used at 50 Mc/s will usually have such a low value of input impedance at 200 Mc/s that it will be very difficult indeed to produce much gain. This means that the mixer noise contribution now becomes a serious factor, and also that induced grid noise has to be considered.

The input conductance due to transit-time damping is proportional to mutual conductance, the square of the frequency and the square of the transit time; as also is the noise current. It is permissible to consider these noise fluctuations to be the noise of the input resistance. It can be seen that the frequency of operation, being a squared term, has a serious effect, and only by the use of specially designed valves, with a suitable transit time, can a low noise factor be achieved, even supposing a low value of R_{eq}. Triodes are in fact most suitable at 200 Mc/s, since they do not suffer from partition noise and can be made to have a good transit-time characteristic. The "Noise Reference Frequency" (defined as the frequency where $R_{eq}/R_t = 1$ and designated f_n) is generally twice as high for a good triode as for a good pentode. It may be shown [1] that the noise factor for a triode when the input arrangements are correctly matched will be :

$$N = \frac{7}{2} + \frac{9}{4} f^2/f_n^2 \quad . \quad . \quad . \quad . \quad (1)$$

While for a pentode

$$N = \frac{7}{2} + 4f^2/f_n^2 \quad . \quad . \quad . \quad . \quad (2)$$

Considering the difference between the values of f_n, the choice of a valve is obviously centred on a triode.

Circuits

It is not necessary to dwell at length on the circuit of the pentode stage for use up to 50 or 60 Mc/s, as it is in most respects similar to that of an intermediate-frequency stage, which will be discussed in detail later. In order to achieve the maximum gain to the mixer grid, a four-terminal coupling network should be used, and band-pass filters have been very common in designs for Band I reception only.

In using a simple triode for the first-stage amplifier the circuit must be very carefully considered. With the simple grounded-cathode connection, feedback through the anode-grid capacitance is the foremost problem, and satisfactory neutralization circuits are not easy to design. The effective input capacitance tends to be increased, and is apt to depend critically upon frequency. Neutralization also becomes too critical if high gain is attempted. One of the first triode input circuits to be used commercially was a push–pull arrangement developed by R.C.A. With this circuit, cross neutralization of a non-critical type is practical, and suitable gain-band-width characteristics were obtained.

Two major disadvantages, however, are (1) the push–pull arrangement is vastly more complicated and expensive than a single-ended one; and (2) the neutralized triodes provide relatively poor isolation between the local oscillator and the aerial.

The grounded-anode triode connection has not been used commercially on a large scale. It requires neutralization, which may be at least as difficult to apply as with the grounded-cathode triode.

The grounded-grid triode connection is, however, very popular. It is an inherently stable arrangement owing to the high degree of cathode feedback. Compared with a pentode, the input and output impedances are, of course, quite low, and for this reason the valve is often made to match the aerial feeder directly when used in commercial receiver designs. This has the disadvantage of making the set rather less selective against image frequency and other similar interferences and does not provide the maximum of attenuation of the local oscillator e.m.f. When matched with a tuned circuit to the aerial feeder and the circuit adjusted for best noise factor, the grounded cathode triode tends to have a low gain. This means that the noise at the input to the second stage has to be considered, and a second low-noise-input stage is really required.

Cascode Amplifier

An amplifier which has been universally used for Band III is the grounded-cathode, grounded-grid configuration, usually referred to as the " Cascode " circuit. The basic form is shown in Fig. 2 (*a*), where two triodes are seen to be connected in series. The second triode may have its D.C. grid voltage fixed by means of the potentiometer and, acting as a cathode follower, holds the anode voltage of V1 very nearly constant. The anode current of V1 still flows in the load formed by the cathode input impedance of V2 but, with the anode voltage constant, the behaviour is that of a pentode having no screen current and no partition noise. From the signal point of view, V2 is a grounded-grid triode, is stable and produces gain with a good-noise factor V1 produces little gain due to the low load, but has a high input impedance —allowing a good matching circuit, which helps the noise factor—and does not essentially require neutralization.

Fig. 2 (*b*) shows a practical circuit in which it will be seen that some inductance (L) has been included in the anode circuit of V1 and a neutralizing capacitor has been added. Although much has been published concerning this arrangement, a brief description is as follows: L is of such a value as to be effective only in Band III (in television) and the neutralizing capacitor C_n with the bypass capacitor C_1 form a bridge circuit with the valve capacities. By careful choice of values it is possible to obtain excellent balance with regard to the feedback and neutralizing reactances, so that optimum gain and noise performance are realized. The purpose of the tuning capacitor C_t will be described later under " Tuners ".

The cascode circuit is described in *Vacuum Tube Amplifiers* (Vol. 18, M.I.T. Radiation Laboratory Series), and figures are given for a first stage in a 2·5-Mc/s band-width amplifier at 180 Mc/s. The cascode arrangement provides a noise factor of 5·5 db as against 7 db for two grounded-grid stages in cascade. Since then special valves of the double-triode type have been developed for commercial use. In

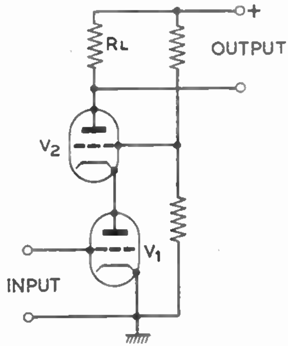
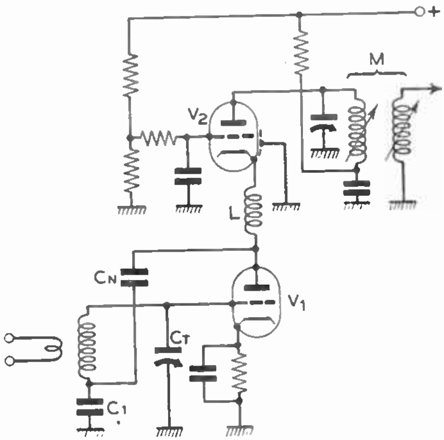
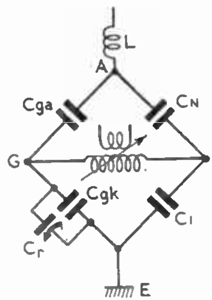


FIG. 2 (*a*) (*left*).—THE BASIC FORM OF CASCODE AMPLIFIER.

FIG. 2 (*b*) (*below, left*).—PRACTICAL CASCODE ARRANGEMENT.

FIG. 3 (*below, right*).—EQUIVALENT CIRCUIT OF THE CASCODE ARRANGEMENT OF FIG. 2 (*b*).





addition to excellent characteristics, special screening is included, so that the anode of V2 (Fig. 2 (*b*)) may be well separated from the rest of the circuit. Needless to say, care must always be taken to continue this shielding outside the valve.

Recently, sharp cut-off tetrodes such as the 6CY5 are being used as R.F. amplifiers in the United States in place of the conventional cascode.

Input Circuits

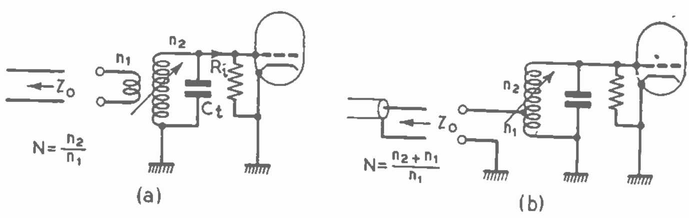
The circuit used to match the aerial feeder to the grid of the first amplifier stage will be discussed from the practical point of view under "Tuners". Certain points of theory are worth noting, however, since the circuit is difficult to design in order to achieve good results at all frequencies. On the other hand, only correct matching will provide the degree of performance necessary to exploit the capabilities of a low-noise-input stage.

Commonly used aerial systems may work into a feeder having a characteristic impedance Z_0 between 75 and 300 ohms, and the problem is to match this to the first valve so that maximum power transfer is achieved with the requisite response curve. This will generally be necessary over the limited band of frequencies occupied by one transmission, with the proviso that the circuit may be easily retuned to another channel by means of a simple change to the transformation network. The valve-input impedance R_i and the dynamic circuit impedance R_d form the load, of which R_i can be considered predominant in most cases. The difficulties encountered due to the dependence of R_i on frequency will be mentioned in detail later; meanwhile it can be seen that between Bands I and III some accommodation must be made for the variation in R_i, which will be of the order of 1 : 16.

Fig. 4 illustrates three single-tuned input circuits which are suitable if the first stage is followed by a band-pass network. The circuit in Fig. 4 (*a*), being a tuned transformer, has the advantage, if correctly designed, of being suitable for balanced or unbalanced cables; in the first case the primary would probably have an earthed centre-tap. It is suitable for use with a cascode amplifier where both ends of the secondary coil are available for connection to the valve and feedback circuit and, in certain types of tuner, enables the best match to be obtained on each channel. Its primary drawback is that it is very difficult to couple tightly enough when ratios above 4 : 1 or 5 : 1 are wanted.

From the coupling point of view, circuit Fig. 4 (*b*) is a slight improvement but cannot accommodate balanced feeder. In tuners of the permeability type it is also difficult to keep the transformation ratio to the value needed.

Circuit Fig. 4 (*c*) is particularly satisfactory for medium ratios if a balanced input is not necessary, and fits well into some tuner arrangements. It also possesses the virtue of being a low-pass filter, and provides excellent attenuation of the local oscillator e.m.f.; although it is



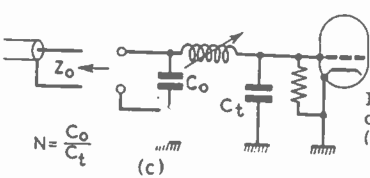


FIG. 4.—SINGLE-TUNED INPUT CIRCUITS: (*a*) TUNED TRANSFORMER; (*b*) TAPPED COIL; (*c*) π-FILTER NETWORK.

often necessary to provide an extra filter for the attenuation of interfering signals at low frequencies.

While double-tuned input circuits have not been very popular in commercial-set designs, they may appear before long for various reasons. Under certain circumstances better matching across the band may be realized, while the noise figure and—generally—selectivity would tend to be improved.

FREQUENCY CONVERSION

The conversion of the incoming signal to the intermediate frequency has most often been accomplished in television receivers with a two-valve circuit. Special frequency changers of the triode-hexode or pentagrid type—as used in radio practice—are generally not suitable at high frequencies and for large band-widths, due to low conversion conductance and high values of R_{eq}. A triode forms a suitable oscillator even at nearly centrimetric wavelengths; while pentode or triode mixers have been used in the same or a separate envelope. In the U.S.A., receivers for Bands IV and V have used crystal mixers due to the inadequacy of available valve types above 400 Mc/s.

Another arrangement popular in this country and in Europe is the self-oscillating mixer—sometimes a triode and sometimes a pentode—which has the advantage of saving valves and components.

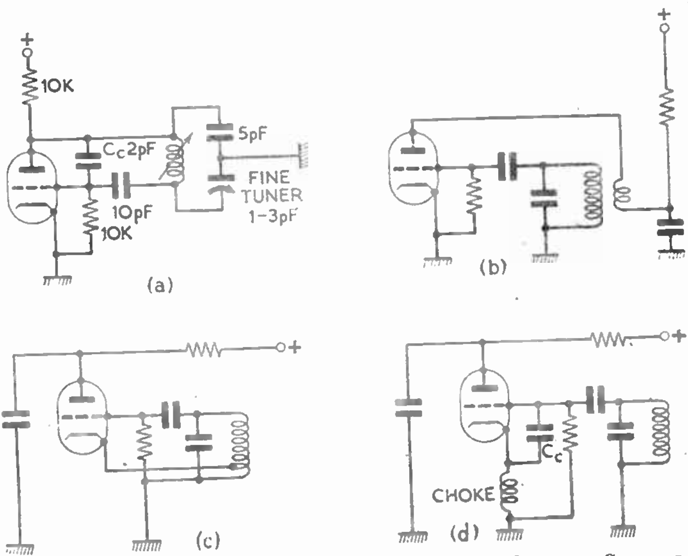


FIG. 5.—POPULAR LOCAL OSCILLATOR CIRCUITS: (*a*) MODIFIED COLPITTS; (*b*) TICKLER COIL; (*c*) HARTLEY; (*d*) MODIFIED HARTLEY.

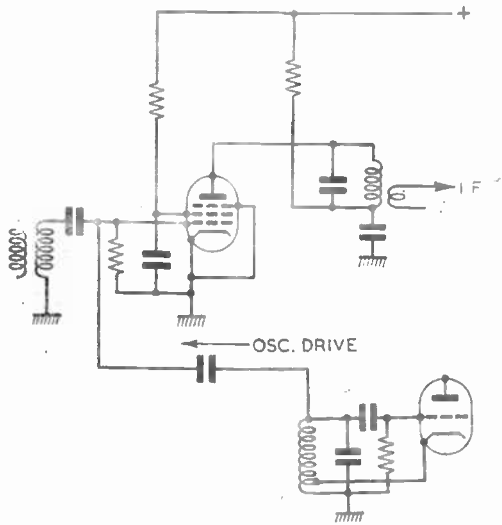


FIG. 6.—PENTODE MIXER WITH SEPARATE LOCAL OSCILLATOR.

Oscillator Circuits

Standard oscillator circuits are, for the most part, used with suitable modifications for the high frequencies encountered. A popular form is shown in Fig. 5 (*a*). This is a modified Colpitts, since the C_{ag} and C_{gk} capacitances form a large part of the tuned circuit. Additional capacitors are often added to obtain stability or for the purpose of fine tuning control. C_c, though only 2 pF, makes an almost unbelievable improvement to the drift characteristic, due to the change of valve capacitances during warm-up. The fact that the coil has only two connections is very helpful in tuner design.

Fig. 5 (*b*) and (*c*) show two circuits using coils having more than two connections. Apart from this disadvantage they have been widely used and have particular merits. Arrangements (*c*) and (*d*) both have earthed anodes. This may be very useful, since the anode becomes a virtual screen, except round the valve lead-out wires, and in this way local-oscillator radiation has been lowered in practical designs. The temperature-drift-compensation-capacitor has been shown in Fig. 5 (*d*), connected between grid and cathode, and could, of course, be connected to achieve the same result on the other arrangements.

The Pentode Mixer

Grid mixing has been found the most suitable with pentodes, a common circuit arrangement being shown in Fig. 6. The oscillator drive is coupled lightly into the pentode-grid circuit, where the bias is self-

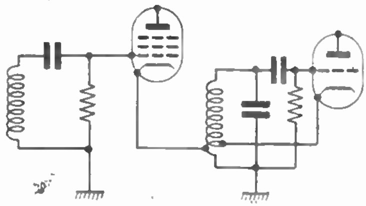


FIG. 7.—ALTERNATIVE METHOD OF COUPLING LOCAL OSCILLATOR TO PENTODE MIXER.

determined by the drive due to the grid leak and coupling capacitor. The arrangement is somewhat critical as to screen potential, and the conditions of operation, in all respects, should be carefully determined experimentally so that maximum conversion conductance (S_c) is obtained. In practice, the gain is seldom better than one-third of the figure obtained with the same valve type used in amplifier service.

An alternative method of coupling-in the oscillator drive is shown in Fig. 7. This has the advantage of removing the effects of the fairly close coupling needed from the input network, and also the oscillator volts to some extent. While suffering from the disadvantage that it is not easy to arrange constant injection over a wide frequency range, it may become popular due to the need to limit severely the effects of oscillator radiation.

The pentode mixer is an extremely satisfactory device in one respect; the lowest television channel is, in some designs, very close in frequency to the intermediate frequency, and the low level of C_{ag} feedback makes the circuit inherently stable, provided that a suitable valve is used.

Offsetting the advantages are the high values of R_{eq} of a pentode in mixer service, the low value of R_i in Band III and the damping caused by grid current. The gain of the previous stage therefore tends to fall appreciably with increasing frequency, and the mixer noise contribution becomes important.

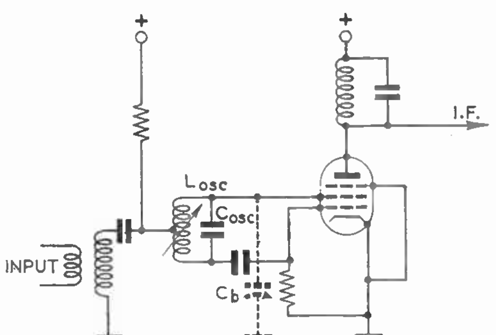


FIG. 8.—SELF-OSCILLATING MIXER.

Triode Mixers

As for radio-frequency amplification, the triode becomes particularly attractive in Band III, where the considerable difference between the input and output frequencies, and the fact that all circuits are tuned, makes neutralization unnecessary. At the lower end of Band I the situation with a triode mixer may be very difficult as the intermediate frequency may be separated by only 3–5 Mc/s from the signal frequency. Even here, the feedback can be overcome by using a capacitive load, but so much of the gain will be lost in this way that performance will be very poor at higher frequencies. In commercial designs the intermediate frequency is either well separated, or the standard forms of neutralization applied. In all other respects the circuits do not differ from those used with pentodes, and further elaboration is unnecessary.

Self-oscillating Mixers

The circuit described below was first publicized in this country in the application reports of Messrs. Mullard Ltd., who, with their associates, were responsible for its development.

The basic circuit shown in Fig. 8 uses a high-slope radio-frequency pentode connected to oscillate between grid and screen. By carefully adjusting the circuit capacitances (a balancing capacitor C_b may be needed for this) the centre tap on the oscillator coil will be at zero oscillator-potential, and the signal is fed in here. Apart from any loss due to the series effect of the oscillator circuit ($L_{osc.}$ and $C_{osc.}$), signal and oscillator frequency components are now modulated on the electron stream, and mixing takes place. The screen voltage and grid current are arranged for maximum conversion conductance, which, it is claimed, compares very favourably with normal grid mixing.

Against the advantages of economy and high S_c must be set the low input impedance caused by grid and screen currents. This may make realization of the full gain of the preceding stage difficult. Further, the accuracy of balance may be poor because :

(a) tuning arrangements of $LC_{osc.}$ are not easy to balance perfectly;

(b) $L_{osc.}/C_{osc.}$ is usually made small to prevent loss of signal, and with a small inductance it may be more difficult to keep balance.

Accuracy of balance is essential to reduce oscillator radiation, and it should be remembered that it is probably not possible for the average service man to check this point. The question of input impedance, balance, etc., tends to make the circuit more or less unusable on Band III.

A development of the self-oscillating mixer is shown in Fig. 9, where the grid leak resistance has been made very large (several megohms) and the coupling arrangement has removed the effect of the other electrode current present in the last circuit. The general problem with an oscillator having a very high grid leak is that it tends to block for long periods or " squeg ". This, too, has been avoided by a novel feedback connection to the return of the signal circuit with the aid of suitably chosen time constants. The balancing capacitor is again present, and must be adjusted for minimum oscillator voltage at the input.

Excellent performance is claimed for this circuit, particularly in regard to noise factor. Once set up, it has many advantages for opera-

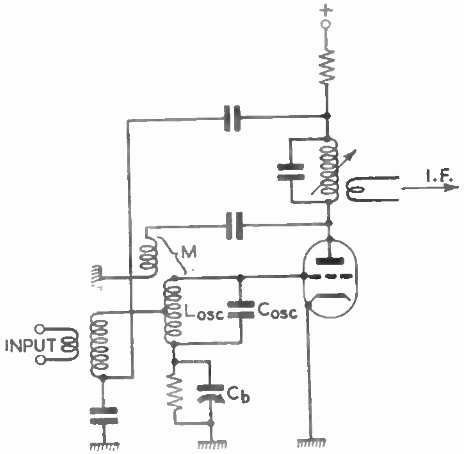


FIG. 9.—ALTERNATIVE FORM OF SELF-OSCILLATING MIXER.

tion on a single-frequency allocation, but does not easily fit the requirements of a multi-channel receiver. It is very suitable at 200 Mc/s, but cannot be used if the intermediate and signal frequencies are close.

Oscillator Stability

It is, of course, essential in any superheterodyne receiver to keep the local oscillator reasonably stable. In a television receiver it should not be necessary to readjust the oscillator tuning control, except when tuning from channel to channel; and the circuit must therefore be frequency stable over both long and short periods. Average sound-

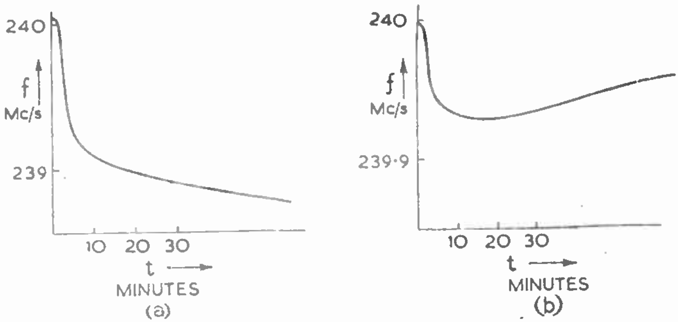


FIG. 10.—OSCILLATOR DRIFT ON BAND III: (a) WITHOUT COMPENSATION; (b) DRIFT REDUCED BY CAREFUL CHOICE OF TEMPERATURE COMPENSATION.

and vision-channel characteristics will probably accommodate a drift of ± 100 kc/s, and it is wise to try to keep well within this figure if picture and sound are not to deteriorate owing to mistuning.

The means of achieving oscillator stability have already been indicated to some extent, and the advantage of partial swamping of the valve capacitances with a suitable capacitor cannot be over-emphasized. Fig. 10 (*a*) shows a typical drift curve for an oscillator in Band III. Most of the change occurs during the first 10 minutes, and is due to the valve reaching operating temperature, following which the general temperature rise in the set causes a further change. In Fig. 10 (*b*) the drift has been reduced overall by about ten times by: (1) the careful choice of a compensating capacitor; (2) making sure that all other components and parts have small temperature coefficients; and (3) in general exercising considerable experimental care.

CHANNEL SELECTORS

Channel selectors fall roughly into two classes. The first—which was in general use in British receivers between 1949–54—is a pre-set arrangement and cannot be readily adjusted by the owner without some inconvenience. The receiver is, therefore, essentially a single-station set, although it can be tuned, when required, to other stations. The second class of receiver has a channel-selector knob available to the owner, and in the U.S.A. has afforded the choice of twelve channels in Bands I and III until the introduction of Band IV and V transmissions. (Eighty-two channels are now theoretically available in America, but there are few designs of selector giving more than partial coverage at the time of writing.) Since 1955, British receivers have been fitted with tuners covering Bands I and III.

Pre-set Tuning

Fig. 11 shows a basic circuit of the radio-frequency, mixer and oscillator stages of a receiver for Band I. Similar arrangements appeared in many manufacturers' products until 1955. L_1, L_2 and L_3 are the signal-frequency circuits, tuning with the valve and stray capacitances in order to give the highest L/C ratio. L_2 and L_3 form a top-capacitance-coupled, band-pass filter, which must be adjusted to peak on either side of the centre frequency to which the aerial circuit is tuned. The local oscillator coil L_4 must be tuned to a frequency higher than the signal by an amount equal to the intermediate frequency.

In some receivers the coils have been permeability-tuned, and the tuning cores have simply been brought out for adjustment during installation. In this case the oscillator is easy to set for maximum sound, but the signal circuits have to be peaked more or less correctly, as otherwise the response curve will be very poor. In these circumstances the band-pass filter has generally been dropped and the signal circuits made flat enough to obviate serious tuning difficulties. This, of course, leads to loss of gain, and the noise factor will be adversely affected. However, not all receivers are intended for "fringe area" reception.

In other designs the four tuning slugs have been ganged together and actuated by a knob-operated, mechanical drive. While there is no

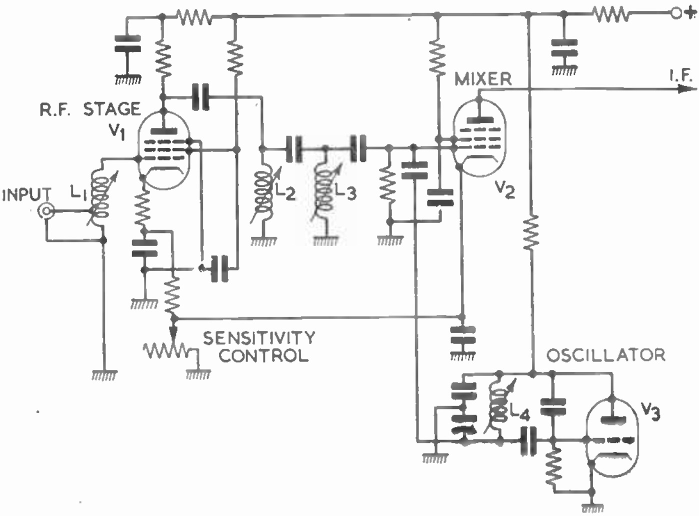


FIG. 11.—TYPICAL "FRONT-END" OF A BAND I RECEIVER.

doubt that this is a method giving superior circuit performance, the ganging problems are serious. The tuning cores must have close permeability-tolerances, and the drive system must be accurate. The coils must be wound to close tolerances and sometimes have a variable pitch. The tracking problem is similar to that encountered in radio sets, and it is generally possible to align the oscillator frequency accurately at each end of the band. The errors at mid-band must be accommodated by having a wider signal-frequency pass-band than would otherwise be necessary.

Another possibility is to have a switch (probably requiring four wafers) to insert new coils for each channel. Although expensive, this method ensures high performance, apart from the effect of the extra capacitance of the switch.

None of these arrangements is particularly complicated, and for this reason most designs have included the parts on the same chassis as the intermediate-frequency amplifier. Considerable complications would, of course, result if Band III reception were required, and above 100 Mc/s the permeability method of tuning has proved very difficult to design. In the higher-frequency region switches, etc., have become very popular and, being most suitable for customer operation, fall into the group about to be described.

Customer-operated Tuners

Where several transmissions are receivable in one location the owner of a television receiver has need to operate the channel selector. This situation exists in most parts of the United Kingdom, and designers are now fitting two-, three-, twelve- and thirteen-channel tuners of a type that can be operated as easily as, or more easily than, the tuning of a normal radio receiver.

In the U.S.A. the majority of tuners have been of the "detent" type. That is to say, the operator has the feeling of operating a switch to move from one station to another, even if the mechanism of tuning is actually not a switch. The American public have expected to see a picture and hear sound immediately upon selecting the wanted channel, but to ensure accurate tuning, a fine oscillator frequency control has been brought out to a second knob. Selector and fine tuner shafts are made concentric.

The rotary wafer switch as the basis of a tuner has been used very successfully in a large percentage of American receivers. Even more popular has been the rotary coil turret, in which the coils required are mounted round the periphery of a rotor drum, with the necessary contacts located at different points. The valves and other parts are located on a chassis with spring stator contacts making connection with the appropriate rotor segment. Other types of tuner have formed only a very small percentage of the total number used.

Fig. 12 gives the basic circuit of a cascode radio-frequency stage, a pentode mixer and a triode oscillator located in the same envelope as the pentode. It can be seen that the input and output sections of the cascode amplifier have been carefully shielded, both inside and outside the valve. Likewise, the internal screen in the triode-pentode helps to prevent accidental oscillator injection.

The turret tuner (in effect) involves no more than the circuit shown, since the coils or transformers appropriate to each channel are presented

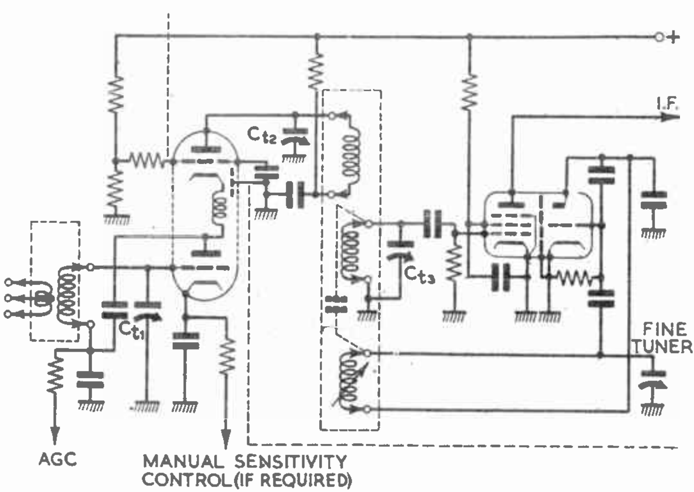


Fig. 12.—"Front-end" of a Multi-band Receiver Suitable for Use with a Turret Tuner.

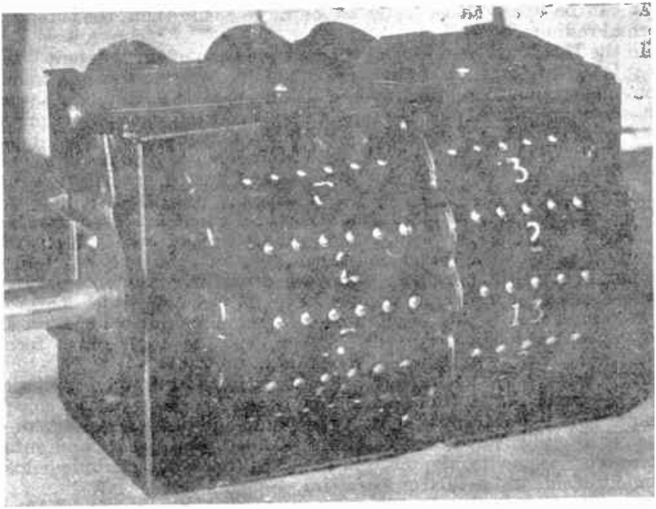


FIG. 13.—A TURRET TUNER.

in turn to the stator contacts. This is inferred in Fig. 12 by the arrows at the end of the coil leads and the dotted line enclosures round the coil sections. Oscillator injection is assumed to be by means of mutual coupling between the coils, but a coupling capacity can be added inside or outside the rotor. It is difficult to allow for individual coil adjustment in a turret and, in practice, the coils are all manufactured to standard inductance values. On insertion in the finished tuner, the stray capacitances are trimmed to the expected value, to enable the circuits to tune accurately. The small trimmers connected to the "hot" ends of the coils are for this purpose, and would generally have a maximum capacitance not exceeding 5 pF. Fig. 13 shows a typical design.

Wafer switches tend to have in themselves fairly high values of stray inductance and capacitance. For this reason, and the fact that the design would become rather bulky, switch tuners normally do not select individual coils. Preferably the tuned circuit is series connected, in suitable sections round the switch, so that the total inductance corresponds to that required for the lowest channel frequency. The result is rather of the form of a transmission line, and the unwanted sections are shorted out. Fig. 14 indicates the principle, although there is some advantage in using both sides of a switch wafer so that each individual section can be shorted. When operating on the highest channel the whole wafer is short-circuited, and a small remaining inductor of the requisite value is tuned to this channel. The coils appertaining to the separate channels are not normally adjustable by the standard methods. since, in any case, they are too small in Band III. The Band III

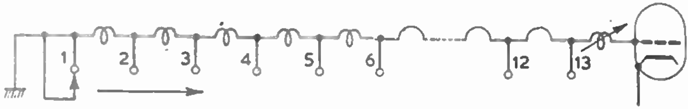


Fig. 14 (*above*).—A Series Connected Switch-tuner Circuit.

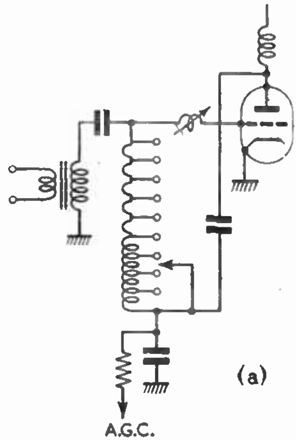
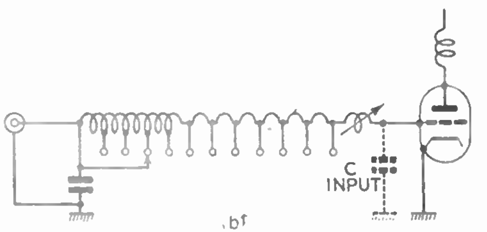


Fig. 15 (*a*) (*right*).—Practical Arrangement for a Switch-tuner Input Circuit with an Input Transformer.

Fig. 15 (*b*) (*below*).—A Switch Tuner with a π-network.



coils can be made in "hairpin" form and stretched as needed. The low-band coils have several turns, but are still conveniently adjustable in this way.

The design of the input circuit becomes difficult with switch tuners because of the cost and bulk resulting from segregating the channels. One method is to have a wide-band input transformer, lightly coupled to the resonant input circuit (see Fig. 15). Here the transformer tends to be troublesome, as it is not easy to design this to work satisfactorily on Band I as well as on Band III. The turret tuner, for all that it has basically the input coil of Fig. 4 (*a*), usually gives more consistent results. An improvement can be made by using a transformer for each band, and a switch wafer for change-over. A better arrangement might be that shown in Fig. 15 (*b*), provided that it could be made suitable for the

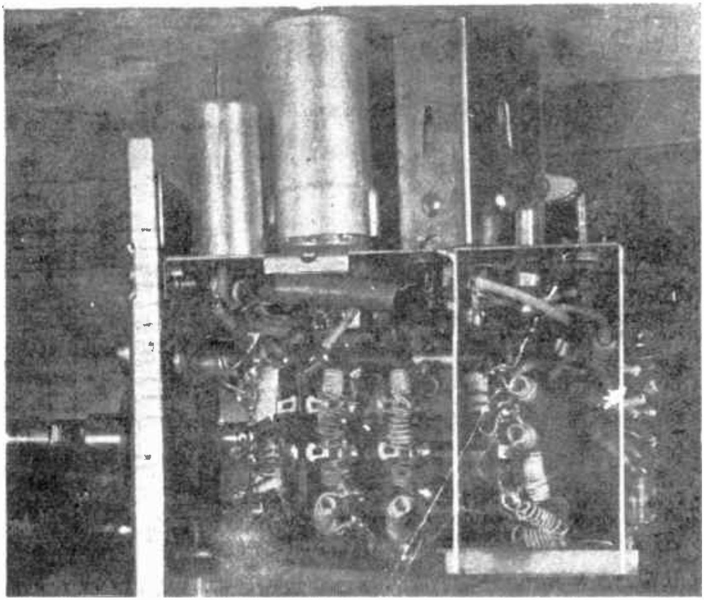


FIG. 16.—A SWITCH TUNER.

cascode amplifier. It may be that double-tuned circuits will provide an answer.

Fig. 16 shows a modern switch tuner.

INTERMEDIATE-FREQUENCY AMPLIFIERS

Choice of Intermediate Frequency

Mention has already been made of the care with which the intermediate frequency should be chosen in the design of a television receiver. The following list indicates the most important considerations :

(1) harmonic interference ;
(2) freedom from direct breakthrough interference ;
(3) second-channel rejection and other spurious responses ;
(4) intermediate-frequency rejection in tuner and mixer stability ;
(5) gain and band-width ;
(6) other response shape considerations.

Dealing with these matters in order, the high-gain intermediate-frequency amplifier of a television receiver is quite capable of producing harmonics of the signals which are applied to its input. In the case of the modulated carrier of a television transmission the power falls away rapidly at each side of the carrier, and any harmonic radiation from the output stages will be most noticeable at multiples of the intermediate-frequency carrier. It is possible for these harmonics to enter the signal-frequency amplifier, and if one is close in frequency to the input signal a beat will occur. This will appear on the picture as a line-pattern, modulating the brightness of the image.

While harmonic interference can be serious, good engineering of the intermediate-frequency amplifier, such as will be necessary to ensure stable operation, usually disposes of the harmonic problem. Harmonics, to some extent, are bound to occur in Band III, and proper design is the sensible solution.

If the highest possible intermediate frequency is used, the number of such combinations will, of course, be reduced, and B.R.E.M.A. has recommended 34·65 Mc/s (vision) and 38·15 Mc/s (sound).

In some locations receivers will be required to operate near to high-power radio-frequency sources, such as short-wave transmitters, radio-frequency heaters, etc. Even though the tuner provides adequate attenuation, there is the risk of interference entering the intermediate-frequency amplifier directly. The solution, except in extreme cases, is a matter of adequate screening, suitable layout and choice of frequency. Screening may be limited by the costs involved, whereas a small alteration in the proposed frequency will probably cost nothing. In this connection due regard to the allocations for amateur transmitters is probably among the more important factors.

A low intermediate frequency naturally implies that the local oscillator frequency will be relatively close to that of the signal. Not only may this imply a poor degree of attenuation of the local oscillator e.m.f., but, for the same reason, the second-channel response may be comparable to the wanted response. Further, the number of oscillator harmonics within a given spectrum will be greater, and the corresponding number of possible spurious responses may cause trouble. Primarily, the second-channel frequencies should be carefully considered in relation to possible sources of local interference, which may often be radiation from other television receivers.

One factor to be set against a high intermediate frequency is that it is more difficult to provide a high degree of attenuation in the tuner as the frequency is increased. In this country frequencies only a few megacycles below Channel 1 have been used successfully, although trap circuits had to be fitted to some receivers in order to remove interference due to aircraft marker beacons. In addition, the use of an intermediate frequency very near to the lowest channel may prevent the use of certain types of mixer circuit, since neutralization may prove impracticable. Fortunately a spacing of 5 or 6 Mc/s is often enough, and triode mixers are very popular in the U.S.A., being neutralized on the lowest channels where necessary.

With the best of modern valves designed for intermediate frequency application, the input impedance does not prove to be an insuperable obstacle up to 50 Mc/s, although more ingenuity in the circuit design is called for in order to achieve results comparable to those attained at the lowest practical frequencies. With present-day techniques, doubling the intermediate frequency (e.g., from 16 to 34 Mc/s) usually shows a slight decrease in the available gain for a given band-width, with a small loss in skirt selectivity. Many designers have thought worth-

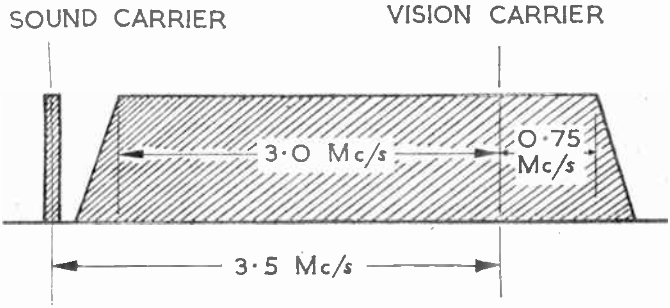


FIG. 17.—VESTIGIAL SIDEBAND TRANSMISSION CURVE.

while the expense of a few further components and the necessary valves in order to more than make good the loss occasioned by the change.

These remarks also apply to the sound channel. A 2 : 1 change in the intermediate frequency produces a rather alarming change in the maximum safe gain available from a given type of valve. Selectivity of the sound channel also calls for some care at the higher frequencies, and the additional tuned circuits provided by the use of a further stage, together with the lower gain per stage—both factors tending to sharpen the selectivity curve—are well worth the extra expense.

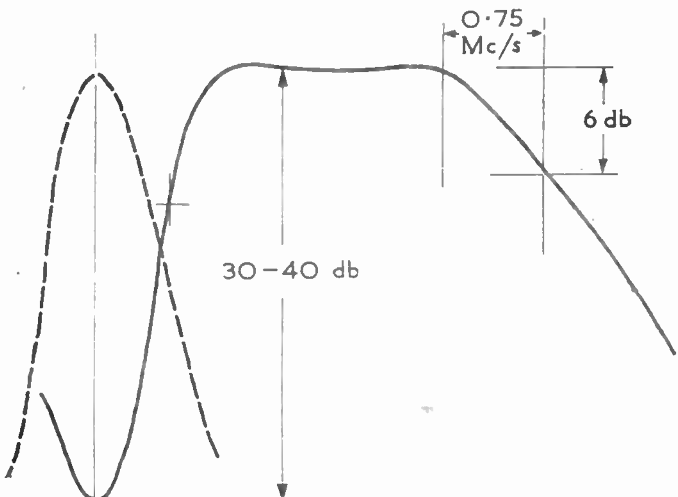


FIG. 18.—MODIFIED RESPONSE CURVE AT DETECTOR OF A VESTIGIAL SIDEBAND TRANSMISSION.

Response Curves

The vestigial-sideband system is now used almost universally for television transmission, and this discussion will, therefore, take no account of double-sideband problems.

In the vestigial-sideband system, the vision carrier with its attendant modulation components is fully transmitted, except for the higher sideband frequencies on one side of the carrier. For the British system —where the upper sideband is suppressed—a transmission characteristic results as shown in Fig. 17. Compensation for this characteristic is required in the receiver, and a suitable response curve is indicated in Fig. 18. The carrier must be 6 db down relative to the top of the curve, and the sloping portion should provide for the British system a linear fall of 8 db/Mc between the points $\pm 0{\cdot}75$ Mc/s. This condition is not easy to satisfy in a practical amplifier circuit, but the alteration of the D.C. component and accompanying low-frequency distortion caused by deviations in the region of 1 db are, fortunately, very difficult to detect.

Correct reproduction of the modulation components between 0·75 and 3·0 Mc/s depends, of course, on a reasonably level response curve. While a perfectly level curve up to 3·0 Mc/s will, in fact, produce an excellent degree of fine detail, it must be remembered that it is necessary to attenuate the response to sound intermediate-frequency signals by at least 30–40 db. The resulting sudden change at this point would produce such severe phase distortion that "ringing" would almost certainly occur. This means that all sharp transients would be reproduced in the manner indicated in Fig. 19. The compromise solution is a gradual fall off in response, making the 3·0-Mc/s point about 6 db down; this is shown in Fig. 17.

The degree of sound rejection must be sufficient to accommodate unusual reception conditions. In fringe areas the strength of the sound carrier may, for example, occasionally exceed that of the vision carrier by about 20 db. This means that 40 db should be considered the lower limit for rejection. The filters employed to ensure this condition tend to give a considerable response outside the band, and this may coincide with the position occupied by an adjacent carrier. In countries where many stations exist it has been found necessary to reduce the response at the adjacent carrier points to 60 db down, or even more if feasible.

Considering again that part of the response curve adjacent to the vision carrier, there is an alternative method of compensating for the vestigial-sideband characteristic. Either the whole or part of the slope required can be inserted after detection. Although at one time this method was thought to be difficult, it has recently been gaining popularity, and excellent results have been obtained.

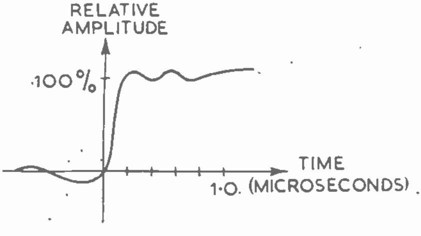


Fig. 19.—Transient Distortion.

Valves for Intermediate-frequency Amplifiers

In the majority of cases the signal at the input to the intermediate-frequency amplifier will be sufficiently large to overcome any random noise contributed by the intermediate-frequency stages themselves. Thus the most important consideration is obtaining the necessary gain and response shape as economically as possible.

The relation between obtainable gain and band-width for a given valve may be determined from the "Gain–Band-width Product" (G.B.), which is directly proportional to mutual conductance and inversely proportional to input and output capacitances.

$$\text{G.B.} = \frac{g_m}{2\pi(C_i + C_0)}$$

The above formula disregards the effects of any stray wiring, capacitances, etc. Valves for intermediate-frequency application of modern design have a slope of about 7–8 mA/volt and a total input and output capacitance of 10–12 pF, giving a figure for G.B. of 100–120. This is an improvement of approximately 30 per cent when compared with the best performance of pre-war valves.

It is also important to choose a valve which will have an adequately high input-impedance at the intermediate frequency in question. If very low it will be the major load into which the previous valve will work, and will thus unduly limit the performance. A figure of merit, which is a useful guide in this respect, is the "Gain Reference Frequency" at which $g_m \times R_i = 1$. Valves can now be obtained with the values quoted in the previous paragraph for g_m and C, and with an input resistance greater than 10,000 ohms at 50 Mc/s and having a value for the gain reference frequency $f_n = 350 - 450$ Mc/s. These figures are satisfactory for intermediate-frequency amplifier design up to 40 Mc/s.

Staggered Tuning

If a multi-stage amplifier is constructed in which the intervalve coupling networks consist of single-tuned circuits, all tuned to the centre frequency (f_0) it will be found that the band-width decreases rapidly with an increasing number of stages, and such a design soon becomes uneconomical. It would also be very difficult to obtain the type of response curve shown in Fig. 17.

This difficulty can be overcome by the principle of "flat staggering", which involves adjusting the Q of the coupling circuits to certain values, and tuning them to slightly different frequencies. This is sometimes done by dividing the amplifier into a number of similar sections having identical adjustments (i.e., pairs; two circuit sections; triplets, etc.), into a number of dissimilar sections or by applying the staggering principle to all the circuits in the amplifier to make, say, a quintuple. As there will be a slight reduction in band-width through cascading similar sections, the last method provides the maximum G.B. figure.

The design of the staggering system commences with a reference circuit which must have constants providing a 3 db loss at the band edges when tuned to f_0 and which will have a G.B. factor with the valve used of $g_m/2\pi(C_{total})$, where C_{total} is the sum of all input, output and

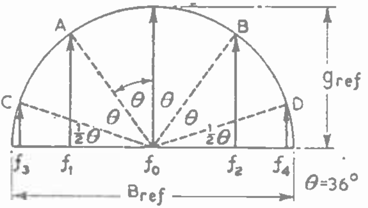


FIG. 20.—CALCULATION OF CIRCUIT IMPEDANCES AND TUNING FREQUENCIES.

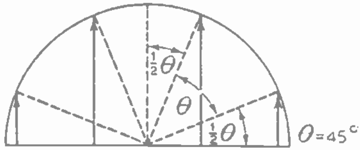


FIG. 21.—EXAMPLE WHERE EVEN NUMBER OF TUNED CIRCUITS ARE USED.

stray capacitances. All the other stages in the amplifier will have a gain equal to a stage having the reference circuit as a load, so that it is at once possible to find the number of stages required provided that the overall gain is known.

For example, with a band-width of 3·0 Mc/s at -3 db points, and a G.B. of 54 (including all stray capacitances) the available gain per stage is 54/3 or 18, and a four-stage amplifier will have an approximate gain of 10,500. This figure is somewhat higher than is usually required but neglects the loss of gain in the last stage due to the detector and the necessary matching.

The circuit impedances and tuning frequencies can be calculated with the aid of the diagram given in Fig. 20. The diameter of the semi-circle is the band-width of the reference circuit $B_{ref.}$ with the centre frequency f_0 at the centre. The circle is divided into n equal parts according to the number of tuned circuits to be incorporated in the amplifier (in Fig. 20 n is 5). The height of the central perpendicular line can be considered to represent the conductance of the reference circuit, and by proportion the values for the conductance—or impedance—of the other circuits may be calculated.

Since the angle θ is known (in this case 36°), then :

$$Z_1 = Z_2 = Z_{ref.} \times \sec\theta \quad \text{and}$$

$$f_1 = f_0 - \tfrac{1}{2}B_{ref.} \times \sin\theta \qquad f_2 = f_0 + \tfrac{1}{2}B_{ref.} \times \sin\theta$$

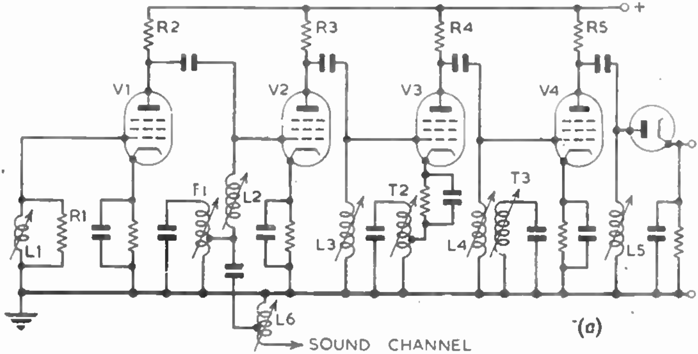
$$Z_3 = Z_4 = Z_{ref.} \times \sec 2\theta \text{ and}$$

$$f_3 = f_0 - \tfrac{1}{2}B_{ref.} \times \sin 2\theta \qquad f_4 = f_0 + \tfrac{1}{2}B_{ref.} \times \sin 2\theta$$

From the above it is seen that the greater the value of n, the larger the value of the impedance of the outside circuits. Even with a quintuple, the values for Z_3 and Z_4 with $B_{ref.}$ equal to 3·0 Mc/s will often be so large that no extra damping is required beyond the valve-input impedance. This would mean that seven circuits could not be used.

In the case where an even number of circuits are to be used—a more common condition—the reference circuit is not actually used in the amplifier as in the quintuple, although it does, of course, appear in the design in the same way (see Fig. 21).

Fig. 22 (a) and (b) shows a simplified circuit diagram of a staggered amplifier with five tuned circuits and the response curves of the circuits, separately and combined. Damping for the coils L1–L5 is obtained with the resistors R1–R5 in parallel with the valve-input impedances.


R2
R3
R4
R5
+
V1
V2
V3
V4
F1
L2
T2
T3
R1
L1
L3
L4
L5
L6
SOUND CHANNEL
(a)

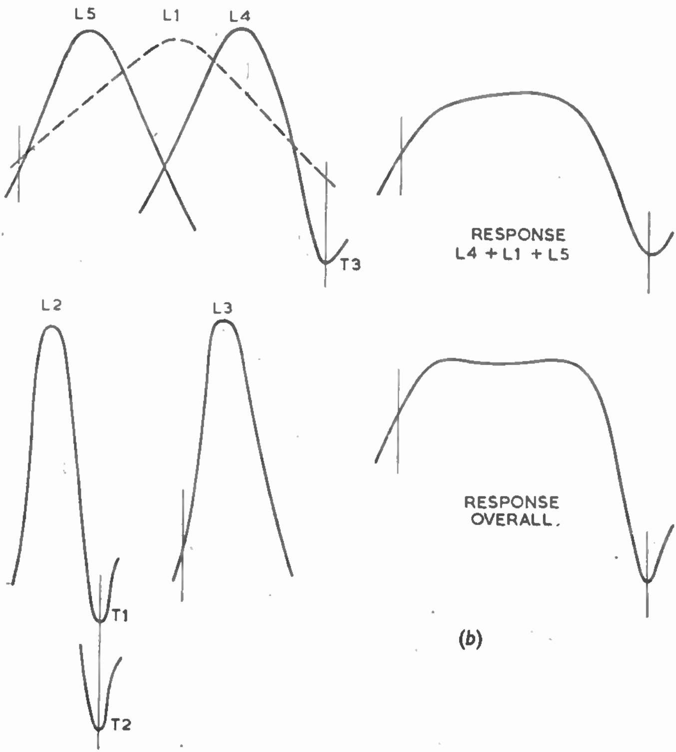

L5
L1
L4
T3
RESPONSE
L4 + L1 + L5
L2
L3
RESPONSE
OVERALL.
T1
T2
(b)

FIG. 22 (*left*).—TYPICAL T.R.F. FIVE-CIRCUIT ARRANGEMENT USING STAGGERED TUNING: (*a*) CIRCUIT; (*b*) RESPONSE CURVES.

It is probable, however, that the resistance values would, in practice, be too large to be placed in series with the anodes. In this case choke coupling with the damping resistors in the grid circuits could be used. Another method which is popular is the use of double-wound transformers. In order to obtain the tight coupling necessary to avoid a secondary resonance, the coils are wound in bifilar fashion.

The provision of the large attenuation needed at the sound-channel frequency is accomplished with trap circuits. These can be high-Q, tuned circuits carefully coupled to the signal circuits in order to give a "suck out" at the sound intermediate frequency, or placed in the cathode lead of a valve to produce degeneration at resonance. Traps T1 and T3 are examples of the first type. Q values of the order of 200 or more are desirable in order to obtain good results, and of course the maximum amount of natural attenuation due to the signal circuits themselves is of great assistance. There is no law governing the choice of frequencies for a particular circuit, but it is a good plan to tune the output circuit close to the carrier so that the grid of the last valve will not be easily overloaded.

Band-pass Filters

The G.B. factors for double-, triple- and quadruple-tuned, band-pass filters in the unequal Q-case are as follows[3]:

$$\text{Double-tuned, G.B.} = g_m/2\pi\sqrt{C_iC_0}$$
$$\text{Triple-tuned, G.B.} = g_m/2{\cdot}31\pi\sqrt{C_iC_0}$$
$$\text{Quadruple-tuned, G.B.} = g_m/2{\cdot}5\pi\sqrt{C_iC_0}$$

This shows that although the response of a single stage ca be improved with the aid of a more complex network—especially in regard to skirt response—there is a noticeable reduction in the available gain. Compared to the stagger-tuned amplifier, which had a maximum gain per stage determined by the total of the capacitances, the band-pass-coupled amplifier is dependent upon the geometric mean of the capacitances. This could lead to an advantage of 2 : 1. The G.B. factor of 54 previously quoted might reasonably be expected to rise to 75 or 80 and, except for the fact that the band-width falls off as the result of cascading, considerable economy might result.

In practice, the filters are often slightly different in each stage, and some of the advantages of staggering can be realized. Fig. 23 (*a*) and (*b*) gives a basic circuit, and shows the method of combining the responses. The remarks applying to damping and traps in connection with Fig. 22 are the same in this case.

Gain Control

It is necessary to control the gain of a television intermediate frequency amplifier in order to deal with the large variations in signal strength which the receiver will be expected to accept. In the case of manual gain-control, the slope of the valves can easily be regulated by varying the grid-bias potential, and the most common method is to

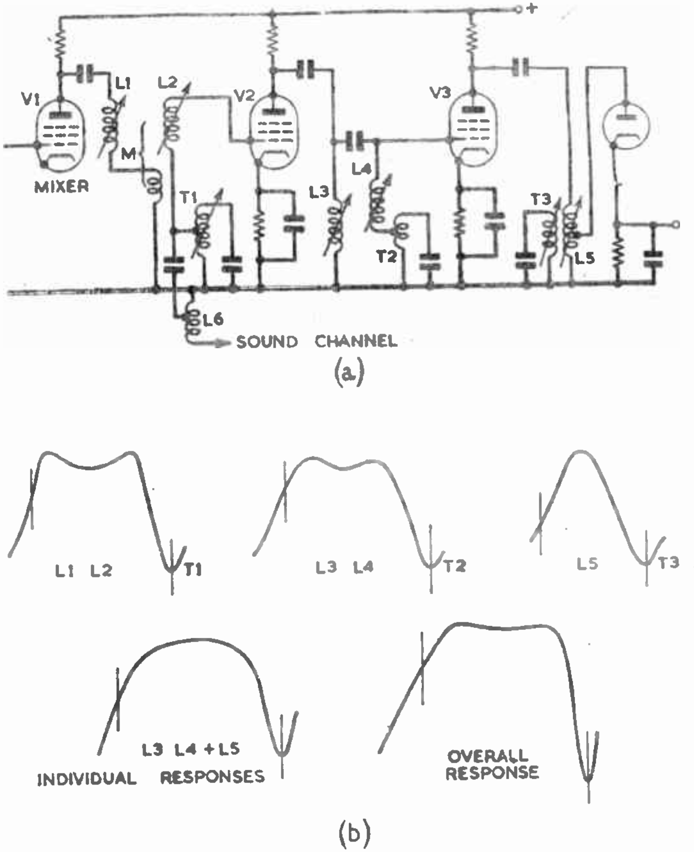


Fig. 23.—Two-valve Amplifier for a Superheterodyne Receiver: (*a*) Circuit; (*b*) Response Curves.

place a variable resistor in series with the cathode leads of the controlled stages. Unfortunately the input impedance varies rather widely with the bias voltage, and compensation has to be provided. The modern valve used in intermediate-frequency amplifiers can be compensated most satisfactorily by inserting a small resistance in series with the cathode; the value for any particular circumstances being found most easily by experiment, since the same value will not necessarily restrict the changes in input resistance and capacitance to the same extent. Unfortunately the effective slope will be reduced by 25–30 per cent, but the stability of the amplifier—as a whole—is usually improved, the effects of valve changing lessened and the nominal value of input impedance improved.

It is general practice not to control the last valve in the amplifier, as the input level is sufficiently high to demand the full grid-base being maintained at all times. The typical, high-slope pentode is not over-suited to operation near cut-off, and if variable-mu valves of comparable performance were available they would almost certainly be used. The maximum range to be expected with a short grid-base valve is not more than 30 db. It therefore becomes difficult to apply an A.G.C. circuit to, for instance, the amplifier of Fig. 22. In a case of this type the sensitivity control applied to the radio-frequency stage must be carefully set to avoid risk of overunning the intermediate-frequency control range.

Sound Take-off and Traps

Although at one time it was general practice to amplify both vision and sound signals in the first, and perhaps second, vision intermediate-frequency amplifier stages, it has been found that this arrangement does not favour producing the optimum response curve. This is because to obtain the best results at the sound frequency, the necessary sound rejection had to be produced in one or two stages. In the best of modern practice, the mixer is followed by a fairly complicated filter which provides a fairly high degree of rejection and a sound take-off point with the highest possible sound level, relative to vision. This has the advantage of reducing the sound signal in subsequent vision circuits to an extent where cross-modulation is most unlikely to occur even under the most unfavourable circumstances.

Most trap circuits giving high sound rejection are found to be tuned slightly away from the sound-carrier frequency, due to the combination resonance with the signal circuit. If the grid of the first sound intermediate-frequency stage is connected to the first sound trap, then the response curve of the sound amplifier will be slightly asymmetric, due to the off-tune, trap resonance. In many cases this is permissible, although the degree of asymmetry should not be too pronounced. If a first sound-tuned circuit is coupled to a vision circuit or trap, the coupling must be adjusted so that the sound circuit does not prejudice the degree of rejection. As a result, the coupling is usually so light that no more signal can be obtained than with the previous method.

In recent designs, use is often made of a "bridged-T" filter which is resonate at the sound I.F. and which produces a high degree of rejection by reducing the coupling between an anode circuit and the grid of the succeeding stage. Interstage coupling of this type provides a rounded top curve, useful for reducing rings and overshoots which may arise with flat-topped curves.

THE SOUND CHANNEL

The gain to be supplied by the sound intermediate-frequency amplifier depends, of course, upon the signal level at the take-off point and the detector level required. This is often a little higher than in the vision channel, and with a probable 3–6 db loss due to the take-off, the sound gain (assuming no common stages) must be at least 6 db higher than the vision channel. Further, the possible discrepancies in some locations

B B

between received carrier strengths makes it desirable to provide more gain in one channel than the other, so that with a suitable control circuit the receiver can accommodate more easily a relative variation of ±6–10 db. When the optimum gain has been achieved with the vision channel it is found that extending the sound sensitivity by a further 6 db is a considerable asset in fringe areas.

The band-width of the sound channel must be sufficient to allow a small amount of oscillator drift (say, ±50 kc/s), since this problem cannot otherwise be overcome. Selectivity—in the general sense of radio practice—is not required, since the number of interfering signals in the vicinity are relatively few and are separated by at least 1·5 Mc/s. An increase of band-width beyond 50 kc/s is certainly possible, and has been found very desirable from the following points of view :

(*a*) oscillator drift can be ignored if kept within the above limits;

(*b*) the wider the sound channel, the better the response to transient interference, and thus better limiting is possible;

(*c*) tuning is far less critical, and the oscillator may be adjusted to favour the picture reproduction.

In the discussion on the vision channel, it was shown that four valves having a G.B. factor of 54, including stray capacitances, could produce a gain of 10,500 with a 3·0-Mc/s band-width. If each sound stage were to have a band-width of about 0·5 Mc/s, then the stage gain would be 100, and it appears at first glance that the sound channel need only consist of two or three stages. In practice, television high-slope pentodes possess sufficient anode-grid capacitance to cause instability if such a stage gain were attempted at the higher intermediate frequencies. It is common therefore to find the same number of valves in the sound and vision channels, and the band-width may often be greater than 500 kc/s, even after allowing for the shrinkage due to circuits in cascade. An overall gain figure as high as 10,000 is seldom found, since the tuner gain is usually higher than would be indicated by this figure.

Coupling Circuits

The coupling networks in the sound channel need be only of the simplest type, and the bifilar coil system, already mentioned, is often used. This design has the advantage of making it possible to maintain a low impedance in the grid and anode circuits. Choke and capacitor coupling are a little more expensive, while with resistance–capacitance coupling to the tuned circuit, the resistor should never be placed in the grid return. Even quite small time-constants will cause blocking in the presence of impulsive interference, and this will sharply reduce the effectiveness of any type of interference suppressor.

Sound A.G.C.

The standard method of automatic gain control found in radio receivers is now almost universally found in television sound channels, although delay circuits are occasionally dispensed with. As will be shown in the discussion on detectors, the value of the load is kept very small, and with it the impedance of the A.G.C. line. This is primarily to overcome blocking due to interference.

Audio Output

Although no special precautions need be taken with the audio stage of a television receiver, it is worth noting that the transmitter is modulated by audio frequencies higher than 10 kc/s, so that substantially better quality than is customary with medium-wave radio reception is possible. Owing to the severe economic limitations imposed upon the more popular television receiver designs, it is true to say that this advantage is seldom, if ever, realized.

Large output powers are not generally required, so that good use can be made of negative feedback to produce a satisfactory response. In this connection many sound-interference suppressors are preferably divorced from the feedback loop if satisfactory operation is to be achieved without considerable complication.

DETECTION

The diode detector has been almost universally used in radio receivers for some considerable time and—except for a few early designs—the diode is used in both vision and sound channels in television receivers. During the last few years the germanium diode has become very popular, this being mainly due to the fact that it requires no heater supply, is small and convenient to use, and has lower self-capacitances. On the other hand, it has a permanent "back-resistance", and thereby is entirely ruled out in certain special applications. For the general-purpose detector in a television receiver, it has been found necessary to specify carefully the type of crystal in terms of detection efficiency in order to achieve results comparable in consistency to those of the standard valve diode.

The Video Detector

In the vision channel it is important to maintain reasonable efficiency in order that the last valve in the intermediate-frequency amplifier shall not be required to provide too much output and that a large amount of gain is not lost. Because of the particular circuit difficulties, efficiencies of the order of 50–60 per cent are considered satisfactory. It is also important to prevent the loss of the higher modulating frequencies in any intermediate-frequency filter following the detector, and to keep the tuning and response of the last intermediate-frequency circuit free from any variation due to the detector.

Fig. 24 (*a*) and Fig. 24 (*b*) show two typical video detector circuits: valve diodes could, of course, be used instead of crystal diodes if desired. While there are many possible connections (for example, the earth point may be moved round, or either positive or negative output may be required), the shunt circuit is seldom used and Fig. 24 (*a*) is the more common. It will be seen that the filter for attenuating the intermediate frequency has been terminated in the load resistance. This is in series with a peaking coil which improves the video response. The filter frequently has two sections, if necessary, because the initial capacitor in series with the tuned circuit and the crystal diode cannot—in the interest of good video response—be made large even compared to the capacitance of the diode.

The small value of this capacitance is one of the primary reasons for low efficiency: another is the internal resistance of the diode itself.

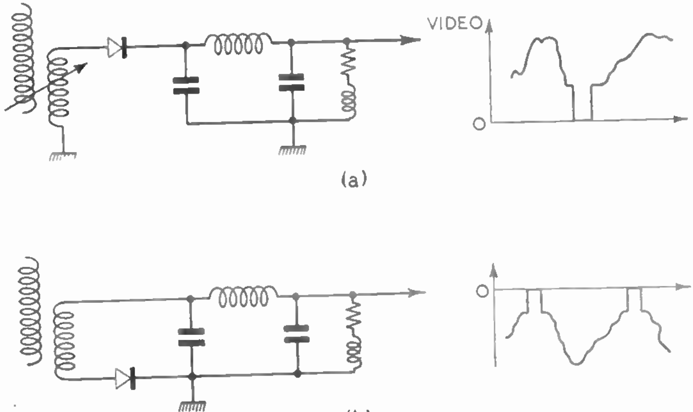


FIG. 24.—TYPICAL VIDEO DETECTOR CIRCUITS.

Values of load resistor above about 5,000 ohms are not practicable for video detection, so that thermionic and germanium diodes are specially made with forward resistances between 100 and 300 ohms. Such a germanium diode suffers from a very low back resistance, but this disadvantage is offset by the low self-capacitance. The damping of the tuned circuit is dependent upon the diode load and the efficiency. It can be expressed :

$$R_d = \frac{R_1}{2\eta}$$

where R_1 is the load resistance and η the efficiency.

This impedance damps the tuned circuit between the valve and the detector; and since it is usually too low for the requirements of the amplifier, the detector has to be coupled to a tap on the coil. The ratio obtainable by this means, without loss of an objectionable nature, is seldom greater than 2 : 1. The alternative, if a greater ratio is required, is to add capacitance. The germanium diode tends to exhibit a slight capacitance variation with different input levels; the transformation, together with any additional capacitance, prevents any effect upon the circuit tuning.

The germanium diode has a more linear voltage–current characteristic at low levels than the thermionic diode. This is a considerable advantage, since non-linearity affects the rectification efficiency and alters the value of R_d. When using a thermionic diode it is often considered advisable to add a fixed damping resistance so that the circuit response does not depend solely upon the diode.

The detector can, of course, be coupled to the last intermediate-frequency stage by means of a band-pass filter. The damping may be such that it is difficult to design a suitable filter, although in practice many receivers have used such a circuit. In one receiver where the detector (a germanium diode) was fed from a quadruple-tuned filter, it was found that the crystal characteristics were too variable for the response curve to be relied upon without very carefully checking the crystals in advance.

Apart from susceptibility to damage by overheating (particularly when soldering) and from the passage of high currents, the germanium diode has proved to be very satisfactory and has, apparently, a very long life.

The Sound Detector

In television practice the problems associated with the sound detector are very similar to those encountered in the vision detector. In order to preserve reasonable band-width—so that undue lengthening of interference pulses will not occur—the load resistance is proportioned accordingly; values between 10,000 and 20,000 ohms are common. To restrict the radiation of the intermediate frequency from the output, a filter of the type used with the vision detector (but without the peaking coil) is customary. The severe problems of self-capacitance, diode internal resistance and linearity of characteristic are less troublesome, since the detection efficiency is higher. A "general-purpose" version of the germanium diode can therefore be used, while a thermionic diode of the low-impedance type is also very satisfactory.

In order to achieve the maximum gain from the last sound intermediate-frequency stage, the detector may be tapped down the intermediate-frequency coil. Again, it has been found that ratios greater than 2 : 1 or 3 : 1 can lead to loss of efficiency, especially when a bifilar-wound transformer is used.

RADIO-FREQUENCY AND INTERMEDIATE-FREQUENCY MEASUREMENTS

While the subject of measurement is discussed at considerable length in many books dealing with electronics, less attention has been given to describing some of the methods used for receiver testing. Television-receiver design and manufacture is now at such an advanced state that if all forms of production testing were included a work of great length could be written. This section will, therefore, deal with certain aspects only as they affect the design engineer.

Alignment

SIGNAL GENERATOR METHOD. The correct alignment of narrow- and wide-band amplifiers can well be one of the most important factors in the design, production and maintenance of television receivers. The time-honoured method involves the use of an output meter and a signal generator possessing accurate frequency calibration, low-leakage and a suitable attenuator. With a single-peaked amplifier (e.g., the sound channel) the generator is set to the centre frequency, and all the circuits are tuned for maximum output. The attenuator is adjusted as necessary; and on completion of the adjustments the readings of attenuation and output provide a measure of the sensitivity of the amplifier.

The wide-band amplifier is rather more complicated; in the stagger-tuned type, for example, there may be a separate frequency for each circuit. If these frequencies are known, then each circuit may be dealt with in turn. For experimental alignment, it is customary to connect the generator to the grid of each stage in turn, working backwards from the output stage. It is then possible to gain some idea as to whether or not feedback is present; this is because the tuning is likely to be altered as a result of feedback. When connected to the input, the generator may be adjusted to various frequencies within and without the pass-band and a response curve plotted. This is usually in the form of the input signal required to produce a certain output figure.

In production and maintenance the generator may usually be connected to the input of the amplifier throughout. The alternative to a signal generator is a unit containing oscillators operating at the required frequencies and producing equal output levels. They are then switched into circuit in the order required.

THE SWEEP GENERATOR. The wide-band amplifier with band-pass filters may be more difficult to adjust with the means described above, and a type of signal generator sometimes known as the "wobbulator" is found to be a great advantage. In this instrument the frequency of the oscillator is continually swept through the passband in a sawtooth or sinusoidal manner with respect to time. By connecting a synchronized oscilloscope to the amplifier output it is possible to show the exact shape of the response curve (to a linear and *not* to a logarithmic scale), and the result of every adjustment is clearly seen. In order to have some idea of the exact frequency at one point on the curve, a marker "pip" is necessary. Some commercially built sweep generators incorporate this facility, but if not, the simplest method is to loosely couple a signal generator to the sweep-generator output. A beat then occurs when the swept frequency coincides with the fixed frequency, and this can be seen as a small "blip" on the oscilloscope trace. Reducing the band-width of the oscilloscope amplifier to about 20–30 kc/s will make the marker beat occupy a smaller portion of the trace, and thereby allow greater accuracy. In production set-ups the marker is often supplied from a crystal oscillator.

The sweep generator displays the conditions in the amplifier so clearly that almost any investigation is greatly facilitated. The method is also applicable to video amplifiers, providing the generator is capable of covering the range.

Noise Measurements

Absolute noise measurements are best made with the aid of a proper noise diode, with the appropriate supplies connected to the input of the amplifier. This process is liable to be difficult or rather tedious when only comparative results are needed, and a signal generator capable of being 100 per cent modulated by a square wave will provide the necessary information. An oscilloscope, having a known band-width, is connected to the detector output, and the signal generator—set to the carrier frequency—is adjusted so that the square wave seen on the oscilloscope appears as in Fig. 25. The negative-going peaks of noise on the top of the square wave must just lie on the same level as the positive peaks from the base-line. The setting of the signal-generator attenuator is then taken. This test can be carried out rapidly, and

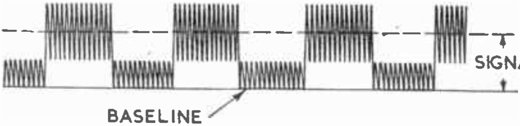


FIG. 25.—SQUARE-WAVE ADJUSTMENT FOR NOISE MEASUREMENTS.

will be reasonably accurate provided that the characteristics of the signal generators and oscillators used (if more than one) are known.

Oscillator Drift Measurement

Since it is desirable to prevent the local oscillator drifting by more than about ±50 kc/s in 200 Mc/s, care must be taken in choosing the method of measurement. Essentially a stable frequency is required for comparison with the oscillator; if an accurate crystal-controlled source is not available, it is perhaps wise to tune to a station and measure the drift of the sound intermediate frequency with a good heterodyne wavemeter. All the measuring equipment should be allowed to warm up for an hour or more before the test is commenced.

When examining the effect of various components in the oscillator circuit it is a good plan to mount the oscillator in a well-lagged oven and to stabilize the temperature over a long period. The temperature can then be altered by any given amount and the resulting drift measured.

VIDEO AMPLIFICATION

In certain pre-war receiver designs the cathode-ray-tube modulation potential was obtained directly from the detector. This was largely made possible by the fact that some of the picture tubes then made could be fully modulated by a signal of about 15 volts peak-to-peak. Since then the trends in cathode-ray-tube design have called for higher voltage levels, and even a detector of the anode-bend type would be difficult to design to give with any factor of safety an output of 50 volts peak-to-peak. It has therefore become customary to interpose a video amplifier having one or two stages between the cathode-ray tube and detector.

Design Considerations

In the case of the single-stage amplifier, the gain and band-width depend upon the "gain–band-width" (G.B.) factor as mentioned in the section on intermediate-frequency amplifiers. Without compensation, the gain will fall by 3 db when the reactive part of the load becomes equal to the resistive part. Unfortunately the stray capacitances encountered (due to the fact that it is seldom possible to mount the amplifier close to the base of the cathode-ray-tube) are often several times the value of the valve-output capacitance, and a gain of more than 5–8 times is difficult to obtain with normal intermediate-frequency-type pentode valves.

With the British system, the band-width over which there should be negligible loss of gain is 3·0 Mc/s, and with an assumed total capacitance of 20 pF the load resistance for a 3-db loss at the highest frequency will

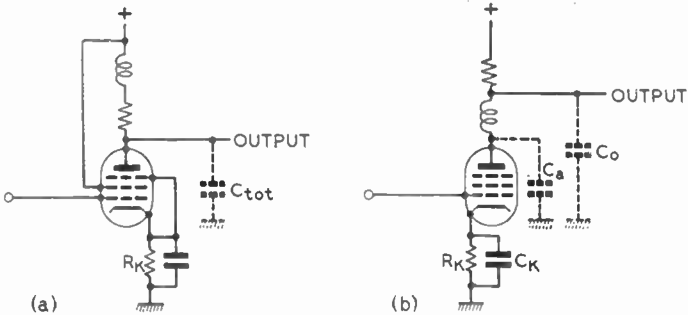


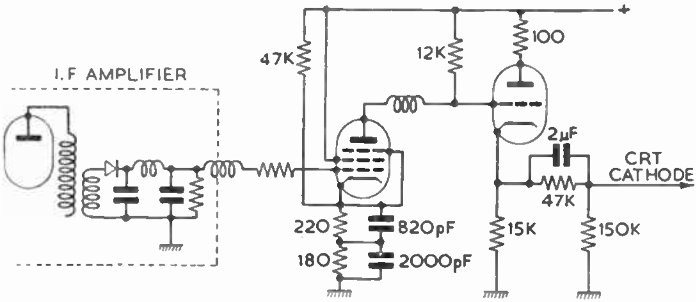
Fig. 26.—Video Amplifier Circuits.

be about 2,700 ohms. For an output of 50 volts peak-to-peak nearly 20 mA peak-to-peak would have to be provided by the valve; this is impracticable with an ordinary intermediate-frequency pentode without using the extreme width of the grid base and accepting the ensuing non-linearity. This difficulty can be overcome by inserting resistance in the cathode lead. The resulting degeneration, while lowering the gain, linearizes the valve characteristic; and shunting the resistance with capacitance affords a rising gain characteristic. Beyond the point where $R_k = 1/\omega C_k$ it becomes difficult to compensate for the time-constant in the anode circuit without inductance. By these means, and with inductive compensation in the anode circuit, it becomes possible to raise the anode load to a value perhaps twice that mentioned above: the swing required is then well within the capabilities of the valve.

The typical circuit will appear as in Fig. 26 (*a*) or (*b*). Best results are obtained with the value of R_k approximately equal to two or three times $1/g_m$ and the anode compensation of Fig. 26 (*b*). With a typical pentode, having g_m equal to 7·0 mA/volt, with an anode load of about 4·7 kΩ (about the maximum workable value) and a cathode resistor of 330 ohms, the gain would be of the order of 10 : 1.

In practice, the stray capacitance into which the video amplifier has to work may well be somewhat variable, and to allow an adequate safety factor the band-width of the amplifier may have to be increased. Further, if the capacitance is variable, accurate compensation will not be possible. In a two-stage amplifier the second stage need not contribute to the gain at all (e.g., a cathode follower), but it will have served its purpose if the majority of the stray capacitance can be isolated from the anode load and compensation network. The use of a cathode follower places such a small capacitance load on the amplifier that the resistance can be increased to 10–15,000 ohms, and gains of 25 can be reached. This arrangement makes it possible to correct for the vestigial-sideband characteristic in the video amplifier. Fig. 27 illustrates a circuit together with the frequency response obtained.

If the second valve in a two-stage circuit is made an amplifier, then negative feedback may be incorporated; this again provides freedom from the effects of possible variations in the stray capacitances. It has



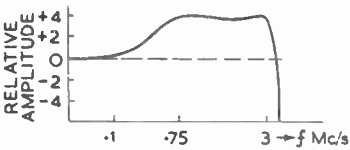


Fig. 27.—Two-stage Video Amplifier Using a Cathode Follower.

the disadvantage that it is no longer easy to D.C. couple the stages, and the D.C. component will have to be reinserted later, or left out altogether. This point will be discussed below under " D.C. Component ".

The D.C. Component

The vision carrier is D.C. modulated, i.e., in the British system, the strength at any instant is proportional to the brightness of the transmitted image except during the periods reserved for synchronizing information. During the horizontal synchronizing pulse the carrier is zero, and following this at 30 per cent of its "peak white value" for a brief period known as the " back porch ". This level corresponds to black, and is kept as stable as possible at the transmitter. In the receiver it is important, if the picture is to be a faithful reproduction, to keep the black level at a potential corresponding to the cathode-ray-tube beam cut-off whatever the signal amplitude. In practice, the signal amplitude is set and kept constant by the gain-control circuit, while the black level is adjusted by variation of the cathode-ray-tube bias. Between the detector, which provides an output voltage depending directly upon the carrier strength, no coupling capacitors are used, or the waveform would establish itself at its mean level. This would cause the bright and dark scenes to appear at a constant brightness but with varying modulation. The black level can be reinserted following an A.C.-connected amplifier by means of a " D.C. restorer " or a " clamp circuit ". The D.C. restorer tends to be unsatisfactory when large and rapid changes occur in the picture content; this is due to the fact that the time constant of the coupling circuit must be long compared with the lowest frequency components transferred. The clamp arrangement can be ideal, and is used extensively in transmission equipment, but is very costly for receiver applications; also, it has a tendency towards unsatisfactory operation in the presence of impulsive interference. The D.C. connected amplifier has therefore become the most popular design in this country.

Reference to Fig. 27 will, however, show that the coupling circuit between the cathode-ray tube and the cathode follower will involve a loss of the D.C. component. This is very common, an approximate D.C. loss of 30 per cent being used. This is incorporated for the purpose of overcoming certain problems encountered when the full D.C. is applied to the picture tube. In the first place the regulation of the high-voltage power supply used in the typical receiver is seldom better than 10 per cent, and the rapid change from an almost black picture to an almost white picture will produce objectionable effects such as changing picture size, flutter, etc. Without vision automatic gain control the changes in signal strength in the fringe area also cause trouble if the full D.C. component is used, since the brightness control will have to be readjusted continually. For these, and other reasons, it has been found best to apply between 50 and 75 per cent D.C.

SYNCHRONIZING CIRCUITS

The synchronizing circuit of a receiver has several functions to perform. First the synchronizing pulses must be separated from the composite waveform, and the vertical pulses separated from the horizontal pulses. The time-bases then have to be locked to the synchronizing information so that the retrace occurs immediately following the commencement of the appropriate pulse. This latter function can be fulfilled by triggering—here the actual pulse is coupled to the time-base and initiates the flyback—or as in the various flywheel sychronizing systems in which the speed or phase of the time-base is compared with the synchronizing pulses in a discriminator, and a control signal developed which corrects the speed when necessary.

The Synchronizing Separator

Removing the synchronizing pulses from the composite waveform is a matter of amplitude discrimination with reference to the base-line, and is simply accomplished by the circuit of Fig. 28. The video waveform is fed to the grid of a valve (preferably a pentode) by means of a coupling capacitor and a grid-leak which is returned to cathode. The waveform must be in the direction shown in the diagram in order that the tips of the synchronizing pulses cause grid current to flow; the signal is thus restored to the cathode potential. By restricting the screen voltage, the grid-base of the valve can be shortened so that only the synchronizing pulses lie above cut-off and cause plate current to flow. The output obtainable then depends upon: (1) the screen voltage (which is fixed according to the required grid base; (2) the anode load; and (3) the anode supply voltage should this be comparable to the knee voltage of the valve.

The anode load of a synchronizing separator must be chosen with some reference to band-width; and for the rise time of the horizontal pulses to be reasonable, a value higher than 10,000 ohms is unsatisfactory. The anode voltage may, with advantage, be chosen so that the knee of the anode voltage current characteristic is reached at

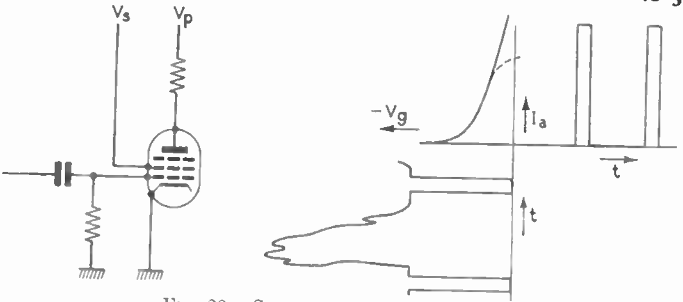


FIG. 28.—SYNCHRONIZING SEPARATOR.

the pulse tips : this is indicated in Fig. 28 by the dotted line. This causes any noise which may be present to be limited and, to a certain extent, removes any imperfections due to poor restoration. With the type of valve used in the intermediate-frequency amplifier, good separation may be achieved down to a video level of 20 volts peak-to-peak, provided an output of 40–50 volts across 10 kΩ is satisfactory. Two-valve synchronizing separators are occasionally used when the conditions are more stringent. As their design depends to a large degree upon the particular conditions involved, generalization is impracticable.

Field Synchronizing Pulse Separation

In a receiver operating on an interlaced vertical scanning system it is very important to separate the vertical (field) synchronizing pulses entirely from the horizontal (line) pulses so that the vertical time-base synchronizes correctly at the half-line point every alternate retrace.

Separation of the vertical pulses can be effected with a resistance–capacitance integrator as shown in Fig. 29. The complete synchronizing waveform is fed to the input " A " of the integration circuit. The charge due to the horizontal pulses at the output " B " is small, and leaks away entirely by the time the next pulse arrives. During the vertical pulse period, however, the charge time is long compared to the discharge time, and the voltage rises steadily with each pulse until the horizontal pulses occur again, when, after a few line periods, it falls to the mean level. Unfortunately there is a difference between alternate fields, since one vertical pulse arrives at the end of a line, and the next at the centre of a line. The capacitor will not be completely discharged at the centre of the line so that in every alternate field the integrated vertical pulse will be larger by this amount. The effect is to cause false initiation of the retrace, and perfect interlace is not achieved. In the American and C.C.I.R. television systems the vertical pulses are preceded and followed by a series of " equalizing " pulses of short duration and at half-line intervals. These pulses allow the same charge to be reached at the commencement of each vertical period, and the interlace is not disturbed.

In order to overcome the lack of " equalizing " pulses in the British system, a non-linear integrator has been used, the basic circuit of which is shown in Fig. 30. From the anode of the synchronizing separator a resistance-capacitance integrator is connected to ground via a diode or

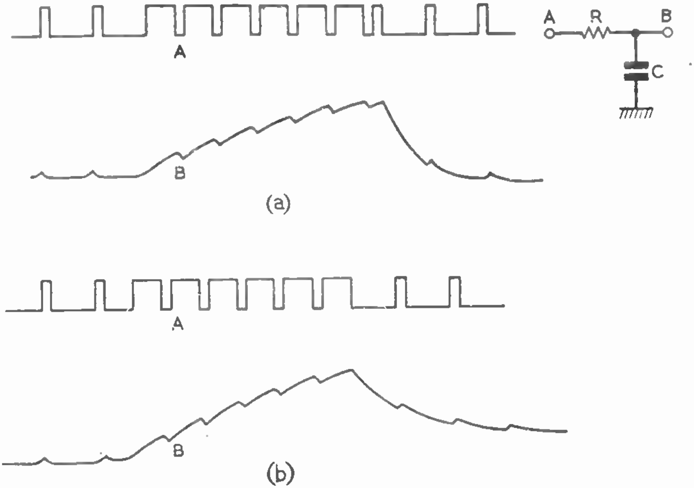


FIG. 29.—RESISTANCE-CAPACITANCE INTEGRATOR.

semi-conductor. In the absence of signal a small steady current flows through the diode and the resistance to ground, and due to the low impedance of the diode most of the anode–ground potential appears across the resistance and capacitor. When the signal is applied, the diode tends to cut off during the short horizontal pulses, and the voltage across the time-constant falls slowly; only to be quickly restored to the original value when the pulse terminates, and the diode becomes conducting again. During the vertical pulse period, the diode is cut off for the majority of the time, and the waveform appears across RC. The virtue of the circuit is that the charge is quickly restored after the horizontal pulses, and that the shape and amplitude of the alternate vertical pulse trains appearing at the output can be identical. To remove the residual horizontal information, this circuit is usually followed by a clipper, and by a slight amount of additional integration.

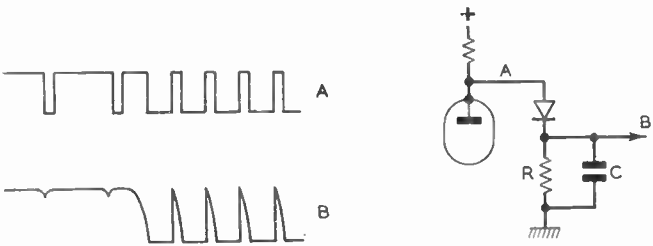


FIG. 30.—NON-LINEAR INTEGRATOR.

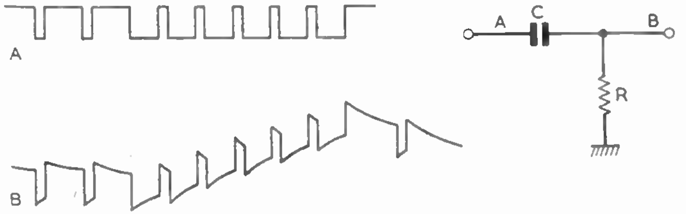


Fig. 31.—Effect of Differentiation.

The vertical pulses can also be separated by differentiation. In Fig. 31 the effect of a differentiating circuit of suitable time-constant is seen : the raised portion of the waveform would, of course, have to be removed with the aid of a clipper. Provided that clipping is performed properly, poor interlace is not likely to occur ; this is because the retrace is initiated by the sharp edge of the first pulse.

Triggered Synchronizing of the Time-bases

Some points in connection with the coupling of the synchronizing circuits to the time-base oscillators are worth noting. First there must not be appreciable coupling between the oscillators. Fig. 32 shows one form of the circuits already discussed where the oscillators are con-

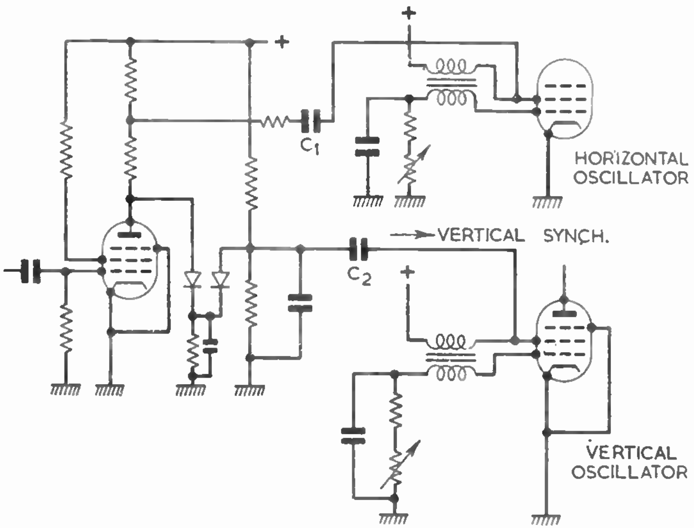


Fig. 32.—Arrangement of Circuits for Triggered Synchronization.

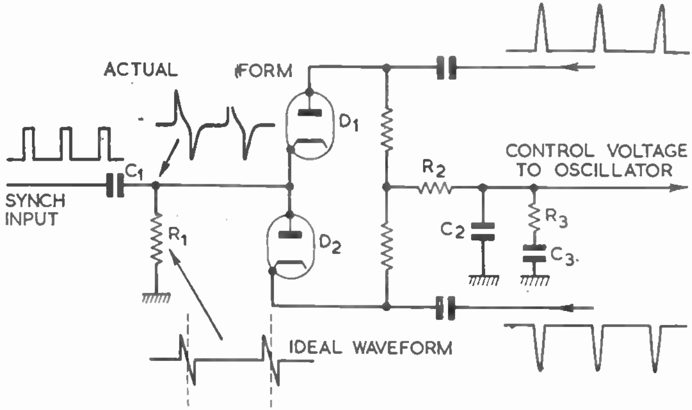


FIG. 33.—TYPICAL AUTOMATIC PHASE-CONTROL DISCRIMINATION CIRCUIT.

nected via the capacitors C1 and C2. C1 is usually chosen so that a tendency toward differentiation occurs, and the leading edge of the horizontal synchronizing pulse is accentuated. Sometimes a resistance is placed in series and/or the anode load is tapped so that the oscillator voltage is restricted as far as possible from entering the vertical synchronizing circuit. If the pulse generated by the vertical oscillator interferes with the horizontal oscillator, the latter may synchronize in an unstable manner during the vertical retrace period and will also—and this is objectionable—take some time to re-establish its proper timing. In this respect the two diodes in the vertical synchronizing circuit shown are most helpful, and the impedance of the potentiometer network associated with the clipper should preferably be as low as possible. Where multi-vibrators are used instead of blocking oscillators, the cathode-coupled circuit has the advantage of allowing synchronization to occur without feedback from the oscillator appearing in the synchronizing circuits.

"Flywheel" Synchronization

As already mentioned, the basis of flywheel synchronization is an oscillator whose frequency or phase is compared with the synchronizing pulse waveform in some form of discriminator which produces a control potential to correct any error that may exist. Many practical circuits have been developed, the majority of which are most correctly described "Automatic Phase Control Circuits". The design of these circuits can be very complicated, and in order to avoid devoting an undue amount of space to this subject two commonly-used circuits will be described briefly.

The first circuit, shown in Fig. 33, demands an input of positive-going synchronizing pulses to enable the control to be applied to the first grid of a cathode-coupled multi-vibrator; in this case the speed will be increased by a negative change in the bias. The synchronizing waveform is first differentiated, the time-constant, C1–R1, being chosen so that the nearest approximation to the ideal waveform shown is achieved. If the polarity of the input is negative, a differentiating transformer can be used and a change of polarity achieved at the same time, but the ideal waveform will be more difficult to obtain in this way.

The time-base is required to provide an output to the discriminator of short pulses of positive and negative polarity, having approximately equal amplitude. This is a simple matter with a transformer-coupled, magnetic-scanning system, since such pulses are produced during the flyback period. To obtain pulses in balance, a special winding may be added to the transformer or a small balancing transformer connected to one of the normal transformer taps.

The positive " gating " pulses applied to the anode of D1 and the negative pulses applied to the cathode of D2 cause both diodes to conduct heavily until a charge nearly equal to the pulse amplitudes is built up across the coupling capacitors, whereupon the diodes will be entirely cut off except at the tips of the pulses. The current due to the conduction flows through the coupling resistors, with the result that if the resistor values, capacitor values and the pulse amplitudes are all balanced, the potential existing at the input at the instant of conduction will appear at the junction of the resistors. If it is now assumed that the conduction, due to the peaks of the gating pulses, occurs at the exact time denoted by the dotted lines through the centre of the differentiated synchronizing pulse in the case of the ideal waveform, then the potential at the junction of the two coupling resistors will coincide with the mean potential of the input waveform. If, however, the gating pulses arrive a little early, they will coincide with a section of the input waveform more positive than the mean level, and the output potential fed to the oscillator will be positive; causing the speed to fall and the gating pulses once more to coincide with the centre of the input pulse-train.

Since the information from the discriminator comes only in very short bursts, the output of the discriminator is integrated by the time constant R2–C2, the value chosen being usually equivalent to several lines. During the vertical synchronizing pulse interval, the timing of the differentiated synchronizing pulses and the gating pulses cannot coincide and, though the gate pulses appear at a time when the input is roughly at mean potential, there is usually some disturbance of the horizontal oscillator timing. After this disturbance the phase correction effected by the control voltage may need to be considerable; and a state known as " hunting " may occur in which the timing varies about the correct position in an oscillatory manner and may never reach the desired state. For this reason the damping circuit consisting of a critical time-constant R3–C3 is connected in parallel with C2.

The main advantage of the circuit described above is that, providing the gate pulses are larger in amplitude than the input waveform, the discriminator does not operate in any way between successive synchronizing pulses. Thus the effect of impulsive interference is extremely small. In practice, the advantage over triggered synchronizing is considerable.

The second circuit, Fig. 34, operates in a similar way except for the fact that the gating is effected by the synchronizing pulses, fed from the phase-splitter V1, and the reference waveform is provided by the time-base, in the form of a sawtooth waveform. The method of operation is the same as in Fig. 33. The potential existing at the junction of the

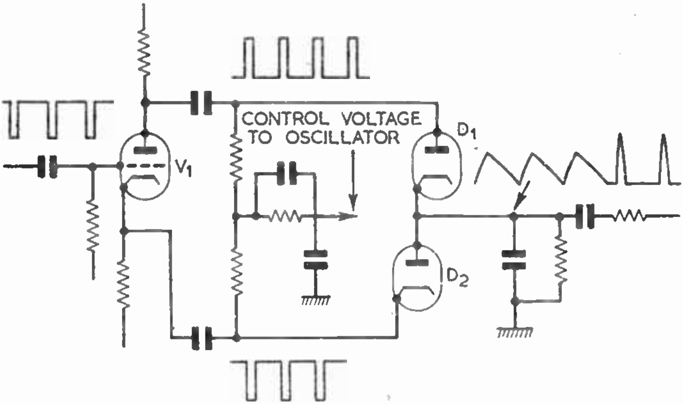


FIG. 34.—AN ALTERNATIVE CIRCUIT FOR AUTOMATIC PHASE CONTROL.

diodes—a small part of the short flank of the sawtooth—is seen at the junction of the coupling resistors due to the current drawn by the diodes. In this circuit, however, any interference pulse coming from the synchronizing separator, if comparable in amplitude to the synchronizing pulses, will cause the diodes to conduct, and a false signal will be applied to the oscillator. In this arrangement an alternative circuit is shown for the smoothing of the control voltage; and this performs the same function as R2–C2, R3–C3 in Fig. 33.

With such a control system, if the best immunity is to be obtained from impulsive interference, it is desirable to use an oscillator of the "flywheel" type, i.e., one which tends to run at the right speed in the temporary absence of control. Oscillators are dealt with in detail later, but one particular point should be mentioned here. If the speed of the oscillator is made to depart from the correct value (for instance, by operating the speed control), the control circuit will correct for the maladjustment until the limit of its capabilities are reached. When the oscillator is manually returned to the original speed—assuming that the synchronization finally failed—it will be found that the control circuit does not cause the oscillator to lock until very near the correct frequency. The difference between the correct frequency and that at which the oscillator can be locked is known as the "pull-in" range, and is determined by: (1) the frequency control voltage characteristic of the oscillator; (2) the signal amplitudes in the discriminator; (3) the smoothing-time constant; and (4) the flywheel characteristics of the oscillator. The narrower the "pull-in" range, the greater the freedom from interference, but, on the other hand, the circuit will be more liable to lose synchronism due to a slight alteration in the horizontal synchronizing frequency at the transmitter, or to drift and the like, in the receiver. In the U.S.A. values of 100–200 c/s have been widely used, but in this country a number of manufacturers have considered 800 c/s suitable. This is due to the fact that the mains frequency is subject to alteration, and this, in turn, alters the synchronizing pulses at the transmitter.

The two circuits so far mentioned are unsuitable for controlling a line oscillator of the cathode-coupled multi-vibrator type. Where the line generator is a blocking oscillator, however, the control voltage is applied via a reactance valve. A typical circuit is shown in Fig. 35, the control voltage being derived from a discriminator circuit as shown in Fig. 33.

A system using a different type of discriminator is shown in Fig. 36, and has also been used in a number of models. In this, a pentode, V2, acts as coincidence detector. V1 is biased so as to pass on, amplified, only the negative portions of the differentiated synchronizing pulses. These are applied to the screen grid of V2, a series of pulses corresponding to the line flyback period being fed to the control grid. The mean anode current of V2 varies with the phase difference between these two sets of pulses, a control voltage being developed at the anode to control an oscillator, V3 and V4, of the cathode-coupled multi-vibrator variety.

The cycle of operation of this system is:

(1) line-oscillator frequency falls;
(2) line-scanning time increases;
(3) line-flyback pulse is delayed;
(4) degree of coincidence between flyback pulse and synchronizing detector is reduced;
(5) anode current of coincidence detector falls;
(6) anode voltage of coincidence detect[illegible]ises;

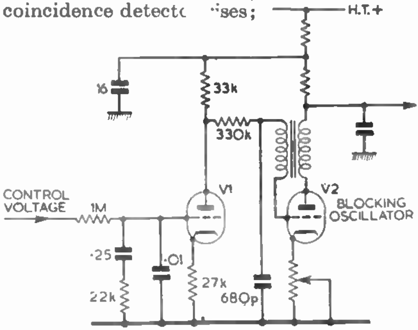
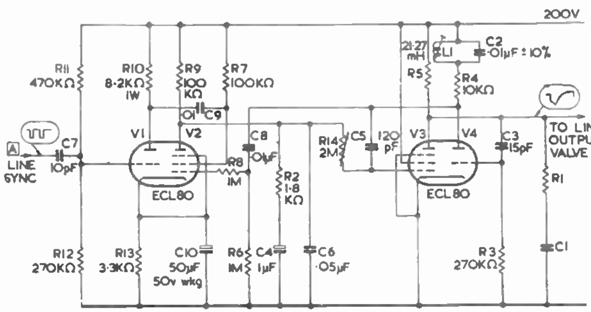


FIG. 35.—REACTANCE VALVE (V1) CONTROLLING A BLOCKING OSCILLATOR (V2).

FIG. 36 (*below*).—USE OF A COINCIDENCE DETECTOR. (*Mullard Ltd.*)



(7) line-scanning time falls;
(8) line-oscillator frequency rises.

For an initial rise in line-oscillator frequency, the converse of the above stages occurs.

A simpler form of flywheel synchronization that has recently become popular is shown in Fig. 37. This circuit is generally designed around a single double-triode valve such as the ECC82 or 12AU7. V1b forms a saw-tooth generator, the frequency being roughly determined by L1–L2. The winding L3, contained in the same special can, which in appearance resembles an I.F. transformer, is resonated by the 0·015-μF condenser to the line frequency and is excited into sinusoidal operation by mutual inductive coupling, and this helps to stabilize the oscillator frequency. The sine wave is superimposed on the normal wave-form at the grid of the oscillator, and the resulting wave is applied to the grid of V1A, which forms a control or "comparator" valve. This control valve is biased so that only the tips of the sinusoidal peaks are conducted. Simultaneously, to the cathode of this valve are applied the line-synchronizing pulses, of negative polarity. The composite pulse, consisting of the peaks of the sinusoidal line-frequency element and the amplified line-synchronizing pulses, appears in the cathode circuit. The duration of the composite pulse, and hence the average current passing, depends upon the phase relationship of the two waveforms, and by integrating the current pulses a potential is built up across the 0·01-μF condenser proportional to the duration of the pulses—a longer pulse being passed by V1A when the time-base generator is running slow compared with the incoming synchronizing pulses, and vice versa. This potential is then used to control the speed of V1B by varying its bias. The 100k, 0·01-μF and 4·7k, 0·5-μF components form an "anti-hunting" network by providing a long time-constant in the cathode of V1A. A moderate change of oscillator frequency can also be effected by shifting the cut-off point of V1A by altering its anode voltage, and this fact is made use of to provide a line hold control.

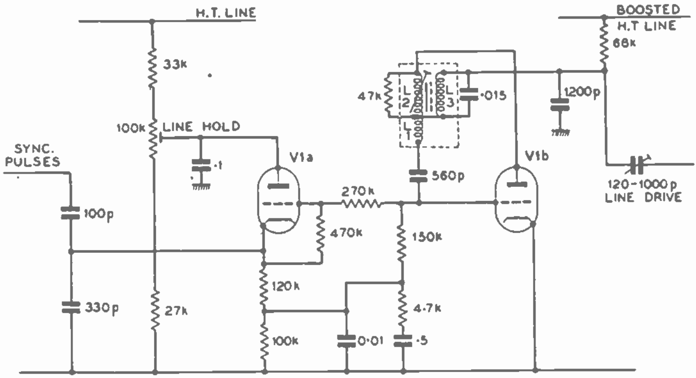


FIG. 37.—USE OF A TRIODE COMPARATOR VALVE ("SYNCHROGUIDE").

TIME-BASE OSCILLATORS

The Blocking Oscillator

The basic circuit of a blocking oscillator is shown in Fig. 38 (*a*). In this version a triode is used, anode and grid being tightly coupled together with an iron-cored transformer. The grid return is taken to earth via a variable time-constant. The circuit constants are so chosen that the oscillation due to the transformer coupling is so great that the charge built up across the grid time-constant completely blocks the valve after the first half-cycle. The blocked state continues until the charge has leaked away sufficiently for the valve to conduct, when the cycle repeats. A sawtooth waveform of the type required for driving a scanning circuit is thus available across the resistance-capacitance circuit. In order to make the flyback of the sawtooth rapid, the inductance and self-capacitance of the transformer are so chosen that the oscillation cycle is very rapid when compared with the repetition period of the sawtooth. This sometimes presents a problem in the horizontal oscillator, where the flyback time must be faster than one-tenth of the trace time.

The sawtooth waveform across the grid time-constant is not quite linear, but this may be corrected by returning the grid resistor to a course of positive potential. Unfortunately, it is not then easy to alter the amplitude of the sawtooth waveform without having a considerable effect upon the speed. For this reason the pentode circuit of Fig. 38 (*b*) may be used. The heavy pulses of plate current flowing

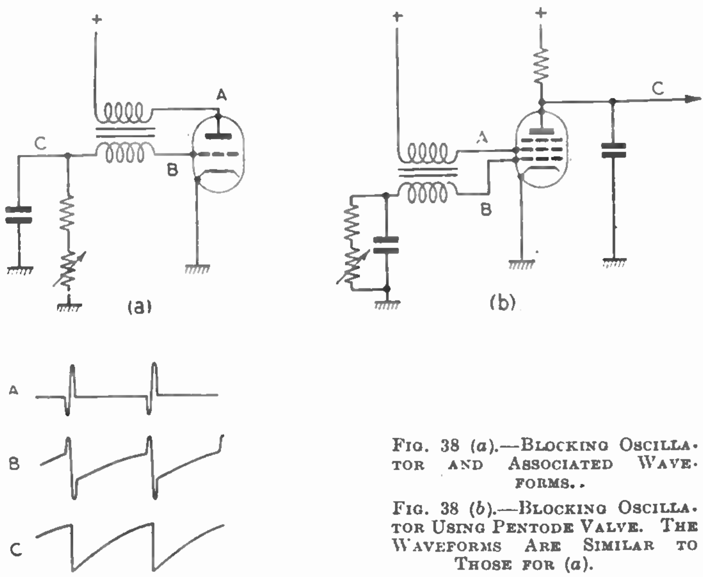
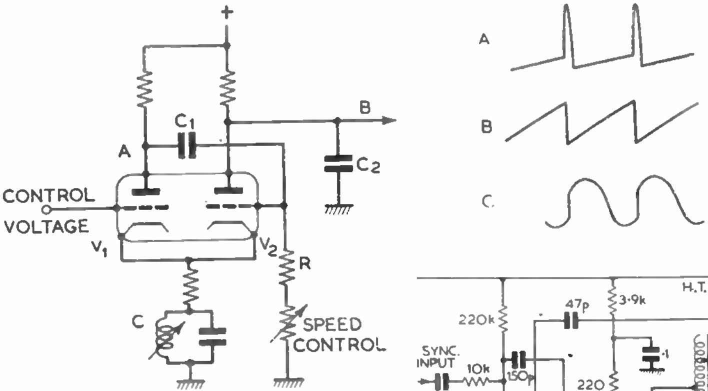


FIG. 38 (*a*).—BLOCKING OSCILLATOR AND ASSOCIATED WAVEFORMS.

FIG. 38 (*b*).—BLOCKING OSCILLATOR USING PENTODE VALVE. THE WAVEFORMS ARE SIMILAR TO THOSE FOR (*a*).



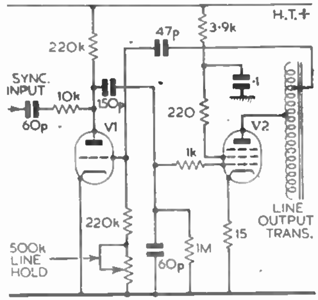


Fig. 39 (*top*).—Cathode-coupled Multi-vibrator and Associated Waveforms. (*below*) A Common Variant of the Basic Multi-vibrator, Eliminating the use of a Double Valve.

between grid and screen when the oscillation occurs cause the capacitor from anode to earth to be discharged; provided that a sawtooth amplitude small compared to the available H.T. supply is required, the resultant waveform is fairly linear. The amplitude may be altered without altering the speed by variation of the anode resistor or capacitor.

Synchronization of the blocking oscillator is relatively easy, and synchronizing pulses of suitable polarity may be injected at the grid or anode. The synchronizing pulse may also be injected in series with one of the transformer windings, or, alternatively, a special winding can be added to the transformer.

The major disadvantage of the blocking oscillator is its lack of suitability for use with automatic phase-control circuits. In the first place, the control must be exerted via the grid resistor, through which sufficient grid current flows to upset either of the discriminator circuits described in the previous section. A cathode follower has therefore to be used or, if more gain is required, a D.C. amplifier. In any case the blocking oscillator, in simple form, will vary considerably in frequency with small changes in the H.T. voltage. With triggered synchronization this is a difficult problem, and, in the case of automatic phase control, the multi-vibrator has been found, in this respect, vastly superior.

The Multi-vibrator

The cathode-coupled multi-vibrator [4] has already been mentioned in connection with automatic phase control, and a simplified circuit is given in Fig. 39. While necessitating the use of a double valve, the transformer of the blocking oscillator is eliminated, and the two circuits are thus comparable in cost. The valve V1 conducts for the majority of the time, and is only cut off during the retrace time. This is indicated by its anode voltage waveform given in sketch A. This waveform is restored at the grid of V2, which acts as a discharge valve and produces a sawtooth waveform at its anode, across C2.

The speed of the oscillator is controlled by the time-constant C1–R, and R is usually made variable for the purpose of manual control. Care has also to be taken to select the circuit constants so that the right retrace time is obtained. The tuned circuit is adjusted to produce the waveform shown in sketch C. The constants of the tuned circuit have a considerable effect upon the "pull-in" range and, in this respect, the design must be carefully arranged to provide the desired result. It will be seen that the waveform across the coil is chosen so that the oscillator "fires" at the point of maximum slope Detuning of the coil from this frequency (which is lower than the oscillator repetition frequency) causes the "pull-in" range to be asymmetrical.

In the circuit shown, the charge circuit connected to the anode of V2 has some effect upon the frequency when the values are altered for the purpose of adjusting the amplitude of the output waveform. A pentode may therefore be used for V2 for the same reason, and in the same way, as in the case of the blocking oscillator.

Vertical Deflection

In the British system the sawtooth current waveform required in the deflection coils has a repetition frequency of 50 c/s. At this frequency the coils represent a largely resistive load, and, in the majority of present designs, will have a value of up to about 50 ohms. Higher impedances have been reached, but the diameter of the wire used becomes so small that such coils are not considered feasible for mass production. The sawtooth current-waveform is derived from an output valve operating in a similar manner to an audio output stage; since the optimum load for such a valve will be several thousand ohms, a matching transformer is required.

If a sawtooth voltage-waveform of perfect linearity is fed to the grid of a triode valve, or to a pentode valve which allows sufficient current swing, the current waveform in a purely resistive load will not be linear; this is, of course, due to the curvature of the valve characteristic. In the case of the frame-output stage, the transformer also produces some distortion in the waveform owing to the saturation characteristic of the iron-core; however, the effect upon the sawtooth is in the opposite direction to that caused by the valve. Since the resultant scan must be linear—or at least so nearly so that the spot at any instant is within 1 or 2 per cent of its expected position—it is not surprising that linearity is the major problem in designing the circuit.

The sensitivity of the scanning coils is, of course, of the utmost importance, and must be known before the design of the output stage can commence. At the present time saddle coils of orthodox design are in the majority, although another type having saddle-shape coils wound in an iron or ferrite yoke resembling a motor stator, and known as the "castellated yoke" is also very popular. The choice of the type of coils is very largely a personal one, and depends upon such factors as availability, raster shape, edge focus and sensitivity, etc. The choice of the exact number of turns, and hence inductance and resistance, can be made so as to provide the best space factor in the particular winding under consideration. It should be remembered, however, that the higher the impedance achieved, the better, in general, will be the performance of the coupling transformer. Coils having an impedance of about 10 ohms will generally require a sawtooth current of 0·5–1·0 ampere, peak-to-peak, to obtain full deflection; the actual figure varying according to the particular deflection angle, and the final anode voltage applied to the cathode-ray tube. Further information on deflection yokes will be found in Section 24, "Cathode-ray Tubes"

Choice of Valve

Both triode and pentode valves have been successfully used for vertical deflection, although the pentode is almost universal in this country at present. In the case of the pentode it must be realized that the maximum voltage swing is fixed by the knee of the anode voltage–anode current characteristic and the available H.T. line voltage. In many suitable valves the knee will prevent the anode potential falling below 80 volts, so that with the commonly used H.T. line of 190 volts the swing must not exceed 110 volts. Assuming 10-ohm coils needing 0·5 ampere, peak-to-peak, the valve must therefore supply a peak-to-peak swing of 22·7 mA. The perfect transformer could thus have a ratio of 22 : 1, no resistance and a very large inductance. In point of fact this is, of course, not possible; furthermore, the transformer also suffers from core saturation.

In practice, a cathode-ray tube with a 70° deflection angle and a final anode voltage of 12–14 kV has been scanned by a 3·5-watt valve with a peak-to-peak anode current of the above order. Using different coils—and, of course, a different transformer—an 8-watt audio-output valve can provide adequate scan for 70° and 90° tubes with final anode voltages of up to 17 kV.

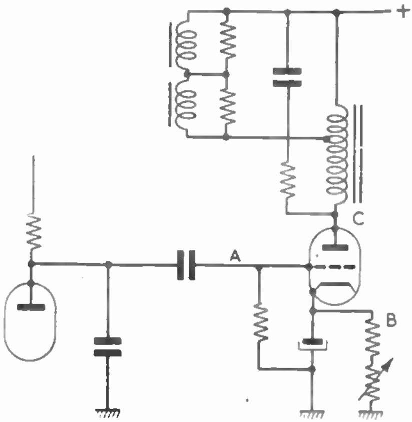
The Transformer

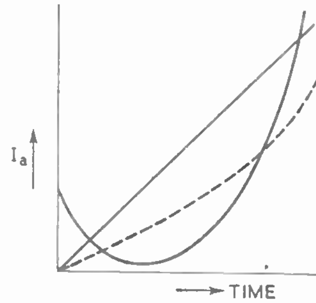
With a very large transformer having substantial primary and secondary inductances and with care being taken (perhaps by adjusting the gap between the laminations) to prevent saturation of the core, a scan of suitable linearity can be obtained by causing a nearly sawtooth current-waveform to flow in the primary winding. Greater efficiency may be achieved by using the auto-transformer connection, considerable space being saved with lower resistance and higher coupling.

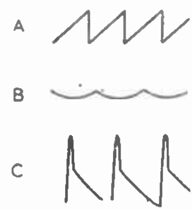
By relating the transformer constants to the load, the energy stored in the transformer inductance may be made to contribute to the deflecting current following the flyback. According to the conditions chosen, the anode current-waveform will lie between the limits indicated in Fig. 40. The straight line represents the sawtooth current condition with large transformer inductances. The other condition will provide the lowest possible mean current with small inductances. Between the two, another mode of operation is found where the peak current required is minimized. In general, a compromise mode is to be preferred. The circuit conditions are rather less difficult to meet, while it would be unwise to allow the stage, however it may operate, to produce less than about 20 per cent overscan at full output.[5]

Fig. 40 (*right*).—Three Modes of Operation for Vertical Scanning.

Fig. 41 (*below*).—Simple Method of Linearity Correction.







Linearity Correction

Because of the variations which occur in valve characteristics, transformer constants and circuit parameters in general, some arrangement must be made for adjusting the linearity of the scan between reasonable limits. Fig. 41 shows perhaps the simplest possible circuit; this is most suitable for use with triode output valves. The input waveform at the grid is made as near as possible to a perfect sawtooth. The bias between grid and cathode is varied in order that the part of the anode-current grid-voltage characteristic used may be chosen so as to counteract the inverse distortion due to the transformer and input waveform. If the reactance of the cathode decoupling capacitor, when compared to the resistance, is appreciable, the waveform may also be used to assist in the correction.

The chief disadvantages of this system are the occasional combinations of characteristics requiring such extreme adjustment of the control that grid current or cut-off is reached. The rapid flyback of the input sawtooth, followed by a very low anode current at the commencement of the trace, may cause the energy stored in the transformer inductance to produce an overshoot or ring. This is limited by the capacitor and resistor connected in series across the transformer primary, and these affect critical damping. The two resistors connected across the scanning coils are for the purpose of damping the circuit at the horizontal-scanning frequency. The vertical circuit may otherwise be excited during the horizontal retrace, and a slight oscillation of the lines in the vertical direction would be visible at the left-hand side of the raster.

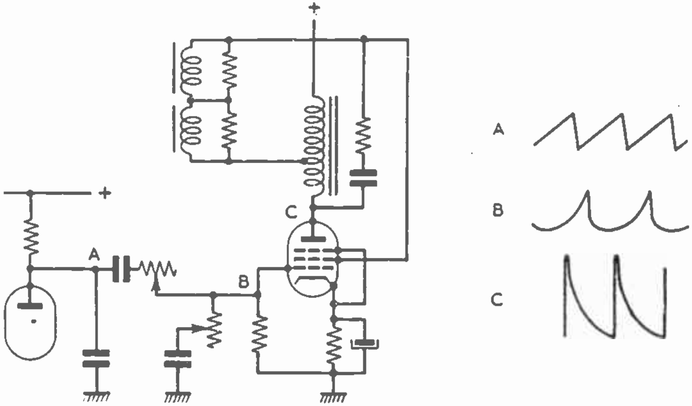


FIG. 42.—LINEARITY CORRECTION SUITABLE FOR A PENTODE VALVE.

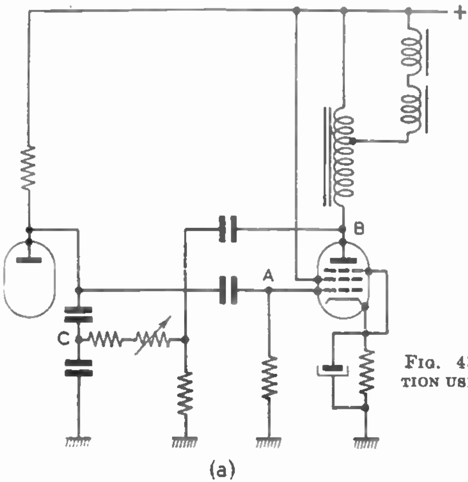
Fig. 42 shows a circuit most suitable for a pentode valve: here the circuit constants call for a slight parabolic component additional to the best sawtooth waveform which can be generated by the oscillator. The circuit is particularly applicable to the mode of operation represented by the dotted line in Fig. 40. The series-connected potentiometer is for the purpose of adjusting the scan amplitude; a degree of residual resistance is presupposed. The potentiometer and capacitor connected from the valve grid to earth distort the waveform, as indicated by the waveform sketch B. Although the controls are rather inter-dependent, this circuit can provide excellent linearity, and is to be highly recommended for mass-production.

The last type of correction circuit to be described is shown in Fig. 43; this is a voltage feedback arrangement. The aim of the circuit is precisely that of the previous example; it has the advantage that the possibility of microphony is lessened. The output waveform, suitably distorted, is fed back to the grid circuit, where it is combined with the grid waveform.

Unfortunately this type of arrangement causes loss of amplitude should the scanning coil resistance increase appreciably as the receiver attains its working temperature. This effect has been offset by various means, including the insertion of a temperature-sensitive resistor in the feedback loop.

Horizontal Deflection.

At the horizontal-scanning frequency, the deflection coils (which will probably be similar or identical to those used for vertical deflection) will represent a predominantly inductive load. Practical values will lie between 10 and 40 mH. If the resistive part of the impedance is negligible, then the voltage waveform across the coils, when a sawtooth current is flowing, will be the differentiated current waveform: i.e., a constant value during the trace, and a pulse during the flyback; Fig. 44 (a) waveform B. The steady voltage across the inductance



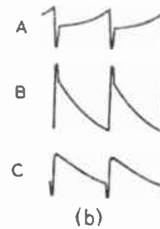


FIG. 43.—LINEARITY CORRECTION USING VOLTAGE FEEDBACK.

during the trace is given by $E = L \times I/T$, where I is the peak-to-peak current and T is the trace time. The presence of resistance causes a sawtooth component to be added to the voltage waveform as shown in Fig. 44 (a) waveform C.

It is important to remove all resistance from the inductance during the flyback, and for this reason the valve to which the inductance is connected is cut off during this period. (This applies whether the inductance is the coils or the transformed coil inductance.) In practice. the deflection coils cannot be made without some self-capacitance, and the sudden reversal of the current in the inductor causes the undamped resonant circuit to be excited at its natural resonant frequency. This oscillation would continue throughout the entire trace if the circuit did not have some losses and the valve did not conduct again. In any case the overswing must be prevented altogether. In the past this was performed by critically damping the circuit with resistance. This, of course, meant that a great deal of the energy stored in the magnetic field at the end of the trace was dissipated in heat.

The Efficiency Diode

The use of a diode valve—known as an efficiency diode—to recover the energy soon became general. The operation may be explained with the aid of Fig. 44 (b). At the time T1 a linearly increasing current commences to flow in the load, a suitable waveform being applied to the pentode grid. A steady voltage is therefore maintained across the inductance, which, for optimum results, will be near to the knee voltage. The diode bias (which must be of low impedance) is adjusted so that the diode is just about to conduct. At the time T2 the pentode anode current is rapidly cut off, and I_a and V_e commence to oscillate at the natural resonant frequency. However, at the termination of the first half-cycle, the voltage across the inductance reaches the diode bias V_d

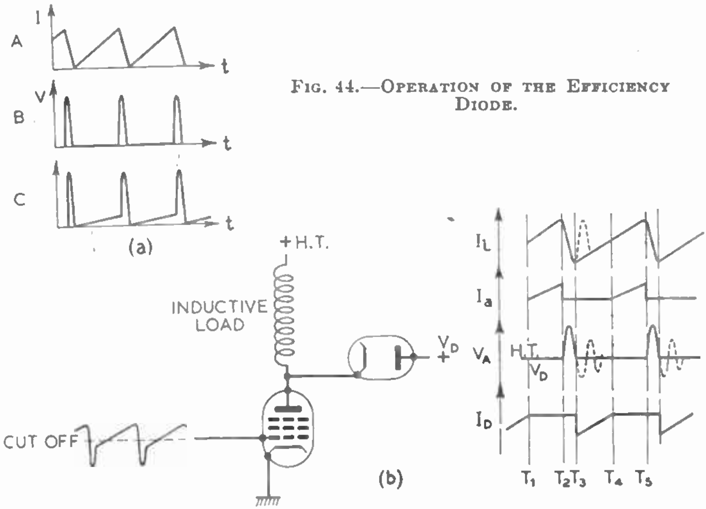


FIG. 44.—OPERATION OF THE EFFICIENCY DIODE.

and is prevented from moving farther. If the circuit has no losses, then the diode current at the instant T3 is equal to the pentode current just before the instant T2. The inductance is therefore connected between the H.T. line and its original potential and, since $E = L\,di/dt$, it follows that the current then changes linearly as between the instants T1 and T2. The H.T. line, therefore, does not need to supply any energy until the instant T4. Thus the energy stored in the magnetic field at the end of the trace is returned to it. In practice there will be some energy lost in the resistance, and this has to be made good by the pentode. A negative pulse is added to the grid sawtooth so that there is no chance of the valve drawing any current during the flyback period when the anode potential is very high.

The Booster Diode

An improvement in the above arrangement may be effected by connecting the diode as a booster diode. This circuit is shown in Fig. 45, in which the inductance is drawn as the (normally-used) transformer with an inductive load. During the second part of the trace, where the valve is in conduction, the anode current is drawn through the diode and a steady voltage exists across that part of the transformer between the pentode anode and diode cathode. The current therefore increases steadily and, since the two sections are coupled together, a steady positive voltage exists between the diode cathode and the top of the transformer, and consequently across the capacitor This voltage depends upon the turns-ratio of the transformer, relative to the bottom section. The value of the capacitor is made large so that

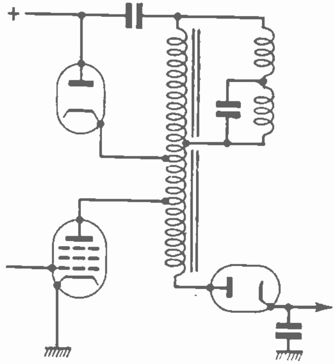


FIG. 45.—BOOSTER DIODE.

the voltage remains constant during the flyback. When the first half-cycle terminates, the transformer-tap potential tends to become negative with respect to the H.T. line, and the steady potentials are resumed once more.

It is thus seen that the voltage available across the transformer is greater than that of the H.T. line less the valve knee voltage, and a very great improvement in efficiency is obtained over the previous arrangement. In order to obtain the high voltage required for the final anode of the picture tube, use is made of the positive-going pulse across the transformer during the flyback. The transformer is extended by the necessary amount at the " hot " end and is coupled to a diode rectifier. The heater supply for this valve must be at the E.H.T. voltage, and is most conveniently obtained from a small winding on the transformer core.

It will be seen in Fig. 45 that a capacitor has been connected across one of the line coils. This is because the self-capacitance of the windings with respect to earth is not evenly distributed and, since the coupling between the coils is poor, it is possible for a small oscillation to occur, excited by the sudden damping exercised by the diode. The small capacitor is used to balance these strays.

The Tuned Leakage Inductance Transformer

As screen sizes become larger it is necessary to increase the scanning angle to keep the depth of the tube to a reasonable size and so ease the problem of cabinet design. 110-degree tubes are now often used, particularly in portable receivers. In addition, it is usually necessary to increase the E.H.T. applied to the tube to maintain, or increase, brightness per unit area of screen. Thus, not only has the electron beam to be bent or deflected through a greater angle, but the beam is stiffer and so more difficult to bend and more power has to be provided by the time-bases. Increased efficiency, particularly in the horizontal scanning, is highly desirable if big increases of input power to the time bases are to be avoided.

One method of reducing scan power requirements is by mounting the deflection coils closer to the beam, and this has led to the use of narrower necks on 110-degree tubes and the introduction of the B8H base.

More efficient magnetic materials have become available, and line transformers using tuned leakage inductance transformers are often used.

If the leakage inductance between the primary and E.H.T. overwind tunes, with the effective capacitance across it, to a frequency about 2·7 times the resonant frequency of the effective primary inductance and effective primary capacitance, then the peak voltage on the anode of the line output valve and cathode of the boost diode valve are reduced by about 20 per cent, while the voltage on the E.H.T. rectifier anode is increased. Where line circuits are limited in output owing to the restriction of voltages that can be safely applied to the valves, then the tuned leakage transformer is one way of obtaining greater outputs. At this 2·7: 1 ratio, the ringing voltage of the leakage inductance is zero at the end of the flyback period and does not continue during the scan. There are therefore no ringing bars or striations on the rastor due to the leakage inductance between the primary and E.H.T. over-wind.

Linearity

Careful adjustment of the circuit constants can bring about reasonable linearity without any correction arrangements. However, the resistance in the transformer and coils causes an inherent distortion, and, since the sawtooth current has to be slightly modified for the scanning of flat-face tubes, a combination of sawtooth and parabolic component has to be added to the scanning current. Two methods are shown in Fig. 46. In the first a tuned circuit is connected in series with the diode and two sections of the transformer. The L/C ratio is chosen with regard to the coil ratio so as to provide the necessary amplitude, and the coil is tuned for the waveform required.

Alternatively, a suitable impedance whose value varies according to the current flowing may be connected in series with the coils. This is most conveniently a saturable reactor, and may consist of a coil adjusted with a small ferrite core which is saturated by a permanent magnet to a critical degree. In order to prevent oscillations, the correction coil may be critically damped as shown in Fig. 46 (*b*).

A linearity control which does not introduce ringing, and in addition is cheap and effective, uses a shorted turn coupled to the line-scan coils. During the line scan an E.M.F. will be induced in the short-circuited turns, and this will produce a current round the turns which will change exponentially during the line scan. This current will in turn produce a flux opposing the main flux of the line-scanning coils, and by choosing a suitable time constant for the short-circuited turns the resultant flux can produce a substantially linear line scan. Better than 5 per cent non-linearity is readily achievable.

In practice, the two short-circuited turns are each positioned between the line scanning coils and the neck of the cathode-ray tube and are adjusted in position for optimum linearity of scan and then fixed. Unless the scanning components are changed, this makers' adjustment can remain set for the life of the receiver.

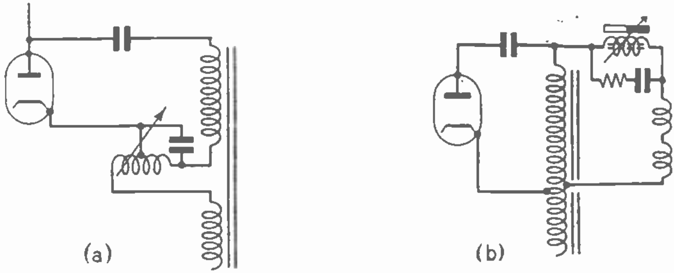


FIG. 46.—LINEARITY CORRECTION WITH BOOSTER DIODE.

Width Control

In mass production it is essential to be able to adjust the scan amp· tude by about ± 10 per cent. This may be effected by placing a sma variable inductor in parallel with a section of the transformer, o alternatively, in series with the coils. In the latter arrangement it i once again necessary to connect a damping circuit across the inducto It must be borne in mind that at the limits of variation these method cause mismatch. Both arrangements are, of course, capable o development in several ways.

Valves

Special valves have been developed for the line-scan circuit. In the case of the output pentode a hig! peak current cathode rating is required—of the order of 250 mA for 70° deflection. During the flyback period the peak voltage swing on the anode may rise to 5 or 6 kV: for this reason the anode lead is usually brought out to a top-cap connection, with special insulation internally. The peak current is unlikely to be obtained without the passage of grid current, and this rating should also be fairly high. Then again, following the flyback period, the anode may swing negative to earth due to poor coupling between the sections of the transformer; under these circumstances screen emission must not occur.

The booster diode must have a higher peak current rating than the output valve, and a similar peak-inverse-voltage rating. Since the flyback pulse is applied to the cathode of this valve, very special insulation is required between cathode and heater; or else the heater will have to be supplied from a suitably insulated transformer. Several modern booster diodes intended for this application have the cathode brought out to a top-cap connection.

In some receivers the E.H.T. diode must stand a peak-inverse-voltage of up to 20 kV, and many precautions have to be taken. The heater will usually have a low voltage rating (1–2 volts) and be coated with the emitting material, or—if a cathode tube is used—will have a rating of about 6 volts. In either case the heater current is very low. One construction very popular overseas is similar to the booster diode: this has a glass base with top-cap anode. In this country, however, the leads have usually been brought out through the ends of the glass envelope for direct connection. This type of construction may largely disappear as voltage ratings increase.

The Transformer

A full discussion of transformer design is outside the scope of this section, but some of the considerations involved must be mentioned. The iron and copper losses at the horizontal scanning frequency are liable to be so large that special materials and techniques have had to be developed. For the core only the very highest grades of laminations can be used, or alternatively compressed dust-iron or ferrites must be used. In recent years the use of ferrite cores has become practically universal on account of the high initial-permeability and the very small losses. The material, however, is easily saturated, and for this reason it is common practice to use butt-jointed or gapped constructions. The self-capacitance of the transformer is also important if a sufficiently rapid flyback is to be achieved; in this respect the small number of turns needed with a ferrite core is of great assistance. Both wave and interleaved winding techniques have been used, although it is found that a wave-winding gives the best result for the section between the pentode anode and the E.H.T. connection. A high self-capacitance in this section soon limits the maximum performance.

Self-oscillating Scanning Circuits

By various means it is possible to introduce feedback in such a way that the horizontal output circuit oscillates. One method is to connect the grid and screen of the pentode as a blocking oscillator, with the screen forming the anode. The circuit has to be carefully proportioned in order that the desired current-waveform will flow in the pentode anode circuit, and synchronization must be of the triggered type unless a very low impedance output from an automatic phase-control circuit can be made available. As this type of circuit is most generally used for the sake of economy, triggered synchronization is usually adopted.

Alternatively, the waveform from a tap on the output transformer may be fed to an additional valve, where it is amplified and suitably shaped by discharging action into a time-constant circuit, and then coupled to the grid of the output valve. There are variants of both methods—with and without the extra valve—but it is seldom possible to produce a scan of the same degree of linearity and with the same efficiency as when using the standard circuit.

The Desaturated Transformer

Alternatively, an arrangement has been employed whereby the magnetizing flux in the transformer is cancelled out. This method has certain advantages, since it is an aid to efficiency and also allows the core section to be substantially reduced. Fig. 47 shows the basic circuit; the transformer is split at the cathode-

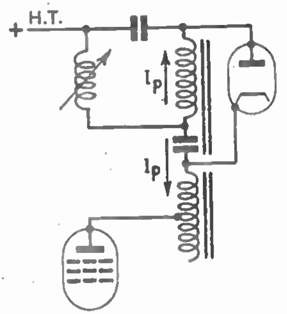


FIG. 47.—DESATURATED TRANSFORMER.

tap of the diode and a coupling capacitor inserted. The H.T. feed is through a large choke which can conveniently be made the width control. It is seen that the pentode plate current has to pass through both sections of the transformer but in opposite directions; the D.C. component can thus, to a large degree, be cancelled out.

A later development is to replace the large choke by the scan coils, fitted with a linearity and width control in the form of a shorted-turn sleeve.

SPECIAL FEATURES

Vision Automatic Gain Control

For some time British television receivers did not incorporate any form of automatic gain control. Following the recommencement of transmissions in 1946 automatic gain control for the sound channel soon became popular, and followed standard radio practice fairly closely. Delay circuits were occasionally incorporated.

In the fringe areas it was long apparent that because of the signal fading experienced some form of vision automatic gain control would prove to be a great boon. Furthermore, a quick-acting circuit might be expected to reduce the characteristic flutter caused by reflections from passing aircraft.

Briefly, two different approaches are possible. The first [6] is to develop a control voltage peak rectifying or "averaging" either the vision carrier or the waveform after detection, with time-constants large enough to remove any low-frequency components. This voltage is fed to the grid returns of the intermediate-frequency and radio-frequency stages, with some method of controlling the gain of this loop, e.g., a potentiometer serving as contrast control. Unfortunately the D.C. modulation of the vision carrier causes the control bias to vary according to picture content. The gain is thus increased considerably when changing from a bright to a dark scene and *vice-versa*. Nevertheless, the scheme was found to be a considerable advantage in very weak signal areas provided that the level of impulsive noise was low enough to prevent the control circuit from backing off and is widely used; one arrangement being shown in Fig. 49.

The second approach is to measure the picture black-level and to control the receiver gain accordingly.[7] The black level (representing 30 per cent modulation in the British system) is kept constant at the transmitter, and is radiated for a few microseconds following the line-synchronizing pulse and for a few lines following the vertical pulse. By using this system the receiver gain is kept independent of the picture content.

If the black-level measurement is made during each "back-porch" period, then the circuit may be made sufficiently quick acting to deal with aeroplane effect, but special care has to be taken to make the arrangement free from interference.

In Fig. 48 a negative-going video waveform is seen to be provided (conveniently) from a cathode follower. This is coupled to the anode of a diode to whose cathode is fed a train of large narrow, pulses, also negative-going. These pulses are best obtained from a local source, e.g., the line-output transformer—and must be narrower than the "back porch" period and must occur at the centre of this period.

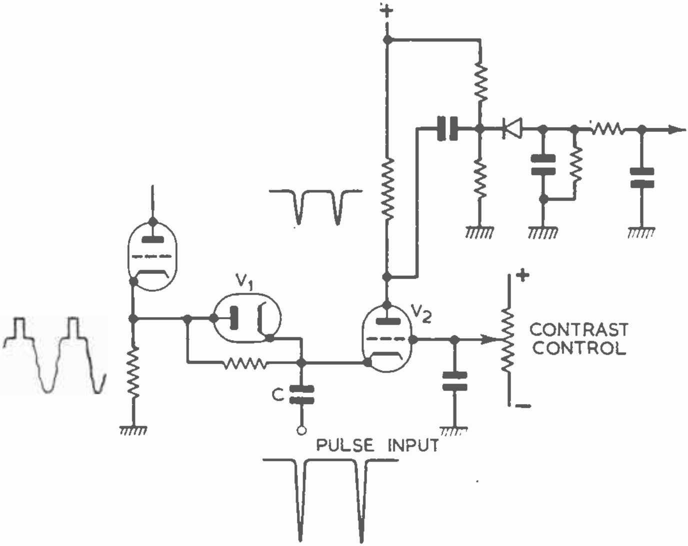


FIG. 48.—AUTOMATIC PICTURE CONTROL.

If the timing is not correct it is necessary to delay or to advance the synchronizing, or the pulse train, by some suitable means. The pulses must be larger in amplitude than the video waveform.

The diode acts as a restorer, and the tips of the pulse train are restored to the potential existing at the cathode of the cathode-follower. This potential, if the timing is right, will be the black level. If the signal should increase, then the potential of the restored pulse train will, therefore, fall with respect to ground and vice-versa. The bias applied to the grid of the valve V2 is so arranged that the tips of the pulse train just cause anode current to flow, so that across the anode load a similar train of pulses exists. However, when the signal strength alters and the restored pulse train moves its potential with respect to ground, the pulse train in the anode circuit of V2 alters in amplitude accordingly. It only remains to rectify this information and so produce a D.C. control voltage. This is done with the aid of the rectifier and smoothing components shown.

This circuit is very insensitive to impulsive interference. Noise pulses at the input are, as shown, in the direction to cut-off the diode. Consequently when the "gating" pulse and a noise pulse coincide, the diode will not conduct and a small amount of the charge on the coupling capacitor C will be lost. At the next pulse it is, of course, highly probable that no noise pulse will occur, and the disturbance due to ignition interference is, therefore, very small.

In some models the gated A.G.C. circuit samples the positive-going video waveform at the cathode of the vision demodulator. With this

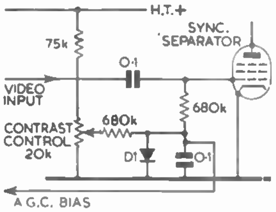


FIG. 49.—MEAN LEVEL A.G.C. SYSTEM SAMPLING SIGNAL AT GRID OF SYNC. SEPARATOR. D1 IS A DELAY DIODE.

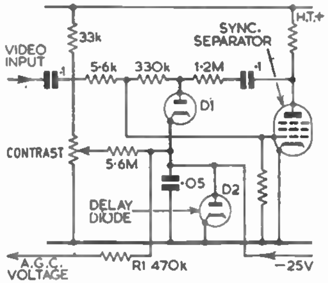


FIG. 50.—A COMMON FORM OF "SYNC. CANCELLED" VISION A.G.C. SYSTEM.

system, a positive-going gating pulse is required. This has been obtained from the anode circuit of a line multi-vibrator, and by differentiation of the signal at the grid of the synchronizing separator. In the latter case the A.G.C. circuit has the advantage of being completely independent of the time-bases.

All but a very few Band I/III receivers incorporate some form of vision automatic gain control. The different strengths commonly encountered between Band I and Band III signals make this feature most desirable in order to eliminate the need for considerable readjustment of the sensitivity and contrast controls when switching between stations. Some designers also incorporate independent, pre-set sensitivity controls to adjust the gain on each Band.

"Sync. Cancelled" A.G.C.

A further common alternative A.G.C. arrangement, known as "sync. cancelled" A.G.C., is shown in Fig. 50. The video wave-form, with positive-going synchronizing pulses and negative-going picture information, is fed to the anode of D1. Negative-going synchronizing pulses from the synchronizing separator are also fed to the same point. The synchronizing pulses are thus cancelled out and, through the D.C. restoration action of the grid circuit of the synchronizing separator, a waveform the most positive portions of which correspond to black level exists at the anode of D1. D1 is a peak detector, and changes in the black level result in a suitable bias voltage across R1.

A number of variations of these systems have been used. One disadvantage of the arrangements described is that the gain is less than unity. Several circuits have been evolved to provide greater gain, though these have as yet found little application.

Impulsive Interference

Interference due to ignition and similar causes becomes very serious where the received signal is weak. The pulses as generated are very fast transients, and it therefore pays to keep them as short as possible in the receiver, so that in both picture and sound channels they will represent the smallest possible amounts of energy. To this end great

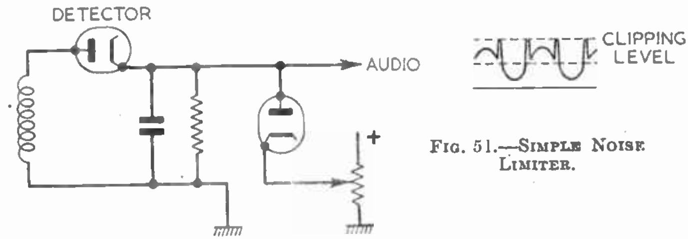


Fig. 51.—Simple Noise Limiter.

care should be taken to see that even though some of the amplifier stages may be driven into grid current, no blocking occurs.

At the output of the sound detector it is possible to obtain good clipping of the noise pulses by a simple diode clipper, but, as Fig. 51 shows, the amplitude of the noise pulses is still comparable to the modulation. Further, with this system, it is possible to clip the audio information.

A great improvement can be obtained with a " following limiter," which depends for its action upon the difference in rise time between the highest audio frequency to be passed and the noise pulse. In Fig. 52 the detector output is coupled to the cathode of the second diode, which by suitably choosing the values of the resistors R1 and R2 is just allowed to conduct. These are usually of the order of several megohms. The value of the capacitor C is chosen so that the diode will continue to conduct when the highest audio frequency is being passed. When a noise pulse occurs, however, the second diode is cut-off and the capacitor starts to charge up to the H.T. potential, at a rate governed by the time constant R1–C. Before long the noise pulse ceases, and the remaining waveform contains only a small residual " blip ". If desired, this effect may be smoothed out by the addition of a small degree of integration following the capacitor C.

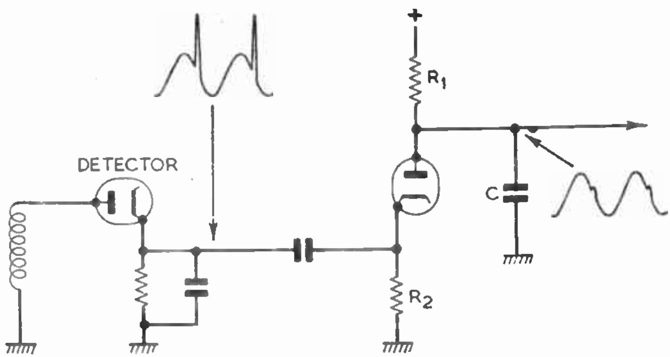


Fig. 52.—" Following Limiter " Circuit.

Unfortunately this circuit is not effective in the vision channel; this is because the rise time of the noise pulses and of the highest vision components are approximately the same. Consequently, limiters tend to be variants of that shown in Fig. 51. An alternative method is the use of a technique known as " Black Spotting ". Since the amplitude of the noise pulses at the detector is very much larger than that of the signal it is possible, by the use of a pre-set clipper, to obtain an output of noise pulses only. These are amplified and fed at a suitable phase to the grid of the cathode-ray tube when modulation is applied to the cathode; or to the cathode of the tube when modulation is applied to the grid.

While this system can be most effective, it is usually difficult to adjust the circuit so as to provide optimum results at other than one contrast setting; fading may also cause poor performance. Furthermore, the fed-back noise pulse arrives slightly later than the one from the video amplifier, so that the transient leading edges are left unsuppressed.

V.H.F./F.M. Reception

The introduction of V.H.F./F.M. sound broadcasting, with the higher-quality, interference-free reproduction possible, has led most manufacturers to incorporate, in some of their television sets, facilities for receiving these programmes. Three sets of Band II coils are fitted in the tuner unit and a double diode or matched pair of germanium diodes is incorporated as an F.M. discriminator, usually a ratio detector (see Section 14). A typical arrangement using a double diode is shown in Fig. 53. V1a also acts as sound demodulator when the set is switched to television reception, V1b then being out of circuit. D1 is the interference-limiter diode and is not used during F.M. reception. TC1 is adjustable to give maximum A.M. rejection on F.M.

Owing to the high I.F. (usually 38·15 Mc/s) it is difficult on F.M. to obtain a high order of A.M. rejection and to provide good adjacent-channel selectivity. To overcome these difficulties, some models bring into circuit on F.M. an extra I.F. stage, while others now incorporate either a separate 10·7-Mc/s I.F. strip for F.M. reception or a dual-channel I.F. strip using techniques similar to those found in combined A.M./F.M. sound receivers. Another system is to use dual conversion to provide a lower second I.F. for the F.M. channel.

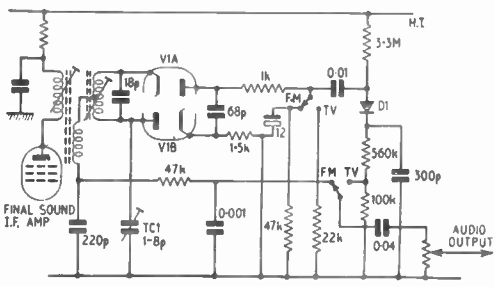


FIG. 53.—SOUND DETECTOR STAGE INCORPORATING FACILITIES FOR V.H.F./F.M. RECEPTION.

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D. H. F. (*Revised by C. H. B.*)

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16. COMMERCIAL HIGH-FREQUENCY RADIO LINKS

16. COMMERCIAL HIGH-FREQUENCY RADIO LINKS

Long-distance communication circuits may be established in several ways, the method chosen depending upon the type of country and the sea distance to be spanned, whether intermediate junctions are required and, of course, capital outlay and cost of maintenance.

The chief methods used are: (1) multi-channel line system; (2) multi-channel very-high-frequency (V.H.F.) system; (3) high-frequency (H.F.) radio links.

High-frequency radio links have the following advantages:

(1) The capital cost is relatively low.

(2) The system can be made very flexible, transmitters and receivers can be switched to different circuits according to traffic loading.

(3) Installation is relatively speedy.

(4) A large number of radio links can be terminated in central transmitting and receiving sites, so that the provision of good maintenance facilities is simplified.

The main disadvantages are lack of secrecy, liability to jamming and the influence of ether disturbances.

OVERALL SYSTEMS

High-frequency communication systems may be divided broadly into two groups, telegraph and telephone, each type of system requiring its own specialized equipment.

Telegraph

One end of a typical telegraph communication system is shown in a simplified form by Fig. 1; the other end, of course, is similar.

The transmit teleprinters may be either the automatic type fed with punched tape or the manual type. In either case, each teleprinter provides a keyed D.C. output which is used to key one of the tones of a multi-channel voice-frequency equipment. The tone keying may be either " on–off " or frequency shift.

The output of the multi-channel, voice-frequency equipment, which, in the case illustrated, consists of six different keyed tones, is taken from the central telegraph office by line to the transmitting station. This connecting line could, if necessary, be replaced by a V.H.F. radio link.

At the transmitting station the various tones are separated and converted into D.C. impulses to key the various transmitters. The power output of each transmitter is taken to a directional aerial which is beamed on the distant receiving point.

Signals from the distant stations are received at the receiving station by means of directional aerials, and the outputs of the receivers are taken to a multi-channel, voice-frequency equipment for transmission to the

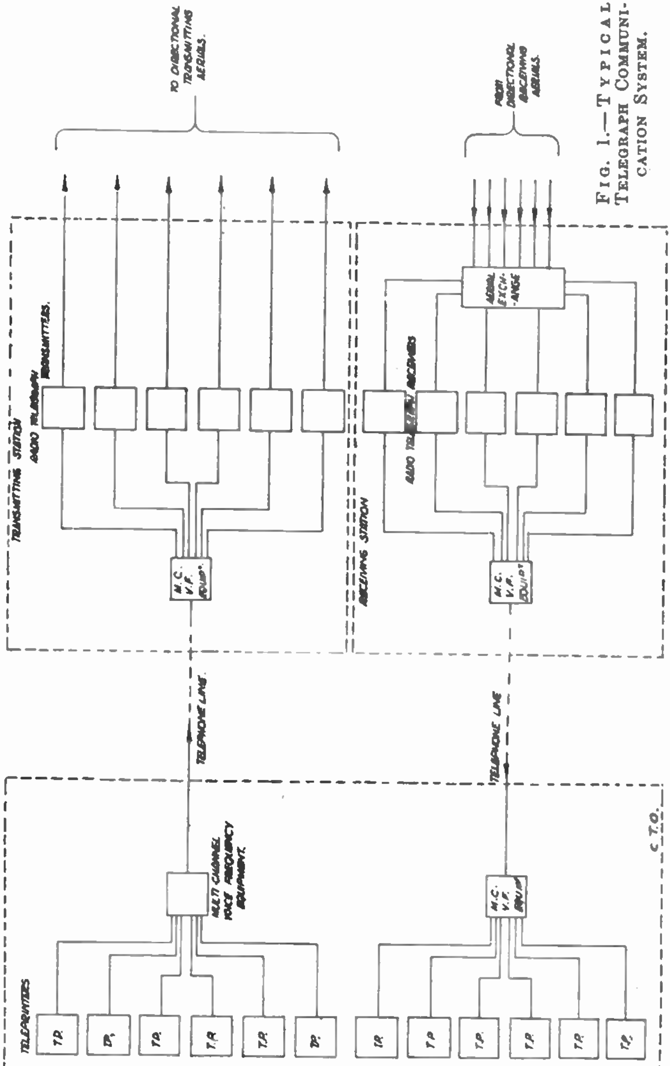


FIG. 1.—TYPICAL TELEGRAPH COMMUNICATION SYSTEM.

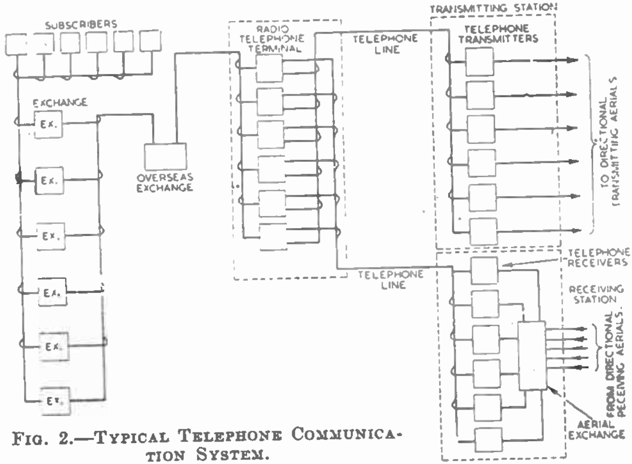


FIG. 2.—TYPICAL TELEPHONE COMMUNICATION SYSTEM.

central telegraph office. There, the messages are printed, the teleprinter slip being pasted on to the usual forms for delivery. In the case of organizations conducting a large amount of telegraphic correspondence, messages may be delivered via a further teleprinter line link, with resultant economy in time and cost.

The distance between the receiving and transmitting station should be

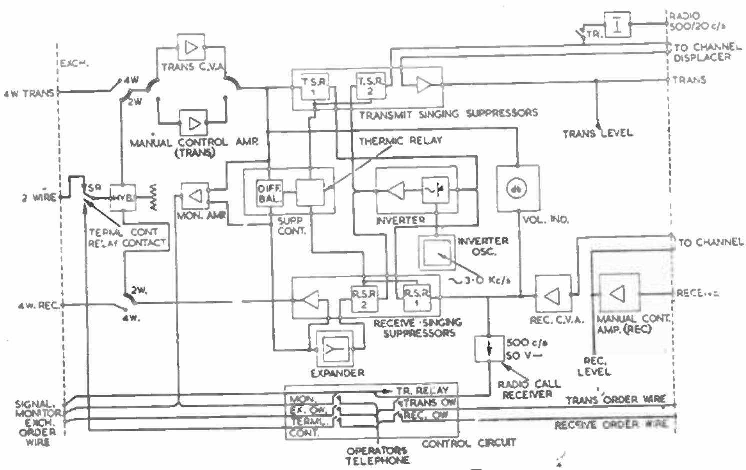


FIG. 3.—RADIO-TELEPHONE TERMINAL.

such that no interference is caused by the transmitters. This obviously depends upon such factors as frequency separation, aerial directivity and transmitter power.

The receiver station should be positioned on a site which is free from electrical interference from power lines, electrical machinery and domestic apparatus. A clear, flat space for aerials—with room for expansion—is also a necessity.

The central telegraph office is usually in a main communication centre.

Telephone

Fig. 2 illustrates one end of a typical telephone communication system, the purpose of such a system being to enable telephone conversations to be carried out over long distances where direct line connection is impracticable.

A telephone call from a subscriber is routed to an overseas or international exchange, and from thence to the radio telephone terminal (R.T.T.). The R.T.T. is a technical control point, which may be semi-automatic in character, where the incoming two-wire circuit is split to provide two two-wire connections to the radio transmitter and receiver respectively. Another function of the radio telephone terminal is to prevent the received signal being re-transmitted and so causing instability in the radio link. This is best achieved by ensuring that the radio circuit is operative in one direction only at any instant.

A block diagram of a radio telephone terminal is shown in Fig. 3.

TYPES OF EMISSIONS

Complete information concerning types of emissions is given in the Final Acts of the International Telecommunication and Radio Conference, Atlantic City, 1947, but the following is a summary concerning high-frequency applications. See also Table 1, page 16–26.

In the past, most radio telegraphic communication systems have used on–off keying, " carrier on " indicating a " mark " and " carrier off " indicating a " space ". The effect of this is to present the receiver with an intermittent signal, which, if integrated over a period of time, is in the space condition longer than in the mark condition; and this results in a low value of average power. If the keying is reversed, so that a mark signal is represented by carrier off and a space by carrier on, an increase in average power results; it has been found in practice that reversed on–off keying shows an advantage over the more conventional method.

Most new installations are using frequency-shift keying (F.S.K.): that is, the mark and space conditions are indicated by different carrier frequencies. With this method the transmitter is radiating continuously so that the receiver has an unbroken signal to deal with.

In practice there is considerable divergence in the values of frequency shift used; they vary from 100 to 1,200 c/s. Various attempts have been made to standardize the shift values, the latest recommendation by C.C.I.R. is as follows:

(1) That it is too early to standardize the actual values of frequency shift, but that every effort should be made to achieve this as quickly as possible for emissions using only two frequencies, one for mark and one for space; that to assist in this, the characteristics shown below should be used as far as possible.

(2) That the value of the frequency shift employed should be the lowest compatible with the maximum telegraph speed regularly used, the propagation conditions and the equipment stability.

(3) That for frequency-shift systems working on two conditions only (i.e., single-channel and time-division multiplex systems) and operating between about 3 Mc/s and 30 Mc/s the preferred values of frequency shift are 200, 400 and 500 c/s.

(4) That the values 140, 280 and 560 c/s may be used provisionally, but 560 c/s should not be adopted for new systems.

(5) That the value of the frequency shift should, if possible, be maintained within ± 3 per cent of its nominal value and, in any case, within ± 10 per cent.

(6) That for circuits using the Morse code, the higher frequency should correspond to the Mark signal and the lower frequency should correspond to the Space signal.

(7) That for circuits using the International Alphabet No. 2 code with start–stop apparatus, the higher frequency should correspond to the start signal (Position A) and the lower frequency to the stop signal (Position Z).

(8) That for telex circuits using the International Alphabet No. 2 code directly on the radio circuit the higher frequency should correspond to the C.C.I.T.T. "free-circuit condition" and the lower frequency to the C.C.I.T.T. "idle-circuit condition".

(9) That for channels of a 7-unit ARQ system (which are referred to in the Annex to Recommendation No. 167 as directly keyed channels e.g., channel A of a two-channel system), the higher frequency should correspond to the code elements shown as letter A and the lower frequency to the code elements shown as letter Z. For the channels which are to have reversed keying (e.g., channel B of a two-channel system) the higher frequency should correspond to the code elements shown as letter Z and the lower frequency to the code elements shown as letter A.

The advantages of frequency-shift keying over on–off operation may be summarized as follows:

(1) Transmitter keying is easier, especially for the higher powers.

(2) For a given keying speed the transmitted spectrum is smaller.

(3) The receiver has an unbroken signal to deal with, but the effect of this is offset to a certain extent by selective fading. A low value of shift is useful in this respect.

(4) The adjustment of the receiver gain controls is less critical.

(5) The receiver A.F.C. is operative in the rest condition.

For point-to-point telephony circuits single-sideband (S.S.B.) operation is replacing double-sideband (D.S.B.) operation.

In the case of a double-sideband transmission with 100 per cent modulation, half the power is concentrated in the carrier and a quarter of the power in each sideband. As the carrier can be replaced by a local oscillator at the receiver, and only one sideband need be transmitted to convey intelligence, it is obvious that the system is wasteful both in power and frequency spectrum.

A single-sideband system is one in which only one sideband is transmitted, and the carrier is usually suppressed to be 16 db below the peak sideband power.

An independent sideband transmission is one in which separate intelligence is radiated on each sideband and the carrier is suppressed to be 26 db below the peak sideband power.

The gain of a single- over a double-sideband system may be summed up as follows:

(1) The carrier is reduced to small proportions, and practically the whole of the radiated power is concentrated in the sideband Thus, for the same signal-to-noise ratio at the receiver, the transmitter power may be reduced. An alternative, and more usual way, of expressing the improvement is to consider the increase in signal-noise ratio at the receiver for the same transmitter peak power. This improvement is 6 db.

(2) The receiver pass-band is halved, giving a further improvement in signal-to-noise ratio of 3 db.

(3) By the provision of a steady local or reconditioned carrier, overmodulation is eliminated when the carrier level is reduced by selective fading. This benefit is a reduction in distortion which cannot be expressed in decibels but is a very important contribution towards the intelligibility of a telephone conversation.

It is advantageous to use a single-sideband receiver for the reception of a double-sideband transmission, for although a decrease of 3 db signal-to-noise ratio results due to the loss of one sideband, this is offset by the reduction of the distortion due to selective fading and also by the possibility of choosing the sideband which is more free of interference.

TELEGRAPH KEYING SPEEDS

It is usual to express telegraphic keying speeds in terms of the shortest modulation (marking or spacing) element, since this indicates the highest keying frequency that the overall system must accept. The unit adopted is the baud, and is equal to a keying speed of one modulation element per second. A Morse dot is a mark element followed by a space element of the same duration, and is regarded as a cycle. If, then, 50 dots are keyed in a second, the transmission would consist of 100 modulation elements, and the speed would be 100 bauds. The letter V consists of 10 modulation elements, so that 10 successive V's per second also corresponds to a keying speed of 100 bauds.

Traffic speed may also be expressed in words-per-minute, and the convention is to equate 100 w.p.m. to 80 bauds or 40 c/s.

TELEGRAPH CODES

Morse

In high speed Morse traffic the radio transmitter is fed with punched tape. The contacts of the Wheatstone transmitter may supply the

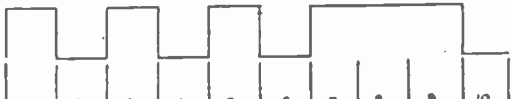


FIG. 4.—BUILD-UP OF THE LETTER V IN MORSE CODE.

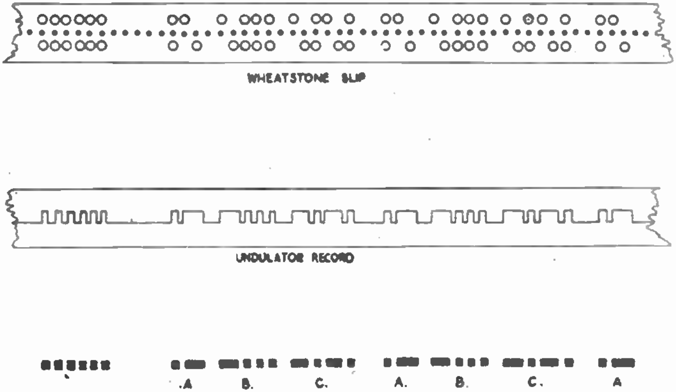


FIG. 5.—PERFORATED AND UNDULATOR SLIP.

radio transmitter directly with polar D.C., or, if a long line is involved, then a tone source would be keyed by the Wheatstone, and converted back to D.C. impulses at the radio transmitter.

Two operators are usually employed to punch tape, so that the transmission can be sustained at high speed without a break.

At the receiver the message is converted back into D.C. impulses, which are fed into an undulator which records the traffic on a paper tape.

Alternatively, the message may be fed into a Morse re-perforator which provides punched tape similar to the original Wheatstone tape. The punched tape is fed into a Morse printer, which reproduces the original message in normal printed characters.

Under very favourable radio conditions, traffic speeds up to 300 words per minute can be attained, but in practice a speed of about 200 words per minute is regarded as very satisfactory. Speeds between 80 and 150 words per minute are the most usual.

The advantages of the high-speed Morse system are :

(1) The speed can be varied to suit radio conditions.

(2) A good operator can read an undulator tape when the signal strength is poor or interference is experienced.

The main disadvantages are :

(1) Labour costs are high, due to the number of highly skilled operators required, or in the case of automatic printing two pieces of complicated machinery are necessary.

(2) A Morse radio circuit cannot be fed into a land-line system without conversion to standard teleprinter keying. Of course, if Morse keying is used on the line system, then this objection does not arise.

(3) In the case of automatic printing a good signal-to-noise ratio at the receiver output is required.

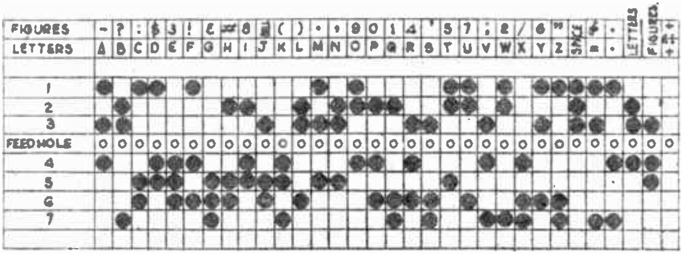


FIG. 6.—R.C.A. SEVEN-UNIT CODE CHART.

Five-unit Teleprinter

This system has largely replaced the Morse system. Its main advantage is that the decoding of the message is almost entirely automatic, so that little labour is involved, and even that may be comparatively unskilled. Another advantage is that a radio link can be extended on a land line, no code conversion being necessary at the junction point.

The disadvantages are :

(1) A good signal-to-noise ratio is required at tho receiver output in order to give an error free record.

(2) It is possible for an impulse of noise to cause the teleprinter to print a wrong character. Errors of this kind cannot always be detected. The carrier shift may be falsely operated, which causes a completely unintelligible record to be printed, so that a re-transmission is necessary.

(3) The transmission speed is comparatively low.

Seven-unit Teleprinter

This code is similar to the five-unit code, except that each character necessitates seven instead of five units for its expression. A sample of the code is shown in Fig. 6, and it will be noticed that each character consists of four spaces and three marks. The printing mechanism at the receiving end is so designed that it indicates when an incorrect code combination is received. For instance, a noise element could add an extra mark pulse, so that the received character would consist of four marks and three spaces. Alternatively, a fade could remove one mark pulse, so that the character would consist of two marks and five spaces. It is possible, of course, for a character to be affected by a noise pulse and a fade so that an incorrect character would be printed. This possibility is a rare one, and further codes have been devised to meet it, but they are not in commercial operation. The main advantage of the seven-unit code, then, is that it is error detecting, but against this must be offset the fact that conversion machinery is necessary for extension to a five-unit land-line system.

A number of American systems use this code.

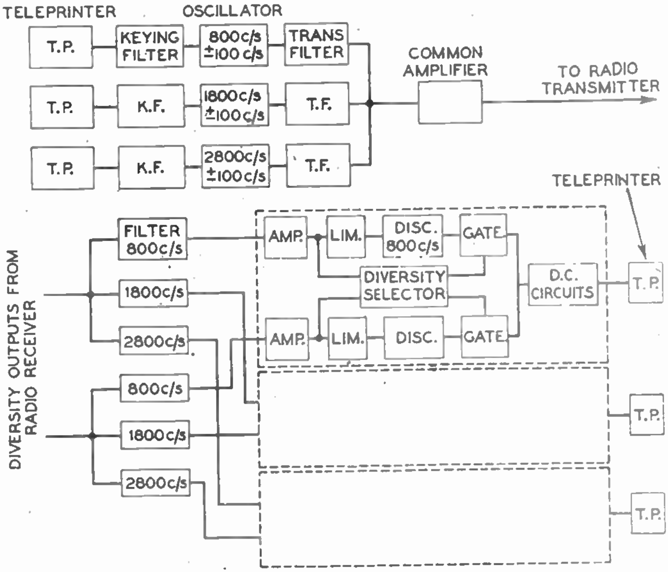


Fig. 7.—F.M.V.F. Block Diagram.

CHANNELLING

A number of schemes have been devised so that more than one telegraph channel may be operated on a radio link. Broadly speaking, the schemes fall into two categories, time division and frequency division.

Time Division

In the case of time division, each channel keys the radio transmitter in turn, and accurate synchronization is maintained between the two ends of the link. The advantage of the system is that each channel is radiated at full power: the disadvantage is that the baud speed is increased in proportion to the number of channels, so that the amount of distortion that can be tolerated is proportionately reduced. Two systems are in considerable use: these are the Cable and Wireless system of Double-current Cable Code, and the R.C.A. Multiplex system, both have been described in detail by Smale.

Frequency Division

The block diagram of a frequency division system for three channels is shown in Fig. 7. Each tone is frequency modulated, and the combined output is used to modulate a single-sideband transmitter.

At the receiver end double-diversity single-sideband reception is used to combat the effects of selective fading.

An alternative method is to use a separate receiver for each channel, the pass bands of the receivers being about 500 c/s.

Four-frequency Diplex System

This is a channelling system which is now being used extensively, and has the merit that it may be applied reasonably easily to existing transmitters and receivers.

It is designed for two-channel operation, the maximum keying speeds being 200 bauds on channel 1, and 100 bauds on channel 2, if the two channels are not synchronized. Channel 2 keying speed may be increased to 200 bauds if the channels are synchronized. The following coding system is recommended by C.C.I.R.

Emitted frequency	*Channel* 1 (*V*1)		*Channel* 2 (*V*2)	
	Teleprinter	*Morse*	*Teleprinter*	*Morse*
f (highest frequency) .	A	Mark	A	Mark
f_3	A	Mark	Z	Space
f_2	Z	Space	A	Mark
f_1 (lowest frequency) .	Z	Space	Z	Space

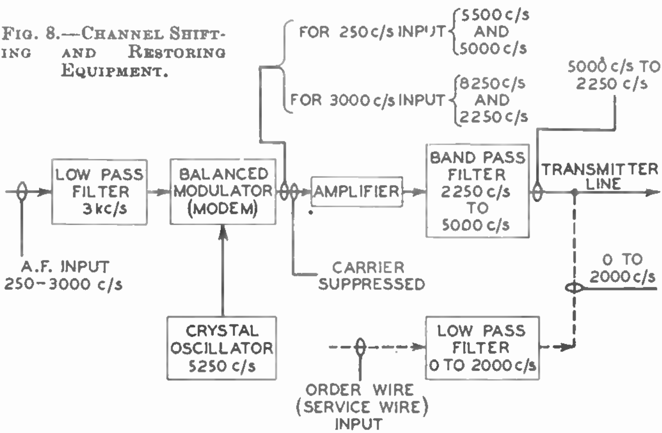
When f_1, f_2, f_3, f_4 designate the emitted frequencies, the spacings between adjacent frequencies $(f_4 - f_3)$, $(f_3 - f_2)$, $(f_2 - f_1)$ being equal.

A represents the start signal of the teleprinter.
Z represents the stop signal of the teleprinter.

Telephony Channelling

Single-sideband transmitters are usually designed to accept modulation frequencies up to 6 kc/s; whilst this is desirable for high-quality music transmissions, it is unnecessary for commercial speech trans-

FIG. 8.—CHANNEL SHIFTING AND RESTORING EQUIPMENT.



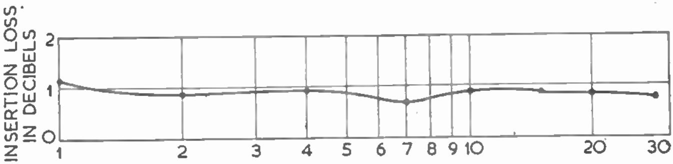


Fig. 9.—Aerial Transformer Response Curve.

missions, which are usually limited to frequencies between 250 and 3,000 c/s. It is common practice to accommodate two commercial speech channels in the 6-kc/s spectrum, and this is accomplished by channel-shifting equipment as shown in Fig. 8.

In the example shown, the traffic-speech channel of 250–3,000 c/s is transferred to the band 2,250–5,000 c/s, and the channel is automatically inverted in the process. The lower speech spectrum of 0–2,000 c/s is available as an order wire, and by this means the operators at the overseas exchange can arrange further telephone calls without interruption of paid time on the traffic circuit.

If the crystal oscillator is set to 6,250 c/s and the band-pass filter tuned to accept 3,250–6,000 c/s, then the order wire can cover 250–3,000 c/s and would be suitable for use as a traffic channel; this is now current recommended practice.

AERIAL EQUIPMENT

Directional aerials are essential for point-to-point working, the principal types are the Franklin or Koomans beam arrays, the rhombic

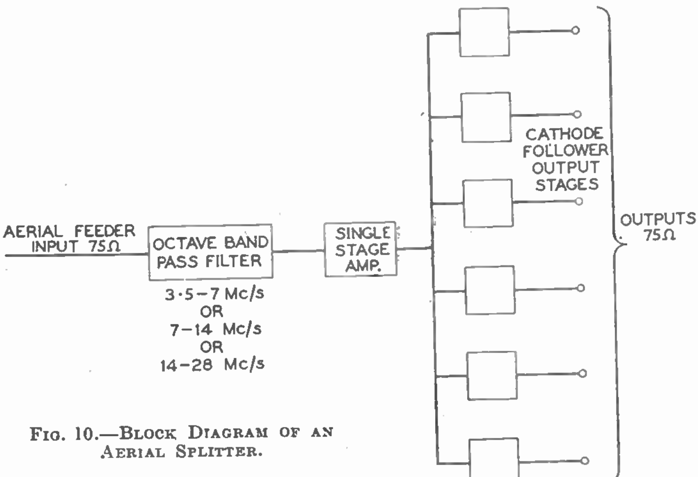


Fig. 10.—Block Diagram of an Aerial Splitter.

and the horizontal array of dipoles (H.A.D.). As the beam aerial is expensive to construct, modern installations tend to favour the rhombic or the horizontal array of dipoles.

Feeders

Aerials must be erected at such a distance from the receiver building that electrical noise, due to motors, relays and the like, is not picked up. It follows also that the feeder system should be non-receptive. In the case of a telephony receiving station which does not require electric motors or relays on its monitoring equipment, the feeder is usually of the open-wire type of 600-ohm impedance. Inside the building, it is more convenient to use 75-ohm concentric cable so that a 600 to 75-ohm matching transformer is necessary at the junction of balanced feeder and concentric cable.

The usual practice for telegraphy stations is to mount the step-down aerial transformer close to the base of the aerial and to run concentric feeder for the whole distance to the receiver building. Best screening is obtained when the feeder sheathing is solid : copper, lead or aluminium being used for this purpose. For the sake of convenience, flexible braided cables are sometimes used inside the receiver building, some loss in screening efficiency results, but this can be minimized by the use of double braiding.

Aerial Exchange

In order to provide flexibility the aerial feeders and receiver feeders are usually terminated at a jack field: by means of flexible plug cords it is then possible to connect any receiver to any aerial. This flexibility can be further increased by the use of an aerial splitter. This device enables several receivers to be connected to one aerial with a minimum of loss and intercoupling between receivers. The block diagram of an aerial splitter is shown in Fig. 10.

COMMERCIAL RECEIVER DESIGN

Telegraph Receiver

A typical telegraph receiver designed for the reception of on–off and frequency-shift signals is illustrated in Fig. 12 and shown in block schematic form in Fig. 11. It is designed for connection to two similar aerials spaced at least five wavelengths apart, so that there is only comparatively little chance of the signal fading on both aerials simultaneously. It follows, then, that the equipment basically consists of two receivers, means being provided to select automatically the stronger output at any instant.

The signal-frequency coverage is 3–27·5 Mc/s, and is provided by three separate amplifiers, each with its own crystal oscillator. The three radio-frequency amplifiers cover 3–6·5 Mc/s, 6·5–14·5 Mc/s and 14·5–27·5 Mc/s respectively. Normally, one of each is fitted to provide continuous coverage.

It will be noted from the block diagram that the amplifiers can be selected as desired, so that, as a radio link employs at least two frequencies for day and night working, the amplifiers may be permanently

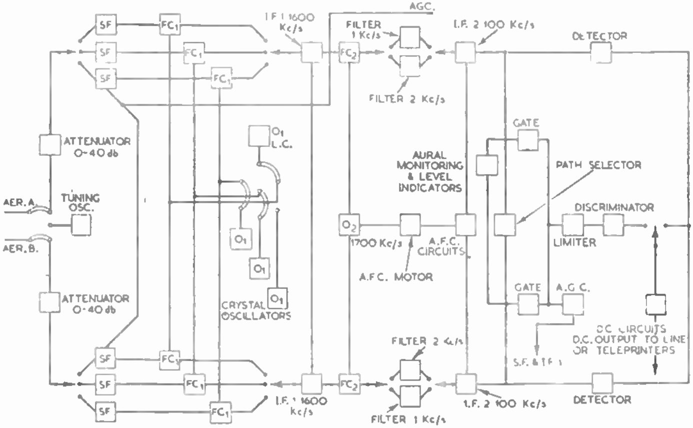


FIG. 11.—SIMPLIFIED BLOCK DIAGRAM OF A TELEGRAPH RECEIVER.

tuned to the required frequencies, and frequency changing then becomes an instantaneous operation.

Should two of the required frequencies be covered by one unit, then that unit could be duplicated and one of the others omitted. A variable-frequency oscillator may be substituted for either of the crystal oscillators. The first intermediate frequency is 1·6 Mc/s in order to provide a good image signal protection, the main selectivity being provided by the second intermediate frequency of 100 kc/s. The second frequency-change oscillator is variable over ± 3 kc/s to provide a fine tuning control: automatic frequency control is also applied to this oscillator.

For on–off keying, the intermediate-frequency output of each receiver path is taken to a diode detector, and combination of the two signal paths is effected by providing a common load to the detectors. The D.C. circuits following the diode load serve the following purposes:

(1) The signal is limited to give a constant amplitude.

(2) Filter circuits are introduced which accept the keying frequencies and reject the short impulses produced by noise.

(3) The width of the signal pulse can be adjusted so as to remove the distortion introduced by the various filters or originating from the radio transmitter.

(4) A polar current output is produced capable of operating a teleprinter or undulator.

For frequency-shift keying, the stronger signal is selected by means of a path selector, limited in amplitude and fed to a discriminator. The D.C. output from the latter is used to operate the same D.C. circuits as are used for on–off keying. The selected signal is also used to operate the automatic gain control circuits, which in turn control the gain of the radio and the first intermediate-frequency amplifier stages.

Radio-frequency Amplifier

To reduce the effects of cross-modulation and overloading from strong signals, the radio-frequency amplifier is preceded by an attenuator,

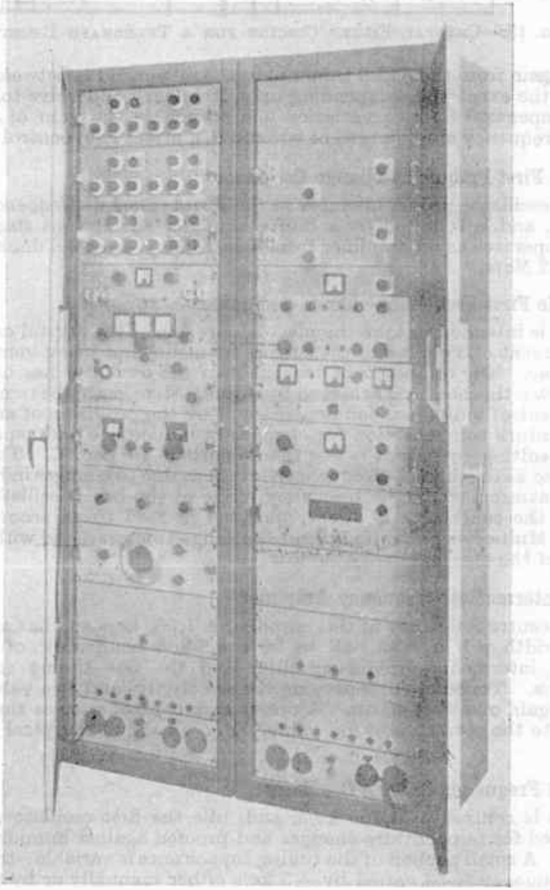


FIG. 12.—TYPICAL TELEGRAPH RECEIVER.
(*Marconi's Wireless Telegraph Co. Ltd.*)

variable in 10-db steps from 0 to 40 db. Two stages of gain are provided, the gain of the second stage being controlled from the A.G.C. line. For maximum efficiency, each circuit is independently tuned.

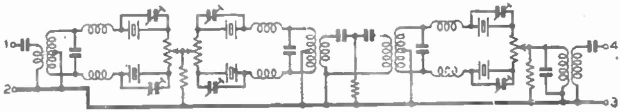


FIG. 13.—CRYSTAL FILTER CIRCUIT FOR A TELEGRAPH RECEIVER.

The gain from the aerial terminal to the mixer grid is between 40 and 50 db, the exact value depending upon frequency and valve tolerances. To compensate for this variation and to enable the gain of the two radio-frequency amplifiers to be equalized, a preset gain control is fitted.

Crystal First Frequency-change Oscillator

The oscillator proper operates at the series resonant frequency of the crystal, and is followed by a buffer amplifier and output stage. The latter operates as an amplifier from 4 to 16 Mc/s and as a doubler from 16 to 32 Mc/s.

Variable First Frequency-change Oscillator

This is intended to take the place of any one of the crystal oscillators in the event of a crystal of the correct frequency not being immediately available. The temperature coefficient of the oscillator has been kept very low: this has been achieved by careful attention to the temperature coefficient of individual components and by the provision of automatic temperature compensation for changes in inductance and capacitance. The resulting coefficient is less than 5 parts in 10^6 per ° C. The unit is sealed so as to eliminate frequency variations due to changes in humidity of the atmosphere. The frequency range of the basic oscillator is 2–4 Mc/s: the calibration is direct, and may be read to an accuracy of 1 kc/s. Multiplying circuits extend the range to correspond with the full range of the high-frequency circuits.

First Intermediate-frequency Amplifier

The centre frequency of this amplifier is 1,600 kc/s and has a nominal band-width of 8 kc/s, this allows for the 2-kc/s band-width of the wide second intermediate-frequency filter and the fine tuning control of ± 3 kc/s. Tuned circuits provide the selectivity, and two valve stages give a gain of about 40 db. A preset gain control enables the gain to be set to the correct working value. Fig. 14 shows a typical response curve.

Second Frequency Change Oscillator

This is centred at 1,700 kc/s, and, like the first oscillator, is compensated for temperature changes and proofed against humidity variations. A small portion of the tuning capacitance is variable: this allows the frequency to be varied by ± 3 kc/s either manually or by the automatic frequency control motor.

Second Intermediate-frequency Amplifier

Two crystal filters centred at 100 kc/s provide the main selectivity. The pass bands are 1 and 2 kc/s, and are selected by a switch. The 2-kc/s band-width is intended for use when receiving modulated C.W.

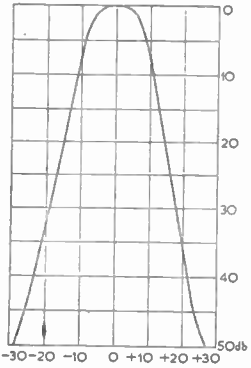


FIG. 14 (*above*).—TELEGRAPH RECEIVER FIRST INTERMEDIATE-FREQUENCY RESPONSE CURVE.

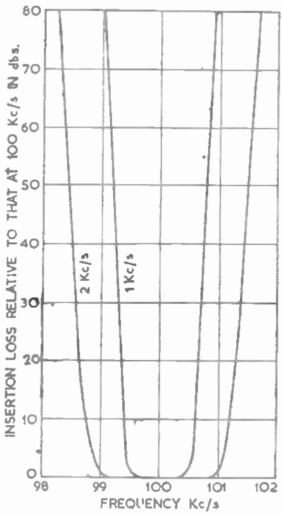


FIG. 15 (*right*).—CRYSTAL FILTER RESPONSE CURVES.

or diplex signals, the narrow filter is suitable for normal on–off and frequency-shift signals.

The circuit diagram of a crystal filter is shown in Fig. 13, and it will be seen that it consists of three identical half-lattice sections connected in tandem. Each section consists of a transformer with its secondary centre tapped to earth forming two arms of a bridge which is completed by a crystal in series with an inductor and a resistor. Additional resistors are connected to earth, and provide a matching network. The filter is designed to work between 75-ohm impedances. The response curves of the two types of filter are shown in Fig. 15: these are independent of temperature over the range 20–60° C. The filters are hermetically sealed against the effects of humidity and corrosion.

The filters are followed by three valve stages with normal transformer coupling, the gain being about 46 db. A preset gain control is fitted. The output stage is capable of handling 1·5 watts of signal without overloading.

Detector

A diode-type detector is used for on–off keying, and is capable of handling a signal amplitude of 400 volts without overloading. As the level necessary to operate the D.C. circuits fully is 4 volts, it will be appreciated that a signal 40 db above threshold can be handled without overloading and without the help of the A.G.C. circuits.

D.C. Circuits

The block diagram of the D.C. circuits is given in Fig. 16. After rectification, the signal is amplified and limited. It then passes to a

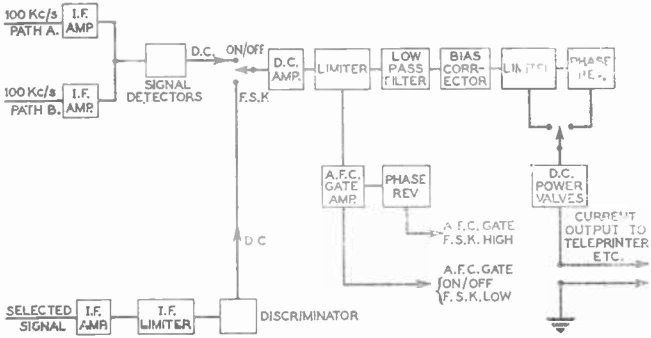


Fig. 16.—Block Diagram of the D.C. Circuits of a Typical Telegraph Receiver.

filter which removes the higher frequencies due to noise. As the filter distorts the signal somewhat, it is necessary to introduce a bias-corrector circuit which enables the operator to select the point in the waveform at which the durations of the mark and space signal are equal. This control could be preset, as its setting is independent of traffic speeds, but in practice it is made variable so as to cater for variations in transmitter or other bias. The signal is finally squared by further limiting. Fig. 17 (*a*) shows the waveform before filtering, Figs. 17 (*b*) and 17 (*c*) after filtering, and Fig. 17 (after final squaring.

For frequency-shift signals, the selected signal at 100 kc/s is limited and passed to the signal discriminator, the output of which operates the D.C. circuits as in the case of on–off keying.

Automatic Gain Control

The A.G.C. circuits are fed from the signal path-selector, and control stages in the signal-frequency and first intermediate-frequency ampli-

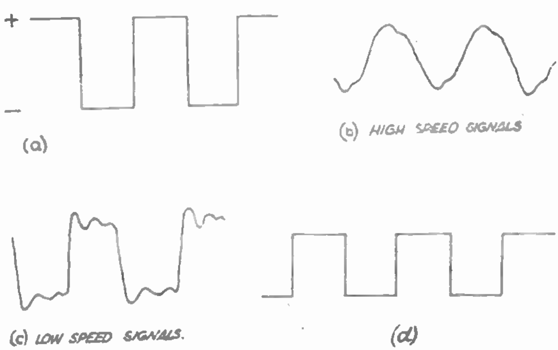


Fig. 17.—Waveforms Before and After Filtering.

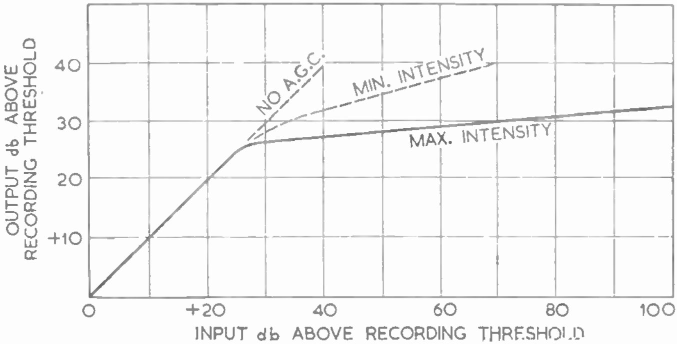


FIG. 18.—TELEGRAPH RECEIVER A.G.C. CHARACTERISTIC.

fiers. With on-off keying, owing to the intermittent nature of the signal, it is difficult to make the A.G.C. action fast enough to follow rapid fading and yet not distort the individual characters. It is the practice therefore to make the A.G.C. intensity and time constant both variable so as to cater for different keying speeds and fading conditions

Very little A.G.C. intensity is used for on-off keying, as it has been found from experience that it is preferable to design the receiver to have a good overload characteristic. Considerably more A.G.C. intensity can be used when receiving frequency-shift signals.

Fig. 18 shows the A.G.C. characteristic, and it will be noticed that the action is delayed by 25 db from the recording threshold.

Automatic Frequency Control

Ideally it would seem that the speed of correction of an automatic frequency control system should vary according to the error frequency, this would mean that the correcting motor would rotate rapidly for a large error, slow down as the receiver is brought into tune and stop the instant that the tuning is correct. In practice, this cannot be achieved, since it is not feasible to design a synchronous type of motor which will start when a frequency corresponding to a large error, say 500 c/s, is suddenly applied.

Experience has shown, however, that a fast correcting speed can be an embarrassment, since intermittent interfering signals and noise would rapidly mistune the signal. It has also been found that to obtain accurate tuning the system must be free from hunting or overshoot. The automatic frequency control scheme adopted in this receiver meets these requirements.

The block diagram is shown in Fig. 19.

Signals at the final intermediate frequency are fed to a gate valve which is cut-off during periods when the control signal is not present. In the case of on-off signals the gate is closed during spaces so that noise does not operate the automatic frequency control. For frequency-shift signals, the control should be from the signal which is transmitted during rest periods.

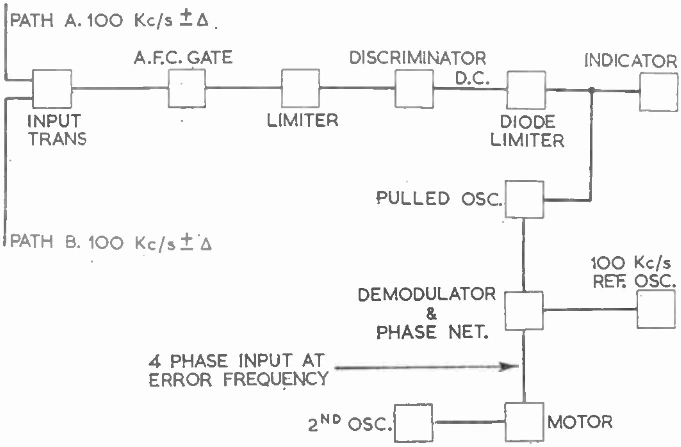


FIG. 19.—BLOCK DIAGRAM OF A TELEGRAPH RECEIVER A.F.C. SYSTEM.

Following the gate valve, the signal is limited and then passed to the automatic frequency control discriminator, which is centred at 100 kc/s for on–off signals and offset by a manual control from the mid-intermediate frequency by half the shift value for frequency shift. The discriminator output is limited and used to pull the frequency of a crystal oscillator. This is achieved by varying the mutual conductance of a reactance valve connected to the crystal oscillator via a $\frac{1}{4}$-wave network. The maximum amount the crystal oscillator is pulled is about 20 c/s, and corresponds to an actual signal error of 50 c/s: larger signal errors have no further effect, due to the action of the discriminator limiter.

The output from the pulled oscillator is compared with a reference oscillator at 100 kc/s and a four-phase output at the difference frequency obtained and applied to an impulse-type motor. The latter is mechanically coupled through reduction gearing to the second frequency-change oscillator.

Monitoring

Aural monitoring is provided by beating the second intermediate frequency signal of either path against an oscillator. For tuning purposes the oscillator is set to 100 kc/s, correct tuning being indicated by zero beat in the case of on–off signals and a continuous uniform note for frequency-shift-keyed signals. The oscillator can be offset by 1 kc/s to give a readable on–off signal.

A pair of simple valve-voltmeter circuits provide visual indication of the path levels.

Tuning Oscillator

This is not part of the receiver circuit proper, but is introduced to facilitate the tuning of the signal-frequency stages. It can be patched into either receiver path, and provides a steady source of signal, so that it is then possible to tune each circuit accurately to give maximum output as indicated on the path-level meters.

Operational Controls

After the receiver has been set up for operation on the three chosen frequencies, the front doors may be closed. Apertures in the doors provide access to the operational controls, which are reduced to the absolute minimum, and comprise: (1) frequency-selector switch; (2) aerial attenuator; (3) fine tuning control; (4) A.F.C. on–off switch; (5) overall gain controls; (6) bias control; (7) monitoring jacks; (8) path level and bias meters.

Summary of Performance

The following summary of the characteristics of the telegraph receiver already described is given as a general guide to the performance of this class of receiver.

General.—Pre-tuned or continuously variable double superheterodyne telegraph receiver capable of immediate selection of three spot frequencies.

Service.—Double-diversity reception of telegraph signals either C.W. on–off 200 bauds, or frequency-shift key 200 bauds, 400–1,000 c/s shift.

Frequency Range.—3–27·5 Mc/s in three bands.

Sensitivity to Record.—C.W. on–off 0·3 μV at 27·5 Mc/s with 2 kc/s passband.

Image Signal Protection.—117 db at 3 Mc/s; 68 db at 15 Mc/s; 70 db at 27·5 Mc/s.

Stability.—Crystal first oscillator 1·5 parts in 10^6 per ° C. Variable first oscillator 5 parts in 10^6 per ° C. Variable second oscillator 5 parts in 10^6 per ° C.

Oscillator Radiation.—The maximum level at the aerial terminal due to the first oscillator is 10 μV.

Second Intermediate-frequency Selectivity.—

1 *kc/s filter*: 1 kc/s wide at 1 db attenuation; 1·2 kc/s wide at 10 db attenuation; 1·9 kc/s wide at 70 db attenuation.

2 *kc/s filter*: 2 kc/s wide at 1 db attenuation; 2·3 kc/s wide at 10 db attenuation; 3·5 kc/s wide at 70 db attenuation.

Automatic Frequency Control.—Capable of following drifts up to $\pm$3,000 c/s with a resultant error of not more than 10 c/s.

Overloading.—An adjacent channel signal, spaced at 2 kc/s from the working frequency for the 1-kc/s pass band, and at 60 db above the wanted signal will not cause blocking in the receiver, provided that the wanted signal does not exceed 100 μV.

When using the 2-kc/s filter, the same result is obtained from a signal 3 kc/s off tune.

Output.—Maximum of 30–0–30 mA into earthed load not exceeding 1,500 ohms.

Telephone Receiver

General

The block diagram of a single-sideband receiver intended for point-to-point telephone working is given in Fig. 20. In appearance it is similar to the telegraph receiver.

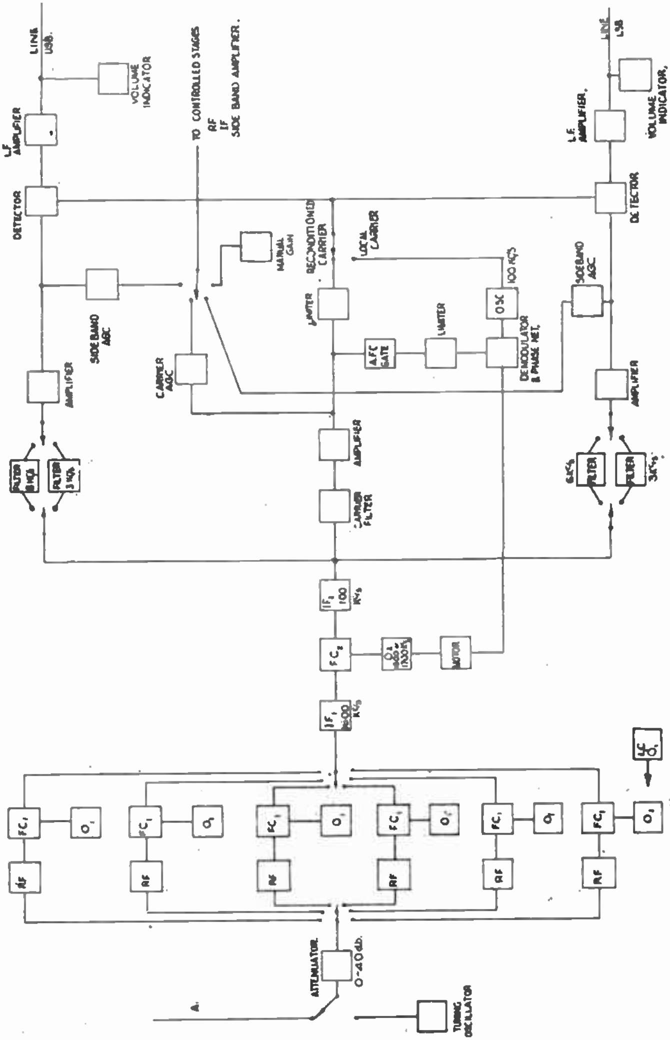


FIG. 20.—TELEPHONE RECEIVER: SIMPLIFIED BLOCK DIAGRAM.

The performance of single-sideband telephone systems is such that it is generally agreed that there is little practical advantage in adopting diversity reception. This equipment follows the current practice, and is a single-path or non-diversity receiver.

The radio-frequency amplifiers and first-frequency change oscillators are similar to those used in the telegraph receiver described, but a total of six pre-set frequencies is provided.

Good image signal protection is obtained by centring the first intermediate-frequency amplifier at 1·6 Mc/s, and the sideband and carrier separation is carried out at 100 kc/s. The second-frequency change oscillator may be controlled manually or by the automatic frequency-control circuits to the extent of ±3 kc/s. The carrier is used to operate the A.F.C. and A.G.C. circuits, and is reconditioned so as to provide an alternative to the local carrier.

First Intermediate-frequency Amplifier

The mid-band frequency is 1·6 Mc/s, the pass-band is 18 kc/s at the 1 db attenuation points to accommodate two sidebands of 6 kc/s each and the fine tuning spread of ±3 kc/s. An overall gain of approximately 45 db is obtained from two stages. A total of ten tuned circuits is necessary to obtain the desired linearity over the pass band.

Second-frequency Change Oscillator

A convention has been adopted in single-sideband working whereby the upper and lower sidebands are interchanged about the 10-Mc/s radio-frequency point. To cater for this inversion the second oscillator may be set to either 1·5 or 1·7 Mc/s.

Temperature compensation and protection against humidity effects are provided, and the frequency may be varied ±3 kc/s either manually or by automatic-frequency-control action.

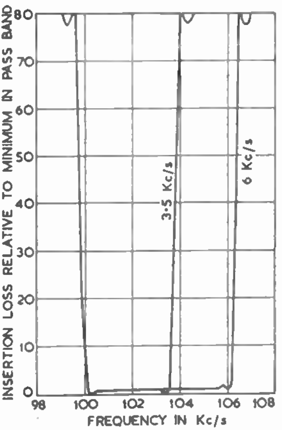


Fig. 21.—Upper Sideband Filter Response Curves.

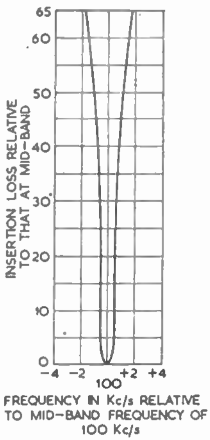


Fig. 22.—Carrier Filter Response Curves.

Sideband Amplifiers

Alternative band-widths of 3 and 6 kc/s are provided for each sideband. The selectivity is provided by crystal filters which are temperature compensated and sealed against the effects of corrosion and humidity. Typical response curves of the upper sideband filters are given in Fig. 21. The lower sideband filters have a similar performance, which is the mirror image about 100 kc/s.

A common single-stage amplifier precedes the filters, and separate three-stage amplifiers follow the filters.

Carrier Amplifier

The pilot carrier is separated from the sidebands by means of the carrier filter, which is only 60 c/s wide at the 2-db attenuation points. This narrow band-width is essential for two reasons, first: it enables a good signal-to-noise ratio to be obtained from the attenuated carrier, and secondly, it offers maximum rejection to the low-modulation frequencies of the sidebands.

Demodulation

The detectors, which are of the balanced type, are supplied with a constant carrier level 20 db above the peak sideband level. Two sources of carrier are available, for whilst it is generally preferable to use the local carrier, on occasions when the signal is subject to spurious frequency modulation the reconditioned carrier is to be preferred.

Automatic Gain Control

This presents a problem in single-sideband receivers, and the present solution is not too satisfactory. As the sideband level is variable, there is no alternative but to operate the A.G.C. system from the carrier. During periods of selective fading, in addition to the normal comparatively slow fading, the carrier will be subject to rapid fades, and to smooth out this effect the A.G.C. system is given a very long time constant. Some variation in sideband level is bound to result, and to offset this it is the usual practice to include constant-volume amplifiers in the radio-telephone terminal.

If one sideband signal consists of a constant-level frequency-modulated tone, as in the case of facsimile transmissions, it is then more satisfactory to operate the A.G.C. system from the sideband. A switch is included in the receiver enabling the A.G.C. circuits to be operated from the carrier of either sideband at will.

Automatic Frequency Control

When using the local carrier for demodulation purposes, any error in tuning has the effect of altering the modulation frequencies. It will be appreciated, then, that only a small tuning error can be tolerated, especially in the case of high-quality music transmissions.

With reconditioned carrier working a reasonable tuning error does not affect the quality of the audio-frequency output, since both sidebands and carrier are mistuned to the same amount, and the difference frequency is unaltered.

The A.F.C. system adopted in this receiver ensures that tuning is accurate to a fraction of a cycle per second.

The output from the carrier amplifier is taken to a limiter via a gating stage which is open when the signal exceeds a predetermined amount: this prevents the operation of the automatic-frequency-control circuits by noise during periods of deep carrier fades. Following the limiter is a demodulator, which is also supplied with an output from the local carrier oscillator. A four-phase output is obtained from the difference frequency between the received carrier and the local carrier, and is used to operate a four-phase impulse-type motor, which in turn corrects the tuning of the second oscillator.

Line Output and Controls

The audio-frequency output from the demodulator is fed to a two-stage amplifier which has low harmonic and frequency distortion characteristics.

As in the case of the telegraph receiver, the controls necessary for normal operation have been reduced to a minimum. These are: (1) aerial attenuator; (2) frequency selector switch; (3) fine tuning control; (4) A.F.C. on–off switch; (5) A.F. gain controls; (6) volume indicators and carrier level meters.

Summary of Performance

General.—Pretuned or continuously variable double superheterodyne S.S.B. receiver capable of immediate selection of any one of six spot frequencies. Single-path reception of D.S.B., S.S.B., or I.S.B. telephone signals.

Noise Factor.—6–8 db, according to frequency.

Frequency Range.—3–27·5 Mc/s.

Image Signal Protection.—117 db at 3 Mc/s; 68 db at 15 Mc/s; 70 db at 27·5 Mc/s.

Stability.—Crystal first oscillator 1·5 parts in 10^6 per ° C. Variable first oscillator 5 parts in 10^6 per ° C. Variable second oscillator 5 parts in 10^6 per ° C.

Oscillator Radiation.—Less than 10 μV at any frequency measured at the aerial terminal.

Second Intermediate-frequency Selectivity.—Sideband filters: Pass band 100–3,500 c/s. 100–6,000 c/s.

Discrimination against frequencies more than 520 c/s outside the pass band is greater than 75 db.

Carrier filter: Pass band ± 30 c/s.

A.F.C.—Up to ± 3 kc/s drift corrected with a residual mistune of less than 1 c/s.

Overall Frequency Response.—Variation in output for constant-amplitude input signals is less than 3 db throughout the pass band.

TABLE 1.—TYPES OF EMISSIONS USED FOR HIGH-FREQUENCY RADIO LINKS

Type of Modulation	*Type of Transmission*	*Supplementary Characteristics*	*Symbol*
Amplitude modulated	Telegraphy without the use of modulating audio frequency (on–off keying)	—	A1
	Telegraphy by the keying of a modulating audio frequency or audio frequencies or by the keying of the modulated emission	—	A2
	Telephony	Double-sideband, full carrier	A3
	Telephony	Single-sideband, reduced carrier	A3a
	Telephony	Two independent sidebands, reduced carrier	A3b
Frequency modulated	Telegraphy without the use of modulating audio frequency (frequency-shift keying)	—	F1

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F. W. J. S.

17. BROAD-BAND RADIO SYSTEMS

17. BROAD-BAND RADIO SYSTEMS

Introduction

Systems having large-channel band-widths are playing an ever-increasing part in modern communications networks. They range from the simple television link [1,2] used for connecting outside broadcast cameras with broadcasting stations, to the transcontinental network [3] capable of carrying either television channels or many speech channels.

Permanent point-to-point links for television are important because of the limited area coverage of television broadcasting stations, and since the cost of television programmes is high, it is desirable to have simultaneous broadcasting over a national circuit. Furthermore, live programmes, particularly of sport, have a great appeal. An alternative to permanent-link stations is to have overlapping broadcast areas served by different frequencies with receivers at each station to hand on the programme. Unfortunately, a high density of broadcast coverage is seldom economically justified, and high-power stations are not usually sufficiently free from distortion to permit much handling in this way.

Many hundreds of telephone channels often have to be transmitted over a trunk route between large cities. Until recently these trunk circuits have been provided by cables. However, over routes where no cable ducts exist, it seems that a radio system can be manufactured and installed more rapidly than a cable system, and a radio system can show a great saving in raw materials such as copper and lead.[4] Other factors, such as the lack of roads and power supplies for isolated radio stations sited on hills, can put cables at an advantage because they can carry their own supplies for repeaters.

It is the general aim that the large trunk radio networks should have transmission characteristics suitable for either television or multi-channel telephony, and should conform to the standards used in co-axial cable networks. The latter handles telephony channels assembled in frequency division multiplex.

Such radio systems use frequency modulation in preference to amplitude modulation because greater linearity, better signal-to-noise and more stable transfer levels are obtainable. Amplitude linearity is most stringent for multi-channel telephony, and usually the number of telephony channels carried is less than would occupy the band-width necessary for television.

As large radio-frequency channel band-widths of the order of 16 Mc/s are used, the largest of these systems have to operate in the micro-wave bands. Consequently, the separation between stations is limited to optical ranges, and with current practice this averages a little under 30 miles. To cover great distances a number of repeater stations are used between the terminals. There are two types of repeater station, the first is essentially a receiver and transmitter connected together at the base-band frequency (that is the television or multi-channel assembly frequency). The second amplifies and changes carrier frequency without returning to the base-band frequency. The first type is used if some channels have to be taken-up or dropped-off at the repeater station, otherwise the second type is preferred, as it introduces less distortion.

The frequency bands that are generally available for broad-band radio relays systems are: 1,700–2,300 Mc/s; 3,500–4,200 Mc/s; 5,925–8,500 Mc/s. In addition, the band 890–940 Mc/s, or part of it, is available in many countries for medium band-width systems.

Other methods of modulation and multiplexing are also in use. Time division multiplex is attractive for radio transmission, as performance is less dependent upon the circuit variations usually associated with radio equipment. As the radio-frequency band-width per speech channel is large, such systems are seldom designed to carry more than twenty-four channels, and perhaps sixty channels is an economic limit. Radio circuits that have linearity just adequate for television can be used successfully for the transmission of several high-grade music channels, or about twenty-four speech channels by the use of pulse position modulation. A great disadvantage with pulse systems is that they do not readily integrate with existing cable circuits. Many thousands of channel miles are giving good service in self-contained systems.

Pulse code modulation is particularly suitable for long-distance networks not involving cable links, or under conditions when the radio stations cannot be sited to obtain good carrier-to-noise ratios. The signal-to-noise ratio in the speech circuit is substantially independent of the carrier-to-noise ratio after a certain minimum is exceeded, and is also independent of the circuit length. So far, these special advantages obtained by Pulse Code Modulation have been extensively made use of only in data transmission; and even then mostly by the modulation of a sub-carrier rather than by direct modulation of the radio carrier.

In some parts of the world, particularly in North America, where some 150,000 broad-band channel miles are in service, radio relay engineers are becoming faced with the problem of frequency-spectrum conservancy. In the next few years this is likely to cause increased interest in single-sideband modulation. Without using frequency-compression techniques, single-sideband transmission can reduce the required frequency spectrum by at least five to one as compared to the present-day frequency modulation demands. Furthermore, there are interesting possibilities in unit construction. However, development problems are difficult because the linearity problems are severe.

Some multi-channel systems operate in the V.H.F. bands, but these bands are narrow and congested. Therefore, the number of speech circuits per radio channel is usually limited to about forty-eight, and the accepted performance is frequently lower than that obtained from microwave systems. However, the separation between stations may greatly exceed optical ranges, often by as much as two to one. Equipment can be built with much lower mechanical precision than that required for the centimetric bands. Thus the use of the V.H.F. bands is attractive when capital outlay has to be kept to a minimum, and when long hops are essential for geographical reasons.

Some very attractive low-cost radio-relay equipments have been made for use in the 450- and 900-Mc/s bands. They have many of the advantages of a true micro-wave system and yet are mainly designed around "lumped-circuit" techniques. A capacity of 120 speech channels is not unusual. The available frequency bands are narrow, and in many countries are not very well protected from interference.

STANDARDS OF PERFORMANCE

Television

The subject of transmission performance for long-distance transmission of television is receiving much international attention.[6] It is complicated by the several standards of definition in vogue; a summary of these standards is given in the documents of the C.C.I.R.[7]

For monochrome television of 405-line definition the following extracts from [7], together with additional comments, indicate the standards so far accepted in the United Kingdom.

(1) The input and output impedance of the transmission circuit should be 75 ohms unbalanced, and have a return loss of not less than 30 db at any frequency between 10 and 3,000 kc/s.

(2) The normal transfer level is about 1 volt peak-to-peak. For transmission to cable the video signal has positive polarity (i.e., its amplitude will increase in proportion to the brilliance of the corresponding point of the picture), and the D.C. component is not transmitted.

(3) The slope of the output-voltage to input-voltage characteristic for a nominally zero gain video-to-video channel should be within the limits 0·9–1·1 over the maximum range of voltage due to the picture signal. The synchronizing signal amplitude, nominally 30 per cent of the total signal amplitude, should be between 21 and 33 per cent of the total signal amplitude.

(4) Stability of overall transmission circuit.

(*a*) Short-period variations (e.g., 1 second) not greater than $\pm 0{\cdot}3$ db.

(*b*) Medium-period variations (e.g., 1 hour) not greater than $\pm 1{\cdot}0$ db.

(*c*) Long-period variations (e.g., 1 month) for unattended circuits not greater than ± 2 db.

(5) *Signal-to-noise Ratio.*—The tolerable signal-to-noise ratio for:

(*a*) Random uniform noise—35 db.

This is expressed as:

$$\frac{\text{Peak-to-peak amplitude of picture signal (black to white)}}{\text{Quasi-peak-to-peak amplitude of noise}}$$

A further allowance must be made for fading, as it is usual that this figure of 35 db is required to be met for not less than 99 per cent of time.

It is generally assumed that the quasi-peak-to-peak amplitude of random noise is seven times its r.m.s. voltage.

Thus,

$$\frac{\text{Black-to-white signal (peak-to-peak)}}{\text{Noise voltage r.m.s. (flat)}} = 52 \text{ db}$$

For random noise in a unit band-width which increases with frequency at a rate of 6 db/octave, 8 db greater noise can be tolerated.[4] Such an allowance can be made for the noise appearing at the output of a frequency-modulation system.

Thus,

$$\frac{\text{Black-to-white signal (peak-to-peak)}}{\text{Noise voltage r.m.s. (tilted 6 db/octave)}} = 44 \text{ db}$$

(*b*) Periodic noise—40–55 db, depending upon frequency.
(*c*) Impulsive noise—35–40 db, if there is not more than one pulse in a 10-second period.

(6) *Speed of Transient Response.*—Using a square waveform with a rise time not exceeding 0·05 microseconds, the rise time from 10 to 90 per cent of the overall amplitude should not exceed 0·16 microseconds.

(7) *Amplitude/frequency Characteristic.*—See Table 1.
(8) *Phase/frequency Characteristic.*—See Table 2.
(9) *Limits for Overshoot and Echoes.*—See Table 3.

TABLE 1.—AMPLITUDE/FREQUENCY CHARACTERISTIC

Frequency kc/s	*Up to 500*	500–1,000	1,000–2,000
Variation from nominal (db)	±1·	±1·5	±2·5

TABLE 2.—PHASE/FREQUENCY CHARACTERISTIC

Frequency (kc/s)	200–2,000	2,000–2,500	2,500–2,800
Variation from nominal Group Delay in microseconds (see Appendix I)	±0·15	±0·25	±0·5

TABLE 3.—LIMITS FOR OVERSHOOT AND ECHOES USING A SQUARE WAVEFORM WITH A RISE TIME NOT EXCEEDING 0·05 MICROSECONDS

Time after Response has Reached 50% Ideal Amplitude	*Limits of Rapid Variation of Response, % of Ideal Amplitude*
0·2–0·5 microsecond . . .	±4
0·5–1·0 microsecond . . .	±1
1·0 microsecond or longer . .	±0·5

Transmission standards for colour television have not been specified. In the United States and Canada broad-band radio-relay systems carry colour-television programmes using a technique which allows existing television receivers to accept the signals for monochrome reproduction. The transmission standards for colour transmission are much more severe than for monochrome, and the following summary indicates the order of the requirements that would be imposed upon a transmission system linking the camera to the broadcasting station for a 525-line system:

(1) *Amplitude response*: D.C. to 4·3 Mc/s response ±2 dB relative to response at 200 kc/s.

(2) *Group delay characteristic*—

(*a*) The group delay variation over the band of frequencies of D.C. to 4·3 Mc/s should not vary by more than ±100 milli-microseconds for a test frequency of constant level.

(*b*) A group delay variation in any part of the baseband, exceeding ±1·0 milli-microseconds should not be produced by a change of signal level from zero to peak value.

Similar compatible colour signals to European (625-line) standards would require a greater band-width, about 5 Mc/s, but the Group Delay requirements would be about the same.

Multi-channel Telephony

It is generally recognized that trunk radio circuits should comply with the agreed international recommendations for co-axial cables.[8,9,10] The detailed characteristics for radio circuits are still being studied internationally,[9,10] and recent international meetings left no doubt that radio and cable links would have to operate together in one network.

Noise from all causes is limited to 10,000 micro-micro-watts psophometric (see Appendix II) at a point of zero relative level (1 mW) at the receiving terminal of a 2,500-km circuit.[8] This limit must not be exceeded for more than 1 per cent of the time during the busy hour. Of the 10,000 micro-micro-watts, 2,500 are allocated to the multiplex terminal equipments and the remainder is made up at the rate of 3 micro-micro-watts per km.

Over a distance of 2,500 km a typical circuit would be required to make accessible channels assembled in supergroups (sixty channels) or groups (twelve channels) at nine intermediate points.

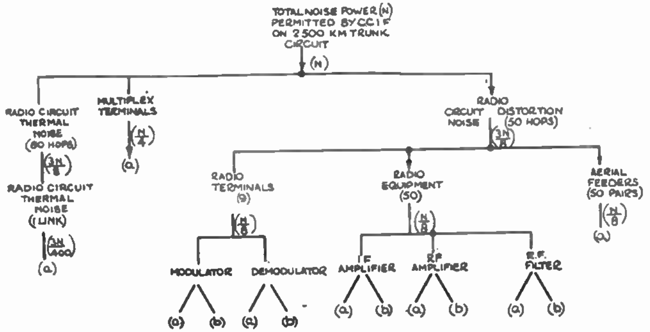


Fig. 1.—Distribution of Noise in a Typical Broad-band System.

(*a*) Noise adding on power basis, including circuit; (*b*) noise tending to add on voltage basis. N is the psophometric noise power at a point of zero relative level (1 mW) in an audio channel equal, to 10^{-8} watts.

Fig. 1 shows a typical analysis for the contributions of circuit and distortion noise from various units of a radio system, with speech channels in frequency division multiplex, frequency modulating the radio carrier.

TABLE 4.—FREQUENCY DIVISION MULTIPLEX CHARACTERISTICS

Number of Channels	24 (*2 groups*)	60 (*1 super-group*)	120 (*2 super-groups*)	180 (*3 super-groups*)	240 (*4 super-groups*)
Channel test level, r.m.s. (dbm) (Zero relative level)	0	0	0	0	0
Multi-channel level, r.m.s., long-term average (dbm)	−1	+3	+6	+7·5	+9
Multi-channel level, r.m.s., exceeded 1% of the time (dbm)	+7·5	+9	+10·5	+11·5	+12
Random noise test signal level, r.m.s. (dbm)	+7·5	+9	+10·5	+11·5	+12
Sine-wave capacity required for multi-channel signals (r.m.s.)					
V. ratio $\frac{\text{Sine-wave capacity}}{\text{Channel test level}}$ (p)	9·5	10	11	12	13
(dbm)	+19·5	+20	+21	+21·5	+22
Frequency band (kc/s)	12–108	60–300 300–552	60–552	60–804	60–1052

There are no universally agreed characteristics, but the figures given in Table 4 are representative of values generally accepted by engineers. Recent information accumulated by some telephone administrations indicates that the multi-channel level quoted above may be slightly high for speech circuits. However, for systems carrying data channels and compressed speech the loading can become higher.

The choice of channel test level is purely arbitrary, and a level of 0 dbm is chosen for convenience. The long-term average r.m.s. multi-channel level is of little value when considering system loading; it will be seen that the average power increases proportionately to the number of channels. The multi-channel r.m.s. level, which is only exceeded for 1 per cent of the time under heaviest traffic loading, is taken as the important factor when noise due to inter-modulation effects is considered. This level is related to the fluctuating characteristic of speech mean power levels, but must not be confused with instantaneous power. The figures quoted are 2 db higher than those obtained by Holbrook and Dixon [13] for speech. This gives some margin for carrier breakthrough, voice-frequency telegraphy loading and pilot signals. For test purposes random noise can be used to simulate a multi-channel signal (see later under "Overall Performance" of F.D.M. transmission), and it is considered that a reasonable test level is that of the r.m.s. multi-channel level which is only exceeded for 1 per cent of the time. The C.C.I.R. recommends a noise level of $(-15 + 10 \log_{10} N)$ dbm when the number of channels (N) exceeds 240. This level is lower than would be suggested by the above table, when the number of channels is less than about 1,000.

The sine-wave capacity is also extremely important, because it defines the maximum voltage, or frequency excursion, that the transmission circuits must handle; the figures quoted are obtained from the results given in reference [13], but a further margin of up to 6 db is not unusual.

References [8] and [14] are relevant.

The impedance at which circuits are connected at base-band frequencies is normally 75 ohms, unbalanced, except when only the groups between 12 and 552 kc/s are used. Then the impedance may be 150 ohms balanced. If co-axial cable circuits are attached to the input or output of the equipment, it is necessary to minimize reflections, and a minimum return loss of 20 dB is usually specified for all frequencies within the band. When the impedance is 75 ohms, the relative input and output power levels, per channel, to the radio equipment are recommended to be −52 and −15 dbm, respectively..

TYPICAL FREQUENCY-MODULATION SYSTEMS

In the simplest arrangement alternate links operate on alternate frequencies, f_1 and f_2, see Fig. 2. The GO and RETURN channels may use the same pair of frequencies, but these are staggered to make the transmit and receive frequencies different at any one repeater station. Sometimes it is not possible to site the stations to avoid interference between the go and return channels, even when the radiated fields are cross-polarized. A "zig-zag" route can also lead to interference between links separated by two or more repeaters. In such cases additional frequencies are required, and it is not unusual to find that as many as twelve frequencies are required in some systems.

The C.C.I.R. have recommended a preferred R.F. channel arrangement for up to Six go and Six return channels each accommodating 600 speech channels, and operating at frequencies above 1,000 Mc/s. This arrangement places the go channels in one group and the return channels in another. The inner channels of each group are separated by a guard band of 213 Mc/s, and the alternate channels within each group are radiated with vertical and horizontal polarization respectively. The frequency separation between these channels is 29 Mc/s. If additional R.F. channels are required, these would be placed half-way between the channels, separated by 29 Mc/s. It should be mentioned that most of the systems that have been installed have not used this frequency plan. Nevertheless, it is indicative of good practice.

Almost all broad-band systems for television, and large numbers of speech channels, use frequency modulation. Most of the fundamental factors are discussed in reference [15].

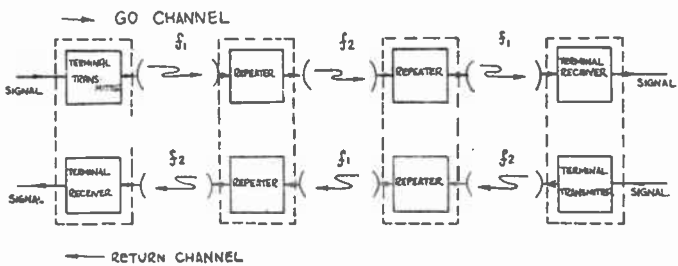


FIG. 2.—FREQUENCY PLAN FOR A BROAD-BAND RADIO SYSTEM.

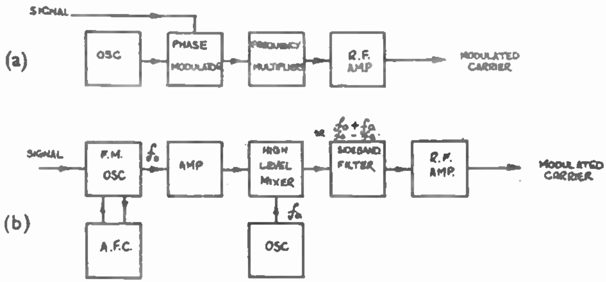
Terminal Transmitters

Although sufficient linearity is readily obtained for television transmission, it becomes one of the major problems when the number of speech channels exceeds a few hundred. A single television link requires a stability of not greater than several hundred kc/s, but a long multi-channel speech system may require a frequency stability of 50–100 kc/s.

Fig. 3 shows block diagrams of typical transmitter terminals. Phase modulation (Fig. 3 (*a*)) of a low-frequency carrier, followed by frequency multiplication, can provide good linearity and adequate frequency stability without elaborate circuits. Difficulties, however, arise in suppressing unwanted frequencies, especially if any frequency conversion is necessary. Phase modulation is not satisfactory for television transmission. Thus this process tends to be restricted to the transmission of fairly small numbers of speech channels, and is of particular importance for V.H.F. and lower U.H.F. systems.

Frequency modulation at a frequency lower than the required radio frequency and translation by a frequency-mixing process to the required band is sometimes used in micro-wave systems.[3,16] It has the advantage that the modulator can be carefully designed for operation on a suitable frequency, and the final frequency simply determined by a separate oscillator and filters.

Velocity-modulated oscillators of the Klystron type can give reasonably linear frequency modulation by varying the reflector or resonator voltage relative to cathode.[1,2,17,18] In simple television links the oscillator usually feeds the aerial directly. For permanent systems radio-frequency amplification and isolation of the oscillator from the



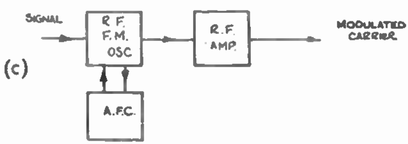


FIG. 3.—TYPICAL TERMINAL TRANSMITTERS.

(*a*) Phase Modulation.
(*b*) Frequency Modulation at a Low Frequency and Frequency Translation.
(*c*) Frequency Modulation at Radio-frequency.

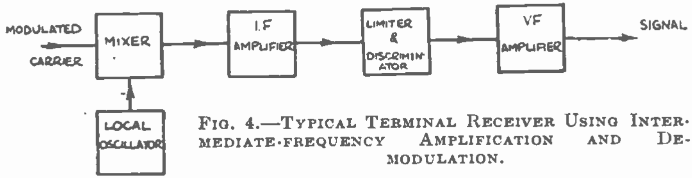


FIG. 4.—TYPICAL TERMINAL RECEIVER USING INTERMEDIATE-FREQUENCY AMPLIFICATION AND DEMODULATION.

aerial is obtained by triodes,[3,16] or travelling wave tubes.[18,26] The output power is usually of the order of a few watts.

Frequency-modulated oscillators are not usually inherently stable in frequency, and automatic frequency control circuits apply correction voltages, to reduce the frequency error. These automatic-frequency-control circuits are fundamentally conventional, but problems are introduced by the broad-band characteristic of the signal. For multi-channel F.D.M. signals the frequency deviation, due to the modulation, has an average value of zero. Consequently, a discriminator followed by a low-pass filter with a cut-off just below the modulation band will have an output voltage proportional to the error frequency. For television, the output of the discriminator is only "gated" during the period at the bottom of the synchronizing pulse.

Both phase and frequency modulation characteristics are used. However, fixed noise and distortion noise usually limit the useful range of pre-emphasis, with an F.M. characteristic, to less than 10 db, and it is common to find no pre-emphasis in the large capacity systems. Frequency deviation for a tone at channel test level ranges from 35 to 200 kc/s, r.m.s. for systems of 24–240 speech channels or more.

Terminal Receivers

Most of the present-day receivers use receivers of the heterodyne type. Local oscillators are of two types: (*a*) crystal-controlled oscillators with fairly high orders of multiplication; and (*b*) frequency-controlled micro-wave oscillators with reference cavities.

With the development of low noise travelling-wave tubes receivers are now being built with low distortion and much improved noise factors. They are simpler to maintain, and have more freedom from spurious responses, but there are selectivity problems.

It can be anticipated that the parametric amplifier will eventually replace the first travelling-wave tube, because of its lower noise factor.

Repeaters

There are two types of repeater station:

(i) Back-to-back terminal transmitters and receivers. These are used where access to the modulation band is necessary for dropping and taking-up speech channels.

(ii) Non-demodulating (or "through") repeaters. Since no modulation, or demodulation, process is used, less distortion is

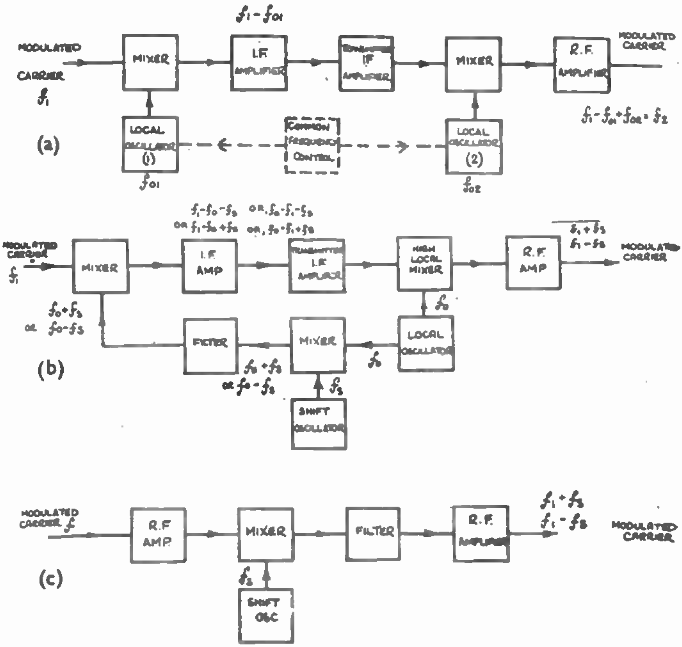


FIG. 5.—TYPICAL NON-DEMODULATING REPEATERS.

(a) Simple frequency shift repeater using intermediate-frequency amplification.
(b) Frequency shift repeater using intermediate-frequency amplification and in which the shift frequency is generated by a separate oscillator.
(c) Frequency shift repeater using radio-frequency amplification.

introduced by these repeaters than by the back-to-back terminals. Fig. 5 shows typical non-demodulating repeater arrangements.

At a repeater it is usual to receive on one frequency and transmit on another. Fig. 5 (a) illustrates how this can be accomplished when most of the amplification is obtained at an intermediate frequency. Frequency errors are cumulative in a system which has repeaters, and cause increased distortion. Therefore in practice with such a simple arrangement as shown in Fig. 5 (a) great frequency stability of the local oscillators would be required. If the local oscillators at a repeater are both tuned high, or both low, with reference to their respective channel frequencies, then the frequency shift introduced by the repeater is the frequency difference between the two oscillators. One way to determine accurately this difference frequency is to control both oscillators to a common frequency standard, such as the consecutive harmonics from a crystal. This introduces errors due to frequency control, and without frequency changing in the reference chain does not allow free choice of channel frequencies. A variation which overcomes this disadvantage controls the first oscillator to a reference, and the second oscillator to maintain a constant difference frequency with the first.[19] Fig. 5 (*b*) shows an alternative in which the shift frequency is added to that of the local oscillator. As the shift frequency is usually fairly small (35–200 Mc/s) and can be generated by a crystal oscillator, little frequency error is introduced.[16]

Except for the transmitter intermediate-frequency amplifier, the high-level mixer and the frequency shift circuits, a repeater includes units of design similar to those used at the terminals.

Fig. 5 (*c*) illustrates the possibility of a repeater in which all the gain is obtained at radio-frequency.[20] With travelling-wave tubes the reflex technique has been successfully employed. This saves one amplifier tube by using the same tube to amplify the signal at two different frequencies, that is before and after the mixer. In this way a repeater has been built using only three tubes.

Limited channel-drop and insert facilities may be provided at "through" repeater stations by branching part of the amplified signal into a receiver, and modulating the shift oscillators. In this case, all the stations are essentially operating on an omnibus circuit and superimpose new signals on existing ones. Consequently, such drop and insert facilities are usually restricted to the lower part of the base-band, for example up to 108 kc/s. This would provide service, supervisory, remote control and twenty-four speech circuits.

RECEIVED CARRIER POWER

Allowance must be made for all the losses between the transmitter and input circuit of the receiver.

$$P_r = P_t - A_{ft} - A_{lt} + G_{at} - A_p - A_f + G_{ar} - A_{lr} - A_{fr}$$

where P_r = received power in db relative to 1 watt;
P_t = transmitted power in db relative to 1 watt;
A_{ft} = transmitter output-filter loss in db (see "Microwave System Design");
A_{lt} = transmitter feeder loss in db (see "Microwave and V.H.F. System Design");
G_{at} = transmitter aerial gain over isotropic radiator, in db (see "Microwave and V.H.F. System Design");
A_p = normal free space propagation loss between isotropic radiators, in db (see "Path Loss");
A_f = fading allowance, including additional path losses, in db (see "Path Loss");
G_{ar} = receiver-aerial gain over isotropic radiator, in db (see "Microwave and V.H.F. System Design");
A_{lr} = receiver feeder loss, in db (see "Microwave and V.H.F. System Design");
A_{fr} = receiver input-filter loss, in db (see "Microwave System Design").

PATH LOSS

Path loss receives much attention when a high-grade system is being designed, because economy demands the greatest possible spacing between repeaters. At the same time an uninterrupted high standard of performance must be guaranteed.

It is conventional to express the noise level of high-grade communications systems as the value not exceeded for 99 per cent of the total time. It has been frequently accepted that the fluctuating level of circuit noise caused by changing path losses in a radio system should conform to this convention. However, a difficulty arises due to the variations of rate and duration of occurrences of fading through the year. For example, the level of noise exceeded for 1 per cent of the worst hour is very much more pessimistic than the level exceeded for 1 per cent of, say, one year.

In considering the loss of a link, it is usual to take the normal path loss and fading separately. If there are a number of links in cascade, some advantage is gained by the peak noise contributions from each link not being simultaneous, owing to unrelated fading behaviour.

In a long system the increased noise due to fading is not usually of great importance. However, improved reliability and signal-to-noise ratio can be obtained by carrying the same information over two radio circuits operating on different frequencies, and combining the signals at the receivers.

U.H.F. and S.F. Path Loss

The loss between two hypothetical isotropic radiators over a path well clear of obstructions and interference is given by :

$$20 \log_{10} \frac{4\pi l}{\lambda} \text{ db}$$

where l is the path length, and λ is the wavelength (both expressed in the same units).

In addition to this loss there are other causes of attenuation :

(a) *Rain, Mist, Fog, Oxygen and Water Vapour.*[22]—Up to 2,000 Mc/s attenuation due to these causes is insignificant. At 2,000 Mc/s absorption due to oxygen and water vapour will produce about 0·2 db loss over a 30-mile path.

At 4,000 Mc/s the loss caused by rain and fog is slightly greater. Very heavy rain (20 mm./hour) could increase losses by about 1 db over a 30-mile path.

At 7,000 Mc/s heavy rain could cause a loss of 10 db over a 30-mile path.

At 10,000 Mc/s heavy rain could be a really important factor, causing losses of up to about 30 db over a 30-mile path.

Above 30,000 Mc/s losses due to absorption and rain both become excessive.

(b) *Refraction Due to Water Vapour Content and Temperature Changes of the Air.*—Reception is usually steady when the atmosphere is turbulent, but under certain conditions the signal may be guided by variations in dielectric constant with height. Occasionally the beam may be curved in such a way as not to reach the receiver.[23] In one reference [39] it is suggested that the effect of

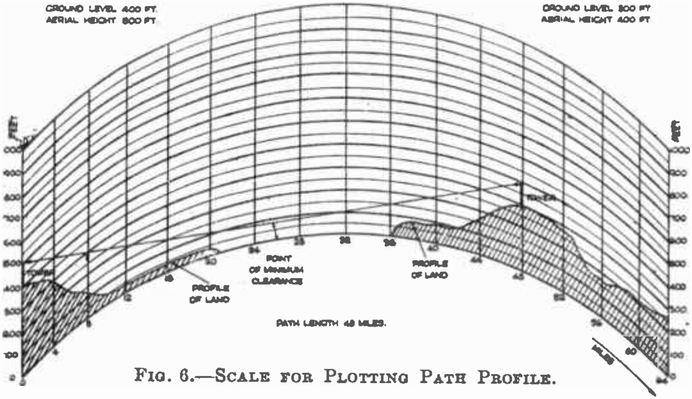


FIG. 6.—SCALE FOR PLOTTING PATH PROFILE.

bending of the radio beam is equivalent to a change of the effective radius of the from " flat-earth " to $\frac{4}{3}$ radius.

(c) *Diffraction and Losses Due to the Wave Grazing an Obstacle.*—It can be shown that the ray is bent by an obstacle in its path, and that losses can occur. Trees and buildings (whose heights are not usually included in survey maps) have often been the cause of the path loss being unexpectedly high. The loss is small with large clearance, and it is common practice to have aerials above the height at which the direct ray and a ray grazing the earth's surface are out of phase (that is having a difference in path length of $\lambda/2$). This is often referred to as the " First Fresnel Surface ", and the relationship between path length and clearance is :

$$h^2 = \lambda/(1/d_1 + 1/d_2)$$

where h is minimum clearance is metres, and d_1 and d_2 are the distances between the aerials and point of minimum clearance in metres. For heights less than this the losses can be expected to increase, for example, with the line of sight grazing a ridge the losses may be 5 db. For greater heights the normal received signal level changes by only about ± 1 db with change of height.

When the path is over a surface which will produce good reflection,[39] such as water, snow, or salt flats, the effect of clearance on signal level is more pronounced and the elevation of the aerials is chosen to optimize the variation. Elevations to provide a minimum clearance of 0·25 first Fresnel zone assuming $\frac{4}{3}$ earth radius, to 1·7 for a flat earth are considered to be about optimum.

(d) *Two-path Interference.*—This is due to the variations of relative phases and amplitudes of the signal arriving by two different paths. These signals interfere and produce the characteristic cusp-shaped graph of received-signal strength against time. It has been observed that two-path interference often becomes appreciable only when the direct ray is reduced in strength. Multiple rays can result from the cause in (b).[23]

Fig. 6 shows the usual method of examining the route.

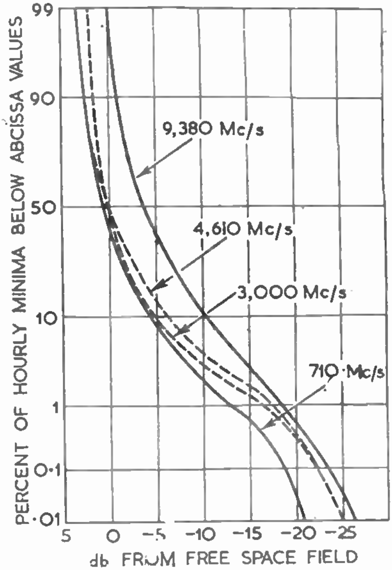


FIG. 7.—RELIABILITY OVER A CLEAR LAND ROUTE OF 40 MILES ON VARIOUS FREQUENCIES.

(*Durkee, 1 August* 1943 *to* 1 *February* 1945)

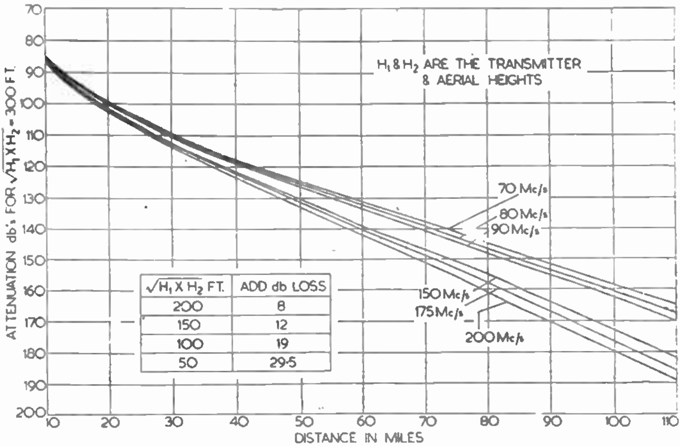
For a smooth earth the height of the aerials in feet necessary to obtain 50 ft. ground clearance is approximately:

$$\frac{l^2}{6} + 50$$

where l is the path length in miles.

Overall variations of path loss have been measured by Durkee[24] for frequencies of 710, 3,000, 4,610 and 9,380 Mc/s, and the results of his tests over a 40-mile clear overland path are given in Fig. 7.

The choice of polarization of the wave has negligible



$\sqrt{H_1 X H_2}$ FT.	ADD db LOSS
200	8
150	12
100	19
50	29·5

FIG. 8.—V.H.F. PROPAGATION LOSSES BETWEEN HALF-WAVE DIPOLES.

effect on fading. So far results have indicated that fading is worse over sea than over land paths.

V.H.F. Path Loss

Systems using the V.H.F. bands frequently operate over ranges in excess of optical clearance. Fig. 8, which is constructed from data due to K. Bullington,[25] gives the general relationship between aerial heights, distance and loss.

Ranges in excess of optical are largely attributable to refraction, which causes the beam to follow the earth's curvature. For convenience engineers have often preferred to think of the beam as straight, and the additional range being obtained by increased radius of curvature of the earth. The effect of variations in refraction are very important in considering the stability of signal level. The average value for the effective radius of the earth is about 4/3 true radius, but variations between $\frac{2}{3}$ true radius and twice true radius have been observed.

Fading is important only for ranges in excess of 25 miles : it is worse for oversea paths than for overland paths. High path attenuation is usually associated with large fading ranges, and ranges of more than 40 db over sea routes and 30 db over land routes are reached.

MICRO-WAVE SYSTEM DESIGN

Broad-band radio systems are relatively new, and development is still rapid.

Modulators

Reflex Klystron oscillators are used to generate a frequency-modulated carrier. Although the output level is about 1 watt, in permanent systems it is usual to connect the klystron into an attenuator (about 17–25 db) followed by a buffer amplifier, or a ferrite isolator, to remove the distortion effects of echoes in connecting cables. Portable television equipments generally have the modulator mounted behind the mirror and use no buffer amplifier.

Frequency modulation at a low frequency is obtained with various types of " phase-shift " oscillators. The mean frequency is often chosen to be the same as the receiver intermediate frequency to allow easy back-to-back tests, and to have a similar transmitter intermediate-frequency amplifier at terminals and repeaters.

Local Oscillators

Klystrons are used for this purpose in the higher-frequency bands, and automatic frequency control voltage is usually obtained from a discriminator and applied to the reflector of the klystron. Mechanical control has also been used. Triodes using quarter-wave (or three-quarter wave) anode circuits and three-quarter (or five-quarter wave) cathode circuits work very well up to about 4,000 Mc/s. The output noise, including microphony, is low. Frequency control is obtained by mechanical means, such as an auxiliary stub and bridge, or moving plunger.

Crystal oscillators working in the range of 10–60 Mc/s, and followed by multiplier stages using tetrodes, pentodes, U.H.F. triodes and klystrons are in current use. To avoid spurious responses low-multiples per stage are used, and good screening and filtering are essential.

Radio-frequency Amplifiers

There are two types of amplifier in general use :

(i) the triode;
(ii) the travelling wave tube.

Triodes have been used up to a little over 4,000 Mc/s.[3, 16] Travelling wave tubes of various types have been made with the object of obtaining a specific characteristic such as a noise factor of better than 10 db, a gain of 25 db or more or a power output of many watts.[18, 26]

Travelling-wave tubes have their beams focused by magnetic means using a solenoid, a barrel-shaped permanent magnet or periodically inverted ring magnets. Permanent magnets make the tubes expensive, consequently many designers prefer the solenoid, which also provides easy adjustment of the field. Electrostatic focusing has also been used. Travelling-wave tubes using electromagnetic focusing have achieved long operating lives, averaging about 15,000 hours.

Radio-frequency Filters

R.F. filters play a very important part in radio relay systems. Pairs of band-pass filters, tuned to different R.F. channels and coupled to a common feeder, make it possible for a transmitter and receiver to share the same aerial. More elaborate sets of filters make it possible for many go and return channels to use the same broad-band aerial. There are considerable problems in coupling a number of filters together, and there are several ingenious arrangements using R.F. hybrids and circulators which give some isolation between branches.

At 1,000 and 2,000 Mc/s co-axial stub radio-frequency filters are very convenient. For 4,000 Mc/s and above waveguide filters are more suitable.

Filter losses usually amount to less than 1 db, but may be slightly greater if one aerial is shared by several transmitters or receivers.

Mixers

Up to the present co-axially mounted silicon crystals have been used in receiver mixers of the co-axial and waveguide types. An overall noise factor, including the intermediate-frequency amplifier, of between 12 and 15 db is typical.

Balanced mixers are sometimes used to reject local oscillator noise, but this does not appear necessary with triode oscillators.

Transmitter mixing is accomplished by triode mixers operating at a fairly high level, travelling-wave-tube modulation (phase or amplitude), or silicon crystals operating at medium levels.

Intermediate-frequency Amplifiers

These have been designed to have band-widths of over 30 Mc/s,[3, 27] but a band-width of about 16 Mc/s is typical. In a receiver without radio-frequency amplification the total gain available between the mixer and the input to the discriminator unit is usually between 60 and 90 db, depending upon the channels carried. This gain normally allows a reasonable fading range to follow the carrier down to peak-noise level (about 30 db), and a margin for loss of gain due to valve ageing (about 1 db per stage).

The choice of intermediate frequency is a compromise between:

(*a*) a high frequency to approach arithmetic symmetry of coupling circuit response and easy image rejection;

(*b*) a low frequency to obtain high input impedances to the valve stages and to have good control over the effective component values. Stability is more easily obtained at low frequencies.

Amplifiers have been designed for mid-frequencies ranging from about 30–150 Mc/s, frequencies of 60 and 70 Mc/s are popular. The C.C.I.R. recommends 70 Mc/s for carrier frequencies above 1,000 Mc/s.

In the present broad-band systems, although frequency modulation is used, it is necessary to have automatic gain control. This is due to the difficulty in designing really effective limiters. In practice, a limiter can only have good broad-band compression of amplitude over a small range of mean input signal level, therefore it is necessary for the automatic gain control to reduce the amplitude of variation by something of the order of 20 : 1.

With frequency modulation, amplitude and phase characteristics of the amplifier are equally important, particularly over the centre part of the band (very roughly 50 per cent) in which most of the energy is concentrated. It is therefore common practice to include phase-equalizer sections in the amplifiers to improve the phase linearity of an amplifier which is otherwise designed to have a sufficiently flat amplitude characteristic.

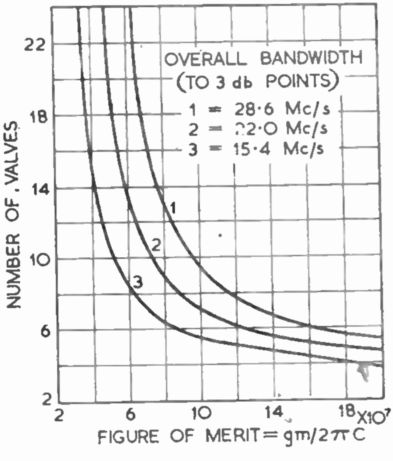


FIG. 9.—NUMBER OF VALVES REQUIRED TO OBTAIN A GAIN OF 90 db AS DETERMINED BY REQUIRED BAND-WIDTH AND FIGURE OF MERIT.

With the requirement of high gain and wide band-width, the figure of merit (see Appendix IV) of the amplifier valves is very important. It can easily be shown that with a given valve and coupling circuit there is a limit to the gain and band-width which can be obtained with single-valve stages in cascade. The limit is set when the addition of valves ceases to increase either the gain or band-width. This is demonstrated by Fig. 9, which shows the number of valves required as a function of figure of merit for an overall gain of 90 db with various band-widths. It is desirable to use a valve with a figure of merit high enough to get away from the near vertical part of the curve. If valves of low figure of merit are used the stability of gain of the amplifier suffers, and often the reliability is poor because of the large number of components and high dissipation of the unit.

The choice of interstage coupling is a compromise between complexity, with difficulty of alignment, and simplicity, with inferior stage gain. Staggered tuning is very common, especially staggered triplets. Staggered tuning is not suitable for high-level stages, due to power-handling problems, and such circuits as compensated band-pass and mutually coupled pairs are used.

Although pentodes and tetrodes are in general use for obtaining most of the gain in the amplifier, they are too noisy for the early stages. Two or more stages of triodes with grounded grids, sometimes combined with stages which have grounded cathode connections, usually form the low noise pre-amplifier.[27,28]

Discriminators and Limiters

For television transmission a compression factor of 10–15 db is usually sufficient, but for multi-channel speech better limiting is necessary. Grid limiters, fast automatic gain control and crystal limiters are used.

By far the most commonly used discriminator for broadband systems works at intermediate frequency, and consists of two tuned circuits fed by separate amplifiers. One circuit is tuned above the centre frequency, and the other below.

The discriminator is followed by a highly stable base-band amplifier, and sometimes the service channels are separated and amplified separately.

Miscellaneous Circuits

It is fairly common practice to originate a tone in the frequency band of 3·3–7 kc/s at the terminal transmitter. This is used as a pilot signal to indicate that the channel is performing correctly.

At each receiver noise is usually measured in a channel at a frequency above, or below, the band required by the television signal, or the multiplex channels. A high level of noise indicates a faulty condition.

Transmitter output power is almost always monitored.

The occurrence of a faulty condition brings into action automatic change-over equipment, which replaces faulty units by reserve units. Other ancillary circuits are provided to "squelch" rapidly a noisy equipment in order to prevent the noise being amplified by the system and causing interference. Similarly, circuits are provided to substitute a new carrier, or pilot, in the event of a failure.

Aerials and Feeders

Paraboloid reflectors, horns and lenses from about 4 to 12 ft. in diameter are used to obtain high gain and directivity. The gain,[29] relative to an isotropic radiator, is given by:

$$10 \log_{10} \eta \left(\frac{\pi D}{\lambda}\right)^2 \text{ db}$$

where λ is the wavelength, D is the reflector diameter (both expressed in same units), and the approximate angle of the beam at half-power points is $\frac{70\lambda}{D}$ degrees (this makes some allowance for uneven illumination).

The illumination factor (η), which is usually between 0·6 and 0·75, defines the efficiency of the launching device in making full use of the projected area of the mirror (or lens) by even illumination and correct wave front. A dipole can be used to launch the signal into the mirror, but better results are usually obtained with a waveguide-launching device which incorporates flares to control the angle of illumination. To avoid excessive side lobes it is necessary for the launching device to reduce the illumination to a low value at the edges of the mirror.

Very accurate matching (to obtain a standing-wave ratio always better than 0·85 and often approaching 0·97) between the feeder and the aerial is essential.

Below 2,000 Mc/s co-axial cables are used for aerial feeders, at 2,000 Mc/s both co-axial cables and waveguides are used, while at higher frequencies waveguides are nearly always used for long runs.

An estimate of feeder losses can be obtained from the following two expressions :

(*a*) Air-spaced co-axial cable with copper inner conductor, the attenuation is approximately

$$0{\cdot}026 \sqrt{f}/D \text{ db per 100 ft.}$$

where f is the frequency of operation in Mc/s, and D is the inside diameter of outer conductor in inches.

(*b*) Copper rectangular waveguide with cut-off frequency equal to 0·7 operating frequency, the attenuation is approximately

$$0{\cdot}04 \sqrt{f}/b \text{ db per 100 ft.}$$

where f is the frequency of operation in Mc/s, and b is the internal width of guide in inches.

Towers and Masts

The stability of the aerial-supporting structure is very important because the high gain antennas used with micro-wave systems have very narrow beam widths. Under high wind conditions most difficulty is experienced with turning moments rather than with vertical deflection.

Self-supporting towers, similar to those used for power distribution, but usually with added struts to give rigidity have most frequently been used. However, guyed masts are very attractive, as they are less expensive, easier to install and have greater rigidity. These guyed masts are often of lattice construction and with triangular or square cross-section. All are equipped with aircraft warning lights and beacons, according to height.

The effect of wind and ice must be considered, and the increased weight and wind resistance allowed for. In arctic zones allowance for ice is usually 2 in. radius for low wind velocities and ½-in. radius for gusts of 120 m.p.h.

Wind loading for various structures is proportional to the surface area and depends upon the direction of the wind; the maximum force is given by: $0{\cdot}0046 \,.\, V^2 \,.\, S$ lb./sq. ft. of projected area, where V is the wind velocity in miles per hour and S is the shape factor of the surface being considered:

Square flat plate	$S = 1$
Tubes and rods	$S = 0{\cdot}67$
Angles (flat)	$S = 1{\cdot}16$
Parabolic reflector (front)	$S = 1{\cdot}2$
Parabolic reflector (back)	$S = 0{\cdot}7$
90° cone	$S = 0{\cdot}75$
60° cone	$S = 0{\cdot}51$

Adequate lightning arresters and ground mats must always be provided.

Power Supplies and Diesel-generator Units

A.C. power supplies are usually voltage regulated to 2 per cent for valve heaters and power packs, and H.T. voltage stabilization is generally required for supplies to oscillators, modulators, travelling-wave tubes and intermediate-frequency amplifiers. In the case of travelling-wave tubes a current-stabilized supply may also be required for the focusing coils. The H.T. voltages vary from about 250 volts for small triodes to over 3,000 volts for some travelling-wave tubes.

Silicon rectifiers are usually preferred because of their long life and efficiency. Their low voltage drop is of considerable advantage in voltage-regulated supplies.

When high reliability of the communications system is essential standby diesel-generator sets can provide a no-break power-supply service. A typical method uses an alternator running as a synchronous motor across the primary supply. Directly coupled to this motor is a large flywheel. A diesel engine, with an electric starter, may be coupled to the alternator and flywheel through a centrifugal clutch. If the primary supply fails, the field of the motor is increased, and it then supplies power to the load from the energy stored in the flywheel. Within about 6 seconds the diesel has started and run up to speed. When the speed equals that of the flywheel the clutch engages. Stand-by battery systems using motor-alternators and transistors are also available.

Supervisory Facilities

It is not unusual to provide elaborate supervisory and remote-control equipment to bring into service stand-by equipment in the event of a fault developing in the active equipment. An alternative to complete sets of stand-by equipment at each station is to have a spare equipped radio-frequency channel available to take over the traffic from any one of several active channels. Since one fault can put a complete section out of action, instead of part of the equipment at one station, this alternative, although it has economic advantages, demands a very much higher standard of reliability than the stand-by equipment for each unit.

V.H.F. SYSTEM DESIGN

Since in many ways the micro-wave multi-channel systems have evolved from multi-channel V.H.F. systems, there is a considerable similarity in the general features.[30,31] However, the smaller number of channels, seldom exceeding forty-eight, and the somewhat less stringent signal-to-noise performance often required, together with the lower carrier frequencies, have tended to make V.H.F. multi-channel systems simpler in design.

Transmitters use frequency or phase modulators. Typical automatic frequency control circuits apply the difference frequency between the modulated carrier and a frequency originating from a quartz-crystal oscillator to a frequency discriminator.

The D.C. output from the discriminator is then used to control the mean frequency of the modulator. In transmitters which use phase modulators the carrier-frequency source is a quartz-crystal oscillator.

Output powers vary between about 20 and 250 watts, and radio-frequency power amplifiers use triodes and double tetrodes. When twenty-four channels are required the standard speech channel spacing is normally used, and the two groups covering 12–108 kc/s are transmitted.

Receivers have radio-frequency band-widths of 0·5–2 Mc/s, and noise factors of 6 or 7 db are typical.

Many types of aerials are in use. As much gain as reasonable dimensions will permit is often the first aim, and the aerials are usually supported by simple guyed towers or masts. Common transmit and receive aerials with good matching at the two separate frequencies are available.

Some typical aerials are six-element Yagis with a gain of 8–10 db stacked in pairs or fours, helical aerials with a gain of 12–14 db and broadside arrays of eight dipoles with a gain of about 14 db.

MEASURING TECHNIQUES FOR FREQUENCY-MODULATION SYSTEMS

Measuring techniques can be broadly divided into two classes :

(i) those for examining overall system performance ;
(ii) those for adjustment and routine inspection of units.

Overall Performance

Television Transmission

One method of measuring group delay over the band of 100 kc/s to 10 Mc/s is described in reference [32]. A 100-kc/s sine-wave is used to generate a short pulse with a repetition rate of 100 kc/s, and with substantial harmonics up to 10 Mc/s. After this pulse has been transmitted through the circuit under test the harmonic components will have suffered various attenuations and delays. In other words, a distorted pulse will appear at the output of the circuit.

The relative delays between each of the harmonics of the distorted pulse and those of a new pulse generated from the received 100-kc/s component are separately measured. At each frequency a minimum output from a selective detector is obtained by adjustment of the amplitudes of the harmonics to be equal but with their phases in opposition. The phases of the frequency components of the new pulse are controlled by phase shifting the 100-kc/s fundamental, and this control is calibrated in delay time, in fractions of a microsecond. Thus by tuning the selective detector to each harmonic in turn, the amplitude and delay controls give attenuation and delay for each frequency, and a graph of the transmission characteristic can be plotted.

An alternative method [33] of measuring group delay uses the principle of an exploring carrier to cover the frequency range of 200 kc/s to 5 Mc/s. This exploring-carrier is modulated by a low-frequency signal (60 kc/s) and transmitted by the circuit under test. A reference 60-kc/s signal is also conveyed by the test circuit, and is used to generate a time-base for application to the X plates of a cathode-ray-tube indicator.

At the receiver a short pulse is generated from the "exploring" 60-kc/s signal, and this is applied to the Y plates of the indicator. Measurements of changes of delay are made from a calibrated phase-shifter in the reference circuit, which is used to position the pulse against a cursor at the centre of the screen.

Frequency response (attenuation/frequency characteristic) can be measured with standard transmission measuring equipment or a variable-frequency signal generator, attenuator and output-level indicator.[34]

Frequency, phase and transient response characteristics can be very rapidly checked by pulse techniques. An example of this [35] uses a sine squared pulse (0·05, 0·1 or 0·34 microseconds duration at half-height). At the output of the transmission circuit the height, duration and overshoot of the pulse is observed.

Linearity,[36] which is expressed as a change of the slope of the input-amplitude/output-amplitude curve, can be measured by superimposing a very small amplitude high-frequency signal (50 kc/s) on a low-frequency saw-tooth or sine-wave (250 c/s) of amplitude swing sufficient to examine the full amplitude range of the system. At the receiver end of the system the variation of the level of the high-frequency signal is a direct indication of change of slope, and may be conveniently displayed (Y deflection) against the large signal (X deflection).

Noise level is often measured as the power falling into narrow bands (4 or 6 kc/s wide) tuned to various frequencies in the base-band. If the levels are then plotted against frequency, the frequency distribution of the noise is apparent.

For the functional testing of links a pulse generator which produces a standard 405-line 50-frame interlaced synchronizing waveform and various video test signals is attractive.[34]

Multi-channel F.D.M. Transmission

Linearity may be measured by the injection of two tones into the modulator and the measurement of the inter-modulation products at the output of the demodulator. Although such a test is frequently called for in a specification, it is not a very satisfactory method of assessing overall distortion and obtaining correct adjustment of a frequency-modulation system. This is due to the dependence of distortion on modulation frequency and the possibility of critical compensation between distortion products from several causes.

An alternative method simulates the multi-channel F.D.M. signal by a random noise spectrum with appropriate loading levels as given in Table 4. The linearity of the system determines the amount of inter-modulation noise produced. This can be measured at different parts of the band by leaving small gaps in the transmitted noise spectrum substantially free of noise. At the receiver the noise level in the gaps is measured. If the transmitted noise spectrum is suppressed, the noise received in the gaps will, of course, be due to circuit noise.

Group delay at base-band frequencies is not usually of importance, but it can be measured by the techniques described for television transmission.

The attenuation/frequency characteristic requires accurate measurement, but standard cable-transmission equipment is satisfactory for this purpose.

Unit Test Equipment

A list of instruments gives a general picture of the test facilities required.

Radio-frequency Circuits, Filters and Aerials

(i) Radio-frequency signal generator with precision attenuator.
(ii) Radio-frequency test receiver.
(iii) Standing-wave line, or slotted waveguide.
(iv) Precision matched loads.
(v) Fixed radio-frequency attenuators.
(vi) Wattmeter.
(vii) Frequency scanner (for circuit alignment).
(viii) Wavemeter.

Intermediate-frequency Circuits

(i) Intermediate-frequency signal generator.
(ii) Intermediate-frequency level detectors.
(iii) Group delay and amplitude response display equipment.
(iv) Intermediate-frequency impedance bridge.

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APPENDICES

I. Group Delay

A transmission circuit will introduce distortion of a signal by conveying the various component frequencies at different velocities. A linear phase characteristic (i.e., phase shift due to the transmission network is proportional to frequency) introduces no phase distortion; it may be shown that all component frequencies of the signal are equally delayed. It is convenient to specify the transmission characteristic by the slope of the phase characteristic (phase versus frequency); this quantity has the dimension of time, and is defined as group delay.

It is usual to express the characteristic in two parts: the constant average delay which introduces no distortion, and group delay variation with frequency which causes distortion. Thus reference is seldom made to the average delay of a transmission circuit, and only the variation of group delay is considered.

II. Psophometric Weighting

The importance of noise to a listener at different frequencies in the audio band varies. The psophometric noise level is the level of flat noise in a given band that would be measured by a psophometer with its frequency response weighted according to the effects of various frequencies on a listener. This weighting is recognized internationally, see references [11] and [12]. The permissible "flat" noise is higher than the psophometric noise level, and the correction factor for a 300–3,400-kc/s speech channel is 3·4 db.

The weighting in Table 5 has been obtained from references [11] and [12].

A useful summary of American practice, which differs considerably from European, is given in reference [40].

III. Noise Power at Input of Receiver

Noise factor is defined as :

$$\frac{P_{si}/P_{ni}}{P_{so}/P_{no}}$$

where $\frac{P_{si}}{P_{ni}}$ is the input signal-to-noise power ratio before being affected by the receiver, and $\frac{P_{so}}{P_{no}}$ is the output signal-to-noise power ratio, including the noise contribution of the receiver.

TABLE 5.—PSOPHOMETRIC WEIGHTING

(1) *Telephone Channel*

C/s	*Gain Relative to Gain at 800 c/s (db)*	*C/s*	*Gain Relative to Gain at 800 c/s (db)*
50	−63	1,000	+ 1
100	−41	1,500	− 1·3
150	−29	2,000	− 3
200	−21	2,500	− 4·2
300	−10·6	3,000	− 5·6
400	− 6·3	3,500	− 8·5
500	− 3·6	4,000	−15
600	− 2·0	5,000	−36
800	0		

(2) *Programme Channel*

C/s	*Gain Relative to Gain at 1,000 c/s (db)*	*C/s*	*Gain Relative to Gain at 1,000 c/s (db)*
60	−32·2	4,000	+8·2
100	−26·1	5,000	+8·4
200	−17·3	6,000	+8·2
400	− 8·8	7,000	+7·3
800	− 1·9	8,000	+5·1
1,000	0	9,000	−0·3
2,000	+ 5·3	10,000	0·7 −

In broad-band systems it is usual to match the aerial and feeder to the input circuit. The input signal (P_{si}) is calculated for matched conditions: therefore the noise (P_{ni}) must be expressed as the power developed in the same load. This power will be one-quarter of the value that could be generated by the aerial resistance.

The power generated by the aerial resistance is

$$P_{nr} = 4KT \text{ watts per c/s band-width}$$

where K = Boltzman's Constant = $1{\cdot}37 \times 10^{-23}$ joules per degree Kelvin (absolute);

T = absolute temperature = 290° Kelvin (assuming temperature is 17° C.).

Therefore, $P_{nr} = 4 \times 1{\cdot}37 \times 10^{-23} \times 290$ watts per c/s band-width
$= 1{\cdot}59 \times 10^{-20}$ watts per c/s band-width
$= -198$ dbW per c/s band-width.

The noise power (P_{ni}) available at the input of the receiver, which is $\frac{1}{4}P_{nr}$, is -204 dbW. The receiver has a noise factor of 12 db, and if the actual noise power is referred to the input of the receiver (i.e., $P_{si} = P_{so}$ in the above definition of noise factor) the noise power

$$P_n = -204 \text{ dbW} + 12 \text{ db} = -192 \text{ dbW per c/s band-width}$$

For a further discourse on Noise Factor see, for example, Section 42 or reference [21].

IV. Figure of Merit

This is

$$\frac{g_m}{2\pi C}$$

where g_m = mutual conductance of valve;

C_t = input + output + stray capacitances of valve and coupling circuit.

J. E. B.

18. RADIO NAVIGATION AND RADAR

18. RADIO NAVIGATION AND RADAR

RADIO DIRECTION FINDING

The chief applications of normal direction-finding equipment are :

(*a*) for fixing the bearings of a ship or aircraft, and
(*b*) for " homing " under conditions of poor visibility, e.g., in fog or at night time.

Radiation

Associated with every inductance through which a current passes is a magnetic field. Likewise an electrostatic field exists between the electrodes of a capacitor. Long before radio communication became a commercial undertaking it had been deduced mathematically that the acceleration of an electric charge will produce an electromagnetic wave in space. This wave comprises an electric field at right angles to a magnetic field. An oscillatory circuit which is capable of radiating energy in this way is known as an open or radiating circuit.

Electromagnetic waves convey energy from the source, either in all directions or confined to certain specified directions according to the design of the radiating system and other factors. The radiation from a vertical wire or loop produces a magnetic field horizontally and at right angles to the direction of the transmitting station and an electrostatic field perpendicular to the surface of the earth, as shown in Fig. 1.

Loop Aerials

It will be seen, therefore, that high-frequency radiation at the receiving station can be detected through either the electrostatic or electromagnetic component. In Fig. 2, the arrow indicates the direction of the transmitter from the receiver and the rectangular coil of wire represents a closed loop forming, with a capacitor, a circuit tuned to the transmitter frequency. In Fig. 2 (*a*) the plane of the loop is perpendicular to the direction of signal transmission and parallel to the wave front, so that the magnetic component induces no current in the loop. The electrostatic component in the ideal case induces equal e.m.f.s in the vertical limbs, and these on one side oppose those on the opposite side, so that the resultant current is zero. In Fig. 2 (*b*) the linkage of the magnetic component of the field with the loop is a maximum, and maximum oscillatory currents are induced in the tuned circuit.

Thus the signal strength in a receiver with a frame aerial, assuming there are no disturbing influences, depends on the orientation of the plane of the loop, and is greatest when this plane coincides with the direction of the transmitting station, decreasing to a minimum and theoretically zero as the loop is rotated through 90°. The direction of

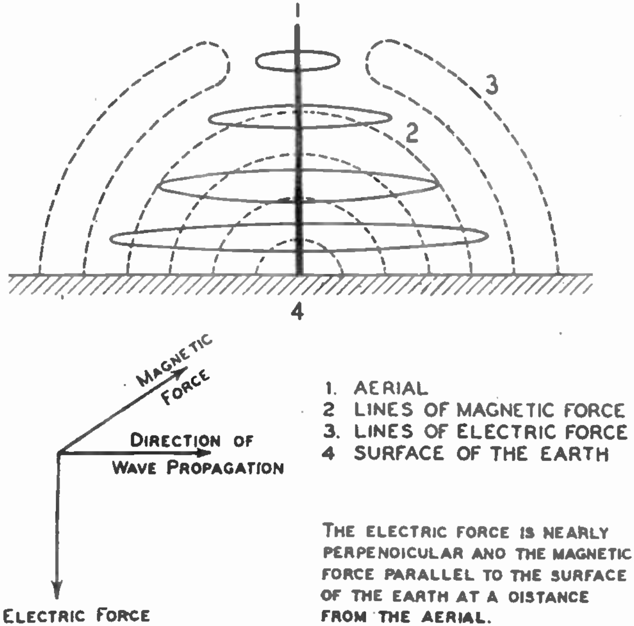


FIG. 1.—RADIATION OF ELECTROMAGNETIC WAVES.

(*a*) Plane of loop parallel to wave front and perpendicular to direction of transmitter. Minimum current induced in loop.

(*b*) Plane of loop perpendicular to wave front and parallel to the line of signal approach. Maximum current induced in loop.

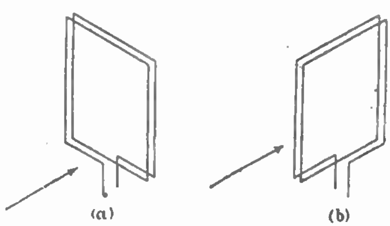


FIG. 2.—LOOP AERIALS.

Arrow indicates direction of approach of signals from transmitter.

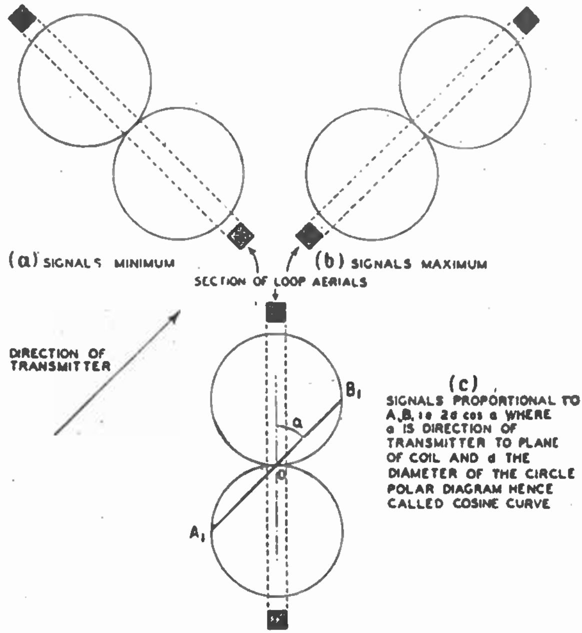


Fig. 3.—Figure-of-eight Polar Diagram of Loop Aerial.

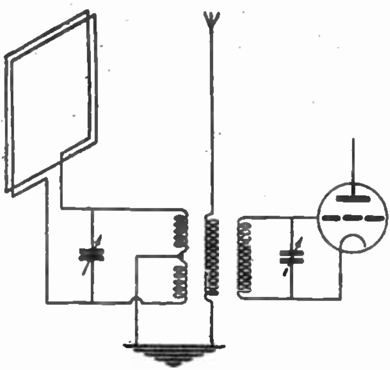


Fig. 4.—Schematic Arrangement of Circuit showing Principle of Sense Determination with Open and Loop Aerials.

the signal approach can, therefore, be determined by finding the position of the loop in which signals are either a maximum or a minimum.

In practice, the point of minimum signal strength is used for determining signal direction. When signals are weak, it may be possible to turn the frame through an appreciable angle about the minimum point without any signals being heard at all. To assist in deciding where the true zero is, it is usual to take what are called swing bearings, i.e., matching the intensity on either side of the indefinite minimum, the midway position between the equal readings being the true zero.

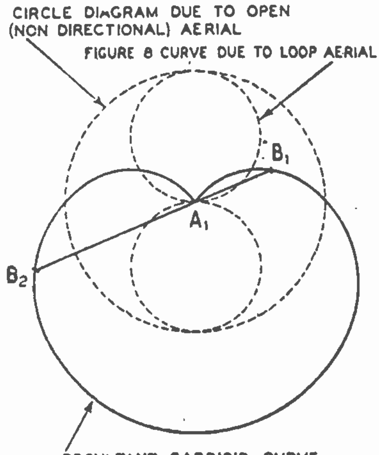


FIG. 5.—CARDIOID DIAGRAM.

Bearings

Theoretically, the relation between signal strength and frame orientation can be depicted by a figure-of-eight polar diagram (Fig. 3), in which the double chord drawn through the common tangent point is proportional to the signal intensity when the plane of the frame makes an angle, say X, to the direction of signal approach.

If we imagine the line of centres of the polar diagram to coincide with the plane of the frame aerial and to rotate with it about the point O, the signal strength will be proportional to A_1B_1. Thus, if the frame is turned till the signals are at maximum, say X_1 (Fig. 3 (*a*)), and then set so that the plane of the coil is in the line of flight, in which position the signal strength is X_2, then $X_2/X_1 = \cos a$, where a is the bearing of the transmitter station from the line of flight. More simply this information can be obtained by attaching a pointer to the frame, so that it moves over a fixed circular scale graduated in degrees.

Such observations with a frame aerial, assuming the absence of other errors, will determine only the direction of the transmitter, but will give no indication as to whether it is in front of or behind the receiver.

Sense Determination

The ambiguity which is thus possible is overcome by the use of an open or non-directional aerial in conjunction with the frame aerial and feeding into the same receiver. The principle of this combination is shown in Fig. 4. An open aerial, being non-directional, produces a circular polar curve, and can be made to circumscribe the figure-of-eight by adjusting the signal strength relative to that received from the frame aerial. If, now, the signals due to the open aerial are exactly in phase with those due to one limb of the figure-of-eight, those in the opposite limb will be in anti-phase. The two curves can thus be combined vectorially, resulting in a cardioid or heart-shaped curve (Fig.5).

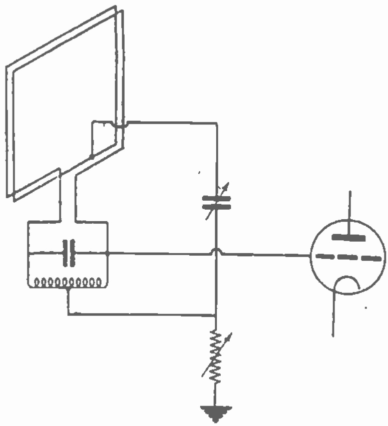


FIG. 6.—PRINCIPLE OF COMBINED OPEN AND LOOP AERIAL.

In principle, therefore, a complete observation consists in determining first the bearing of the transmitter station by means of the frame aerial alone, as previously outlined. This fixes the AB direction. The open aerial is then brought into circuit and signal strength observed first with the frame in one position and then with the frame reversed, either turned through 180° or reversed in connection by means of a change-over switch. These readings will correspond respectively with A_1B_1 and A_1B_2, which can be distinguished in direction by their relative magnitude (Fig. 5).

The basic principles of the circuit are shown in Fig. 4, of which many modifications are in use. It will be clear that this principle, though simple, depends in practice for its success on the absence of disturbances which falsify the readings. Practical apparatus for direction finding is often elaborate because it must include accessories for securing the three essentials which the application of these principles demands, viz. : (1) elimination of reception errors; (2) correct phasing of frame and open aerial signals; (3) correct relative signal intensity in the two cases.

The open aerial employed to secure a cardioid diagram is not, as shown in Fig. 4, essentially a separate conductor. The electrical centre of the coil, which does not necessarily coincide with the geometrical centre owing to differences in capacitance of the two sections, can be used for this purpose with an adjustable resistance for securing correct relative signal intensity. This arrangement is shown diagrammatically in Fig. 6.

Errors

Though simple in theory, the use of the frame aerial for direction finding is liable to disturbances from several sources, and correction of possible errors is essential before reliability can be secured. The principle causes of inaccuracy are :

(1) *Aerial, or Out-of-phase, Vertical Effect.* This arises from unequal e.m.f.s being induced in the two vertical limits of the frame owing to slight differences in their respective capacities to earth, producing a distortion of the polar diagram. Vertical error is proportional to the linear dimensions of the frame, while signal strength is proportional to the square of these dimensions, so that its effect is less the larger the frame. The error is independent of the orientation of the frame. Correction of aerial effect can be

secured by the use of a symmetrical design of frame with an earthed electrical centre and by suitable earthed shielding to eliminate electrostatic coupling between aerial and tuned grid circuits.

(2) *Direct Pick-up.* This cause of trouble is avoided by the use of suitably disposed earthed shields.

(3) *Quadrature, or Random-phased Re-radiation from Nearby Conductors.* If large enough, it may so distort the polar diagram as to make it approach a circle and give uncertain readings in any direction. To a lesser degree, it results in an indefinite zero in the polar diagram.

(4) *Displacement Currents, if the Turns of the Frame are not All Co-planar.* The frame is then equivalent to a tall thin coil when viewed edgewise, and this results in a smaller figure-of-eight diagram at right angles to the main one. The net result is again to flatten the zero. The obvious method of correcting this error by using pancake coils results in an increase in out-of-phase vertical effect.

Bellini–Tosi Aerials

To utilize the advantages of a large frame without the inconvenience of a heavy rotating mass, ground stations usually employ one of the modifications of the Bellini–Tosi system (Fig. 7). There is no fundamental difference in this arrangement from that already mentioned. There are two large fixed frame aerials at right angles to one another, and these are connected respectively to two small coils, also fixed at right angles. A third, or search coil, is mounted axially with this latter pair and is free to rotate around them. It is closely coupled to the small fixed coils and is connected to a suitable tuned circuit and amplifier. The search coil serves to determine the direction of the resultant flux through the two small fixed coils, and with them constitutes what is known as a radio-goniometer.

Several conditions must be fulfilled to secure accurate indications with equipment of this type. The aerials must be identical in size and of the same high-frequency resistance, otherwise indications will not be proportional to the signal strength in the two cases. There must be no magnetic, electrostatic or conductive coupling between the loops. Similarly, the coils of the goniometer field must be identical and accurately at right angles.

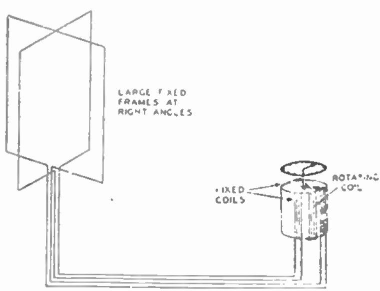


FIG. 7.—PRINCIPLE OF BELLINI-TOSI SYSTEM OF FIXED FRAME AERIALS WITH RADIO-GONIOMETER FOR GROUND STATION DIRECTION FINDING.

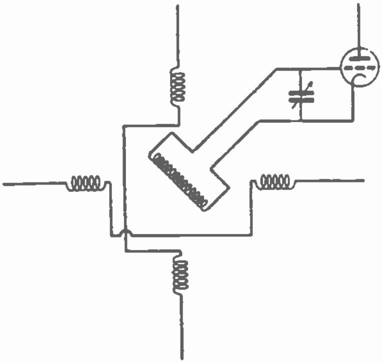


FIG. 8.—SCHEMATIC ARRANGEMENT RADIOGONIOMETER IN BELLINI-TOSI SYSTEM.

Adcock Aerials

So far it has been assumed that no voltages are induced in the horizontal limbs of the frame. This is true only if the waves are travelling horizontally and the wave front is perpendicular to the earth surface. With oblique transmission, as for instance from a plane in flight to a ground station, errors in reading would be produced by the voltages induced in the horizontal limb of the frame, so that the frame zero would not coincide with the position when its plane is at right angles to the line of signal approach.

The chief disadvantage of the Bellini-Tosi system is that it is subject to this type of error and is, therefore, unreliable at night. This is due to horizontally polarized waves being reflected from the Heaviside layer and picked up on the horizontal portions of the loop. From dusk to dawn the reflected wave forms a large proportion of the total signal strength, and thus affects seriously the accuracy of the readings. The Adcock aerial system is designed to eliminate the effect of pick-up on the horizontal limbs of the frame. In principle it consists of two pairs of open spaced aerials with screened horizontal connections so that reception is confined to the vertical aerials. For H.F. working the aerial spacing will be about 20 ft., so that the system is applicable only to fixed ground stations. In application, the principle of the Adcock aerial necessitates many precautions, by suitable disposition of the aerial system or otherwise, to ensure that horizontal limb pick-up is zero, or is balanced out. In some cases the horizontal limbs may be buried and auxiliary circuits employed to remove residual errors.

Cases of considerable difficulty are those where the same site is used as a transmitter and receiver station, though much can be done in the way of eliminating disturbances by the choice of a suitable site for the ground station in respect of power mains, natural objects and other transmitter stations.

Visual Indicators

To relieve the operator of the tedium of aural detection, several types of visual indicators or " radio compasses " are available.

METER TYPE. The amplifier output circuit may be fed to a full-wave rectifier connected to a zero reading microammeter, while a motor-driven commutator serves to reverse both aerial and instrumental connections in synchronism. When the machine is on course, the succeeding impulses through the instrument are equal and opposite and there is no deflection. If the machine strays from its course, the pulses through the indicator are still alternating but unequal in magnitude, so that the instrument shows a deflection to one side or the other, corresponding to the machine's deviation.

NEON-TUBE INDICATORS. Another kind of visual indicator used in the standard type homing receiver consists of two neon tubes of the tune-on class, as employed as visual tuning devices in broadcast radio receivers. Normally, the two tubes are operated under conditions which produce a visible discharge of about half the length of the tube.

CATHODE-RAY TUBES. Recently there has been a tendency for existing forms of indicator to be displaced by the cathode-ray tube. Visual indication of signal strength, phase, direction or frequency can be shown.

The chief advantages are: (1) Since the indication is given by the deflection of an electron beam, the instrument is devoid of mechanical inertia so that voltage variations at radio frequencies which no aural or dial meter could distinguish can be depicted without much difficulty. (2) Since deflection of the spot can take place in two co-ordinate directions, the relation between two variables can be shown simultaneously.

Visual presentation of bearings is, in practice, frequently assisted by the use of a constantly spinning loop or goniometer.

Automatic direction finders for aircraft may employ self-aligning motor-driven loop aerials which rotate automatically until the axis of the loop points in the direction of arrival of the received signal.

RADIO NAVIGATIONAL AIDS

Radio Beacons

The direction-finding equipment carried on board ships, usually either Bellini-Tosi fixed-loop aerials or the simpler rotating loop, is generally used to obtain bearings on the special radio-beacon transmitters in operation in coastal areas throughout the world. Apart from fixed radio-beacons, automatic beacons are also installed in many lighthouses and lightships. Radio-beacons normally operate within the frequency range 255–415 kc/s, and a typical installation would comprise a crystal-controlled transmitter for A1 or A2 emission, with powers ranging from about 20 watts to several hundred watts for maritime use (considerably higher powers are often found in the aeronautical service); a master time switch and an automatic code sender providing a repetitive call sign in Morse at about 7 w.p.m. for station identification, followed by a long dash for direction-finding purposes.

V.H.F. Omni-directional Radio Range

A development of the radio compass is the V.H.F. omni-directional radio range, usually known as VOR, which enables an aircraft to fly along a pre-determined path into an approach zone, and is suitable for furnishing navigational information between the range beacon and aircraft up to 100–150 miles distant. Unlike the normal course beacons, this system provides an infinite number of tracks radiating from the transmitter. The transmitting equipment can be used as part of the radio-telephony communications system.

The transmitter is arranged to radiate two signals whose relative phases vary with azimuth. One signal has a constant phase through 360°, and is termed the reference signal, whilst the second variable signal has a phase varying with azimuth. The variable-phase signal is transmitted from two aerial systems mutually at right angles, and fed so that when the field set up by one system is at a maximum, that of the other is at a minimum. The result is a figure-of-eight radiation pattern rotating at a pre-determined rate of the order of 30–50 c/s. This produces in the receiver, after detection, a sinusoidal voltage.

This system is described in greater detail in Section 19.

V.H.F. Direction Finding

The widespread use of very high frequencies for airborne communications has led to the development of direction-finding equipment for these frequencies, and in fact the accuracy now attainable tends to be an improvement on that of the Adcock medium- and high-frequency systems. The shorter wavelengths involved make it possible to use resonant aerials, and in practice a rotatable pair of spaced vertical dipoles is commonly employed in ground D/F stations. The geological characteristics of the site play a less important role on V.H.F.

Beam Approach and Glide Path

A number of V.H.F navigational aids to approach and landing have come into general use, based essentially on the Lorentz system of two directional transmissions with a sharply defined area of overlapping in which equal signals are received from both transmissions. This area may be denoted by interlocking Morse characters or by modulation at two different frequencies. Results may be presented to the pilot by means of crossed-pointer meters. See Section 19 for further information.

Consol

A medium-frequency system for providing navigational assistance to ships and aircraft, fitted with a standard communications receiver only is the Consol or Sonne. The range of this system is of the order of 1,000 miles over sea. The automatic beacons transmit a series of overlapping lobes of radiation keyed with interlocking dots and dashes.

According to the sector the observer is in, dots or dashes are heard at the beginning of the emission. In the pattern shown, Fig. 10, there are twenty-four sectors. In twelve of these sectors dots are heard at the commencement of the cycle of operation, and in the other twelve sectors dashes. When the observer is on the limit between a dot and a dash sector he hears the dots as well as the dashes, and thus receives a continuous tone, termed the equisignal. The transmission pattern is rotated continuously, with identification signals interspersed. The observer thus counts the dots or dashes that he hears before the equisignal. A position fix can be obtained by repeating the procedure on a second beacon.

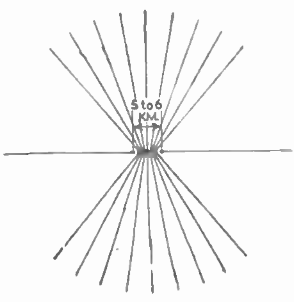


Fig. 9.—The Three Transmitting Stations Used in the Consol System.

At the transmitting station, an automatic transmitter, with special keying

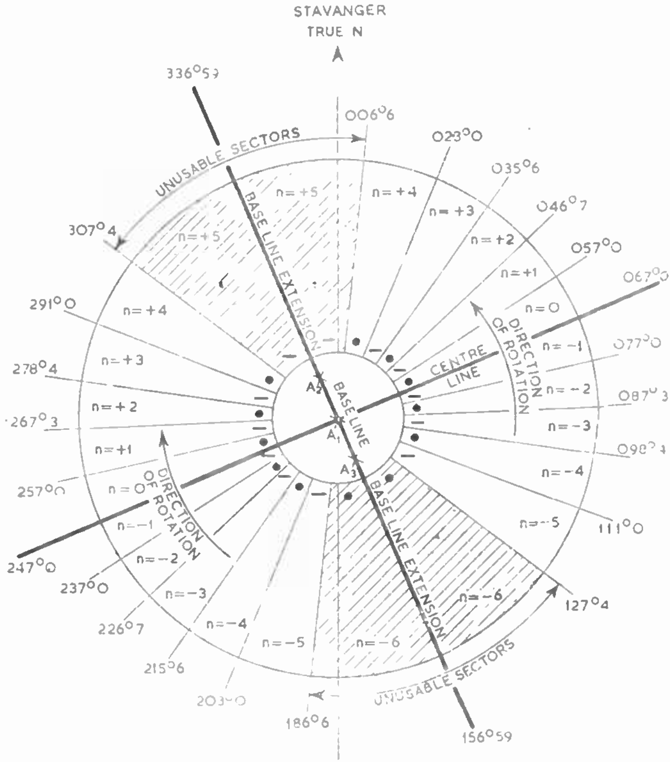


FIG. 10.—TRANSMISSION PATTERN OF THE CONSOL BEACON AT STAVANGER, NORWAY.

Position of Central Aerial 58° 37′ 21″ N., 05° 37′ 49″ E., frequency 319 kc/s.

and phasing circuits, is arranged to energize an aerial array of three mast radiators which are erected in line. A multi-lobe radiation pattern is produced, and the alternate lobes or sectors are characterized by dots and dashes. These are interlocked so that they merge into a continuous tone on the boundary line between the sectors. The radiated pattern is rotated slowly, and the equisignal moves through one sector's width during the course cycle. This cycle, including an identification signal, takes 40 seconds to complete, and the sector width is of the order of 10° and 15°.

Since several sectors may have identical polar patterns, there may be some ambiguity, but in practice confusion is unlikely to arise, since the approximate position is usually known. Positive sector identification

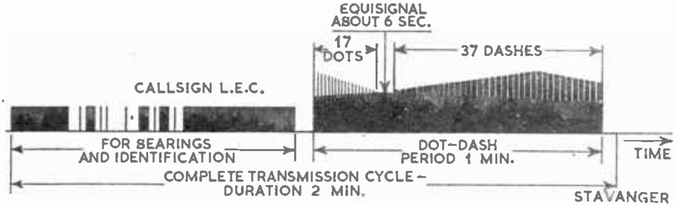


FIG. 11.—DIAGRAM OF THE SIGNAL STRENGTH DURING THE WHOLE TRANSMISSION CYCLE FROM STAVANGER, FOR ONE POSITION LINE.

may be provided by additional apparatus, but this is seldom considered necessary.

The navigator must be provided with tables or a chart overprinted with great circle bearings with the consol station as origin. Extensive networks are in operation in Western Europe. In practice, bearings with an accuracy of $\frac{1}{4}°$ or $\frac{1}{2}°$ are obtainable during daylight. At night the likely error may be of the order of 2°. The transmitters operate within the frequency range 255–415 kc/s.

PULSE-MODULATED SYSTEMS

Pulse-modulated navigational systems were developed both in Great Britain and America during the 1939–45 war. The British system is known as " Gee ", and was intended primarily for aircraft purposes, but has also been used extensively for ship navigation.

The Loran, or American Long-range Navigational Aid, is similar basically to Gee, but operates on a longer wavelength and has a greater range under most conditions.

Developments from the basic system are described in Section 19.

Basic Principles

The operation of both systems is based on determining the difference in the time of arrival at the receiver of radio-frequency pulses transmitted simultaneously or at definite intervals from a series of ground transmitters.

The Gee system depends on the use of hyperbolic or lattice charts.

Two stations A and B are " synchronized " so that they can send out pulses at the same instant; it is obvious that if the aeroplane is equidistant between the two stations, the pulses will be received by the aircraft radio at the same instant. If the aeroplane is nearer to station A than station B, the pulse from A will be received a fraction earlier than the corresponding pulse from B. This difference in time represents the difference in distances from the two sending stations; in other words, the operator can deduce from the indicator that he is, say, 2 miles nearer to station A than to station B. The problem of fixing the position of the area may be resolved by the use of hyperbolic charts.

In these charts any point on a hyperbola has a *fixed difference in distance* from the two focii. Referring to the diagram in Fig. 12, if A and B represent the two focii or sending stations, say 10 miles apart, if

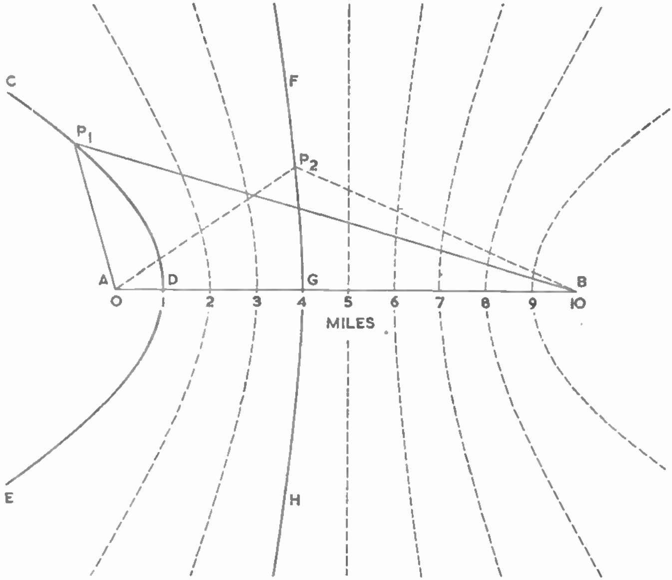


FIG. 12.—PRINCIPLE OF THE HYPERBOLA.

For any point, say P_1, on the hyperbola CDE, the distance P_1B minus the distance AP_1, will always be 8 miles (i.e., the difference factor). Similarly, for any point P_2 on the hyperbola PGH, the difference factor will be P_2B minus AP_2, namely 2 miles.

we draw a hyperbola from position 1, which is distant 1 mile from A and 9 miles from B, every point on that hyperbola will have a difference of 8 miles in its distances from A and B.

Similarly, if we draw a hyperbola through the position 4, which is 4 miles distant from A and 6 miles distant from B, the difference factor for all points on this curve will be 2 miles.

Hyperbolic charts of this kind are prepared for each pair of stations, and can be shown on a diagram as in Fig. 13. If, after taking his first reading, a navigator in an aeroplane at the point P finds that his difference factor for transmitting stations X and Y is, say, 50 miles, he can tell that his position is somewhere on the hyperbola RPQ.

If a third transmitter Z, suitably disposed in relation to the transmitters X and Y, is also sending out simultaneously synchronized pulse signals, the observer is able to obtain a further position line, in a similar way, by measuring the difference in time between signals arriving from, say, transmitters X and Z. Therefore he can deduce that his position will also lie on the line TPS, and his actual position with relation to the three transmitters will be where the two lines, or hyperbolæ, intersect.

Suitable charts to facilitate position plotting can be drawn, consisting of a series of hyperbolæ, each representing a known difference in distance

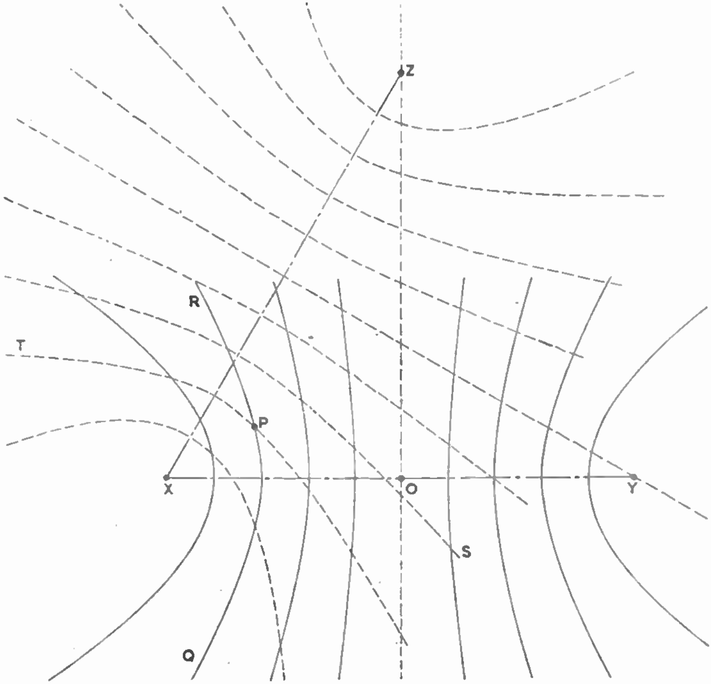


FIG. 13.—TYPICAL LATTICE CHART FOR PULSE-MODULATED SYSTEMS.

between the three transmitters, and these charts can be superimposed on a map of the area served. On this account, the Gee and Loran systems are also termed hyperbolic or lattice systems.

The Gee System

In practice the Gee system consists of a chain of transmitters arranged in groups of three or four, the transmitters being located about 70 or 80 miles apart.

Each transmitter operates on the same radio frequency, usually between 20 and 80 Mc/s (3·75 and 15 metres), and is continuously emitting synchronized radio pulses. A master transmitter is used in each group to control the pulses of the subsidiary or slave transmitters crystal-controlled timing being employed.

The aerials are omni-directional, and the use of the very high frequencies results in the reception of ground waves only, thus eliminating reflection troubles normally associated with the lower-frequency wavebands. They have, however, the disadvantage of optical range, and signals may be seriously impaired if the optical path between the transmitter and the receiver is obstructed.

The equipment carried by a ship or aircraft for the operation of the Gee system consists of a receiver and an indicator of the cathode-ray display type.

In operation, the pulses received from the three or four transmitters can be received and displayed simultaneously on the screen of the cathode-ray tube, and the arrangement permits the time difference of arrival of corresponding pulses from each transmitter to be determined to within a millionth of a second.

Satisfactory reception of the transmitted pulses over sea can be obtained up to 100 miles, but, in practice, reasonable results have been achieved at ranges of 150 miles.

For aircraft, the range depends on the nature of the land between the aircraft and the transmitters, and also the flying height of the plane. Reception is normally quite satisfactory up to 400 miles at '5,000 ft. and up to 200 miles at 5,000 ft.

The Loran System

This system is based on the same principles as Gee, but operates on a radio frequency of approximately 1,900 kc/s (157 metres).

Much greater ranges are possible with Loran, as it takes advantage of the increased range of the ground waves at the lower frequencies, and is not so readily affected by height. There are, however, other troubles introduced which are peculiar to the short and medium wavelengths, e.g., although the range of the Loran system may be twice as far by day as that obtained with Gee when operating over sea, it may not be as good over land, as ground absorption will occur.

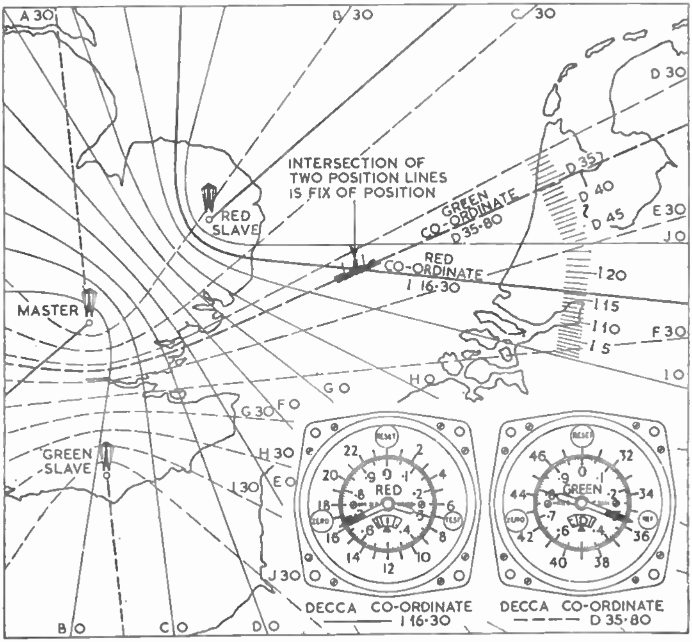
At night Loran waves are readily reflected by the upper atmospheric layers, and the range thereby considerably increased, with the added difficulty for the observer of distinguishing between the direct and the reflected wave.

With Loran, only two transmitters are operated in synchronism, a master and a slave, which may be spaced up to 600 miles apart. Signals from the master transmitter are received at the slave transmitter, and are relayed by the slave after an accurately controlled time delay. To obtain a fix the observer has to take readings from two pairs of transmitters, and the deductions are read off from lattice charts in a manner similar to the Gee system.

A range of about 700 miles by day over the sea can be obtained with this system and twice this distance by night. This range is considerably reduced for the direct wave over land, especially over rough ground, but the reflected wave is, of course, unaffected by the nature of the terrain, and ranges of 300–1,200 miles may be obtained over land at night.

Decca

Among the numerous position-finding systems that were developed during the 1939–45 war, one of the most important was the Decca system, which uses low-frequency continuous wave-phase comparison to provide continuous position fixing. On the transmitting side a chain



· Fig. 14.—The Decca System.

Two sets of Decca co-ordinates and two decometers are shown to illustrate the principle of the system. The "purple" Decca lines and decometer are omitted.

of stations transmits continuously, on frequencies which are in some simple ratio such as 3/2 or 5/4, and at the receiver each set of two signals are multiplied to the same frequency, and their phase difference is measured and indicated on a counter. In practice, a chain may consist of four stations, a "master" and three "slaves" known as the "red slave", "green slave" and "purple slave". Transmission frequencies are of the order of 100 kc/s. The official range of this system is 240 miles from the master station, but in practice fairly accurate results can be obtained at considerably greater distances. A number of chains have been set up in North-west Europe, for both marine and aeronautical use. For air-navigational purposes, only two of the slave stations are normally used. The aircraft equipment comprises three low-frequency receivers with multipliers, and discriminator, which measures the phase difference, and amplifier units with two integrating meters. Continuous reception is essential for accurate position indication.

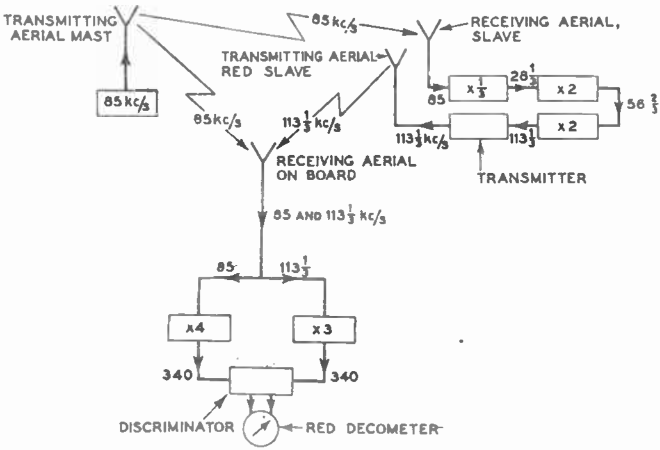


Fig. 15.—Frequency Modification in the Decca System.

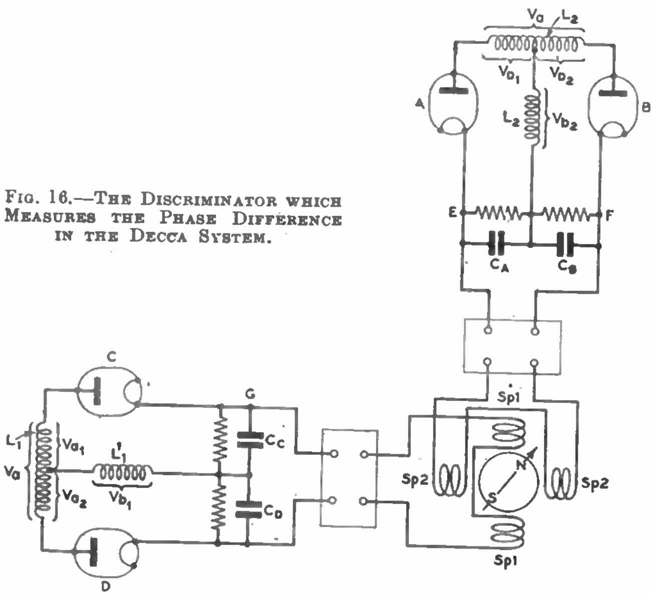


Fig. 16.—The Discriminator which Measures the Phase Difference in the Decca System.

RADAR

Radio detection and Ranging (Radar) is the art of locating the presence of objects by radio means, determining their angular position, with regard to some reference point (relative bearing) and their range. In order to accomplish this, a beam of R.F. energy is directed over some given area in search of a target, by means of a highly directional rotatable aerial. If the beam strikes a target some of the R.F. energy is reflected and a small portion of this reflected energy travels back in the direction of the transmitter.

Information derived from Echoes

If a sensitive receiver, capable of detecting this reflected energy, is arranged to operate in the vicinity of the transmitter, together with some time-measuring device, capable of measuring the extremely short periods of time elapsing between transmission of the energy and reception of reflections (Echoes), the following information can be deduced when echoes are obtained :

(1) Some reflecting body (in radar terminology "a Target") has been found by the beam, as demonstrated by the echo received at the receiver and recorded by the time-measuring device.

(2) It can be shown that the range or distance of the target from the transmitter is proportional to the time interval, measured from the instant that transmission of energy commences, to the instant at which the returning echo is received.

(3) The bearing of the target, measured with reference to the direction of the ship's head, or in the case of a shore station, measured from compass North, is indicated by the angle through which the aerial must be rotated, in order that the centre of the beam may face the target. This is evidenced by the fact that an echo is received and recorded only when the aerial is rotated so that the beam covers the target.

(4) The elevation or height of an airborne target can be obtained, under favourable conditions, by measuring the angle of elevation by which the aerial must be tilted in order that the centre of the beam may face the target, and by simultaneously measuring the slant range thus obtained. The slant range as indicated in Fig. 17 is measured in the same manner as a normal range.

Basic Requirements

The minimum requirements for the basic radar system are therefore:

(*a*) A suitable transmitter.

(*b*) A sensitive receiver.

(*c*) A device capable of measuring intervals of time of the order of a microsecond or less.

(*d*) An aerial system having highly directive properties and capable of being rotated through any desired angle. If designed for use against airborne targets it must be capable of being tilted to an extreme elevation of at least 45°.

An essential *condition of operation* is that the transmitter and receiver must be arranged in such a manner that received echoes can be readily identified with the individual transmissions from which they result.

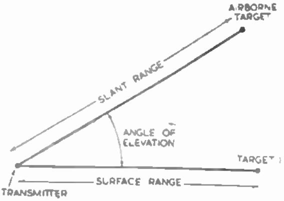


FIG. 17.—TARGET HEIGHT MEASUREMENT.

The above condition is satisfied by pulsing the transmitter so that it generates short, sharp bursts of R.F. of the order of a microsecond or so in duration, with a sufficiently long interval or resting time between each successive pulse to permit all echoes from the extreme limits of the service area time to return before the commencement of the next pulse, as shown in Fig. 18.

The pulse length, or duration of each pulse (in microseconds), and the rate at which successive pulses are repeated (repetition rate or pulse frequency) are determined in design by the performance and duties which a particular radar system is required to fulfil.

Military Applications

Military applications can be generally classified under the following headings :

(a) *Radar Sets for Long-range Warning and Search.* Fixed shore stations or ship stations fitted with these sets keep a constant watch, and search some specific area continuously, in order to give early warning of the approach of hostile aircraft or surface craft. The major requirement for this class of set is maximum range.

(b) *Fire-control or Gunnery Sets.* Radar sets installed for this purpose must be capable of measuring range, bearing and/or elevation with great accuracy. In general, these sets are designed for a degree of precision in measurement which is not practicable for long-range warning sets.

(c) *Airborne Sets.* Light, portable equipment is used by patrol aircraft when searching for enemy targets. The equipment may be designed to detect other aircraft, surface vessels or submarines. The radar equipment may also be used in such cases as an aid to navigation to determine course or position with reference to a home-based beacon.

The main requirements for airborne equipment are that it should be light and compact, and within limits sets must be designed for maximum range, definition and/or precise measurement, according to their particular application.

(d) *Radar Beacon Stations.* These are generally small, compact transmitter–receiver sets, in which the receiver may be actuated by radar transmission from a ship to cause the beacon transmitter to radiate radar signals. This transmission is treated as an echo at the calling station, where it is measured in the normal manner for range and bearing.

Maritime Radar

Shipborne radar is now fitted to many thousands of vessels, and is chiefly of assistance in avoiding collision during fog and for navigational

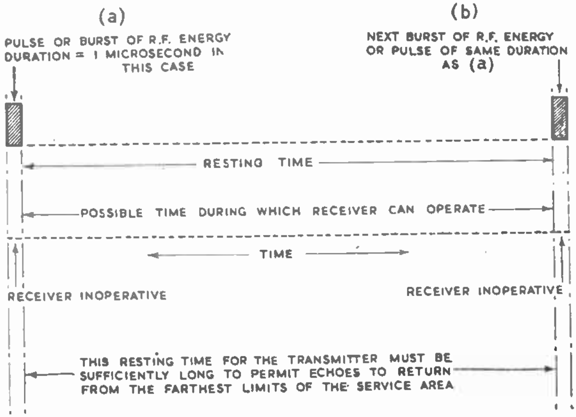


FIG. 18.—RELATION OF PULSE LENGTH TO REPETITION FREQUENCY.

The transmitter and receiver must be arranged so that the echoes can be identified with their transmitted pulses.

purposes in coastal waters. A number of different equipments have been developed by various manufacturers, but the following figures are typical of modern installations. P.P.I. presentation on a 5–12-in. cathode-ray tube with a scan rate of from 20 to 30 r.p.m. Peak radio-frequency powers from 7 to 60 kW. Pulse lengths from about 0·1 micro-second for short-range working, to about 1 microsecond for longer ranges. Most equipments provide a switched selection of ranges from about ½ or 1 mile to about 40–50 miles. Automatic frequency control is often incorporated, and single or double cheese-type scanning aerials are general. Most modern equipments operate on the 3-cm. band (9,320–9,500 Mc/s), but the 10-cm. band (5,460–5,650 Mc/s) is still used in some installations.

Ship-borne radar is not always capable of guiding a ship in fog through a narrow channel to its berth, and for this purpose harbour radar installations have been developed. Such installations may comprise several radar scanners located at various points in the harbour and its approaches with the information presented on a series of P.P.I. screens at a central control point, which is provided with arrangements for the passing of advice—usually by V.H.F. radio-telephony links—to the pilots bringing in the vessels.

BASIC PRINCIPLES OF RADAR

The basic set-up for a radar system is shown in Fig. 19, in block diagram form. This consists of a transmitter or R.F. generator, a sensitive receiver with video output, an indicator which includes a cathode-ray tube and its associated time-base and amplifier circuits, an aerial system including radiator, feeder system and transmitting–

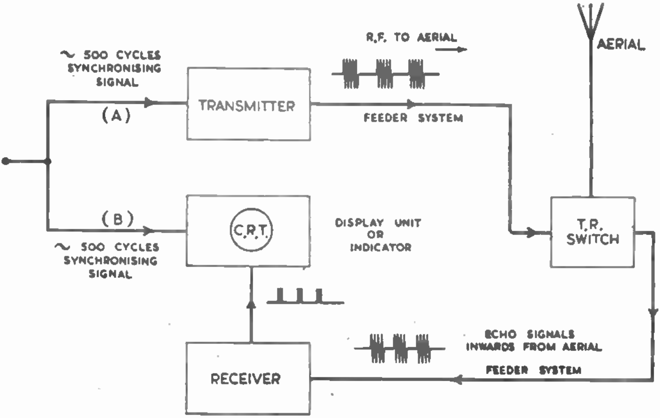


Fig. 19.—Basic Radar System.

A and B are synchronizing signals supplied from, say, the A.C. mains at 500 c/s, to the transmitter unit and display unit, respectively.

A and B establish the repetition rate for the System at 500 c/s, and start the time-base generator in the indicator at the same instant that each transmitted pulse of R.F. commences.

B also times the application of a positive pulse to the control grid of the cathode-ray tube in order to intensify the electron beam from the instant that the time-base generator starts up, until it completes one sweep.

All these operations are repeated 500 times per second.

receiving switch. Finally, the system must incorporate some means for co-ordinating, or synchronizing the following operations :

(*a*) The commencement of each pulse of R.F. energy.

(*b*) The commencement of each sweep of the time-base voltage, i.e., the commencement of each trace.

(c) The brightening up of the trace for the period of each sweep, i.e., increasing the intensity of the electron beam for each sweep duration.

The flyback is timed by the fall of the time-base voltage to zero, and the dimming of the electron beam takes place at the end of each completed sweep of the time-base voltage because the duration of each brightening pulse is made equal to the duration of the sweep or to the time for the completion of each trace. The synchronizing function for (*a*), (*b*) and (c) may be performed by pulses from : (1) a master synchronizing circuit, (2) by a pulse from the transmitter to the indicator, (3) sometimes by a pulse from the indicator to the transmitter. The methods enumerated above are not the only alternatives, but they serve to indicate the principle employed for synchronizing the operations of the units concerned. As more refinements are employed, additional pulses must be supplied from the synchronizing or timing source, in order to perform the additional co-ordinating functions.

No matter how many synchronizing pulses are required, they must all be controlled from the same source.

In some cases it is necessary to render the receiver inoperative during the period of transmission. This can be done by means of a pulse which increases the bias of some of the valves in the first I.F. stage for the duration of the transmitting pulse; alternatively, the H.T. can be removed entirely from one or more valves in the receiver during transmission. This necessity arises when it is desired to receive echoes from targets very close to the transmitter, the reason being that the time between the end of each transmitted pulse and the instant of arrival of the echo is so short that the receiver must be in a state to receive efficiently immediately the transmitted pulse has ceased. If the transmitter employs high power it tends to paralyse the receiver for a short time after transmission ceases, due to static charge accumulated in bias circuits and the like. In these circumstances it is usual to render the receiver inoperative, or partially so, during each transmission period, thereby reducing the *recovery time* and enabling signals from nearby targets to be received. *Recovery time* determines *minimum range*.

Fig. 20 shows the time relationships for operation of the major functions in the basic radar system.

Calibration and Range-mark Generators

Expansion of the basic radar system by the addition of refinements is exemplified by the provision for calibration, which is made in nearly every case.

The method of subdividing the cathode-ray tube trace by means of a superimposed transparent scale is obviously not in itself very accurate. Variations in the time-base circuits, due to ageing or change of values of components, must cause inequalities and change of speed in the rate of deflection of the beam across the screen, and therefore in the trace which it produces. A calibrator circuit is provided for detecting and correcting such changes. In the simplest form of calibrator a generator is caused to start up at the same time as the time-base generator by a synchronizing pulse. This generator produces signals at equal intervals of time, which are in effect equal time subdivisions of the main time-base trace. The signals generated by the calibrator circuits are mixed with the receiver output on the "Y" plates of the "A" display, and appear along the trace as bright marks at *equal time intervals*. If the time-base circuits have altered, these will not line up with the fixed subdivisions of the transparency. In this case provision is made whereby the speed of the time-base circuit can be adjusted in order to bring the calibration marks into alignment with those of the fixed scale. This operation must be performed periodically, and a pulse additional to those shown in Fig. 20 is required from the synchronizing source for the calibration generator.

In many cases, where the A.C. supply is at 500 c/s, this is used as the synchronizing source from which the synchronizing pulses are sent out to the various units which are to be co-ordinated. In other cases, where the frequency of the supply is too low, a master circuit develops a synchronizing signal at the required frequency. This may be done by a Wien Bridge Oscillator, for example. Some forms of transmitter, however, develop their own repetition frequency, and in this case a pulse may be taken from the transmitter to synchronize the other

DEPARTMENT OF DEFENSE

WIRING COLOR CODE FOR ELECTRONIC EQUIP.

CIRCUIT	COLOR
Grounds, grounded elements & returns	Black
Heaters or filaments, off ground	Brown
Power supply, B plus	Red
Screen grids	Orange
Cathodes	Yellow
Control grids	Green
Plates	Bl
Power supply, minus	Violet (purp
A-C power lines	Gray
Miscellaneous, above or below ground returns, AVC, etc.	White

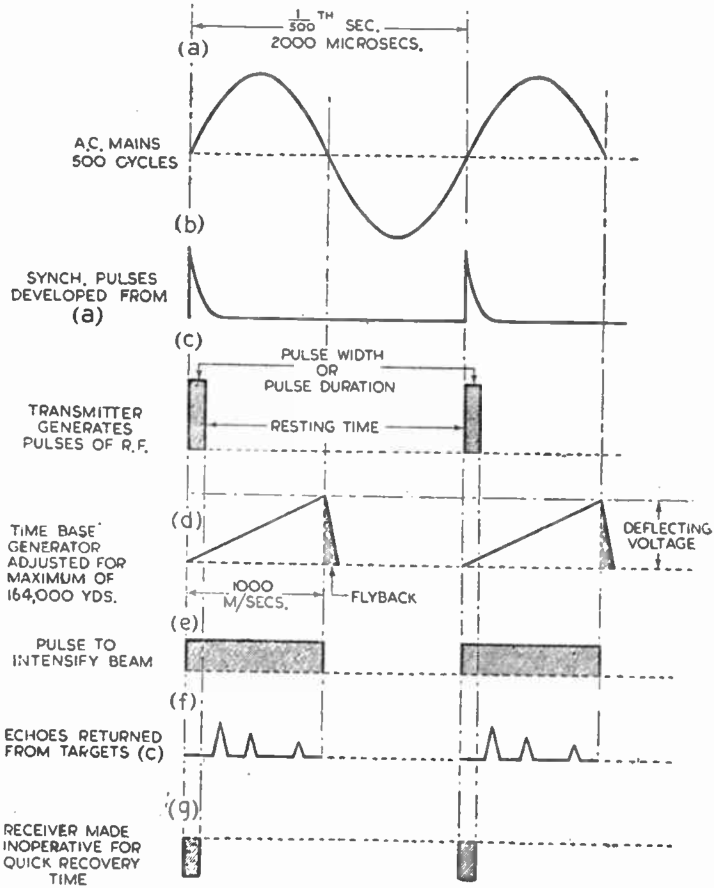


FIG. 20.—TIME RELATIONSHIPS IN BASIC RADAR SYSTEM.

Major units synchronized by external master timer. In this instance, the A.C. supply is at 500 c/s.

units and the necessity for an external master timing device or synchronizer does not arise. Also cases arise where it is necessary to synchronize all the units by means of a pulse from either the sweep generator or from the Range-mark generator in the indicator unit.

Major Units and Constants

The form of transmitter selected for a radar system depends almost entirely upon the frequency which has been selected, and this in turn, as we have already seen, depends upon the application. Standard triodes, due to their inter-electrode capacity, are unsuited to operate in circuits generating frequencies of more than about 200 Mc/s. Special triodes are available for frequencies up to about 400 Mc/s, but for all frequencies above about 400 Mc/s the magnetron is almost universally employed in R.F. generators.

The radar bands are designated as follows :

P band	225–390 Mc/s
L band	390–1,550 Mc/s
S band (10-cm. band)	1,550–5,200 Mc/s
X band (3-cm. band)	5,200–10,900 Mc/s
K band (1-cm. band)	10,900–36,000 Mc/s
Q band (8-mm. band)	36,000–46,000 Mc/s
V band (6-mm. band)	46,000–56,000 Mc/s

The constants of the radar system are pulse width or pulse duration, repetition frequency or repetition rate, average power, peak power and duty cycle.

PULSE WIDTH OR PULSE DURATION. The pulse width or pulse duration is the length of time in each transmission for which R.F. energy is radiated. This is only another way of saying that it is the length of time for which the R.F. generator is operating. The *minimum range* at which a target can be received is determined largely by the pulse width. If the target is so close to the receiver that the echo is returned to it before the transmitter is turned off, the echo will be masked. The maximum limit of pulse width for a particular system is therefore determined by the nearness of the target which the radar set is required to detect. In the case of long-range warning sets, the pulse width may be longer than for short-range, precision sets, especially if the latter are to be designed to pick up buoys and similar navigational marks at close range for navigational purposes.

PULSE-REPETITION FREQUENCY OR RATE. Sufficient time must be allowed between transmitted pulses for an echo to return from the most distant target at extreme range. If this is not done, returning echoes will be obscured by succeeding transmitted pulses. This necessary time interval fixes the highest frequency that can be used for the pulse-repetition rate. It means that the lower limit of the permissible repetition rate for a long-range warning set is below that of the lowest permissible rate for a short-range precision set. When the aerial is rotated at constant speed, as is the case in some applications, the beam of energy strikes a target for a very short time. therefore a sufficient number of pulses must be transmitted in order to ensure a reasonable echo. The persistence of the screen of the cathode-ray tube and the rotating speed of the aerial also affect the minimum desirable repetition rate.

AVERAGE POWER. The power output rating of a valve as given by the manufacturer is determined by its power-handling capacity over a continuous period of time. This is known as *Average Power.*

Peak Power. In radar, transmission takes place in short pulses with comparatively long resting periods in which there is time for the valve to cool. When operated in this manner, a valve can handle a peak power, for very short periods, of many times the value of its average handling power capacity for continuous working. The useful power of a radar transmitter is contained in the radiated pulses, and is therefore termed the *Peak Power* of the system. Since the radar transmitter is resting for a long time compared with its operating time, the average power delivered during one cycle of operation is quite low compared with the peak power available during pulse time.

A definite relationship exists between the average power dissipated over an extended period of time, and the peak power developed during pulse time. The time for 1 cycle of operation is 1/frequency, thus the greater the pulse width the higher the *average power*, and the longer the pulse-repetition time the lower the *average power.*

Therefore :

$$\frac{\text{Average power}}{\text{Peak power}} = \frac{\text{Pulse width}}{\text{Pulse-repetition } \textit{time}}$$

The operation of the radar transmitter can be described in terms of the fraction of the total time that the R.F. energy is radiated. This time relationship is called the duty cycle :

$$\text{Duty cycle} = \frac{\text{Pulse width}}{\text{Pulse-repetition time}}$$

From the foregoing considerations it is clearly desirable to keep the duty cycle small in order that the peak power may be as large as possible.

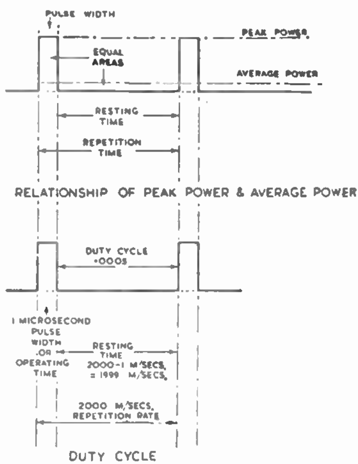


Fig. 21.—Relationship of Peak Power, Average Power and Duty Cycle.

Display Units or Indicators

The three principal forms of display are as follows :

The "A" Display. In this case amplitude is plotted against time on the cathode-ray-tube screen. The time-base voltage is applied to the "X" plates and the receiver output is applied to the "Y" plates, the general form of the display is as shown in Fig. 22.

The "B" Display. In this display the time-base may be applied to the "Y" plates, and a voltage proportional to the bearing of the aerial is applied to the "X" plates. The output of the receiver is applied to the grid of the cathode-ray tube to modulate the brilliance of the beam. In this arrangement the beam scans the whole area of the screen in comparative darkness until an echo is received. The arrival of an echo intensifies the beam, and a bright spot appears on the screen at the position taken up by the beam at that instant (Fig. 23). Since the beam scans the screen under the combined influence of the time-base voltage and the bearing voltage, the point at which the spot appears indicates the range and bearing of the target at that instant, with regard to some fixed reference mark. In this display, therefore, range is plotted against bearing.

THE P.P.I. DISPLAY. The P.P.I. display presents, in polar co-ordinates, a map of the area being covered, with the aerial occupying the centre of the screen. The cathode-ray tube is intensity modulated. The time-base sweep moves from the centre radially outward. The position of the radial time-base sweep is controlled by, and synchronized with, the aerial position through 360° of rotation. The top of the screen represents dead ahead. If the aerial is pointing dead ahead, the sweep moves from the centre of the screen to the top. Likewise if the aerial points 90° from dead ahead, the sweep of the time-base moves from the centre, radially outwards, at an angle of 90° to the right of dead ahead. Thus a polar map is developed on which range is plotted radially against bearing through 360°. This type of display is applied largely to search, harbour control, convoy keeping, ground controlled interception and navigation.

An electromagnetic tube is usually employed, and the display obtained by causing the time-base to start from the centre of the screen outwards, so that each trace extends from the centre to the circumference of a circle the radius of which is made proportional to range. The coil carrying the time-base current is rotated around the neck of the tube by a small motor, the motor being driven in synchronism with the aerial, which is also continuously rotated. If the output of the receiver is applied to the grid or cathode to give intensity modulation, echoes from all targets in the area will appear on the screen as bright arcs, situated at distances from the centre which are proportional to range, and positioned for bearing relative to true north (or some fixed reference bearing) on land, or to the ship's head. The beam may either be brightened by a brightening pulse or suppressed by a suppressing pulse.

In order to augment its usefulness, it is customary to add range markers and some reference bearing mark to the P.P.I. It is often desirable to

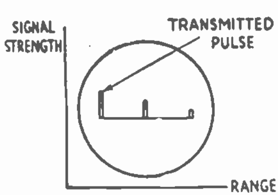


FIG. 22.—APPEARANCE OF THE "A" DISPLAY.

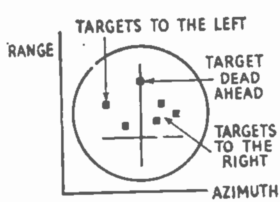


FIG. 23.—APPEARANCE OF THE "B" DISPLAY.

In this display range is plotted against bearing.

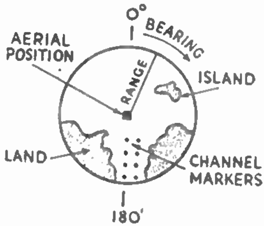


FIG. 24.—APPEARANCE OF THE "P.P.I." DISPLAY.

A type of display which is used for searching, harbour control and navigation.

see nearby signals on a more expanded scale than that used for distant signals, and the P.P.I. usually has several ranges from 1 to 50 miles or more.

The Use of the Cathode-ray Tube

It has already been stated that pulsed radar systems measure range or distance in terms of *time*. This is made possible by the development of the cathode-ray tube as a device for measuring very short periods of time and by the knowledge that all ether waves travel in free space with a velocity of 186,000 miles per second or 328 yd. per microsecond (approximately).

It is evident that the energy reflected from a target travels back in the direction of the transmitter with the same velocity as the transmitted wave travelled outwards to the target, consequently the time taken for the reflected energy to travel back from the target is exactly half the time elapsing between the commencement of each pulse of transmitted energy and the return of the echo.

The range calibration of the cathode-ray tube can therefore be expressed as :

$$\text{Range} = \frac{\text{Observed time for outward and return journey}}{2} \text{ (microseconds)}$$

multiplied by the velocity of ether waves in free space (328 yd. per microsecond).

$$\text{Range (yd.)} = \frac{\text{Observed } \textit{time} \text{ in microseconds} \times 328}{2} = 164 \text{ yd. per}$$

microsecond.

This is the fundamental *time–range* relationship which is used to calibrate the cathode-ray tube.

Accurate Range Measurement

Since a given range may be represented by the total length of the time-base trace, the range of a target may be estimated by observing the fraction of the trace between its commencement and the leading edge of the echo. Estimates of this kind, however, depend upon the judgment of the operator, even when a transparent, subdivided scale is supplied. It is, therefore, usual to provide a circuit which generates

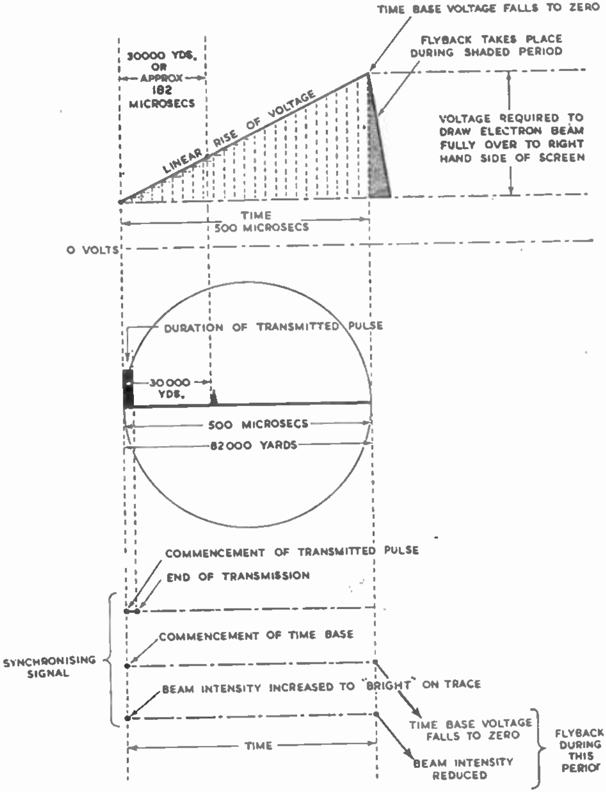


Fig. 25.—A Relation of Range to Form of Time-base Voltage Wave.

This example shows: Sweep time for time-base say, 500 microseconds = 500 × 164 yd. = 82,000 yd. Range of Target = say, 30,000 yd.

under the control of a synchronizing pulse, very short pulses at regular intervals—corresponding in time perhaps to 1,000 yd. These pulses are mixed with the receiver output on the " Y " plates, in the case of an " A " display, and appear as bright marks at intervals corresponding to 1,000 yd. along the trace, or as range circles on a P.P.I. display.

For precise measurement this method is not sufficiently accurate, since it still becomes necessary for the operator to interpolate for intermediate readings. The accuracy with which this is performed varies with the individual. Therefore, when precise measurement is required within a few yards, other means are provided for this purpose. One method, which gives reasonably satisfactory results, is to compare the time-base voltage, measured at the instant of time that the echo is received, with the voltage of a standard, calibrated potentiometer.

Various types of indicators have other refinements. For example, it is possible to select the small portion of trace containing an echo in which the observer is interested, and to expand this small fraction so that it covers the entire width of the screen. This permits examination of the particular echo in detail, and without distraction by other echoes which may be present at the same time. It also has other beneficial results. In the case of a plane on interception duty the observer may be provided with a " B " type display on which, by manipulation of the radar controls under direction of R.T. from the ground, the target can be displayed. In order that the pilot may follow the target without distraction, the observer can select from his display on the " B " indicator the portion containing the target of interest and cause a reproduction of this portion to appear on the display unit situated at the pilot's position. This type of display is termed a selected range display.

Beam Widths

Further consideration of the matter shows that beams which are narrow in the horizontal plane are also desirable for accurate measurement of bearing of surface craft or other objects. A narrow beam offers the advantages of increased concentration of energy and better definition (i.e., the ability to pick out and discriminate between adjacent targets). Why, then, are narrow beams not always used? The answer is given below in the following order :

(1) The width of a beam depends upon the number and disposition of the dipoles and the size and geometry of the reflecting system. In other words, the radiating system, including reflector, must be increased in dimensions when narrow beams are required.

(2) The length of the dipoles is determined by the R.F. carrier frequency employed. For long waves the dipole length must be increased, since $\frac{\lambda}{2}$ dipoles are usually employed in the radiating system.

(3) For long-wave systems, therefore, radiating reflectors become very large indeed. They require considerable space when erected, they are mechanically difficult to support and rotate, and the wind resistance becomes very high indeed.

(4) On the other hand, as the wavelength decreases the range obtainable for a given R.F. power output decreases.

Thus the position arises where long-range sets for warning purposes must employ comparatively long waves, using frequencies of the order of 100 Mc/s in order to obtain the necessary range with reasonable power outputs. The beam in this case is, therefore, generally fairly wide—perhaps 20° or more, because at the comparatively low frequencies used for warning sets, radiating systems for narrow beams would become impracticable on account of the dimensions and weights involved.

For the foregoing reasons, therefore, long-range warning sets usually employ a comparatively low carrier frequency and a fairly wide beam, whilst short-range precision sets make use of ultra-high frequencies of the order of 3,000 Mc/s and upwards—(10 cm. and downwards) in order to gain the increased precision of measurement inherent to narrow beams, the use of which now becomes possible, but accompanied by a sacrifice in range.

At about 3.000 Mc/s, the dimensions of reflectors suitable for producing narrow beams become small enough for installation, support and rotation in mobile radar stations.

When wide beams are used it will be seen from Fig. 26 that the *rate of change* of signal strength in the region of the maximum value at " A " is very small indeed. For example, at the angle θ, OB is very little less than OA, which is the maximum value. This makes it very difficult to find with precision the exact bearing of a target. Accordingly, a system of beam splitting is used in order to improve the discriminating properties of the aerial system.

Beam Splitting

Further inspection of Fig. 26 shows that the *maximum rate of change* of field strength, which is what we require for discriminating purposes, takes place at points C and D, the field strength at these points is 70·7 per cent of the maximum, and they are called the " half-power " points of the lobe or pattern.

Assuming that the target " T " is fixed, and that the beam if rotated to right and left of it alternately through the angles AOA′ and AOA″ respectively (Fig. 27), the patterns would intersect at the half-power points C. This means that when the lobe is switched to the right the signal from the target will be proportional to CO, similarly when the lobe is switched to the left the signal from T will also be proportional to CO. In other words, *the signal from T will be proportional to OC if the beam is switched to right or left of the target through this angle.*

Assuming that the aerial continues to swing backwards and forwards through the angle A′OA″, consider the effect on the echoes received from target T1. When the beam swings to the left, a maximum signal will be received from T1, but when it swings to the right, the

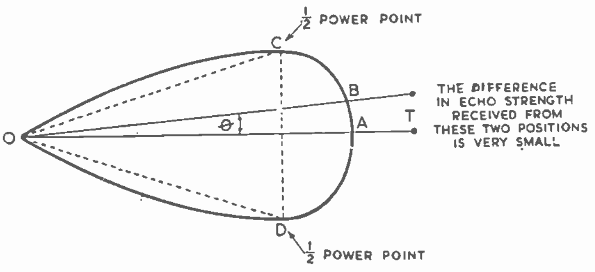


FIG. 26.—BEAM CONSTANTS.

C and D are called half-power points, since the field strength at these points is 70·7 per cent of the maximum OA and because E^2 is proportional to power, i.e., $\left(\frac{1}{\sqrt{2}}\right)^2$.

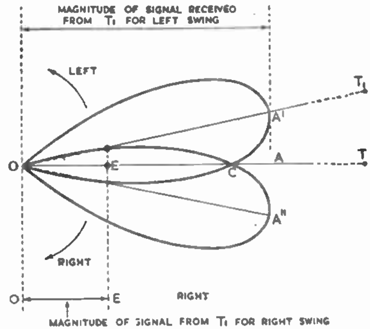


Fig. 27.—Beam Splitting.

Method used to improve accuracy training with wide beams.

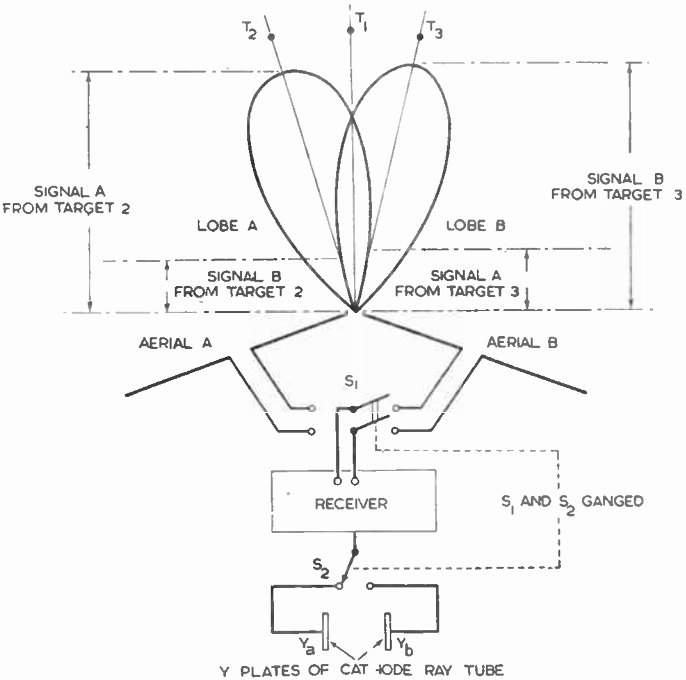


Fig. 28.—Beam Switching.

Schematic diagram of manual switching system using two aerials and one receiver to demonstrate the principle of beam switching.

signal received from T1 falls to a value proportional to EO which is less than CO. This means that the signal received from T1 will be greater for the left-hand swing than for the right-hand swing. These conditions are exactly reversed for the target at T. The conclusion is, therefore, that *equal* signals from right- and left-hand swings are only obtained when the centre of the aerial system exactly faces the target. When the target is in any other positions relative to the aerial the signals received for right- and left-hand swings of the beam are *unequal*. This is the principle generally employed to enable the operator to train the aerial on target when wide beams are used. The field pattern is switched continuously, as the aerial is rotated and signals received from right and left lobes are also continuously compared. When *both* positions of the lobe give *equal* signals the aerial is on target.

Beam switching employs electronic means for switching or displacing the axis of the beam alternately to right and left of the normal axis, so that if the centre of the aerial is held fixed to face the target, the beam itself sweeps right and left across the target and a double lobe is produced.

RADAR AERIALS

The aerial system of a radar equipment usually comprises the following units :

(*a*) The aerial proper, i.e., elements that perform the dual purpose of radiating radio-frequency energy and receiving reflected energy or echoes.

(*b*) The feeder system which joins the output from the radio-frequency generator to the input of the aerial proper.

(*c*) Some form of electronic switching device for isolating the receiver during transmission, and the transmitter during reception.

(*d*) In many cases a beam-switching device for accuracy in determination of bearing.

(*e*) Mechanism for rotating the aerial and controlling its movements, according to operational requirements and the conditions imposed by the type of display employed. Also means for indicating at the operating position the bearing of the aerial at any instant of time.

Requirements of the Aerial System

The aerial system for any radar set must conform to the following requirements :

(*a*) Possess the required directional properties when functioning at the selected frequency.

(*b*) Maximum overall efficiency for transmitting and receiving under condition (*a*).

(*c*) Size and weight must be confined within any limits specified, with due regard to mechanical problems involved in mounting and rotating the structure.

Directional Effects

All practical aerials are directive to some extent. In general, however, the term directive refers to a radiating system which has been designed deliberately to concentrate its radiation in a relatively narrow beam. Directional effects may be obtained by the use of two or more aerials, so placed and phased that the radiated waves from parts of the system add in some preferred direction and cancel in others. A modification of this is a system of radiating elements, disposed in such numbers and so positioned that a common reflector of suitable dimensions causes radiation to take place in the form of a beam.

Parabolic metallic reflecting surfaces may also be used to beam radio-frequency energy in a fashion similar to the beaming of light by a searchlight. This type of beam transmission is most practical when frequencies are sufficiently high (centimetre waves) to permit the dimensions of the reflector to be comparatively small.

General Design

The type of aerial depends primarily upon the frequency employed and also upon the required performance. The required beam dimensions are determined largely by the application, and when this has been decided upon, the type of array and number of elements, and reflecting or directing system, can be selected. The dimensions of the aerial are, of course, determined by the frequency employed.

In general, aerial systems used for long-range warning sets are formed by stacked dipoles, with solid, perforated or rod reflectors. In this case the horizontal or vertical directivity of the beam is determined by the number of dipoles stacked in horizontal and vertical planes. The larger the number of dipoles stacked end to end in width or one above the other in height, the narrower becomes the beam horizontally or vertically.

For wavelengths of the order of 10 cm., parabolic reflectors become practicable in size and weight. For frequencies at which a co-axial feeder is used, a dipole and a reflector are employed to flood the radiator with radio-frequency energy. When a waveguide feeder is used, the

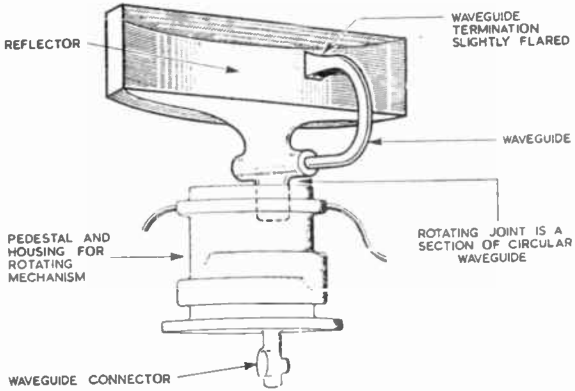


Fig. 29.—Parabolic Reflector Aerial.

R.F. energy issuing from the flared end of the waveguide floods the parabola and is reflected as a beam narrow in the horizontal plane, but wide in the vertical plane.

end of the waveguide may be flared to match the guide to free space, but more often the flare is made comparatively small, and positioned in such a manner that it floods a radiator of design suitable for the production of the required beam. All radiators, whether for long-range warning sets or for precise measurement, are arranged for rotation. Some rotate through a limited arc only, but in the majority of cases they are arranged to rotate through 360°. When a P.P.I. display is used they must be capable of continuous rotation. In addition to rotation in a given plane, aerials designed for radar sets used for airborne targets must also be capable of being elevated.

FEEDER SYSTEMS

The choice of feeder is affected by the general nature of the equipment, type and layout of units, and the working frequency. The last is very often the determining factor. There are three main types of feeder :

(*a*) Open-wire feeder.
(*b*) Co-axial feeder.
(*c*) Waveguide.

The first is only suitable for the longer radar wavelengths, the other two being applicable to microwave frequencies.

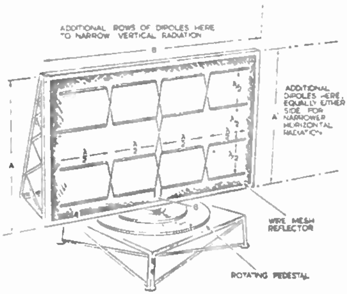


FIG. 30.—A COMMON TYPE OF AERIAL FOR USE WITH LONG-RANGE WARNING SETS, USING STACKED DIPOLES.

Open-wire Feeder

The open-wire feeder simply consists of two parallel wires spaced a suitable distance apart. The wires can be supported on insulators, or use may be made of the "quarter-wave section" which presents a very high impedance across one end when the other end is short-circuited (Fig. 31).

The great objection to the open-wire feeder is the amount of power lost by radiation at the higher frequencies, and since the development of radar has tended towards the use of higher and higher frequencies, this type of feeder may be regarded as obsolescent.

Co-axial Feeder

This occupies an intermediate place between open-wire feeders and waveguides. The co-axial feeder, consisting of two concentric conductors, the inner normally a rod or wire and the outer a tube, is easy to install and does not radiate. The dielectric may be air or some suitable solid dielectric such as polythene. If air is used, either dielectric spacers or quarter-wave stubs are used as support for the inner conductor.

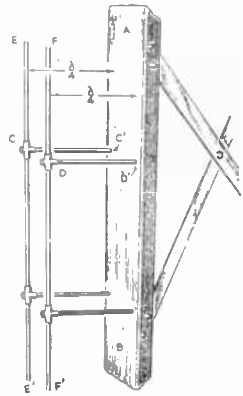


Fig. 31.—Open-wire Feeder.

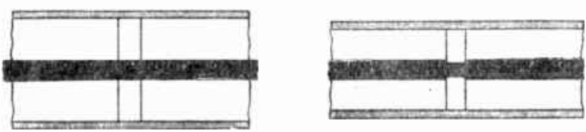
AB is a metal member of a mast or tower, shorting the sections CC′ and DD′. EE′ and FF′ are open-wire transmission lines. This construction is only possible for one fixed frequency.

Since the aerial system is usually required to turn in azimuth and elevation relative to the remainder of the equipment, a rotating joint is necessary. The basic method of construction, employing the quarter-wave principle, is shown in Fig. 34. As the quarter-wave section is open-circuited at CD, the impedance across AB is zero, thus effecting the required connection between the two lengths of outer conductor.

The power-handling capacity of the feeder depends upon the spacing between conductors, and can be increased by increasing the dimensions. In practice, there is, however, an upper limit to the permissible size. If, for any given frequency, the diameter of the outer conductor is increased above a certain size, the feeder is capable of acting as a waveguide, an action which interferes with normal propagation. For wavelengths of about 10 cm. the feeder diameter cannot be made greater than about 2½ in., or a little less. This naturally sets a limit to the power that can be transmitted by co-axial feeder at the higher frequencies.

Waveguides

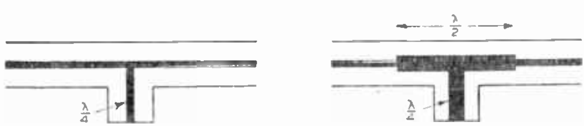
A waveguide normally consists of a metallic tube which may be either circular or rectangular. Subject to certain conditions, such a tube



(a) (b)

Fig. 32.—Co-Axial Feeders.

Inner conductor supported on dielectric spacers.



(a) (b)

Fig. 33.—Co-axial Feeders.

Inner conductor supported by quarter-wave stubs.

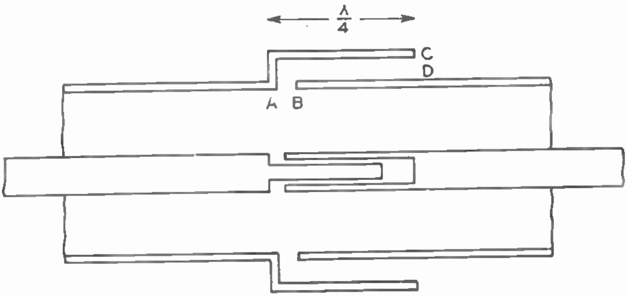


Fig. 34.—Rotating Joint for Co-axial Feeder.

permits electromagnetic energy to travel through it in much the same way, although at a lower velocity, as radiation can travel through space.

The fact that a waveguide must have a minimum dimension which is proportional to the working wavelength explains why waveguides are only practicable for the higher frequencies. For such frequencies, however, they possess definite advantages over co-axial feeders. The absence of an inner conductor results in considerable mechanical simplification, and the power-handling capacity and attenuation are both improved.

The advantages of waveguides may be enumerated as follows: (1) Complete shielding and therefore no radiation loss. (2) No dielectric loss. (3) Copper loss less than that of a co-axial line of the same size operated at the same frequency. (4) Greater power-handling capacity than for a co-axial line of the same size. This arises because power is proportional to E^2, since a waveguide has no centre conductor, the distance between opposite potential surfaces for equal external dimensions is twice that of a co-axial line. This means that the waveguide can handle larger voltages without arcing over, and therefore more power. (5) Finally, construction with a waveguide feeder is simpler than with co-axial line.

DUPLEXING

The parallel problems of (a) avoiding power losses in the receiving circuits during transmission, and in the transmission circuits during reception, and (b) protection of the receiver during transmission, is sometimes solved by using separate aerials for transmission and reception. Since this method is uneconomical, a common aerial is generally employed, together with some form of electronic switch which performs functions (a) and (b) quite satisfactorily.

This switch depends for its action upon the properties of quarter-wave and half-wave sections of transmission line or waveguide when used in conjunction with spark gaps and/or resonant cavities. The system is described under duplexing systems.

The problem of switching is simplified to a certain extent by the fact that the impedance of the transmitter during operation is different from that during resting periods. Fig. 35 shows the principle and simplest form of T.R. switching. T is a junction point in the feed line, where the receiver feeder joins the transmitter feeder, both being connected to the radiator via the third feeder. For purposes of illustration it is assumed that the characteristic impedance of the transmission line, the feed-point resistance of the aerial, the input impedance of the receiver and the output impedance of the transmitter when generating radio-frequency power are all 250 ohms. The transmitter output impedance rises between pulses to 5,000 ohms, and the resistance of the conducting gap is 50 ohms. The pulse from the transmitter divides at the T-joint; part goes to the receiver branch and causes the spark-gap to break down, and the rest goes to the aerial. The breakdown of the spark-gap places a resistance of 50 ohms across the 250-ohm line just a quarter-wavelength from the T-junction. By transmission-line theory, the T-junction end of the quarter-wave line, now terminated in 50 ohms at the receiver end, will exhibit an impedance of 1,250 ohms at the T-junction, and the transmitter pulse has the choice of a 1,250-ohm path to the spark-gap or a 250-ohm path to the aerial. Since the aerial terminates the transmission line in its characteristic impedance, most of the energy passes to the aerial, whilst the remainder is used to keep the spark-gap going.

At the end of the pulse the gap is deionized and receiving signals reach the T-junction. Again there is a choice of two paths. The receiver path has an impedance of 250 ohms (since the spark-gap is now open-circuited). The path to the transmitter is, however, made a half wavelength, and since this is terminated in 5,000 ohms, received signals coming towards the T-junction meet an impedance of 5,000 ohms. Most of the received signal now goes to the receiver, and very little goes to the transmitter.

T.B. or A.T.R. Switch

When the output resistance of the transmitter does not change sufficiently to allow the use of a resonant line alone in blocking receiver signals from the transmitter, a second spark-gap and a second switch, the T.B. (Transmitter Blocker) or A.T.R. (anti-T.R.) switch, are used. In Fig. 36 the transmitted pulse passes down the line to spark-gap 2, and causes it to arc over, the resulting short-circuit is reflected back to the main feeder as a high impedance by the action of the quarter-wave line. During the transmitted pulse, the T.B. switch merely uses a small amount of power in its spark-gap and the T.R. switch functions as before. The T.B. switch is in effect an open-circuit during the resting period because the shorting bar which is a quarter-wave below the gap is reflected by the half-wave line of the T.B. switch, as a short-circuit across the transmitter feed line. This is reflected in turn to the T-junction as a high impedance by the qu ter-wave section of the feed line between the two joints, thus blocking the transmitter channel for all received signals.

T.R. Switching applied to Waveguides

A similar system is used to perform the T.R. function when waveguides are used. Instead of the resonant lines, however, resonant

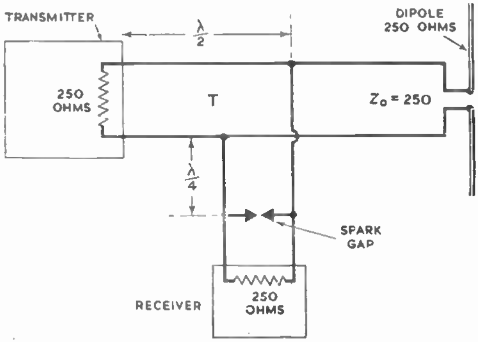


FIG. 35.—ELEMENTARY ARRANGEMENT FOR T.R. SWITCHING.

The pulse from the transmitter divides at the T-joint; part goes to the receiver branch and causes the spark gap to break down, the rest goes to the aerial.

cavities are employed with internal spark-gaps across the high-potential sides of the cavity.

Duplexing at Microwave Frequencies

The same general considerations apply as in the case of longer wavelengths, but as the mixer input circuit for microwave operation is subject to damage more easily than the comparatively sturdy radio-frequency amplifiers which form the input circuit at lower frequencies, special protective measures must be taken. The T.R. cell must be very carefully designed if it is to protect the crystal used for mixing satisfactorily. A gas-filled gap must be used, and since the voltage drop across the gap, once the discharge has started, is nearly independent of the current flowing, the principle of transformer action must be employed in order to keep down the effective value of this voltage as seen from the receiver circuit. This " transformer " is a resonant cavity with a high Q.

AERIAL CONTROL

It is frequently necessary to drive heavy aerials at varying speeds, in either direction of rotation, and to maintain accurate control of their motion. Special D.C. motors must generally be used for this purpose. A simple drive for rotating, under control, a radar aerial, is shown in Fig. 38. The D.C. generator, direct-coupled to an A.C. motor, has its field connected to a potentiometer in such a way that the magnitude and polarity of the voltage applied to the armature of the D.C. motor can be varied, and the aerial can be rotated in either direction, at any speed from zero to maximum.

The basic system has a number of limiting factors. As a rule, it is desired to control the directon in which the aerial points, and the aerial may be required to rotate continuously for normal searching, or to be turned only a few degrees in azimuth to determine the bearing of a particular target. Also the driving system must be capable of supply-

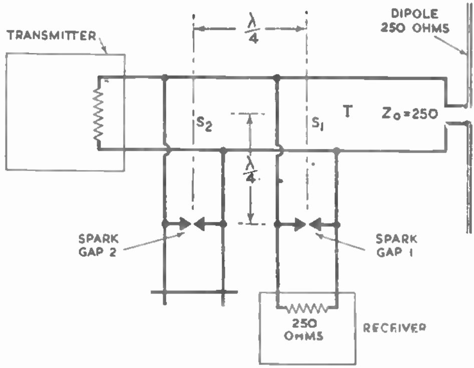


FIG. 36.—ANOTHER FORM OF T.R. SWITCHING.

A second spark gap is employed when the simple T.R. switch affords insufficient protection.

ing sufficient power to make a large aerial rotate in step with the controls in spite of varying wind pressure. Devices used are the magslip or the " Selsyn ".

Such a system is shown in Fig. 39. The field supply for the aerial-driving motor and the D.C. generator is obtained from a small self-excited D.C. generator, driven by an A.C. motor. The D.C. generator field supply from the exciter is regulated by a resistance-bridge arrangement. When the bridge is balanced, the voltage supply to the field of the D.C. generator is zero and the aerial-driving motor is stopped. Any change in the resistance of the balancing arms results in field current to the D.C. generator, the resulting rotation depending upon the magnitude and direction of the change taking place in the electrical equilibrium of the bridge. The shorting contacts are operated by a lever arm L geared to a differential Selsyn motor. Thus the speed and direction of rotation of the aerial are determined by the lever arm, which controls, via the shorting contacts and the bridge, the magnitude and direction of the voltage impressed on the main D.C. generator field.

If the handwheel is turned when the aerial is in the stopped position, G_3 energizes the stator of the differential Selsyn and the rotor DM_1 turns, causing a deflection of the lever arm L. This deflection raises the shorting contacts momentarily to change to a new shorting position. The D.C. generator field is energized, and the drive motor turns the aerial in the same direction as the handwheel was operated. The system G_1M_1 forms a remote indicating system.

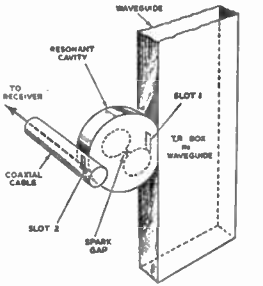


FIG. 37.—T.R. SWITCHING FOR WAVEGUIDE SYSTEMS.

A resonant cavity and spark gap are employed. The T.R. box is coupled to the waveguide by slot 1 and coupled to the receiver via slot 2 and the co-axial line.

Airborne aerials often include stabilizing devices to eliminate distortion due to roll and, sometimes, pitch. In "line-of-sight" stabilization, the scanner base is fastened rigidly to the aircraft, and the beam is held pointed in the right direction by lifting the aerial as the scanner rotates. The scanner of a well-known cloud and collision warning equipment consists of a paraboloid illuminated by a back-fed dipole and reflector, stabilized in both roll and pitch so that whatever the altitude of the aircraft within the limits of $\pm 10°$ in pitch and $\pm 45°$ in roll, the scanner reference platform is horizontal. A gyroscope is used as reference for stabilizing purposes, in conjunction with the servo amplifier unit and stabilizing motors.

Marine Radar Aerials

The aerial is mounted as free as possible of the obstacles on shipboard, such as the funnel, ventilators, derricks, etc., in order to be able to search the horizon freely. It is usually placed on a special mast equipped with a platform. Although high mounting is necessary in order to achieve a large maximum range, it is not desirable from the standpoint of minimum range, as the beam may then pass beyond nearby objects such as buoys.

Airborne Radar Aerials

The size of an airborne aerial is limited by the finite size of the aircraft, and by the possible reduction in aircraft performance resulting

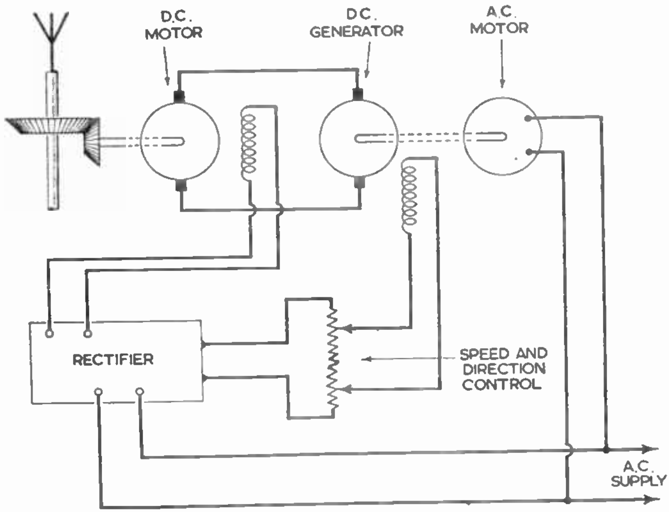


FIG. 38.—SIMPLE AERIAL DRIVING SYSTEM.

Special D.C. motors must generally be used, as the aerial must be driven at varying speeds in either direction of rotation.

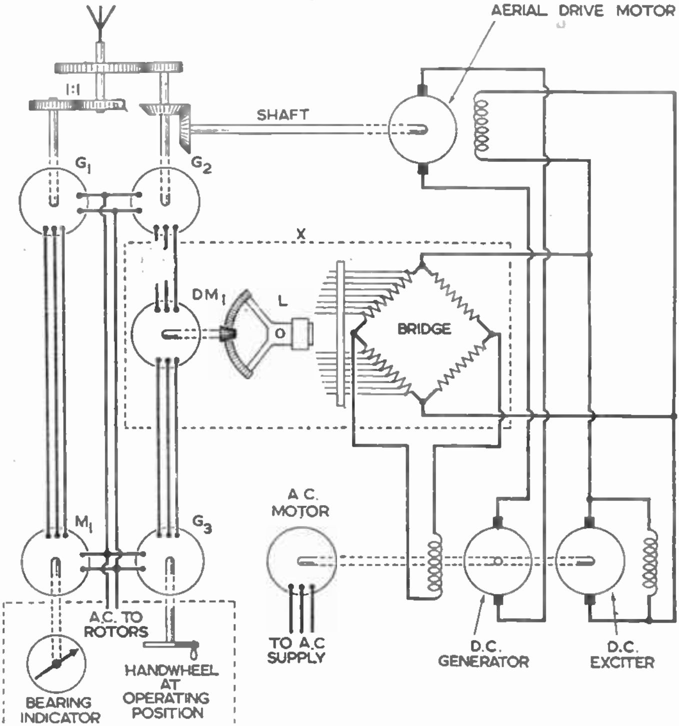


FIG. 39.—AERIAL DRIVING SYSTEM.

This more elaborate system employs a differential Selsyn motor, to which the lever arm L is geared. This system supplies sufficient power to make a large aerial rotate in step with the controls despite varying wind pressure, and allows the aerial to rotate continuously or to be turned only a few degrees by hand for examination of a particular target.

from increased drag. The necessity for a radar view of the ground restricts the possible locations. Aerials are commonly installed in the nose, in the bottom of the fuselage, in the leading edge of a wing or in a separate nacelle mounted above the fuselage. A removable, streamlined installation hung beneath a wing can contain all the radar components except the control panel and indicator.

TRANSMITTERS

Transmitters for radar may be classified under two main headings, viz., transmitters for frequencies up to about 400 Mc/s, and transmitters for frequencies above this value. This first classification arises from the fact that triodes and special triodes may be used in R.F. generating circuits, efficiently, up to about 400 Mc/s; for frequencies above this value, the magnetron is almost universally employed.

The limiting frequency for triodes and special triodes is brought about by:

(*a*) Inter-electrode capacity of the valve.

(*b*) Inductance of the electrode leads.

(*c*) Transit time of the electron.

(*d*) Physical shrinking of external capacitances and inductances to impracticably small sizes.

(*e*) Consequent approach to the frequency limit of the valve itself, i.e., the oscillation frequency that will occur due to its internal capacities and inductances when the electrodes are externally connected in circuit by the shortest possible conductors.

(*f*) Losses due to: skin effect, large capacitance charging currents, eddy current losses in adjacent conductors, dielectric loss in the glass envelope, seals and pinches, energy loss by direct radiation from the circuit.

Special triodes have been employed for frequencies up to 600 Mc/s, but generally in very small transmitters at considerably reduced efficiency.

Classification of Transmitters

Sub-classifications of transmitters relate to the method of pulsing: (1) transmitters included in the first classification, up to about 400 Mc/s, may be either self-pulsed or pulsed by the application of a high-voltage power pulse developed externally; (2) transmitters for frequencies above say, 400 Mc/s, and using a magnetron, are always externally pulsed by the application of a very high-voltage high-power pulse developed externally from the magnetron itself. The magnetron and its associated input and output circuits convert this high-voltage high-power pulse into a pulse of R.F. at the required carrier frequency.

Classification of Pulsing Systems

The important difference in the two forms of pulsing is that when a self-pulsed transmitter is used it develops its own repetition frequency, and the repetition frequency so developed is used to synchronize the other units of the system. The frequency of self-pulsing R.F. generators is not, however, very stable, and therefore the repetition frequency is not sufficiently constant for use with sets required for accurate measurements. It has the advantage, however, of being economical as regards space and weight and is relatively simple to operate. Self-pulsed R.F. generators are therefore used when some instability of the repetition frequency can be tolerated and when lightness, simplicity and compactness are important.

High-voltage, high-power pulses for application to R.F. generators can be developed either at low level or at high level. High-voltage, high-power pulses, developed at low level, can be produced at any desired repetition rate, and the degree of rate stability can be established within very narrow limits of variation. This method, however, employs a large number of circuits, occupies considerable space and adds to the weight of the equipment as a whole. It can be used for very precise measurement, and in cases where a suitable A.C. frequency supply or other means of synchronizing is not available. The block diagram in Fig. 40 is an example of the method employed in producing high-voltage high-power pulses from a synchronizing signal at low-voltage level.

Fig. 41 shows the method of producing high-voltage high-power pulses at high level. In this method the frequency of the discharge of high voltage across the spark gap determines the repetition frequency, and the other units of the radar system are usually synchronized by a pulse from the transmitter when the high-level method is employed. This method has several advantages. In the first place it can deal with pulses of very high power, since a spark gap can be designed to operate at any reasonable voltage. Voltages employed in this system to produce the necessary power may be of the order of 30,000 volts or more, and the peak power developed may be of the order of a megawatt.

The combination of artificial line, together with a rotary discharger, is simple and highly efficient. It suffers from two disadvantages, however, viz.: Some slight instability of the repetition frequency due to the vagaries of the discharge. This is inherent in all spark gaps, because ionization as a prelude to discharge is itself somewhat irregular. Secondly, the waveform of the pulse from the artificial line depends entirely upon the goodness of the artificial line. In this system there is no possible opportunity to correct any defects in the waveform of the high-voltage high-power pulse output before it is applied to the magnetron.

From what has already been said regarding the classification of transmitters and the high-voltage high-power pulses by which they are fed, it is clear that examination of any radar system discloses that its composition must be analysed under two separate headings:

1. The major units comprising the system.
2. The pulsing system by which they are operated and synchronized.

Three classes of pulses can be readily distinguished:

(*a*) Trigger pulses proper.
(*b*) High-voltage, high-power pulses to operate the R.F. generator for definite periods of time.
(*c*) Low-voltage, low-power pulses, also having a definite time duration for various purposes.

Development of Pulses

Trigger pulse waveforms are generally sharp-pointed with steep leading edges; rising almost vertically to some given peak voltage. These pulses are generally used when it is desired to drive a valve into conduction or when it is desired to drive it momentarily into the cut-off state. Synchronizing pulses, when they do not combine a timing function, generally fall within the trigger-pulse classification.

Trigger pulses may be developed from a sine wave by a sequence of shaping and dimensioning operations. These might be, for example, a limiting or slicing operation which roughly squares the sine wave, this roughly squared output may then be improved by an operation performed on it by an overdriven amplifier. The output from this stage is generally a good rectangular wave, with steep sides and a duration of approximately half the time period for one cycle of the A.C. input. The next operation is to develop a voltage wave, the peak value of which is proportional to the rate of change of the input voltage across the resistance in an RC circuit (differentiation). In this way a voltage wave with a sharp-pointed peak and steep leading edge may be developed at a repetition frequency equal to that of the input sine wave. Such pulses are very suitable for synchronizing, with or without amplification, according to requirements.

Low-voltage, low-power pulses may be developed in a variety of ways and are suitable for synchronizing and timing the duration of

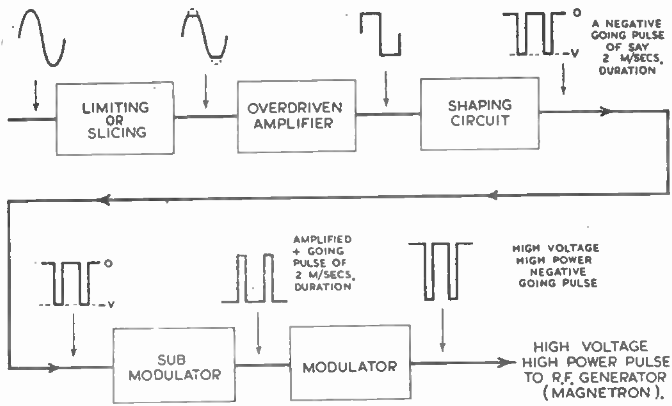


Fig. 40.—Development of High-voltage, High-power Pulses at Low Level.

An example of the method used to produce pulses of this nature from a synchronizing signal at low voltage level.

operation of time-base generators, range-mark generators, strobe generators, brightening pulse generators and various similar functions in the indicator and the receiver circuits.

These low-voltage, low-power, time-control pulses may be developed from a since-wave input by squaring operations similar to those employed in the initial stages of development of the trigger pulse, but instead of a differentiating operation following upon this, the output square wave obtained from the sine wave must be adjusted for time duration. This means that the width of the flat top must be made proportional to any time duration that may be required. This operation can be performed by an amplifier having an input circuit somewhat similar to that which is used for the production of trigger pulses, but the amplifier is biased in a different manner. On the other hand, alternative circuits are available for producing low-voltage, low-power pulses such, for example, as one of the many forms of multi-vibrator circuits, the phantastron, the artificial line, etc.

In general, the pulsing system for any assembly of major units may be laid out in several ways, by using different combinations of pulses and pulse-generating circuits.

The object of the radar engineer is, of course, to lay out the pulsing system in the most efficient and economic manner, hence radar systems differ widely, not only in respect of the types employed for the major units, but also in respect of the entire pulse-system layout, which is designed to be suitable for the requirements of the type of major units employed, and at the same time as efficient and economical as possible.

Beyond the very general statement that circuits and devices for shaping and dimensioning (electrically), the large variety of pulses used

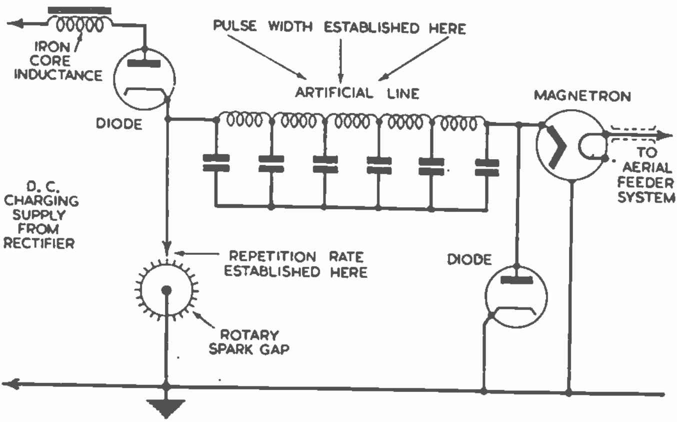


FIG. 41.—DEVELOPMENT OF HIGH-VOLTAGE, HIGH-POWER PULSES AT HIGH LEVEL.

The peak power developed in this system may be in the region of 1 MW.

in radar depend largely upon the deliberate introduction of distortion into amplifier circuits, it is not possible, in the space available, to attempt any detailed description of how these functions are performed. It can, however, be said that a knowledge of the factors which produce distortion in amplifier circuits and the manner in which they operate is essential.

The Self-pulsing Oscillator

In the circuit shown in Fig. 42 the R.F. frequency is determined by the L.C. constants of the circuit. Bias for the valve is provided by grid current which charges C1 through the cathode-grid resistance R1. R1 permits C1 to discharge during the portion of the cycle in which the grid is not positive to the cathode. The result is that the bias on the grid is proportionate to the voltage across the grid-input circuit and stable oscillation conditions exist.

If the time constant of C1, R1 is increased by increasing either C or R the charge on C1 cannot leak off fast enough to follow the fluctuations of voltage caused by irregularities of the electron stream. As a result the charge on C1 grows with each successive cycle, until a point is reached where the voltage across C1 is too high to permit feedback of the amount of energy required from the anode circuit to maintain oscillations. Oscillations therefore cease, and cannot start again until the voltage across the condenser C1 has fallen, by leakage, to a value sufficiently small to pass from the anode circuit the necessary amount of feedback required to start up the oscillatory action of the valve again.

Push–Pull R.F. Generator Circuit

The R.F. generator shown in Fig. 43 is a tuned grid, tuned anode push–pull self-pulsing oscillator, using pairs of quarter-wave sections of transmission line to act as parallel resonant circuits in place of the usual coils and condensers.

The use of quarter-wave transmission lines provides a high "Q" for the oscillatory circuits, and the R.F. frequency stability of the generator is therefore good. Oscillators of this type are usually designed to have as much inductance as possible and to oscillate with the valve internal capacities, the latter providing the channel for the necessary feedback for the maintenance of oscillations. In a push–pull type of oscillator, the inter-electrode capacities are in series and, therefore, their effective capacity is reduced to a minimum. The charging currents are also half the value of that for a single valve.

The transmission lines forming the external resonant circuit are extensions of the internal leads to the valve electrodes. By this arrangement the inductance of the internal leads is used as part of the oscillatory circuits. The "Q" of the circuit is large because the inductance and capacity of the external resonant circuit are distributed, and because the skin effect can be reduced by using conductors of large diameter (with consequent large circumferential conducting area). Resistance can be further reduced by silver plating the tubes forming the resonant lines.

The length of the quarter-wave line in the grid circuit is physically less than the quarter-wave line in the anode circuit, because the internal capacity from grid to cathode is greater than the internal capacity from anode to cathode.

Theoretically the shortening bar has zero potential at its electrical centre. It is therefore convenient to connect the H.T. + to this point. Energy in the anode circuit is coupled to the feeder by the circuit marked TS. This acts effectively as the secondary of a transformer. The two-wire transmission line to the feeder is connected to the coupling line at the point at which the impedance of the coupling line is equal to the impedance of the transmission line.

The synchronizing pulse for the indicator circuits—to synchronize the start of the time-base generator and the brightening of the trace to the start of the transmitter—is taken across R. The by-pass and filter circuits in the filament leads are mainly for the purpose of reducing the effects of negative feedback and the loss of output power that would result from this undesired effect.

Pulse Control

C1 and R1, the values of which control the pulsing action, are shown in Fig. 43 and marked accordingly. The duration of the pulse is determined by the capacity of C1, i.e., by the time required to charge it. This time is affected by any factor that affects the magnitude of the grid voltage such, for example, as the tuning of the primary to the secondary, the value of the H.T. voltage, the tuning of C2 to the coupling line, the position of the transmission line taps and the degree of coupling.

When the transmitter is tuned for maximum power output, therefore, the pulse duration is controlled by the capacity of C1. The pulse-repetition rate can be varied by most of the adjustments that affect the pulse duration. Changing any one of these values causes C1 to be charged to a slightly different voltage, and it therefore takes a different time for C1 to discharge to the voltage at which the valve can again begin to oscillate. The practical method of altering the repetition rate is to change the value of R1, which is usually made variable, in steps, for this purpose.

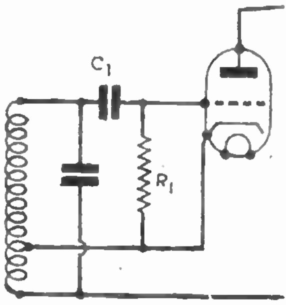


Fig. 42.—Hartley Type Oscillator.

This circuit can be arranged as a self-pulsing oscillator by adjusting the values of C_1 and R_1.

Magnetron R.F. Generators

Fig. 44 shows the input and output circuits for a magnetron R.F. generator. Fig. 44 (*a*) is without pulse transformer. Fig. 44 (*b*) is with pulse transformer. The functions of the pulse transformer are: (1) to match the output conditions of the pulse-forming circuit to the input of the magnetron; (2) to permit the magnetron and waveguide assembly to be located remote from the main operating position, i.e., close to the aerial; and (3) to enable a lower voltage to be generated in the primary circuit, since it can be stepped up to the required value before application to the magnetron.

The output from the magnetron is via the small looped probe which feeds directly into the waveguide feeder system. Matching is performed by the piston marked PN in the sketch.

The magnetron used in modern radar transmitters is of the resonant type and is a development of the split-anode magnetron.

RECEIVERS

Radar receivers are generally of the superheterodyne type. The essential requirements are that they should be sensitive (minimum sensitivity being of the order of 1 μV per metre at least), the signal-to-noise ratio should be high, the band width of the I.F. stage should cover the video frequency range, and all circuits must be designed for minimum distortion.

When automatic gain control is applied to a radar receiver, it is

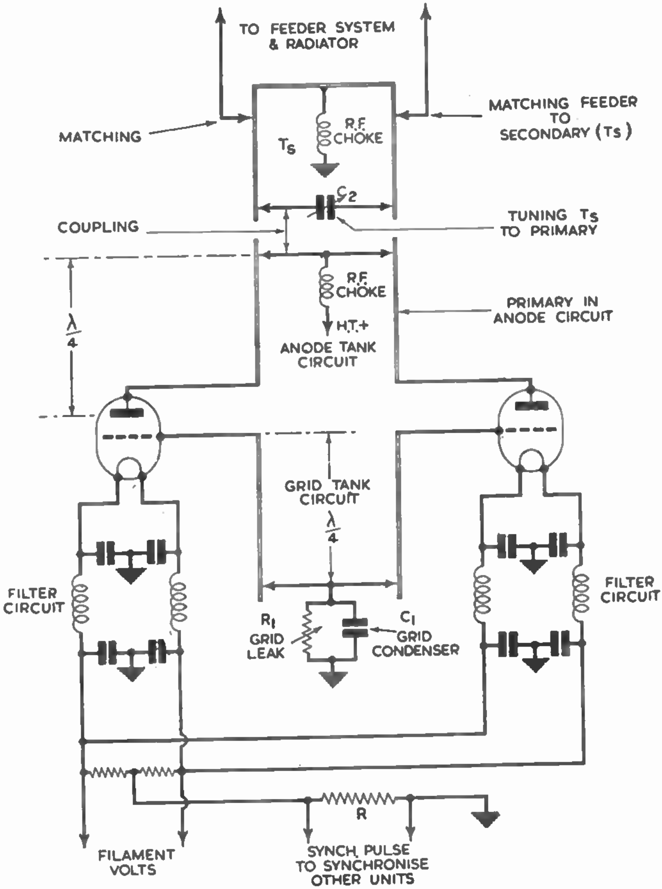


Fig. 43.—Push-Pull R.F. Generator Circuit for V.H.F. Pulsed Transmission.

Tuned grid, tuned anode, push-pull, self-pulsing R.F. generator for use at about 100–200 Mc/s.

essential to be able to apply it to one particular selected signal and not to any others that may be present. This is necessitated by the fact that fading of signals from one direction is generally different from that which may be experienced from others. In order to comply with this requirement, arrangements are provided whereby the operator can move a control to select and track any given echo. Manipulation of the control causes automatic gain control to be applied to the particular echo which is being tracked.

Automatic frequency control is also essential, particularly on precision measuring sets, in which waves of the order of 10 cm. and below are employed.

Any small change in frequency of either the local oscillator or of the transmitter frequency affects the beat frequency and hence the voltage developed across the I.F. input circuit. Signals passed to the I.F. circuit are generally very small in magnitude at ultra-high frequency, and therefore any reduction on the beat signal applied to the I.F input cannot be tolerated. In order to avoid such losses, arrangements are made to maintain the beat frequency constant by automatically adjusting the frequency of the local oscillator, to compensate for any change in beat frequency due either to a change of frequency of the local oscillator itself or to a change in the output from the transmitter.

Essential Requirements

The requirements of a high degree of sensitivity and high signal-to-noise ratio are easily understandable. If the received echo is weak, or if the signal-to-noise ratio is abnormally low, the echo becomes indistinguishable, with any degree of accuracy, from irregularities on the trace caused by receiver noise (grass). It is also obvious that the ability of a receiver to make use of the minimum possible signal determines the extreme range of useful echo signals for any given transmitting power. Thus, it is evident that the efficiency of the entire system depends to a greater degree upon the sensitivity and efficiency of the receiver than upon the power output from the transmitter. In other words, it is easier and more efficient to obtain useful range by improving the performance of the receiver, than to increase the power of the transmitter.

Since the R.F. envelope of energy radiated is rectangular in shape, it follows that the echo which returns takes the same form. It is also clear that a well-defined leading edge is desirable in order that accurate measurement along the trace to the leading edge of the transmitted pulse may be made. The analysis of a square wave shows that it is composed of an infinite number of sine waves of different frequencies and phases, wherefore the frequency band required to pass a square wave undistorted is, in theory, infinite.

In practice, however, it is possible to tolerate conditions which are ess severe, and it is generally accepted that a band-width of from 1 c/s to approximately 2 Mc/s is sufficient to pass a reasonably square wave. This corresponds roughly to the requirements of television. In consequence of this the I.F. stage must be designed to pass a suitable band of frequencies within these limits. A complication arises, however, because the signal-to-noise ratio of a receiver falls as the band width is increased, and rises as the band-width becomes narrower. The result of this is that the best possible compromise must be effected, having regard to the application for which the particular set is designed. For example,

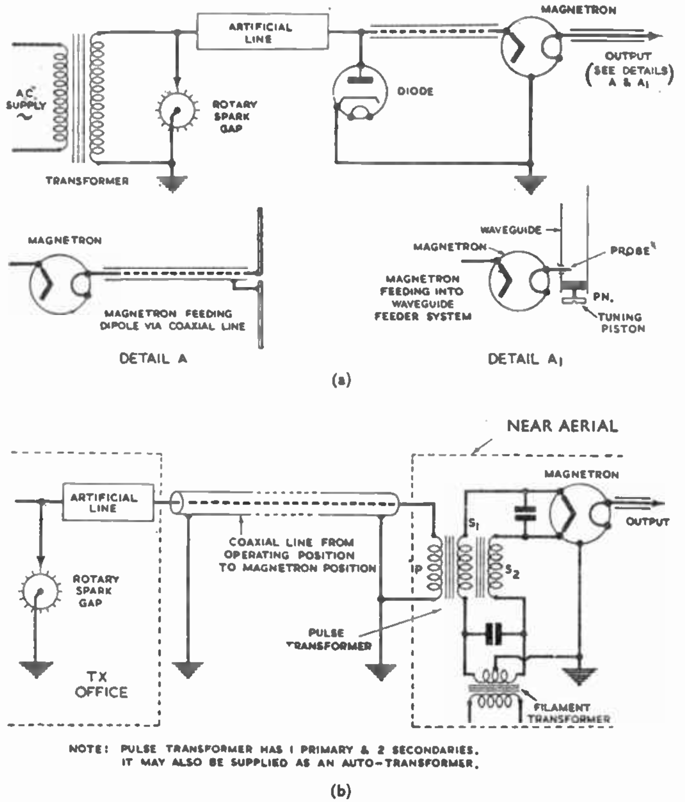


FIG. 44.—INPUT AND OUTPUT CIRCUITS FOR MAGNETRON R.F. GENERATOR.
(*a*) Without pulse transformer.
(*b*) With pulse transformer.

since accuracy of measurement is not the most important requirement for long-range warning sets, a narrower band width can be tolerated in order to increase the signal-to-noise ratio, and so to promote the efficiency of the receiver. When very precise measurement is required, the band-width must be such that an undistorted pulse envelope is passed. In this case the signal-to-noise ratio of the receiver falls, and therefore the need arises for a stronger echo to enable it to be clearly distinguished from the noise (grass). Grass in radar terminology is applied to the fringing of the trace produced by noise voltages in the receiver.

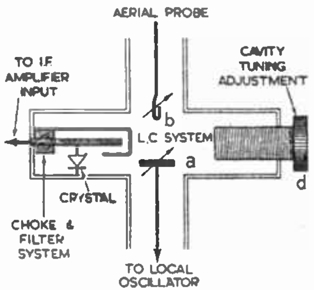


FIG. 45(*a*)(*above*).—CRYSTAL MIXER IN A RESONANT CAVITY.

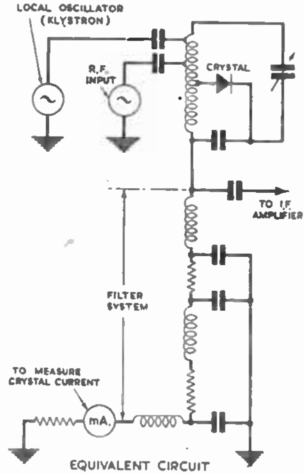


FIG. 45 (*b*) (*right*).—EQUIVALENT CIRCUIT FOR CRYSTAL MIXER IN A RESONANT CAVITY.

From that which has already been said about square waves, it follows that all amplifier circuits in the receiver must be designed to pass a square wave with minimum distortion. The output from the receiver is at video frequency.

In cases where it is necessary to render the receiver inoperative during the transmission period, in order that it may be in a condition to receive echoes from nearby signals immediately after transmission has ceased, a pulse may be provided, either to increase the negative bias on —say—the first two valves of the intermediate stage during the transmission period, or the H.T. may be removed from the anodes of these valves during the pulse time of the transmitter

When the signal-to-noise ratio falls below a ratio of 1 : 1, as it tends to do with increase of frequency, R.F. amplification is abandoned altogether, since it only tends to aggravate the trouble. At the higher frequencies, therefore, the I.F. is obtained by causing the output from a klystron, used as a local oscillator, to beat with the received R.F. Mixing is performed by a crystal operating in a resonant cavity. This arrangement is shown in Fig. 45, where it will be noted that the equivalent circuit shows a resonant cavity acting as a parallel resonant-tuned circuit.

The frequency of the klystron can be controlled, within narrow limits, by the voltage on its repeller electrode, and this is the method adopted when automatic frequency control is applied.

RADAR NAVIGATION

Since the impression made by a P.P.I. picture differs considerably from that made by a conventional navigation chart, it is obvious that special radar maps and charts are of great assistance to the navigator. Admiralty charts for use on radar-equipped ships are available.

Correction for Ship's Motion

In a normal P.P.I. display the ship's position is represented by the centre of the screen, with the top of the picture representing either the direction in which the ship is travelling or, if gyro-compass stabilization is used, a fixed position, which is usually north. Movement of other ships and "fixed" objects are apparent from the trails behind the spots produced on the screen by the various echoes, as a result of afterglow. These trails can give an idea of the speed and direction in which the objects are moving. The movements are, however, all relative to the ship, and true directions and speeds are not immediately apparent; fixed objects will also appear to have an apparent motion directly towards or away from the observer.

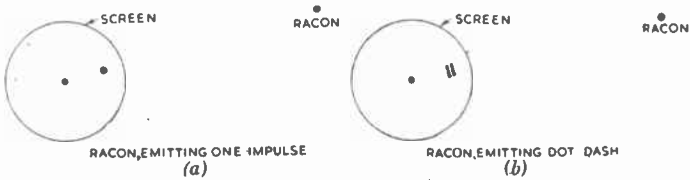
A navigator may often be able to appreciate the general situation more quickly and easily if presented with a display in which fixed objects are shown as such, with moving objects marked temporarily, their position being altered from time to time. Such displays are known under such names as "True Motion", "Track Indication" and "Chart Plan" presentation.

This can be achieved on the radar screen by arranging for the time-base sweep on the P.P.I. (having a north-upwards stabilized picture) to start from some position other than the centre, and arranging for this starting-point to be automatically moved across the screen to correspond with the actual motion of the ship. To carry this out in practice it is necessary to have information on the speed and course of the ship. The former can be obtained from the electric logs commonly fitted on ships, the latter from a gyro compass.

In due course the starting point will run off the screen and must then be re-set to the original starting-point. This may be done automatically or by means of a manually operated switch or push button. If the information fed in represents the ship's speed, as distinct from distance run, some form of electronic integrating circuit must be used to convert speed to distance.

Radar Beacons

Radar beacons, also called responder beacons, transponders and racons, consist essentially of a receiver, which picks up pulses from a radar transmitter, and a transmitter, triggered by the output of this receiver, which emits signals which can be detected by the radar receiver. Because the power transmitted by the beacon is independent of the strength of the radar signal which triggered it, beacon signals may be seen by radar receivers at much greater distances than ordinary echoes Since the beacon gives bearing and range, only one is necessary to determine position. The beacon can also be made to transmit a characteristic signal, for instance "dot-dash". The distinctive group of signals then seen on the screen cannot be mistaken for an echo from a normal target; the first of the series of pulses reaches the receiver at very nearly the same time as would a normal echo and shows the relative positions of beacon and receiver, while the other pulses identify the beacon. See Fig. 46.



(a) (b)

Fig. 46.—Picture on the Screen of the Transmission from Responder Beacons (Racons).

One disadvantage of responder beacons is that if two radar sets are interrogating the same beacon and both aerials are directed at it, each operator will see the beacon's replies to the other.

Experiments are being conducted with beacons which transmit either continuously or according to a certain time-code, but in each case independent of interrogating signals.

For most applications in the microwave region, the frequency of the beacon transmitter is not the same as that of the radar transmitter. If signals from these beacons are to be received, the receiver must be switched to that frequency with a special control, or a second receiver must be used. This system is preferred for aircraft, but for marine use it is usually more useful if the P.P.I. picture and the beacon are visible at the same time. If a radar beacon is to transmit in such a way that every 3-cm. ship's radar can receive its pulses without switching to a special frequency, it will be necessary for the beacon to transmit on a very broad frequency band. Such beacons are being constructed, and may be used more and more in the future.

In marine applications beacons are an important aid to navigation. For inshore navigation and pilotage especially, the advantages of obtaining accurate fixes from shore points are obvious.

As aids to air navigation, beacons are installed both on the ground and on the aircraft themselves. Ground beacons can be used not only for fixing position but also for course indication.

Secondary Surveillance Radar

To overcome the limitations of primary-surveillance radar equipment for long-range air-traffic control, increasing attention is being given to secondary-surveillance radar systems, in which a transponder is carried in the aircraft and automatically retransmits the radar signals received by the aircraft. As the transponder frequency is normally different from the interrogation frequency, the response can be easily distinguished from primary-radar reflections, and the secondary-radar display will be free from ground clutter and cloud reflections. The main disadvantage with the system is that transponder equipment must be carried in the aircraft. By coding the reply from the transponder, positive identification of particular aircraft is possible. In for example the Cossor Type 1 Secondary Surveillance Radar equipment the control and interrogator frequencies are 1,215 Mc/s; control-transmitter peak power 5 kW; interrogator-transmitter peak power 1 kW; transponder-transmitter frequency 1,375 Mc/s; transponder peak power 200 watts. The airborne equipment weighs 30 lb., excluding cables.

" Sarah "

The " Sarah " (Search and Rescue and Homing) apparatus is intended to provide location signals for distressed persons by means of a small radio-beacon transmitter attached to the " Mae West " life jacket. The equipment carried by the operator comprises a radio-beacon transmitter with aerial, a speech unit and battery, with a total weight of 52 oz. The beacon equipment transmits a coded pulse generated in an optimized squegging oscillator with a repeating sequence controlled to provide groups of pulses at a suitable low pulse-repetition frequency. The characterized pulse spacing in each pulse group provides a type of display on the cathode-ray tube of the rescue receiver which permits ready identification of different beacons. By time-sharing methods, a right and left aerial on the search aircraft are arranged to display beacon spikes either to the right or left, respectively, of the vertical reference trace.

RADAR SONDE

Radar Sonde, which has been developed by the Telecommunications Research Establishment working on behalf of the Meteorological Office and by Mullard, Ltd., represents a marked departure from previous radio-sonde methods. Earlier methods, which involved the use of an airborne transmitter continuously radiating and the use of a radar reflector for wind velocity and range measurement, employed several different methods for the transmission of information, including : variation of carrier frequency, variation of audio modulation, chronometric coding and Morse coding.

The new system is designed for the measurement of wind speed, wind direction, height, temperature, pressure and humidity; it is fully automatic, and slant ranges in excess of 100 nautical miles are obtainable, allowing accurate measurements of strong winds at great heights. Pressure, temperature and humidity are telemetered to an accuracy of 0·1 per cent, though the meteorological measuring elements may not be capable of accuracy of this order.

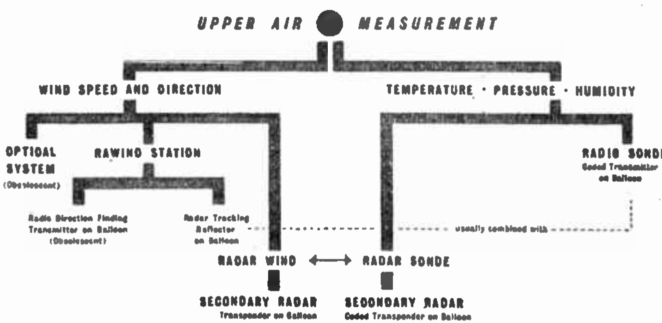


Fig. 47.—Methods of Upper Air Measurement.

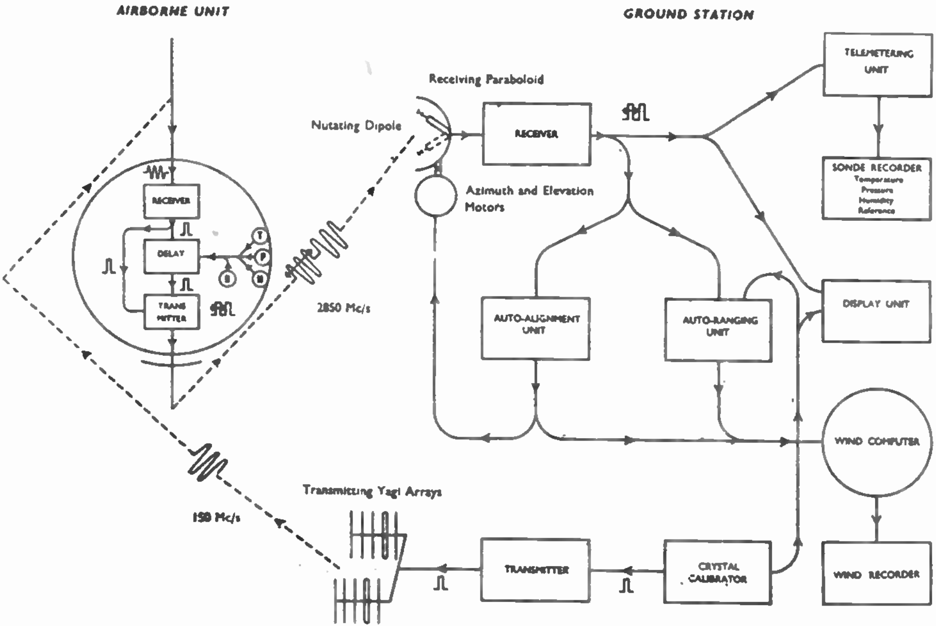


FIG. 48.—SCHEMATIC DIAGRAM OF COMPLETE RADAR WIND AND RADAR SONDE THEODOLITE.
(*Mullard Ltd.*)

Each equipment comprises a fully automatic ground radar interrogator and computer station, which operates in conjunction with airborne transponder units. These units are suspended from balloons capable of ascending to a height of 100,000 ft. The balloon carries a container which holds temperature-, pressure- and humidity-sensitive elements. These are connected to a transmitter so that, when the receiver in the airborne equipment is interrogated, a coded signal is transmitted containing the meteorological information. For radar wind operation only, there are no meteorological units, and the airborne transmitter, upon interrogation, gives off plain answering pulses so that the time interval between sent and received pulses on the ground is a measure of distance. The send and receive frequencies are different, and are selected to suit the differing conditions of operation between the ground and airborne systems. This secondary radar system allows greater ranges to be obtained than could be economically achieved with primary radar. Facilities of this equipment include: (*a*) auto-follow of balloon in direction and range; (*b*) auto-computation of wind speed and direction; (*c*) auto-computation of temperature, pressure and humidity; (*d*) auto-recording of all results.

Operation

The balloon unit is continuously interrogated from the ground station. In the case of radar wind airborne units, the interrogating pulse is received and fed to a modulator which triggers the airborne transmitter. The time interval between interrogating and responder pulses is a measure of balloon range, and the rate of change of this interval is a measure of wind speed.

In the case of radar-sonde airborne units the received signal is fed to a modulator stage, both directly and via a variable delay stage. Thus a single interrogation pulse initiates two responder pulses having a variable time displacement. The delay time is determined by the three meteorological elements, which provide telemetering signals representing temperature, pressure and humidity measurements. A fourth, calibration, element is also provided.

The airborne signals, having been picked up by the ground station, are passed to the receiver and the resultant video signal fed to four units. These are:

(1) Auto alignment unit, which corrects any misalignment of the receiving aerial in elevation or azimuth.

(2) Auto ranging unit, which computes the range of the airborne units (a crystal calibrator providing reference information).

(3) Display unit, which presents the responder pulses on a cathode-ray tube, superimposed upon a calibrated scale. A double trace is provided, giving coarse- and fine-range measurement. The crystal calibrator provides the calibration markers.

(4) Telemetering unit, which computes the meteorological information, which is then recorded by four pens, and the teleprinter.

The outputs from the auto-alignment and auto-ranging units are also fed to the wind computer, which derives the instantaneous wind direction, and to the wind recorder, housing the pen recorders, which present wind-speed and true-height records.

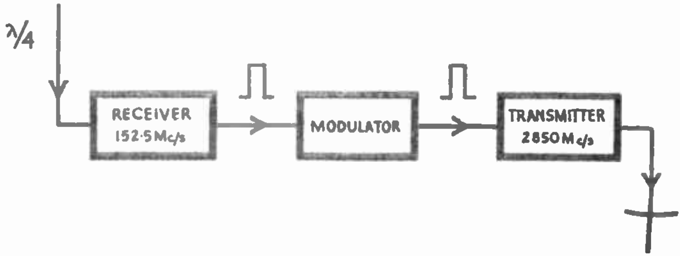


Fig. 49.—Block Schematic of the Airborne Radar Wind Unit.
(*Mullard Ltd.*)

Ground Station

The transmitter radiates through a pair of Yagi aerial arrays with a power gain of eight. These aerials are vertically polarized and mounted on either side of the receiving paraboloid reflector. The transmitter consists of a blocking oscillator, buffer amplifier, modulator and a self-oscillating output stage. The output stage, which operates on a frequency of 152·5 Mc/s, is a disc-seal triode with co-axial cavities, and is pulse modulated to produce 2-microsecond pulses at a nominal peak power of 50 kW. It is triggered by the crystal oscillator. A wave-meter and monitoring oscilloscope are built into the equipment.

The crystal oscillator provides a crystal-controlled pulse at a repetition frequency of 404 c/s for triggering the transmitter; this is obtained by frequency division from an 808-kc/s crystal oscillator. It also provides calibration markers for the range display cathode-ray tube and for computing circuits. At a frequency of 808 kc/s, one cycle corresponds to a target range of 0·1 nautical mile.

The microwave signals from the airborne unit are received by a 5-ft.-diameter paraboloid and nutating dipole, which produce a conical

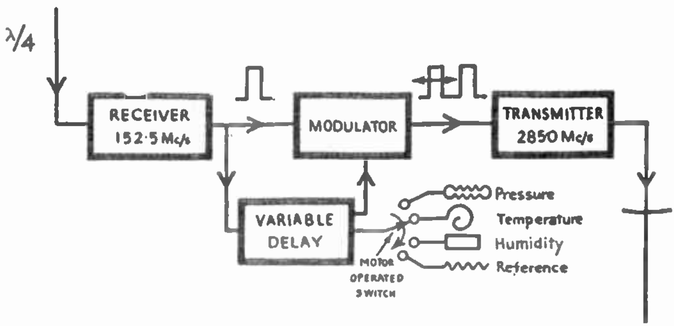


Fig. 50.—Block Schematic of the Airborne Radar Sonde Unit.
(*Mullard Ltd.*)

scan similar to that used in anti-aircraft fire-control radar. Any misalignment of the aerial causes amplitude modulation of the incoming signal at the coning frequency of 11 c/s.

The phase and amplitude of the modulation define the direction and degree of misalignment, and are the error signals in a servo system which corrects the alignment of the aerial. The aerial driving motors operate over a speed range of 1–10,000 r.p.m.

Tacho-generators coupled to the aerial motors provide bearing and elevation rate-of-change signals for the wind computer.

The microwave receiver incorporates a co-axial crystal mixer and a 13-Mc/s intermediate-frequency amplifier. Automatic gain control and automatic frequency control are provided, both being protected from interference by means of strobe gates. The frequency of the reflex klystron local oscillator is motor-controlled, and provision is made for automatic frequency searching and automatic locking on the correct channel. When the received signal-to-noise ratio falls below a predetermined level, indicating the end of the flight, a warning lamp glows.

Airborne Units

The 152·4-Mc/s interrogating pulse is received on a quarter-wavelength aerial, by which the airborne unit is suspended from the carrying balloon. It is fed through a three-stage receiver, employing subminiature valves, which provides an overall sensitivity of 2·5 mV. The video output pulses are fed to a modulator stage which triggers a self-oscillating transmitter working on a frequency of 2,850 Mc/s. This transmitter provides a peak power of at least 30 watts, and uses a small disc-seal triode in a re-entrant cavity, to which is coupled a quarter-wavelength aerial fitted with a counterpoise. The apparatus is maintained at a constant temperature.

The radar sonde unit incorporates in addition to the equipment above (fitted in the radar wind unit), the three meteorological units for the measurement of pressure, temperature and humidity. These incorporate an aneroid capsule, a resistance thermometer and a piece of gold-beater's skin element, respectively.

The mechanical movements of the pressure and humidity elements are converted to voltage changes which, with the resistance thermometer voltage, are sampled consecutively by a motor-driven switch connected to a phantastron delay circuit. The phantastron is triggered by the received pulse, and produces a second pulse with a time delay of between 0 and 1,000 microseconds, which is accurately controlled by the meteorological element. The modulator is triggered by the second pulse, so that a pair of pulses with a variable spacing of up to 1,000 microseconds is radiated back to the ground station.

The pressure element is enclosed in the thermally insulated container. The stabilization of temperature in the telemetering circuit has enabled almost a tenfold improvement in accuracy to be obtained.

All power supplies are derived from 6·5-volt, 2·5-ampere secondary batteries. The H.T. supplies are obtained from a vibrator producing 100 volts at 25 mA and 800 volts at 1 mA.

19. AERONAUTICAL RADIO AND RADAR EQUIPMENT

19. AERONAUTICAL RADIO AND RADAR EQUIPMENT

DESIGN REQUIREMENTS

As far as mechanical design is concerned, airborne equipment must be as compact and light as possible within the limits of the overriding requirement of mechanical strength and electrical reliability.

All airborne electronic equipment designed for civil or military use in the major countries of the world must conform to certain design standards. These standards are laid down by the relevant military authorities or by the body responsible in the country concerned for the licensing of civil aircraft. (Before an aircraft is granted a Certificate of Airworthiness, without which it is not allowed to fly, it and its equipment must conform to certain standards of design and serviceability laid down by the licensing authority. Two such bodies are the Air Registration Board in the United Kingdom and the Civil Aeronautics Administration in the U.S.A.)

It should, therefore, be assumed when designing an airborne radio or electronic equipment that it will be required to conform to the standards laid down by the licensing authorities in the countries where it is expected to be sold. These standards are freely available from the bodies concerned, and as they are subject to periodical modification, they will not be dealt with in detail here. In the United Kingdom the relevant section of " British Civil Airworthiness Requirements " should be consulted for complete information on the standards required. Another useful guide is British Standard G.100 issued by the British Standards Institution. In the case of certain types of equipment whose performance characteristics are the subject of international agreement through I.C.A.O. (International Civil Aviation Organization), certain standards are also published by that body, and should be consulted (for example, such equipments as V.H.F. omnidirectional range receivers, instrument landing system receivers, etc.). In the case of equipment intended for sale in the U.S.A. or for fitting to American-built aircraft, ARINC specifications are also relevant. They are, in fact, worth consulting in any case for the sake of valuable information they contain and because there is developing a wider general tendency for airlines and aircraft manufacturers outside the U.S.A. to specify ARINC standards for radio installations.

ARINC or Aeronautical Radio Inc. is located in Washington D.C., U.S.A. and is a body of which the U.S. airlines and also now many foreign (including British) organizations are stockholders. It exists primarily for the purpose of operating aeronautical radio facilities in the U.S.A. (and some other countries). The Airlines Electronic Engineering Committee (A.E.E.C.) of ARINC produces and publishes standards for many types of airborne radio and electronic equipment.

In general, the standards laid down have a threefold purpose :

(*a*) To ensure that the equipment does not endanger the safety of the aircraft by breaking loose from its mountings, catching fire, etc.

(*b*) To ensure that it will give adequate service under operating conditions. That is, that under certain conditions of vibration, temperature change, pressure change, supply voltage change and humidity, it will not fail mechanically, nor will its electrical characteristics change to such an extent that it will not operate adequately.

(c) To ensure that the systems employed are such that when operated against any relevant ground facility conforming to international standards, it will provide adequate communication, correct navigational information, etc.

Before a new equipment can be considered for use in any civil aircraft, it must be submitted to, and approved by, the licensing authority. Once this approval is obtained, the equipment can be freely used, but should its design be modified, further approval must be sought. The number of the approval certificate should always be displayed on every production equipment.

The fitting of an unapproved equipment in an aircraft will result in permission to fly that aircraft being withdrawn until the equipment is removed.

The requirements for military applications are in general more severe than for civil use.

GENERAL CONSTRUCTION

There is no compulsion to design airborne radio equipment to any particular shape, but if it forms part of the radio installation of British or American civil aircraft, it is unlikely to be acceptable unless it conforms to the standards in use in these countries which are the E.E.A./S.B.A.C. (Electronic Engineering Association, Society of British Aircraft Constructors) and A.T.R. systems respectively.

Both these systems are based on a series of standard case sizes and they are not compatible. They both involve one standard box height coupled to a number of different depths and widths. (ARINC have recently introduced a new case half the normal height to enable very small units to be mounted one above the other.) The British system contains a greater variety of widths than the other as this dimension varies in steps of 1 in. It also includes a racking system made on a modular basis from standard parts. In both systems automatic connection to external wiring is made by back-mounted plugs on the units and there is a special quick release mounting system by means of spigots at the rear and thumb screws at the front. Neither the plugs nor the holding-down systems are compatible.

It is becoming apparent that the British system is likely to be superseded by the American even for British use. A.T.R. system details are published by ARINC in their Specification No. 404.

POWER SUPPLIES

The types of power supply most likely to be met with in aircraft are: (*a*) nominal 14-volt D.C. (now rare); (*b*) nominal 28-volt D.C.; and (*c*) 115-volt, 400 c/s, 3-phase (4 wire).

System (*c*) is the preferred one for all new aircraft, and any D.C. service will be 28 volts rectified from the main supply. Any D.C. aircraft (except perhaps very small ones) are most likely to be 28 volts. In general it is advisable to make provision for optional operation of equipment from D.C. or A.C.

Most D.C. and A.C. supplies are regulated, but equipment should be designed to tolerate substantial variation in voltage (and frequency in the case of A.C.). A nominal 28-volt supply can vary between 18 and 29 volts under certain conditions. Variation of 102–121 volts and 210–420 c/s can occur on A.C. supplies, although regulated supplies are specified to be held to 108–121 volts and 380–420 c/s.

In general, if the main supply in an aircraft is A.C., it will not be permissible to connect a substantial radio load to the D.C. system. Equipment connected to the A.C. supply will be required to conform to certain limits on power factor and phase unbalance. In the case of D.C. supplies one pole is earthed; although there is now almost universal agreement on which pole this should be, it is safer to design D.C. equipment so that it will operate with either side of the supply earthed.

Regulations for switching and fusing aircraft equipment are very stringent, and the designer should make sure that in this and all other respects he has the latest information from all relevant authorities. If the equipment is intended for a known aircraft, the makers should be consulted.

Transistor power converters, which are now becoming fashionable, need special care in design when operated from aircraft supplies, as these can be subject to transients of considerable amplitude arising from the switching of heavy inductive loads.

Effect on Magnetic Compasses

Magnets or magnetic materials in the equipment can cause permanent deviation of magnetic compasses and the flow or interruption of heavy direct currents can also cause deviations which will vary according to whether the gear is operating or not. The former effect can be compensated at the compass, but the latter cannot : one of the tests carried out by the licensing authority during approval of an equipment determines the least distance from the compass at which the gear can be mounted without producing unacceptable deviations under any condition of operation. As the magnetic compass is mounted somewhere just in front of the pilot and the radio station not far behind him, it is quite easy (unless care is taken to reduce stray magnetic fields) to design an equipment which cannot be mounted in the radio station and is, therefore, unusable. One particularly important point to watch is in connection with remote controls for use by the pilot. The on/off switch in such a remote control should not generally be made to switch the D.C. supply for the equipment direct, but should operate through a relay in the main equipment to keep the heavy D.C. out of the neighbourhood of the compass.

Mechanical and Electrical Design

In designing an airborne equipment the main factors to bear in mind are :

(1) Adequate rigidity of chassis and other structural parts to prevent failure through shock and vibration. In this connection it is essential that no part of the assembly shall resonate over a wide range of vibration frequencies, as this will cause rapid failure of apparently strong assemblies.

(2) Rigid fixing of all components—not even the smallest components should be hung in the wiring.

(3) Adequate temperature compensation of tuned circuits, etc.,

to ensure that performance is maintained over a wide range of temperatures.

(4) The effects of reduced pressure on flash-over or corona effects in components operating at high voltages. In general, it can be assumed that civil equipment will not be required to operate at a pressure lower than that equivalent to 10,000 ft. above sea-level.

(5) The effects of humidity—all insulating materials should be moisture resistant to a high degree, and the greatest care must be taken to avoid moisture traps in tag boards, intermediate-frequency transformer assemblies, etc.

The latest issue of DEF.5000, published by the Ministry of Supply, should be consulted.

COMMUNICATION EQUIPMENT

Communication with aircraft is carried out by amplitude-modulated telephony on very high frequencies (V.H.F.) between 118 and 132 Mc/s, and by telephony and telegraphy on high-frequencies (H.F.) between approximately 2 and 20 Mc/s. Arising from the desire to economize in the number of crew carried, the use of telegraphy—with its need for a special operator— is rapidly disappearing, and eventually only telephony may be used for both long- and short-distance communication.

Telephony sets are designed for pilot operation, and their controls are consequently as simple as possible. V.H.F. sets are crystal-controlled, either by a number of crystals cut for particular channel frequencies or by crystal-saving circuits which enable the equipment to be set to any one of the channels covering the range 118–132 Mc/s in steps of 200 or 100 kc/s (eventually these steps will probably be reduced to 50 kc/s). Channel selection is by push-button or rotary switch. The automatic-gain-control requirements for V.H.F. are stringent: without any manual adjustment during operation the equipment must work at distances from the ground station between 100 or so miles and a few hundred yards without over-loading or excessive change in volume. Transmitter power on V.H.F. ranges from a fraction of a watt on light-plane equipments to several watts on main-line aircraft. American equipments use powers up to 30 or more watts, but this is often the cause of long-range interference, and negatives one of the main advantages of V.H.F., whose limited range—compared with high frequency—enables the better segregation of local traffic. V.H.F. communications signals are vertically polarized.

V.H.F. is employed for communication with air-traffic control and airfield control authorities when the aircraft is flying in areas of high traffic density under the control of such authorities. For general and long-distance communication, high frequency is used, and the equipments employed may either be pilot or radio-operator controlled. In the case of pilot-controlled equipments these are again preset to the channel frequencies required for the route to be flown, and channels are selected by push button or switch. In the case of operator-controlled equipments, these are usually crystal-controlled transmitters and continuously tunable receivers. An aircraft flying from, say, London to Singapore may require to operate on as many as 50 high-frequency channels during the journey, and this presents quite a problem in the design of an equipment of the pre-set type. Transmitter powers for long-range high-frequency telephony range up to 150 watts. At present, limiting factors on the power (which could well be higher from the point of view of good communication) are the capacity of the power supply (for supplying equipments of over 100 watts the power unit becomes very large and heavy) and the danger, when the transmitter is fully modulated at high altitudes, of aerial flash-over.

The use of radio teleprinters in the air is also being actively considered. This system has particular application to weather forecasts and other messages which may require to be recorded in the aircraft for future use. At present, such messages have to be passed at dictation speed and taken down by a member of the crew.

H.F. communication at long distance on the aeronautical bands leaves a great deal to be desired; the difficulty of exceeding about 100 watts with airborne A.M. transmitters has led to the development of single-sideband (A3a) equipment which offers a substantial effective power gain (see Section ·7). An example of the latest technique is the Collins 618T H.F. transceiver. This equipment can be set at 1-kc/s intervals throughout the range 2–30 Mc/s (i.e., 28,000 channels), in the air without pre-setting. It has an A.M. output power of 100 watts, but operation is also possible on a single-sideband suppressed carrier with 400 watts peak envelope power. To meet the stability requirements of S.S.B., the frequency error of the receiver is held to less than 25 c/s, and the long-term stability is 1 in 10^6 (plus 2·5 c/s) *per month.*

Selcall

Because of the high noise level and severe overcrowding on the H.F. aeronautical bands, pilots are subjected to a great deal of distraction both from noise and unwanted signals if they keep a continuous listening watch (which is essential).

To combat this problem, a system of selective calling named " Selcall " is now in fairly substantial use. Its characteristics are the subject of I.C.A.O. standards, and there is a relevant ARINC specification.

The system involves the radiation by the ground station of a signal consisting of two pairs of audio tones. These are picked up by the aircraft receiver, to the audio output of which is attached a unit having filters responding only to the combination of four tones unique to the aircraft in which they are carried. When the correct tones are received, the Selcall unit actuates a warning device which informs the pilot that he should listen to the radio.

The Selcall facility can be applied to any air or ground equipment. The four tones are selected from twelve located between about 300 and 1,000 c/s. The tones are radiated two at a time in pulses of 1 second duration ($\pm$0·25 second), and the pulses are separated by a gap of 0·2 seconds ($\pm$0·1 second). The frequency of the tones is held within $\pm$0·15 per cent of nominal. The most common types of receiving filters are electro-mechanical vibrating reeds. The system will work with signal-to-noise ratios as bad as 6 dB adverse and distortion of the tones up to 15 per cent.

NAVIGATIONAL AIDS (SHORT AND MEDIUM DISTANCE)

Radio-range Systems

The object of the " radio range " is to provide the pilot of an aircraft with two, four or more radio beams radiating from a ground station.

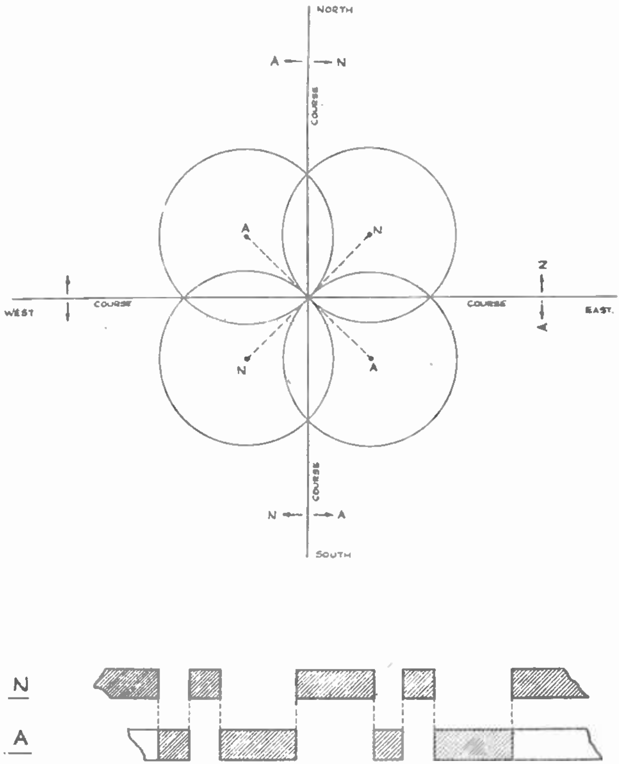


FIG. 1.—MEDIUM-FREQUENCY FOUR-COURSE RANGE: POLAR DIAGRAM AND INTERLOCKING MORSE CHARACTERS.

These beams, in the case of two- or four-course ranges, are so aligned that in following them the pilot is following some recognized air route or "airway". It is particularly important in areas of high traffic density that an easily followed navigational system of much higher precision than "dead reckoning" navigation shall be available.

Four Course Medium-frequency Range

An Adcock aerial system is used on the ground. The centre aerial is fed with a continuous-wave signal somewhere in the 200–400 kc/s range, and the two aerial pairs are fed with another signal whose frequency differs from the former by about 1,000 c/s. The signal to one pair is keyed "A" in Morse and that to the other "N". These characters are arranged to interlock so that if the A and N signals are

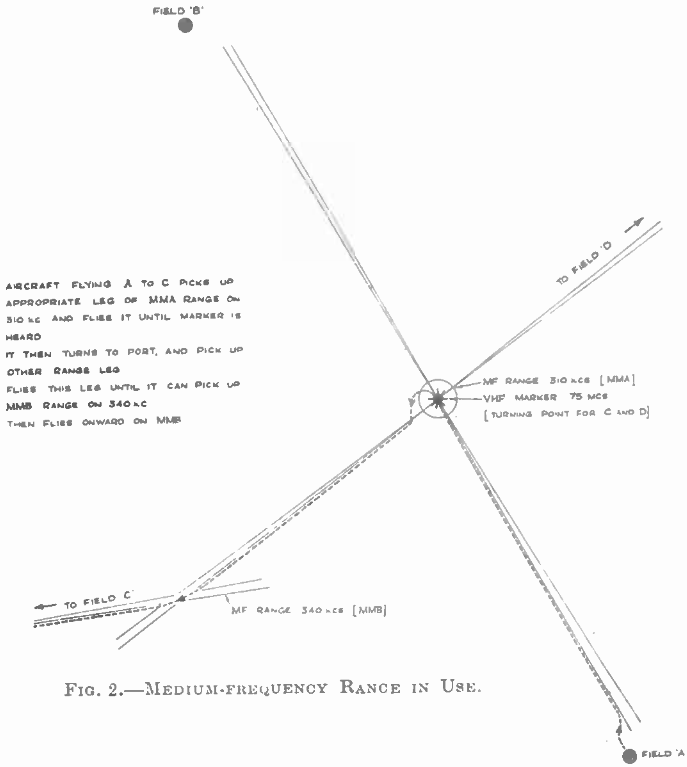


Fig. 2.—Medium-frequency Range in Use.

received at equal strength, a continuous signal will result. The radiation of two different radio frequencies ensures that a 1,000-c/s heterodyne note will be heard on any medium-frequency receiver without using a beat-frequency oscillator.

In a symmetrical Adcock system equal signals will be received along lines bisecting the angles between the planes of the aerial pairs. Thus on these lines a steady note will be heard. Deviation from the lines will produce either an A or N signal. In the simplest system therefore four courses will be set up, at 90 degree intervals. The angles between the courses can be altered by suitably phasing the inputs to the aerials, and the whole pattern can be orientated in any direction by suitable siting of the aerials.

Aircraft Equipment.—Only a radio telephony receiver tuning over the medium-frequency band is needed in the aircraft. On some range systems V.H.F. marker beacons are placed at various places to mark turning points, etc. For reception of these a V.H.F. Marker Receiver will be needed in addition. (See the later section on I.L.S : para. Marker Beacons.)

Ambiguity.—Consideration of Fig. 1 will show that an aircraft flying *north towards* the station will have " A " on the starboard side and " N " on the port. This is equally true of one flying *west away* from the station. Thus the information from a range station of this type is ambiguous. The ambiguity can be resolved by taking a D.F. bearing on the station with a radio compass, but in most cases the pilot knows which side of the station he is and can resolve the ambiguity mentally.

Status.—This type of range is obsolescent in the U.S.A., but is in current use in the United Kingdom and Europe.

Two-course V.H.F. Range (Visual–Aural)

This range represents an early attempt to obviate the ambiguity of the four-course range and to take advantage of the lower noise level on V.H.F. It is obsolete in the U.S.A. and never came into use in the United Kingdom. It required special equipment in the aircraft, where the A–N signal had to be read in conjunction with the position of a pointer on an instrument (similar to an I.L.S. localizer instrument). It will not be further described here.

Omnidirectional V.H.F. Range (V.O.R.)

This American system is now adopted by international agreement as the primary short-distance aid for flying in control zones. It is in use in the U.S.A., but only to a very limited extent in Europe as yet.

The system operates at the moment on V.H.F. frequencies between 112 and 118 Mc/s, and special equipment is required in the aircraft.

Principle of Operation

The station radiates a rotating field so that any receiver within range will receive a signal, amplitude modulated at the rate of rotation (about 30 c/s). At the same time an omnidirectional signal is radiated, amplitude modulated at about 10 kc/s. Thus the receiver will produce two audio signals, one of 30 c/s and the other of 10 kc/s. The 10-kc/s signal is further frequency modulated by a frequency exactly the same as that of the rotating pattern, and phase locked to it so that a receiver due north of the station would receive both 30-c/s signals exactly in phase. Obviously, as the receiver moves round the station, the phase difference between the rotating pattern signal and the reference signal will vary, and all that is necessary to determine the bearing of the aircraft from the station is to be able to measure this phase difference. In addition to the navigational signals, the station is also modulated at low level with Morse identification and with speech if required. These further signals can be heard on the receiver without interrupting the navigation facility, by filtering out the navigation signals in a separate audio-output channel. The V.O.R. signals are horizontally polarized and, therefore, require different aerials from the V.H.F. communication system in the aircraft.

Aircraft Equipment

A block diagram of a typical V.O.R. receiver is shown in Fig. 3. The operation can be followed if we assume that we require to place the

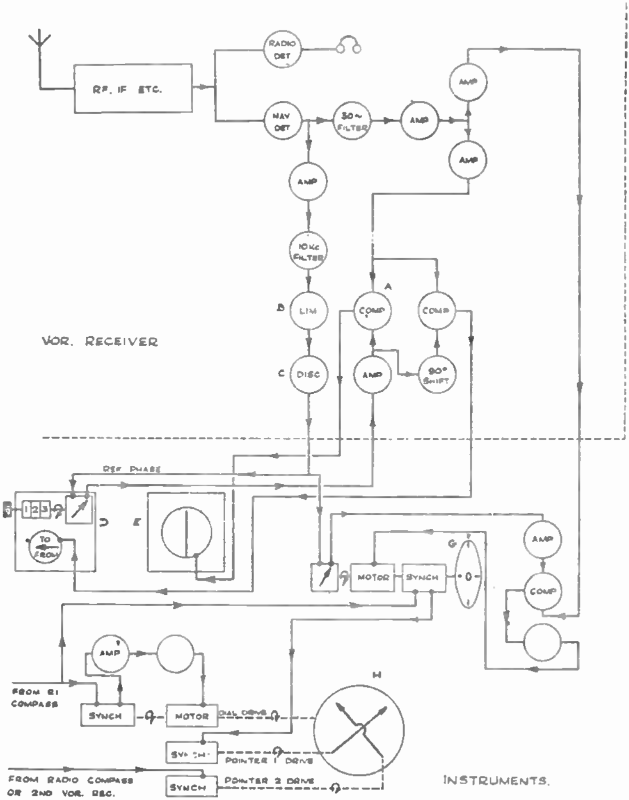


FIG. 3.—REPRESENTATIVE V.O.R. AIRCRAFT SYSTEM.

aircraft on a bearing of 045 degrees magnetic from the station and that the direction of rotation of the pattern is such that the variable phase signal lags 45 degrees on the reference phase along this line.

From the navigation detector of the receiver the 30-c/s signal of the variable phase and the 10-kc/s sub-carrier of the reference phase are passed through band-pass filters and the variable-phase signal is fed into the phase comparator " A ". The 30-c/s reference phase signal is resolved from the 10-kc/s signal via the limiter and discriminator " B " and " C " and passed to the omni-bearing selector " D ", where its phase can be varied through 360 degrees by rotating a dial marked in degrees. The phase-shifted signal is then fed on to phase comparator " A ". This phase comparator is designed to show balance when the two signals fed to it are in quadrature, as it is easier to give an accurate indication of this condition than of coincidence. It is thus necessary to so calibrate the selector that when its shows, for example, 045 degrees, its output is actually shifted 45 + 90 degrees. The output of the phase comparator " A " is rectified and fed to a centre-zero meter " E ", callea the " course deviation indicator ". When this is centred, it indicates that the omni-bearing selector is set to the same reading as the radial bearing from the station on which the aircraft happens to be.

This indication is, however, ambiguous by itself, as there are two conditions of the phase-shifted reference signal when it can be in quadrature with the variable phase. These conditions correspond to the true and reciprocal bearings, and must be resolved. This is simply achieved by shifting the reference phase by a further 90 degrees and comparing it in phase comparator " F " to see whether it is now in phase with the variable phase or in anti-phase. This comparison need not be accurate, as it has only to indicate one of two widely differing conditions. The output of phase comparator " F " actuates a further small meter movement which points either to " TO " or " FROM " as directed by the phase comparator. All that is necessary is to arrange the system of operation of all stations and receivers so that the various phase relation-

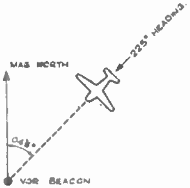
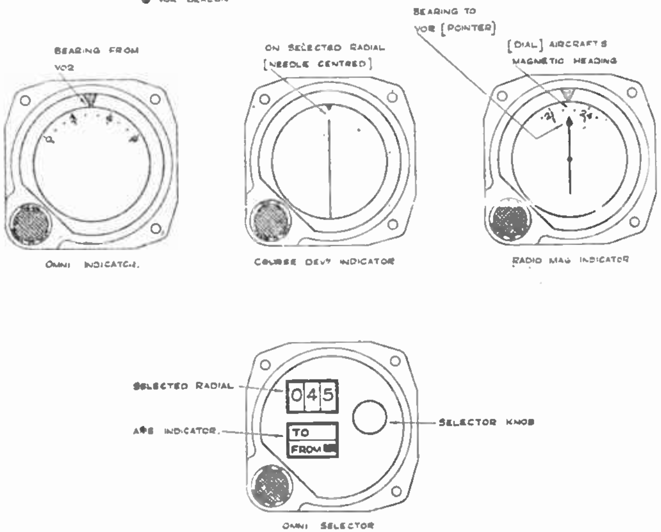


Fig. 4.—V.O.R. Instrument Readings for Aircraft in Position and Heading as Shown (*left*).



ships are such that when the aircraft is, for example, in a position whose bearing is 045 degrees FROM the station, setting the omni-bearing selector to 045 will centre the course-deviation indicator and at the same time make the ambiguity indicator point to FROM. If it is desired to determine the direction *to* a V.O.R. station the omni-selector is turned until the course deviation indicator is centred and the ambiguity indicator shows " TO ".

Further elaboration of the receiver will enable direct indications of bearing from range stations to be obtained without setting an omni-bearing selector. This is accomplished by using the signals produced by a phase comparator to rotate a phase shifter (similar to the omni-selector) via a servo system. The moving element of the shifter rotates until it takes uo a position corresponding to the bearing from the range station which is then indicated on a dial " G " attached to it. Ambiguity is avoided by making the setting corresponding to the reciprocal bearing an unstable condition of the servo-system.

The indications so far described have the advantage that they are not " heading-sensitive ", in other words the reading shown is the same irrespective of the direction in which the aircraft itself happens to be pointing. If heading sensitive indications are required, there is instrumentation possible which feeds the omni-bearing information into a further instrument " H ", where it moves a pointer over a rotating scale whose position is controlled by the aircraft's remote-indicating compass.

Accuracy and Range

The overall system accuracy of V.O.R. (including ground station and receiver errors) appears, on the evidence of operational experience so far, to be of the order of $\pm$2–4 degrees.

The range to be expected is optical line-of-sight plus 15 per cent at vertical angles not exceeding 40 degrees above the horizon as measured from the location of the ground-station aerials. For instance, an aircraft flying at 2,000 ft. can expect adequate signals up to a distance of 70–80 miles from a V.O.R. station.

Distance-measuring Equipment (D.M.E.)

In order to provide a system which will fix the position of an aircraft by reference to only one ground station at one time, it is proposed to associate with each V.O.R. station a responder beacon which, when interrogated by suitable equipment in the aircraft, will provide information on the aircraft's distance from the station. In this way a single V.O.R./D.M.E. station can provide bearing and distance of an aircraft relative to the station site, and this is, of course, sufficient to determine completely the aircraft's geographical position.

The principle of operation is simple, involving only the measurement of the interval between the transmission from the aircraft of an interrogating pulse and the reception of a reply from the beacon. In practice the technique is complicated by the need to provide a direct indication of distance on a meter, together with a method of selecting only the beacon required without danger of a false indication from any other beacon within range.

Distance-measuring equipment has not developed at the same rate as V.O.R. and still presents a number of problems not yet completely solved. For this reason the characteristics at present laid down will be indicated only as a guide to the type of system employed.

The intention is to provide 10 interrogation (air) frequencies spaced 2·5 Mc/s apart between 963·5 and 986 Mc/s. These are channels 1–10. Ten reply (ground) channels at the same spacing between 1188·5 and 1211 Mc/s are numbered 00–90. Combinations of interrogation and reply channels give 100 distinct operating channels from 1 (reply 00 interrogation 1) to 100 (reply 90 interrogation 10). D.M.E. signals are vertically polarized.

Further protection against spurious indications is provided by using double interrogation and reply pulses with 10 different separations. In a currently advocated system these double 2·5-microsecond pulses are spaced from 14 to 77 microseconds in multiples of 7 microseconds, and interrogation and response spacings are paired to give 10 modes from "A" (interrogation 14 reply 77) to "J" (interrogation 77 reply 14). Each D.M.E. channel is paired with a different mode, chosen so that adjacent channels differ in mode as much as possible. For example, *Channel* 1 reply 00, interrogation 1 mode "A" *Channel* 2 reply 00, interrogation 2, mode "D".

Means are also provided for the beacon to radiate at frequent intervals a three-letter identification signal. Several methods of achieving this are currently under investigation.

Accuracy and Range.—D.M.E. range is of the same order as V.O.R. Accuracies of $\pm\frac{1}{2}$ mile or 3 per cent, whichever is the greater, are to be expected with present equipment.

Presentation.—The method of presenting D.M.E. information is by a meter with a roughly logarithmic scale calibrated 0–200 miles. Identification is intended to be presented visually also.

Capacity.—The operational requirement for D.M.E. is that any one beacon shall operate satisfactorily in the presence of 2,000 aircraft in an area defined as a square of 500 miles side if not more than fifty aircraft are tuned to the channel on which the beacon under consideration is operating. This corresponds to a very high traffic density, which for many years will probably only obtain for brief periods at particular locations.

Status

Before the installation of V.O.R./D.M.E. facilities in the U.S.A. was complete, the military authorities adopted a new system called TACAN. Because of the requirement for a common system for military and civil use, TACAN stations were soon being installed on the same sites as V.O.R.

Problems have abounded, and the civil users have understandably been reluctant to discard the V.O.R. facilities for which nearly all V.H.F. receivers now cater.

At the time of writing a hybrid system called VORTAC is being promoted in which the bearing is obtained from V.O.R. and the range from the distance element of TACAN.

At the best V.O.R./D.M.E. or VORTAC cannot give comparable accuracy with such systems as Decca, and a considerable body of opinion holds the view that the accuracy is in fact not nearly good enough for navigation in congested airspace.

TACAN might be expected to give somewhat higher bearing accuracy than V.O.R. (but still lower than Decca) and distance accuracies of both TACAN and D.M.E. are theoretically about equal.

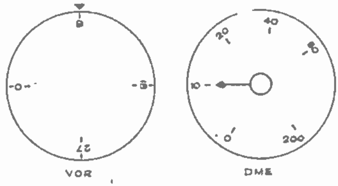
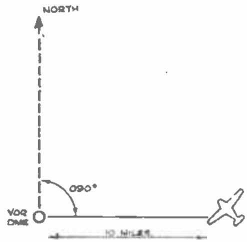


FIG. 5.—V.O.R./D.M.E. FIX.



The R/θ System of Navigation

It will already be seen that the position of an aircraft using V.O.R. and D.M.E. can be fixed by determining "θ", the bearing, and "R" the distance from the beacon. This provides a simple method of navigation as long as it is required to fly along a V.O.R. radial, but if it is necessary to fly along a line which is not a radial, the pilot will have to fly a compass course which he must check at intervals by taking V.O.R./D.M.E. fixes. In order to get round this difficulty it is intended to employ an "R/θ computor" which will automatically compute the data necessary to fly a given track, taking into account the continual change in range and bearing during the passage of the aircraft through the service area of the beacon. Experiments with this type of device continue with moderate success.

Tacan

This is the latest U.S. civil/military common-system short-range navigational aid to which reference has already been made. It provides both bearing and distance information, and its principles of operation are generally similar to those used in V.O.R./D.M.E. The TACAN system provides 126 two-way channels. Air–ground frequencies are in the range 1,025–1,150 Mc/s, and ground–air are in the ranges 962–1,024 and 1,151–1,213 Mc/s. All information is provided by a composite signal of considerably complexity.

Bearing. This information is provided in two stages to ensure accuracy, and both sets of information are provided by a rotating-aerial pattern producing an amplitude-modulated signal at the aircraft. The coarse pattern is in the form of a rotating cardioid pattern giving a 15-c/s modulation. Each time this sine-wave modulation (as seen from due south of the beacon) crosses the zero axis going in the positive direction a reference train of twelve sets of twin pulses (pairs spaced 30 microseconds is radiated. This is used to determine the aircraft's bearing in the same way as is the reference modulation in V.O.R., although the methods differ in detail. Because there are practical limits to the accuracy with which the phase of the 15-c/s signal can be measured, the station also radiates a 135-c/s pattern which goes through the complete phase cycle of the coarser pattern every 40 degrees. This pattern is highly accurate, giving 9 electrical degrees for every one degree of azimuth, but is ambiguous, and the coarse pattern serves to resolve this ambiguity. The fine pattern has its own reference burst which is radiated every time it crosses the 40° zero line, going in the positive direction. This train consists of six twin pulses with 24-micro-second spacing between the pairs.

Range. This involves a pulsed interrogator–responder system basically similar to D.M.E.

Identification. This facility consists of the transmission of pulses at a P.R.F. suitable to give an audio signal, and arranged to form a Morse code identification signal every 75 seconds. A future application of these pulses could be to transmit more complicated information on a digital basis.

Distance Measuring Equipment (TACAN)

To enable the TACAN distance-measuring facility to be used by aircraft fitted with V.O.R. equipment, a supplementary unit called D.M.E.T. has been developed in the U.S.A. which enables the TACAN distance facility only to be interrogated. A typical equipment provides distance information up to 200 miles (with accuracy 0·17 mile or 0·2 per cent of distance measured) and search time 22 seconds maximum. The power amplifier provides 1 kW peak pulse in any one of 126 channels 1 Mc/s wide in the range 1,025–1,150 Mc/s.

HYPERBOLIC SYSTEMS OF NAVIGATION

Several of the systems to be described subsequently depend for their operation not on the determination of the actual distances of the aircraft from two or more ground stations but on the difference of the distances from such points.

It is clear that if we take two points on the ground and measure the distance from the aircraft to each point we can fix its position in one of two places (one on each side of the line joining the two ground points). If we know which side of the line we are, this ambiguity can be resolved, or we can provide a third reference point so placed that any position has three distance readings uniquely corresponding to it.

To carry out this exercise by radio or radar in the aircraft would involve the use of transponder beacons on the ground and interrogator equipment in the air. The use of continuously radiating ground beacons not only simplifies the airborne equipment but also makes the ground facility immune from saturation. With such a system we can no longer determine the actual distance from two stations by measuring the time of travel of a pulse over the double journey from air to ground and back, but we can measure the *difference* in distance from the aircraft to each ground station by measuring the difference in arrival time of two pulses sent out simultaneously by the two stations.

It can be shown that the locus of a point whose distance from two others shows a constant difference is a hyperbola. Other differences will correspond to other hyperbolæ. The line corresponding to no difference in distance will, of course, be a straight line, the right bisector of the line joining the two stations.

Consideration will show that two stations cannot provide a fix by this method but only a position line, anywhere on which the aircraft may be situated. It is, therefore, necessary to provide a third station

FIG. 6.—HYPERBOLIC FIX FROM THREE STATIONS.

so that two intersecting position lines may be set up corresponding to the stations taken in pairs AB, BC.

Further information on hyperbolic systems is given in Section 18.

Gee: A Medium-distance Hyperbolic System

This British system was developed during the late war, and is still in wide use, although not adopted internationally.

A Gee pair capable of providing a position line consists of a "master station" and a "slave" separated by about 70 miles. The stations operate on frequencies of about 20–85 Mc/s with pulse powers of the order of 300 kW. The master station radiates 6-microsecond pulses, which are received direct by the aircraft and at the same time, trigger the slave station, which sends a further pulse to the aircraft. The aircraft thus receives two pulses the difference between whose time of arrival corresponds to the difference in distance from the two stations plus a fixed delay corresponding to the distance between master and slave stations. All that is needed in the air is a suitable V.H.F. receiver and a cathode-ray display for measuring the time differences.

Accuracy and Coverage

The accuracy of a Gee system depends on the precision with which the time differences can be measured, and this is about one-tenth of the pulse width or 0·6 microsecond. The absolute accuracy falls with distance (as the hyperbolæ diverge) from 100 yards at the base line to 5 miles at 450 miles. Coverage depends on altitude, but is considerably better than optical line of sight at low levels. Gee "chains" usually consist of four stations, if all-round coverage is required, for reasons common to all hyperbolic systems which are dealt with below.

Accuracy Limitations of Hyperbolic Systems

The accuracy of position fixing on hyperbolic systems depends on:

(*a*) The closeness of the spacing of the position lines; that is, how many hyperbolæ can be accurately laid down and the way the spacing varies with distance and at different bearings.

(*b*) The angle at which the position lines from two pairs of stations intersect.

We have already seen that the lines diverge with distance, and this factor causes the absolute error to increase while remaining an almost constant percentage of range. A further effect arises as we progress from bearings which are near the right bisector of the base line to those which approach the base line. At bearings beyond about 60 degrees from the normal to the base line the spacing between adjacent hyperbolæ corresponds to larger and larger changes of bearing, and it is consequently considered that accurate coverage of hyperbolic systems is limited to an area on either side of the base line bounded by lines at angles of 60 degrees from the right bisector of the base line. Thus, to obtain complete 360-degree position-line coverage we need three slave stations situated as nearly as possible at 120-degree intervals around the master.

The above arrangement, while satisfying the requirements for position-line accuracy, does not solve the problem of accurate fixes in all locations. The position lines cut at angles sufficiently obtuse to give accurate fixes at all points within short and medium distances, but the angle of cut becomes too acute for accurate working at distances great compared to the base line. The result of this is that the radial accuracy is low and, in fact, it varies approximately inversely as the square of the distance, while the position-line accuracy decreases directly.

The only way to combat this trouble is to increase the base line or to so dispose the stations that the position lines cut more nearly at right angles. Both these methods are used in hyperbolic systems in actual practice.

NAVIGATIONAL AIDS (MEDIUM AND LONG DISTANCE)

Decca

This is a British hyperbolic system suitable for medium- and long-distance navigation but of very high accuracy, enabling it to be used also for short-distance working.

Decca operates on low frequencies (L.F.) and is a continuous-wave system which makes it unique among the hyperbolic systems. Its ground transmitters are of 1 or 2 kW power, and have simple vertical aerials with heavy top loading. Special aircraft equipment is required.

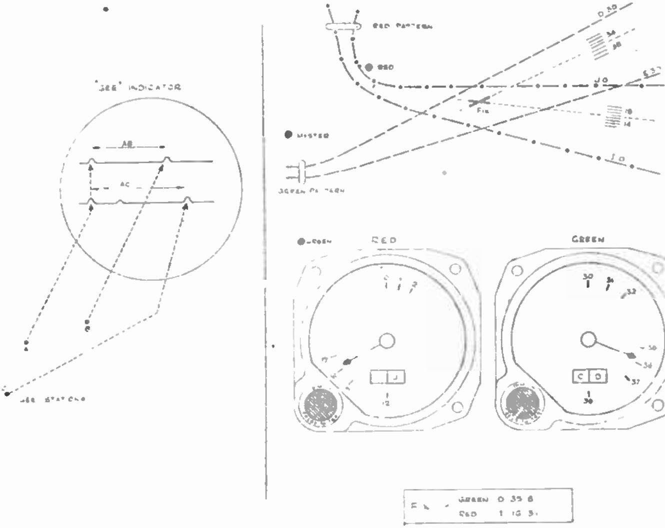


FIG. 7.—COMPARISON OF GEE REPRESENTATION (*left*) AND DECCA REPRESENTATION (*right*).

The basic element of the system is again the two stations, master and slave, separated by 60–70 miles. To reduce the problem to its simplest terms, let us assume that both stations are radiating on the same frequency and that the phase relationship between the master and slave emissions is so adjusted that at the master station the two signals are in phase. Now, if we move along the line joining the stations we shall pass through other points of phase coincidence at one-half wavelength intervals. Extending from these points in either direction from the base line will be hyperbolæ of phase coincidence corresponding to definite differences in distance from the two stations. In between these position lines there will, of course, be many others corresponding to other constant phase relationships. Suitable receiving equipment in the aircraft feeding a phase meter will enable us to determine the phase difference between the two signals. This will tell us that we may be on one or other of several hyperbolæ. (It is obvious that if there are many lines of phase coincidence, there will also be many other lines of identical phase difference.) There is thus a high degree of ambiguity in the basic system, but if this ambiguity is resolved, the accuracy then becomes very high, as will be shown later.

Practical System

In practice, the master and slave stations must be on different radio frequencies or it will not be possible to distinguish between them, and two frequencies bearing some simple relationship such as 90 and 120 kc/s are chosen.

The signals on 90 and 120 kc/s are received on separate receiving channels and multiplied four times and three times respectively to produce two signals of 360 kc/s, whose phases can now be directly compared. It is important that the receiving and converting system shall not disturb the phase relationship.

The position line spacing (i.e., the distance between lines of similar phase relationship) is a function of the *comparison* frequency and not the original carrier frequency. Thus, along the base line, the points of phase coincidence, for example, will be separated by one-half wavelength at 360 kc/s in the case we are considering (i.e., 430 metres).

The higher we make the comparison frequency, the greater will be the phase sensitivity of the system (and also the greater the ambiguity, as the number of "lanes" bounded by position lines of similar phase will be increased).

In practice, a Decca chain consists of a master and three slave stations giving 360-degree coverage. The aircraft carries the necessary receiver and phase meters, and special maps are provided overprinted with patterns corresponding to the position lines generated by the various pairs of stations.

As an aircraft moves to and fro in the service area of a Decca chain, the phase meters will rotate continuously to and fro, and arrangements are made to indicate on each instrument the number of complete revolutions it has made. Thus, each position line can have allocated to it a number made up of so many complete revolutions plus the position of the fine-reading pointer, which rotates once per "lane". Thus, if the instruments are set before take-off to the number of complete rotations corresponding to its position on the ground, the aircraft's position can be read off directly at any time in subsequent flight by noting the reading on the coarse and fine scales and referring to the appropriate lanes on the chart.

This simple system fails through ambiguity if the reception is interrupted or if the aircraft enters a Decca pattern at an unknown point. This problem is countered by "lane identification". The basic principle of this is to alter the characteristics of the system for short periods at intervals, to lay down a coarse pattern with no ambiguity but sufficient discrimination to show which "lane" the aircraft is in.

Accuracy and Coverage

Accuracy is extremely high. Position-line accuracy is ±10 yards on the base line of any pair to about ±100 yards at 200 miles. Accuracy of fix is rather worse than this, depending—as in other hyperbolic systems—on the angle of cut of the lines. At night the system is subject to errors at distances where the sky wave is appreciable, but this aspect is under continual investigation.

If narrow-band receivers are used, the protection from interference is good, but inferior to pulse systems.

Developments of the Decca System

Decca Flight Log

The basic Decca system so far described has the disadvantage operationally that it is not very suitable for pilot operation, needing as it does the reading of several meters and reference to a special map. This is not quite as easy for the pilot as the direct information from two dials that he is so many miles at such a bearing from a known point (as with the V.O.R./D.M.E. system).

The recent development of the "Flight Log" more than solves this problem. Basically the system consists of a specially drawn chart which moves on rollers and over which (at right angles to the motion of the chart) moves a stylus. Both these motions are controlled by the navigational information provided by a conventional Decca system via a convertor unit, which converts the data into the necessary rectangular co-ordinates.

By this means the pilot can actually see his track being plotted as he goes along.

The charts are provided in rolls of fifteen in suitable cassettes: the required chart is automatically selected by turning a dial on the control box.

A further feature is a "memory device" by which, if the aircraft flies off the chart, the stylus will commence to draw its track correctly as soon as it re-enters the area covered by the chart.

Further Refinements of the System

The latest Decca receivers avoid complete loss of indication in the event of the master transmission not being received for a period, by using a stable oscillator driven by the master signal in place of direct use of the master signal. With this type of receiver, the oscillator will continue to run with adequate accuracy during periods of non-reception of the master signal, and navigational information will not be interrupted.

Dectra

This is a projected Decca system for use over long sea routes, and is intended to give a number of tracks between two terminal points which can be flown with the aid of a simple left-right indicator. A further facility is a direct-reading range indicator giving distance flown. The claimed track accuracy at the worst point is ± 5 miles, and the range accuracy 10 miles. Track accuracy will improve rapidly as the terminal points are approached.

The system employs two stations at each end of the route, arranged at the usual Decca spacing, and so disposed that the right bisector of the base line of each pair coincides with the route between the two terminal points.

Each pair lays down a track pattern, and one station at each end forms a pair to give the range information.

As the distance is expected to be up to 2,000 miles, it is possible that after the middle point is passed the reception of the station at one end or the other which is being used as one of the range pair may fail, and this problem is met by using a highly stable (1 in 10^8) crystal oscillator in the aircraft to replace the missing signal. With this method, the range information is generated by one station and the oscillator.

A pilot system is now in use on the North Atlantic; and results so far bear out the claims referred to above. The stations at each end are situated about 80 miles apart, and the radiated power into vertical aerials of 400–600 ft. is estimated at 4 kW in the case of the masters and 2 kW for the slaves. The system works on two frequencies in the region of 70 kc/s, and in common with normal Decca requires little channel space (10 c/s for each channel).

Delrac

This is another projected Decca system. It has not yet been evaluated by practical tests. It is a C.W. phase comparison system operating in the 10–14-kc/s region. Presentation can be either on the Decca flight log or on Decca phase meters. The ground stations would be in master-and-slave pairs, each on a 1,000-mile base-line. Each pair produces a position line, and because of the small channel space needed (5 c/s) several pairs can be used, enabling fixes to be obtained.

Ambiguity is resolved as in standard Decca by producing coarse patterns at intervals; this is done in Delrac by radiating bursts of signal on the basic frequency, followed at discrete intervals by several other frequencies, giving two or three successively coarser patterns. Thus for a two-stage system on a basic frequency of 12 kc/s, the ancillary frequencies would be 16 and $13\frac{1}{3}$ kc/s or for a three-stage system 16, $13\frac{1}{3}$, $12\frac{4}{9}$.

Radiated powers of the order of 10 kW are envisaged, and accuracies of not worse than 10 nautical miles anywhere in the service area are anticipated. Tests so far have shown greater accuracies. It is estimated that twelve pairs of stations could cover the extra surface of the globe.

Consol

This is a medium-frequency continuous-wave system of German origin and requires no special equipment in the aircraft (only a radio compass capable of continuous-wave reception such as is carried by most aircraft).

Each Consol station lays down a series of radial position lines similar to V.O.R. except that the service area of any one station is limited to two opposite sectors 120 degrees wide. Two Consol stations will provide a fix.

The station has three vertical aerials evenly spaced on a base line of about 3 miles, and so fed that two overlapping patterns of narrow lobes are laid down. One pattern is keyed with dots and the other with dashes, and the keying characteristics interlock, so that along the equisignal lines where adjacent lobes overlap a continuous note will be heard. The phases of the inputs to the aerials are altered in synchronism with the keying, so that during the keying cycle each line moves up to take, at the end of the cycle, the position originally occupied by the next one to it. The patterns on either side of the station move in opposite directions, and at the end of the keying cycle they revert to their original positions. Each cycle is preceded by the call-sign of the station and a long dash to warn the user to be ready to start his observations.

Now if we consider an aircraft situated at some point within one of the twenty-four sectors set up by the station, the operator will hear the identification and warning signal followed by, say, a succession of dots which will fade into a steady signal as the equisignal line passes through the aircraft. After the equisignal, dashes will appear.

The angular position of the equisignal line at its passage through the aircraft can be fixed by counting the number of characters heard before and after the steady tone, as their number is fixed and their spacing is locked to the rotation of the pattern. Tables are available showing the bearings corresponding to different counts for each station.

The obvious ambiguity can be resolved by taking a radio-compass bearing on the station, which will determine in which sector the aircraft is flying.

Accuracy and Coverage

Day-time accuracy of Consol is between 0·2 and 0·5 degrees near the right bisector of the base line, decreasing to 0·4–1·0 degrees towards the 60-degree limit of coverage either side of the normal.

Range is 700 miles over land and 1,000 over sea. Night ranges are greater, but errors due to sky wave will arise.

Speed of fix depends on the length of the cycle, which varies from 40 to 120 seconds. Three to five minutes should suffice to get a fix from two stations.

Loran

This American hyperbolic system is primarily intended for long-range navigation over large sea areas. The principle is similar to Gee, but the stations operate on frequencies of the order of 2 Mc/s with 50-microsecond pulses of 100 kW. and a pulse-repetition-frequency (P.R.F.) of the order of 25 per second.

Chains consist of a master and two or more slaves. Adjacent stations operate on the same pulse repetition-frequency and the pulse repetition-frequencies of all pairs in the chain differ (so that the operator can pick out the pair he wants by synchronizing the sweep of his cathode-ray display to the appropriate pulse repetition-frequency). All stations in a chain, except the end ones, are double, as they are each part of adjacent pairs.

The different pulse repetition frequencies make it impossible to get fixes from two pairs at the same time without considerable duplication of apparatus in the aircraft, and consequently a Loran fix takes somewhat longer than Gee.

Accuracy and Coverage

Loran is subject to the same geometrical restrictions as all other hyperbolic systems, but within these restrictions the average error of fix from a three-station chain in the ground-wave region ranges from 300 yards at short distances up to 1 mile at 700 miles. On sky wave errors range from 1½ to 8 miles at distances from 300 to 1,400 miles.

The pulse is considerably rounded, and time measurement involves adjusting amplitudes of two pulses and matching them by superimposition on the cathode-ray-tube display. This is difficult if the pulses are distorted by sky-wave transmission, and is a factor in the lower accuracy achieved in sky-wave operation. Also Loran charts are computed for ground-wave working, and correction must be applied when using sky wave.

Although the frequencies used by Loran are subject to heavy noise and other interference, the system operates well in all parts of the world, as the signals have been found to be readable through interference of much greater strength.

Ranges over land are of the order of only 100–250 miles by day. Over water, day-time ranges are 700 miles or more. At night sky wave increases ranges to 1,000–1,400 miles. Range is more or less independent of aircraft altitude above 1,000 ft.

Sky-wave Synchronized Loran

To take advantage of the great night-time sky-wave ranges of Loran to increase the base line (see "Accuracy Limitation of Hyperbolic Systems"), Sky-wave Synchronized (S.S.) Loran was evolved. This system uses widely separated stations with synchronizing pulses transmitted from a master to slave by sky wave.

The best arrangement is with four stations at the corners of a square of roughly 1,000-mile side. Useful coverage extends over nearly the whole of the 1,000,000 square mile area, and while accuracy is greatest at the centre, it is nowhere reduced below 1¾ miles by the geometry of the system.

Errors due to variations in height of the reflecting layer may reach 8 microseconds, causing an error of no more than 1 mile from this cause at the centre of the area. Further errors arising from unpredictable excursions in path length due to ionospheric storms can reach 5 miles, but these have not been found to obtain for more than 1 per cent of operating time. Serviceability averages 95 per cent of the night, except when reflecting points are in the auroral zone.

L. F. Loran

Loran operating on low frequencies has a coverage of the same order as S.S. Loran at all times of the day. The pulse width is increased to 300 microseconds, and consequently accuracy of timing falls to 4 microseconds. This would be compensated for by the increased geometrical accuracy with the greater base line, but also a system of "cycle-matching" has been suggested which would give timing accuracies of $\frac{1}{10}$ microsecond.

It is interesting to note that pulse-repetition-frequency selection would enable sixteen chains to be operated in one radio-frequency channel of 30 kc/s width and that twelve chains would cover the major part of the globe.

Loran-C (CYTAC)

This system is basically similar to standard Loran (now known as "Loran-A"). It is an American development and is only in experimental use. It differs from Loran-A in the radio-frequency used, which is of the order of 100 kc/s, at which a band-width of 20 kc/s is needed. The pulse coding and methods of measuring at the receiving end are more sophisticated than Loran-A, giving greater accuracy under worse conditions. Ranges of 1,200–2,000 nautical miles on groundwave and 3500 nautical miles on skywave are claimed using pulse powers of 60 kW (radiated).

The preferred station configuration is master and three slaves located at the corners of a square of 750 nautical miles side. With this configuration, fix accuracies of 1 nautical mile are predicted. It is estimated that by pulse-rate coding, more than enough stations to give world coverage could be accumulated in a few kilocycles band-width, (the 20 kc/s mentioned above would accommodate 288 stations; only fifty are estimated to be needed). Errors due to propagation conditions are said to be low, and resistance to interference to be high because of the refined pulse techniques used.

Navarho

This is an experimental American R (or Rho)–theta system of long-range navigation. It is basically a one-site system which distinguishes it from most other long-range systems.

The azimuth element is provided by three aerials spaced at the corners of an equilateral triangle of sides equal to 0·36–0·4 wavelengths (at the operating frequency of 90–110 kc/s). These aerials are fed sequentially in pairs with co-phased equal-amplitude power. The received signals are compared vectorially in the aircraft to extract the bearing information. The distance element is provided by phase comparison between a highly stable oscillator (1 in 10^9) in the aircraft and a ground transmitter to which it has been synchronized before departure. As long as the stability of the oscillator in the aircraft is of this high order, any phase changes between the two signals will be due to the change in distance from the ground station caused by the motion of the aircraft. In practice, the ground station transmits four pulses every second. These are 170 msec. in duration and separated by 80 msec. The first pulse is for synchronization and for phase comparison for the distance element. The remaining three are transmitted in sequence from the three aerial pairs and provide the azimuth element.

The predicted accuracy of Navarho is 5 nautical miles for distance and 2 degrees for azimuth. Ranges of 1,500–2,000 nautical miles are expected. Early flight trials have produced figures worse than the above.

Although the band-width used is small (20 c/s), the azimuth element needs good signal-to-noise ratio to provide reasonable accuracy; radiated powers of the order of 40 kW are thought to be needed. Reduction of the band-width of the receiver to as low as 1 c/s would improve the operation and is practicable.

The considerably better accuracy of the distance element would suggest a rho–rho system using distance information from two or more stations to provide a fix. This would, however, remove the one-site aspect and turn it into just another hyperbolic system.

The system has no ambiguities of operational significance. R–theta systems for long-range must inevitably be of lower accuracy than, say, hyperbolic systems, because of the difficulty of measuring theta to sufficient accuracy.

Omega

This is an American very-long-range hyperbolic system operating in the V.L.F. band (10–14 kc/s). The system is experimental only.

The principle of the system is basically similar to Decca in that phase comparison is used to determine distance from any two stations. Spatial ambiguity attendant on this system is resolved by transmitting also on a secondary frequency (removed by about 500 c/s) from each station. Estimated accuracy is 1 nautical mile. No thorough evaluation has yet been carried out.

Radio Web (Radio Mesh)

This is a French experimental system. It operates on the principle that if two stations radiate signals on different but close frequencies a moving line of phase coincidence will travel between them. This line has the form of a hyperbola with the stations as its foci. In practice, the information-carrying frequency can be the modulation frequency.

By using four stations at the corners of a quadrilateral, a system of intersecting position lines can be defined. If the product of the scanning-period times the scanning speed is less than the distance between the stations there will be no ambiguity of equiphase lines. In the experimental system four stations are situated at the corners of a rectangle of about 100 km. side. Since the position is recalculated by the equipment every second, failure of the signal will not produce any error when reception is resumed, because each determination of position is quite separate and independent.

The tests so far have been carried out on frequencies where the ground-wave propagation is very poor (1,600–1,900 kc/s), which resulted in very low signal strengths. Errors under these conditions have not exceeded 2 miles, and it is expected that on lower frequencies these results could be considerably improved.

The area covered would be increased greatly by correct choice of carrier frequency and modulation-frequency difference. A 4 to 1 increase in the distance between stations could be achieved without ambiguity with the present modulation frequencies.

THE RADIO COMPASS

The radio compass consists essentially of a receiver tuning between approximately 100 and 1,750 kc/s associated with a loop aerial which is driven by a servo system to align itself automatically to the bearing of any station to which the receiver is tuned. Sense signals from a separate omni-directional aerial are fed into the system to enable the loop system to reject the reciprocal bearing.

The loop position is fed back to an instrument with a 360-degree compass scale, on which the bearing of the station with relation to the heading of the aircraft is displayed. Often two radio compasses operating two concentric pointers on one instrument are used to enable simultaneous cross bearings to be taken on two stations.

The radio compass is used with either known broadcast stations or special beacons, such as "locator beacons", used in conjunction with the instrument-landing-system (see later).

Accuracy of radio compasses is of the order of 1 or 2 degrees. They are subject to the usual polarization and night-effect errors common to direction-finding (D.F.) systems. Ranges up to several hundred miles may be expected, according to the power of the ground station.

SYSTEMS INDEPENDENT OF GROUND FACILITIES

Methods of navigation independent of ground aids are attractive from both the civil and military points of view, and a great deal of research has been devoted to this subject.

The main classes into which such devices fall are as follows: (*a*) dead reckoning; (*b*) celestial; (*c*) inertial; and (*d*) radar.

The first involves estimation of the aircraft's present position by calculation based on its assumed speed, heading and drift since the last time at which its position was known. The system is of low accuracy, because it is usually based on information of low accuracy, particularly in relation to true ground speed, which has to be deduced from indicated air speed and such information as is available on the wind velocity.

Celestial navigation is very accurate and involves calculation of the position of the aircraft from measurements of the apparent position of various stars and other heavenly bodies. To make use of the system a navigator must be carried, although automatic star-tracking systems have been proposed. Also adequate visibility of the heavens is essential but not always possible except at good altitudes.

Inertial navigation is effected by measuring the acceleration of the aircraft in various directions relative to its fore-and-aft axis and using this information to compute automatically its movements and hence its present position.

System (*d*) is the only one which really falls within the scope of this chapter and is represented at the moment by the various Doppler systems of navigation.

Doppler Systems

Doppler is not in itself a complete system of navigation, but it provides accurate information on the true velocity of the aircraft relative to the ground, which is the main imponderable in dead-reckoning navigation.

General Principles

In essence, the system consists in projecting a radio signal towards the ground below the aircraft and measuring the Doppler shift in frequency of the reflected signal arising from the motion of the aircraft relative to the ground. The magnitude of the shift is given by the expression:

$$d = \frac{2vf}{c} \cos \alpha$$

where d is the Doppler shift; v is the velocity of the aircraft; f is the frequency of the signal; c is the velocity of propagation; and α is the vertical angle between the direction of motion of the aircraft and the direction of propagation of the signal.

In normal flight the velocity of an aircraft will have two components—a major one in the direction of the fore-and-aft axis of the machine and a minor one, at an angle to the other, due to the wind. The actual track of the aircraft will be along the resultant of these two vectors.

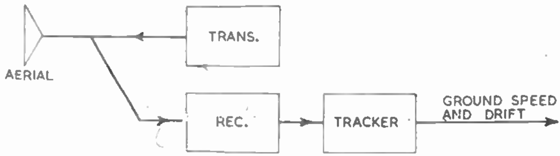


FIG. 8.—BASIC DOPPLER SENSOR.

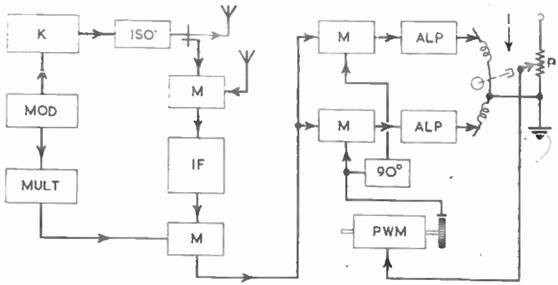


FIG. 9.—DOPPLER T/R AND TRACKER UNITS.

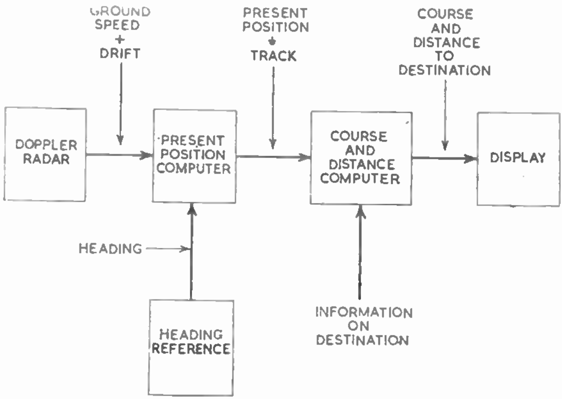


FIG. 10.—DOPPLER NAVIGATOR SYSTEM.

The drift angle, represented by the angle between this resultant and the heading of the aircraft, can be determined by using two Doppler beams inclined symmetrically to either side of a vertical plane passing through the machine's fore-and-aft axis.

The drift can either be computed from the difference in the Doppler shifts on these two beams, or this difference can be used to rotate the aerial system in azimuth until equal shifts are obtained. The angle assumed by the aerial system under these conditions is the same as the drift angle.

Having determined the forward velocity relative to the ground and the angle of drift, it is only necessary to know the aircraft's true heading in order to define its true speed and direction relative to the ground. The heading information can be obtained from the aircraft's compass.

A further necessary refinement is the setting up of one or two beams projected backwards relative to the first two. This provides an additional source of information which can be used to render the system less sensitive to changes in altitude and also to mitigate the effects of pitching.

Since the beams are not infinitely narrow, and since the reflecting surface is rarely of a uniform nature, the Doppler shifted signal received is never on one frequency only but tends to spread over a small band. It is necessary to determine the centre of this frequency spectrum in order to fix the average Doppler frequency, and this is achieved by what is generally known as the frequency tracker system.

To reduce errors arising from aircraft attitude still further than is possible by the backward–forward beam (or Janus) system, it is necessary to introduce pitch stabilization of the aerial system or, alternatively, to introduce a correction in the Doppler data derived from information about the aircraft's attitude. In either case the attitude information is already available in the aircraft by instruments normally carried.

Practical Systems

The simplest Doppler system will display ground speed and direction only, but most practical systems employ additional computing equipment to enable automatic calculation of the aircraft's position to be achieved.

Such a system is depicted in much-simplified form in Fig. 8. The basic configuration of the Doppler radar shown in this figure is further elaborated in Figs. 9 and 10.

Doppler radars fall into two main categories: C.W. and pulse. Practical systems have been evolved using both methods, although C.W. presents some problems concerning effective transmitter–receiver isolation.

A modern C.W. system is represented by the Marconi AD2300 series which uses frequency of 8,800 Mc/s with a C.W. power of 1 watt and aerial beam width (at 3-db points) 4 degrees, depression angle 67 degrees, broadside angle 12 degrees. Instrumental accuracies obtainable with the AD2300A (sensor only) include: ground speed 0·5 per cent ±3 knots; drift angle ±0·25 degree; distance flown 0·5 per cent. With Types AD2300B and AD2300C, which include computers, position accuracy can be indicated to within 1 per cent of distance flown when heading is known to within ½ degree. The basic inputs to the computer are ground speed and drift angle from the Doppler element of the system, heading from the aircraft's gyro-magnetic compass or other heading reference, and true airspeed from the aircraft's instrument system. The computed and other outputs are fed to a navigation display panel, while those required by the pilot are fed to his flight-instrument system and to the automatic pilot if desired.

In most, if not all, existing military Doppler equipments, the operation is directly monitored and controlled by a specialist member of the crew. This makes it possible to cut down the amount of electronics; in a civil aircraft these tasks must be performed automatically. The AD2300 series provides automatic search, lock-on (acquisition) and memory. The automatic memory comes into action whenever the Doppler signal-to-noise ratio is below an acceptable level. Automatic search is designed so that the spectrum is continuously monitored to ensure that the correct signal is selected. It is of particular importance in a civil Doppler navigator that the equipment should "fail safe".

A representative pulse Doppler is the Decca Type 61, which forms the Doppler element of the DIAN system in which a Decca Flight Log or other display can be run at will from either Doppler or Dectra. The power is 20 watts peak with a 50 kc/s P.R.F. and a 4-microsecond pulse width. The aerial beam width is 4½ degrees, with a depression of 67 degrees and a broadside angle of 10 degrees. The equipment includes a memory system which continues the computation during any interruption of the Doppler signal using the last known heading and airspeed of the aircraft.

APPROACH AND LANDING AIDS

Ground-controlled Approach (G.C.A.)

This is a precision radar system enabling the position of an aircraft, in range, elevation and azimuth, to be determined on the ground with respect to the correct point of touch-down on the runway. As no special apparatus is required in the aircraft beyond either a V.H.F. or high-frequency communication set, the system will not be described here. The pilot controls the aircraft entirely on verbal instructions from the ground controller.

Instrument Landing System (I.L.S.)

This system, which is internationally standardized as the primary approach and landing aid, requires extensive special equipment, both on the ground and in the aircraft. It relies entirely on the interpretation by the pilot of data presented to him by automatic equipment on the ground. No ground operators are involved.

The system has three basic functions which give the pilot data on:

(*a*) his position in elevation relative to the correct landing path;
(*b*) his position in azimuth relative to the correct course to fly to line up with the runway;
(*c*) the moment of his passage over three salient points on the approach path.

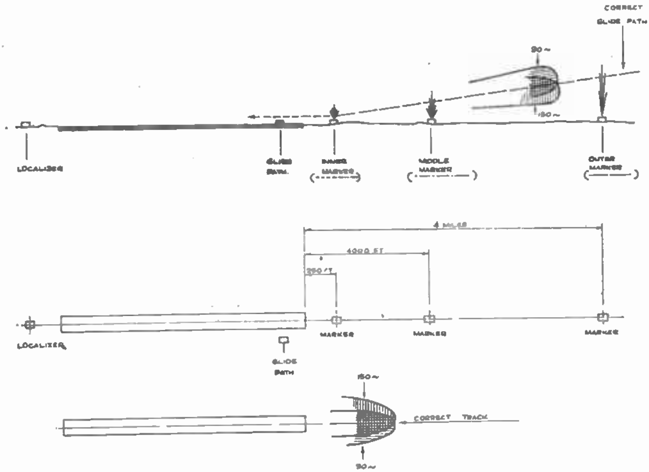


FIG. 11.—INSTRUMENT LANDING SYSTEM LAYOUT.

A further ancillary to the system, the locator beacon, enables the pilot to approach from a distance (using a radio compass) to a point where he can pick up the transmissions giving him data on (*a*) and (*b*).

The Locator Beacon

This is a low-power medium frequency beacon radiating an M.C.W. (A2) Morse identification signal and providing a continuous signal for operating a radio compass. By homing on this beacon with his radio compass the pilot arrives at a point where he can pick up the signals from the I.L.S. localizer.

I.L.S. Localizer

This comprises a transmitter and aerial system located about 750 ft. beyond the " stop " end of the runway. Two signals are radiated from separate aerial systems. One is modulated at 90 c/s and the other at 150 c/s. The two aerial patterns intersect so that the equisignal line is down the centre of the runway. The localizer receiver in the aircraft covers the frequency range allocated to the service (108–112 Mc/s). It feeds both the 90- and 150-c/s signals to a centre-zero instrument (also the course-deviation instrument for V.O.R.), and when this instrument's needle is centred the pilot knows that he is correctly aligned with the runway.

The localizer signals are horizontally polarized (as for V.O.R.), and the system is always arranged so that the 150-c/s pattern is on the right as the aircraft faces the approach end of the runway. Coverage is such that usable signals are obtained up to 2,000 ft. altitude at 25 miles over a 20-degree sector centred on the runway centre line. Cover is also provided up to 17 miles at 2,000 ft. in any other direction, and up to 1,000 ft. all over the immediate vicinity of the landing area. The course sensitivity is such that a 3-degree deviation will result in a full-scale deflection of the I.L.S. pointer.

The localizer can be modulated with speech, and also radiates a two- or three-letter Morse identification signal. These facilities are available without interruption of the navigational facility.

Often, V.O.R. receivers are arranged to act also as I.L.S. Localizer receivers.

I.L.S. Glide Path

This operates on frequencies between 328·6 and 335·4 Mc/s, and needs a separate receiver in the aircraft.

The transmitter and aerial system are located beside the runway near the "touch-down" point, and in this case the 90- and 150-c/s patterns are arranged so that the equisignal line corresponds to correct angle of descent for the aircraft. The line can be adjusted to lie between 2 and 4 degrees above the horizontal, and usually passes over the "touch-down" point at about 20 ft.

The signals are again applied to a centre zero instrument whose pointer is horizontal when in the centre position. This instrument is incorporated in the I.L.S. localizer indicator, and its pointer is at right angles to the localizer pointer. The system is always arranged so that the 150-c/s area is below the glide path.

Glide path coverage is 10 miles over an 8-degree sector in azimuth about the runway centre line.

Marker Beacons

These are used to mark specific points on the approach path to give the pilot some information on his passage along it.

Each beacon is a low-power, amplitude-modulated transmitter on 75 Mc/s. The aerial systems are arranged to beam signals upwards into the approach path over a fairly narrow angle.

The designations of the I.L.S. beacons are :

Outer Marker : 4 miles from runway threshold.
Modulation—400 c/s, two dashes per second.
Middle Marker : 2,000–4,000 ft. from runway threshold.
Modulation—1,300 c/s; alternate dots and dashes.
Inner Marker : 250 ft. from runway threshold.
Modulation—3,000 c/s; six dots per second.

The pilot hears the distinctive signals in his headphones as he passes over the beacons, and at the same time different coloured lights flash on the instrument panel near the I.L.S. cross-pointer instrument.

The aerial patterns are so arranged that with an aircraft flying on the normal approach path at 90–100 m.p.h. the signals will be audible for times ranging from 6 seconds for the outer marker to 3 seconds for the inner.

For reception of I.L.S. markers or airway markers (which all operate on the same radio frequency) a special receiver is required in the aircraft. This is usually a simple single-channel, crystal-controlled superheterodyne with a system of audio-filters, rectifiers, relays, etc., to operate the signal lamps.

A Typical I.L.S. Receiving System

A typical I.L.S. equipment will usually consist of a V.O.R. receiver (which is commonly designed so that, when it is tuned to the I.L.S. localizer frequencies, circuits suitable for the 90/150 c/s I.L.S. signals are brought into operation) together with a Glide Path receiver and a Marker Beacon receiver.

The Glide Path receiver will usually provide twenty crystal controlled channels at 300 kc/s intervals in the range 329·3–335 Mc/s. The I.F. band-width will be of order the of 135 kc/s at the 6 db point and 500 kc/s at 60 db. The A.G.C. should hold the output substantially constant for inputs between 30 and 100,000 μV. Gain uniformity from channel to channel should be such that the on-course output will not change by more than ± 2 μA nor the deflection sensitivity by more than ± 3 μA.

Deflection sensitivity will be such that, with an average signal of several hundred microvolts, a difference in strength between the 90- and 150-c/s modulations of not more than 2 db will produce an output to the cross-pointer indicator of the order of 80 μA.

A Marker Beacon receiver is usually a straight T.R.F. receiver, crystal-controlled on one channel. The I.F. response will be of the order of 40 kc/s band-width at 6 db and 200 kc/s at 60 db. Automatic gain control should hold the output within 6 db for inputs between 200 and 100,000 μV. Sensitivity should be such that 100 μV will give adequate lamp indication.

MISCELLANEOUS NAVIGATIONAL SYSTEMS

Gee Track Guide

Gee is not a direct-reading system and is, therefore, not suitable for pilot operation, but a Gee system has been proposed which makes use of the fact that it would be quite practicable to make a direct-reading device of the "left–right" indicator type which would enable a pilot to fly along any selected Gee hyperbola: since this would merely involve detecting a constant time difference.

By re-arranging the disposition of the stations this facility could be made to produce an approach aid of longer range than I.L.S. and greater precision than V.O.R.

The system involves placing a Gee master and one slave about 10 miles apart so that the right bisector of the base line lies on the continuation of the runway. It can be shown that the Gee position lines can be regarded as straight lines radiating from the centre of the base line for all distances over three times the length of the base line (i.e., over 30 miles in this case).

Working from the known resolution of the system, it can be seen that there will be about thirty readily distinguishable tracks in the 60 degrees on either side of the normal to the base line. These could very well take the place of some of the radials of the V.O.R. system, but will afford much greater accuracy.

Seven of these tracks will also actually pass within 1½ miles of the centre of the base line, and one, of course (the line of no time difference), will pass exactly through the centre.

Since each line corresponds to a definite constant time difference, it would be easy to make the pulse spacing operate a centre zero meter through suitable circuits.

Using transmitters of 1 kW peak power, the range would vary from 50 miles at 1,000 ft. to 145 miles at 10,000 ft. Track accuracy would be of the order of ± ¾ degree.

Shoran

This system has nothing in common with Loran, as its name might suggest, but is a very accurate American system used for aerial survey work, and employs direct measurement of distance to two known points by using an interrogator in the aircraft and two transponders on the ground.

Teleran

This is a projected American system for determining the aircraft's position by ground radar and transmitting the data to the pilot, together with any other useful information by a television link.

WARNING DEVICES

Cloud and Collision Warning

One of the most serious natural dangers which beset aircraft is cloud of the cumulo-nimbus type, in which ascending and descending air currents may reach velocities of 30,000 ft./minute. At night it is quite possible to fly into such clouds without warning, but it was found several years ago that they could be detected by centimetric radar.

Such a system has been developed, and consists of a radar scanning a wide sector ahead of the aircraft, the scanner being stabilized against both roll and pitch.

This radar will give warning of tropical cumuliform clouds of dangerous size at distances of up to 40 miles. It also gives some protection against collision with other aircraft, which it will detect at distances up to 14 miles.

A further facility is terrain-clearance indication—the equipment will detect high ground within its field of view and give adequate information on its distance and the approximate relation between its height and the altitude of the aircraft.

Radio and Radar Altimeters

The barometric altimeter in an aircraft is set before take-off to read the actual height of the airfield above sea-level. This setting only holds good when the atmospheric pressure is the same as it was when the altimeter was set. If the aircraft flies to an area where the pressure at sea-level is different, it will not read correctly. Furthermore, it only shows height above sea-level—not above the ground.

In an attempt to provide an altimeter that would not only be immune from atmospheric change but would also indicate actual height above the ground the radio altimeter was developed.

The earliest successful system (which is still in use) involves a transmitter–receiver which radiates from a small dipole on one wing a signal varied sinusoidally between 420 and 460 Mc/s at a rate of 120 c/s. This signal is reflected from the ground and picked up by the receiver via an aerial on the other wing. The received signal is compared in frequency with the signal passed direct from transmitter to receiver and, because of the time taken to effect the journey to the ground and back, its frequency will be different from that being received direct from the transmitter.

The audio difference frequency is proportional to the altitude. It is averaged over a number of sweep cycles, and the result is applied to a meter calibrated in feet. The equipment has two ranges 0–400 ft. and 0–4,000 ft. Accuracy is 5 ft. + 5 per cent of the altitude for the low range and 50 ft. + 5 per cent of the altitude for the high range.

For high-altitude work the radar altimeter exists. This is a very simple radar system giving distance only. It uses the same type of aerials as the radio altimeter.

One of the factors which makes it still impossible to land an aircraft in zero or very low visibility (even with G.C.A. or I.L.S.) is the fact that it is impossible to measure its height with sufficient accuracy (±1 or 2 ft. is essential), and many types of radio altimeters have been projected for this purpose.

Phase Altimeters

One proposal on which some development has been carried out is the phase altimeter. In this system a carrier of the order of 3,200 Mc/s is modulated at 492 kc/s and the phase-shifted modulation in the reflected signal compared with the original modulation with a "Decca" type phase-comparison meter. The modulation frequency chosen produces 1 revolution of the phase meter per 1,000 ft.

The modulation frequency is crystal controlled to 1 part in 10,000; as 10 ft. in height corresponds to about 3 degrees phase shift, measurement should be possible to better than 10 ft.

This is not good enough for blind landing, but the whole system is basically more suitable for general altimeter work than the two previously mentioned, which can often be subject to gross errors, due to installation difficulties and the difficulty of correct adjustment during maintenance.

Modern F.M. Altimeters

The altimeter which has shown great promise of low-altitude accuracy is the latest F.M. type.

In this, a transmitter of about 500 mW power on 4,200–4,400 Mc/s is swept over a 100-Mc/s band. The reflected signal is picked up on a receiver and, after rectification in a balanced mixer, a signal whose frequency is proportional to height is obtained. This is further processed to obtain a D.C. output proportional to height.

A built-in check on operation and calibration is provided by the ability to connect a delay line of accurately known characteristics between transmitter and receiver aerials.

Versions of the system described above have been used in the United Kingdom to give the altitude information for completely automatic landing of aircraft. The accuracy of this equipment is said to be

3 per cent of indicated height; it is capable of resolving height differences of 2 ft.

Radar Transponders

To improve the operation of the ground radar equipment of Air Traffic Control, by enabling positive identification of aircraft echoes through precipitation or at the limit of range, transponders can be carried in the aircraft. These are interrogated by the ground facility and transmit replies which can be coded, if desired, to convey additional information.

AIRCRAFT AERIALS

In earlier days of aircraft radio the aerial consisted of a long, weighted wire which could be paid out or reeled in by a hand-driven winch. This type of aerial was used on larger aircraft until the end of the late war, and was very convenient for high-frequency working, as its length could be adjusted to suit the frequency being used.

The trailing aerial was often supplemented by a fixed aerial running from a mast, forward, to the tail fin. On the faster aircraft, this eventually replaced the trailing aerial, and is still used on the slower aircraft of the present day. There are difficulties attendant on the use of the fixed aerial, arising from its short length, which necessitate heavy loading on the lower portion of the high-frequency range. At higher altitudes, because large voltages are built across the loading coil, problems concerned with flash-over at the feed end of the aerial are likely to be

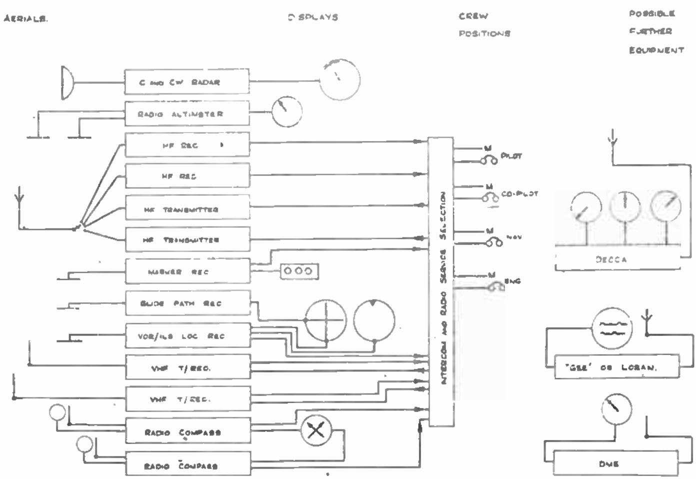


FIG. 12.—TYPICAL MAIN-LINE AIRCRAFT RADIO AND NAVIGATION INSTALLATION.

experienced, and this is a major obstacle to the increase of transmitter power.

For V.H.F. operation the most common aerial for many years has been the standard Air Ministry " whip ", a tapering steel rod mounted on a strong bakelite base and forming a simple quarter-wave vertical aerial. For the I.L.S. localizer and glide path and marker service, small dipoles are in use.

Radio compass aerials, originally unstreamlined, screened loops about 18 in. in diameter, soon became modified to smaller loops in streamlined housings, stood off from the fuselage on short pedestals. For sense aerials, two types have been in general use: a longer version of the V.H.F. whip or short horizontal wire aerials, stood off a few inches from the fuselage on short posts.

With the advent of jet aircraft it has become impossible to tolerate the drag of even small carefully streamlined aerials; " suppressed " aerials for all radio services are in existence, and considerable development is being devoted to this branch of the art.

Suppressed Aerials

(a) *For Radio Compasses.*—These commonly consist of small dust-iron-cored loops fitted in recesses in the fuselage and covered by flush-fitting insulating plates. Specialized design has resulted in these aerials being comparable in performance with the older type mounted clear of the skin.

(b) *For V.H.F. Services.*—For all V.H.F. purposes (I.L.S., V.O.R., communication, etc.) many types of aerials are in use, according to the type of aircraft. Two main problems affect the design of suppressed V.H.F. aerials: the difficulty of attaining the efficiency of externally mounted aerials and the necessity to obtain all-round coverage, especially for communication and V.O.R. aerials.

It is well known that a " slot " or cavity in a conducting surface can be excited in the same way as a conventional dipole, and this type of aerial is widely used; sometimes in highly specialized forms.

In other cases areas of the tail fins are made of insulating material and have various shaped metal-foil radiators fastened to them and suitably connected to act as aerials.

Another technique is to make the wing-tip or a portion of the leading edge of the main plane or tail out of insulating material and house within it a folded dipole or other similar radiator.

(c) *For High-frequency Working.*—The problem here is to obtain a radiator of sufficiently large dimensions to reach any sort of efficiency at the frequencies used (especially at the lower end of the band). The technique usually takes the form of exciting a large portion of the aircraft, such as the main plane or the fuselage, by means of a coupling coil. The problem is complicated by the difficulty of getting adequate coupling over the band, and provision usually has to be made for tuning of the coupling and also for variable matching. Although there are considerable difficulties in the suppressed high-frequency aerial technique, it has the advantage in high-altitude aircraft that the whole of the aerial-coupling system can be pressurized (as all parts are inside the aircraft), and this removes a great deal of the trouble with flash-over on high power at high altitudes.

D. H. C. S.

20. RADIO ASTRONOMY AND SATELLITE COMMUNICATION

20. RADIO ASTRONOMY AND SATELLITE COMMUNICATION

TECHNIQUES OF RADIO ASTRONOMY

In 1932 Jansky discovered that there are radio waves reaching the earth from outside it. The incoming radiation has the characteristics of random noise. Since then the contours of intensity over the sky have been plotted at different frequencies; in addition to an intense belt of emission along the Milky Way, many localized sources have been found, one of these being the Sun. The study of these naturally produced radiations has become known as radio astronomy.

At the present time observations are confined to wavelengths which can penetrate the earth's atmosphere. The micro-waves are affected by molecular absorption and the long waves are reflected out by the ionosphere. The limits are approximately 1 cm. to 30 m.

Quite apart from the intrinsic interest of the subject, practical applications include:

(i) Information on solar–terrestrial relations, leading to increased understanding of radio-wave propagation.

(ii) Information on refraction and absorption by the ionosphere.

(iii) Knowledge of the sky noise level.

(iv) Use of radio sources for navigation independent of visibility.

The techniques of radio astronomy are aimed at particular aspects of the general problem of determining the intensity and polarization of the incoming radiation as a function of direction, frequency and time.

It is already known that:

(i) The distribution of emission over the sky consists of a background radiation, superimposed on which are numbers of discrete sources which vary in size from seconds of arc upwards. The background is not resolvable into discrete sources. It has a maximum value in the plane of the Milky Way and towards the centre of the Galaxy, and is thought to originate largely in the interstellar gas.

(ii) There is a continuous spectrum of radiation over the whole observable range of wavelengths, although the intensity decreases with wavelength. Line emission of monatomic hydrogen at 1,421 Mc/s has been observed, and there are other spectral lines (notably those of deuterium at 327 Mc/s and the hydroxyl radical OH at 1,666 Mc/s) which may be observable.

(iii) Except for the Sun and Jupiter, no intrinsic variations of intensity with time have been found for any source. There are variations due to the ionosphere, notably at frequencies near the critical frequency and also fluctuations (analogous to the "twinkling" of stars) of the intensity and direction of sources due to irregularities in the F-layer.

(iv) Again excluding the Sun and Jupiter, the radiation is in general randomly polarized. Some plane polarization has been detected, and it is important for theoretical reasons to investigate this.

(v) The sun is a source of particular interest, since it is the most accessible star. The solar corona emits thermal radio radiation, and the variation of the shape of the radio sun with wavelength gives information about electron density and temperature in the corona.

It is a highly variable source at times when there are sunspots on the disk. The Sun is then said to be active. Circularly polarized radiation has been observed at these times.

Current lines of work in radio astronomy include.

(i) Classification and correlation of the radio emission associated with solar activity.

(ii) Surveys of the distribution of the background radiation over the sky.

(iii) Measurements of the positions, intensities and angular sizes of discrete sources.

(iv) Study of the 1,421-Mc/s hydrogen line, involving measurement of the intensity and Döppler shift of radiation in different directions.

The basic equipment required for radio astronomy is evidently a highly directive aerial and a low-noise receiver. In addition, there are such special requirements as wide-band spectrometers at metre wavelengths, narrow-band spectrometers at 1,421 Mc/s, polarimeters, etc.

General Principles

Consider the energy from a discrete source ΔE between frequencies f and $f + \Delta f$ falling on an area ΔA in a time Δt ($\gg 1/(\Delta f)$). The flux density S of the source at frequency f is then defined as:

$$S = \frac{\Delta E}{\Delta f \,.\, \Delta A \,.\, \Delta t}$$

The usual units are those of the M.K.S. system, i.e., watts per square metre per cycle per second (watts . m.$^{-2}$. (c/s)$^{-1}$).

Typical flux densities at $f = 100$ Mc/s are $S = 10^{-22}$–10^{-25} watts . m.$^{-2}$ (c/s)$^{-1}$ for discrete sources, though when active the Sun can reach 10^{-19} watts . m.$^{-2}$. (c/s)$^{-1}$.

For an extended source (or the background) we can specify the energy coming from a particular direction (θ, ϕ). Suppose that from an area subtending a solid angle $\Delta\omega$ the flux density is ΔS, then the brightness $b(\theta, \phi)$ at frequency f in that direction is

$$b(\theta, \phi) = \lim_{\Delta\omega \to 0} \frac{\Delta S}{\Delta\omega}$$

The units of $b(\theta, \phi)$ are watts per square metre per (cycle per second) per steradian (watts . m.$^{-2}$. (c/s)$^{-1}$. ster^{-1}). It is evident that the flux density of a source is the integral of the brightness due to the source over the solid angle subtended by it, i.e.,

$$S = \int b(\theta, \phi) d\omega$$

For any quoted value of these quantities the polarization must be specified. If the polarization is random, then the total flux is twice that received by a linearly polarized aerial.

The brightness is often given as a "brightness temperature", T_b, defined as the temperature of a "black body" which would give the same flux as that observed. This is convenient, since it may be shown that a matched aerial situated in an enclosure of constant temperature T delivers a thermal noise power P per unit band-width equal to that available from a resistance at temperature T, i.e.,

$$P = kT$$

where P is the noise power per unit band-width; k is Boltzmann's constant $= 1{\cdot}38 \times 10^{-23}$ joules per ° K.; T is the absolute temperature. The receivers used in radio astronomy may be calibrated in terms of the thermal noise power from a resistor at the input, and in this case it is possible to proceed (without knowledge of the band-width) from the output signal to the "effective aerial temperature" (see later), and hence the sky brightness.

The relation between b and T_b is given by Planck's Law

$$b = \frac{2hf^3}{c^2} \cdot \frac{1}{\exp(hf/kT) - 1}$$

where h is Planck's constant $= 6{\cdot}623 \times 10^{-34}$ joules secs. and c is the velocity of light $= 2{\cdot}9972 \times 10^8$ m.sec.$^{-1}$.

At radio wavelengths $hf \ll kT$ and Planck's Law reduces to the Rayleigh–Jeans approximation

$$b = \frac{2kT_b f^2}{c^2} = \frac{2kT_b}{\lambda^2}$$

where λ is the wavelength.

If a source in a direction (θ, ϕ) has a flux density S we can define an "effective area" of an aerial as

$$A(\theta, \phi) = \frac{2P}{S}$$

where P is the power absorbed per unit band-width. (The factor 2 assumes random polarization.)

For the general case

$$P = \tfrac{1}{2}\int A(\theta, \phi) \, . \, b(\theta, \phi) d\omega$$

Let the maximum value of $A(\theta, \phi)$ be A_0 and the mean be

$$\bar{A} = \frac{1}{4\pi}\int A(\theta, \phi) \, . \, d\omega$$

The "directivity" $D(\theta, \phi)$ of the aerial $= \dfrac{A(\theta, \phi)}{\bar{A}}$

and $$D_0 = \frac{A_0}{\bar{A}}$$

It may be shown by a thermodynamic argument that

$$A = \lambda^2/4\pi$$
$$\therefore \quad A_o = D_o\lambda^2/4\pi$$

The effective solid angle of reception is given by Ω_o (say) where

$$\Omega_o = 4\pi/D_o = \lambda^2/A_o$$

and the total number of beam-widths in the sky is

$$1/\Omega_o = D_o/4\pi$$

The "effective aerial temperature" T_a is given by

$$T_a = \frac{1}{4\pi}\int D(\theta, \phi) \,.\, T_b(\theta, \phi) d\omega$$

We note that if $T_b(\theta, \phi)$ is a constant T_b then

$$T_a = T_b \text{ as would be expected,}$$

also for a small source situated in the direction such that $D(\theta, \phi) = D_o$, and having a temperature T and subtending a solid angle $\Omega(\ll \Omega_o)$

$$T_a = \Omega D_o T/4\pi = T \,.\, \Omega/\Omega_o$$

If the aerial is connected to the receiver by a feeder of temperature T_o and fractional power loss α, then the effective temperature at the receiver input is T_e where

$$T_e = \alpha T_o + (1 - \alpha) T_a$$

(i.e., T_e is the temperature of a matched resistor giving the same noise power).

Aerials

Aerials [1, 2] that have been used in radio astronomy include:

(i) *Half-wave Dipole.* This is not sufficiently directive by itself (directivity = 1·6), but end-fire arrays such as the Yagi can have

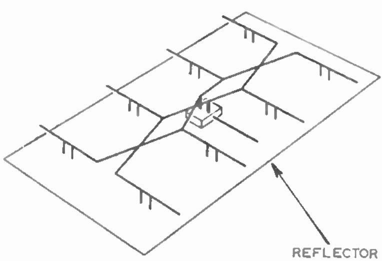


FIG. 1.—A BROADSIDE ARRAY.

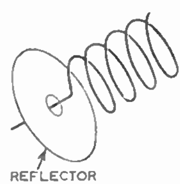


FIG. 2.—A HELICAL AERIAL.

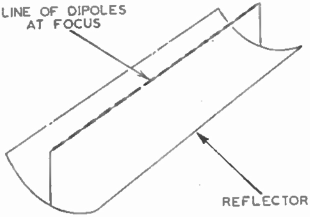
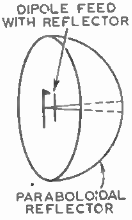


FIG. 3 (*left*).—A PARABOLIC TROUGH AERIAL.

FIG. 4 (*right*).—A PARABOLIC REFLECTOR WITH DIPOLE FEED AND REFLECTOR.



directivities ≃10. Yagis are light, but narrow-band and difficult to adjust. A dipole over a reflecting surface has a directivity of 3·2.

(ii) *Broadside Arrays of Dipoles.* See Fig. 1. A number of dipoles mounted above a reflector form a useful aerial at metre wavelengths. For a small number of dipoles n, the directivity is $\simeq 3{\cdot}2\,n$. For a large array the effective area is close to the physical area. Disadvantages are the large number of connectors and the difficulty of changing frequency.

(iii) *Helical Aerials.*[3] See Fig. 2. Helical aerials detect circularly polarized radiation incident along the axis of the helix. They may nevertheless be used for randomly polarized radiation, since this can be considered as made up of two components rotating in opposite directions. Helical aerials are wide-band (≃2 : 1 in frequency).

(iv) *Rhombic Aerials.* These are also wide-band, but the directivity is not high.

(v) *Horns.*[4, 5] The horn is cumbersome at long wavelengths, but much used as a feed at the focus of paraboloidal reflectors used at centimetric wavelengths. The horn is valuable for absolute measurements of flux density, since the directivity may be calculated theoretically to within an accuracy of a few per cent.

(vi) *Parabolic Reflectors.* These may take the form of: (*a*) a parabolic trough Fig. 3 with a string of dipoles along the line focus, or more commonly (*b*) a paraboloid of revolution with either a dipole and reflector or a horn feed at the focus (Fig. 4). The great majority of radio-astronomy aerials for high-frequency work are

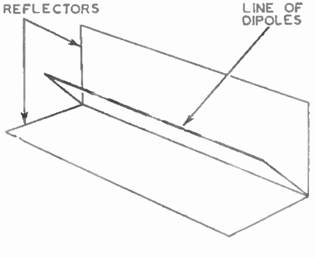


FIG. 5.—A CORNER REFLECTOR.

of this latter type. The wavelength of operation may be easily altered by changing a single feed. A disadvantage is that the effective area is only about 50 per cent of the actual area. Quite apart from the physical movement of the reflecting surface, the beam may be moved up to about three times its own width by displacement of the feed point.

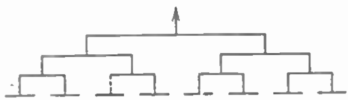
(vii) *Corner reflectors*. See Fig. 5. These are not very directive, but are cheap and simple to make.

(viii) *Slotted Wave-guides*. These have been used to give a fan-shaped reception pattern at short wavelengths.

Feeders

The requirements for feeders in radio astronomy are low loss and phase stability. At centimetric wavelengths waveguides are used, and at the longer wavelengths usually transmission lines. Flexible co-axial cables with polythene insulation are cheaper, but have more loss than air-spaced co-axial lines. Twin-wire lines have very low loss, but the electrical length alters when they get wet, so that arrays using such feeders may be unserviceable in conditions of rain, dew and frost (see Sections 12 and 13). We have seen that the effective temperature at the input of the receiver depends on the temperature T_e of the feeder. It may therefore be necessary to provide feeders with thermal insulation (e.g., by burying them) to maintain T_e constant. If a number of aerial elements are to be connected together it is often desirable to make the electrical path from each the same as in Fig. 6. This ensures that the

FIG. 6.—EQUALISATION OF FEEDER PATHS.



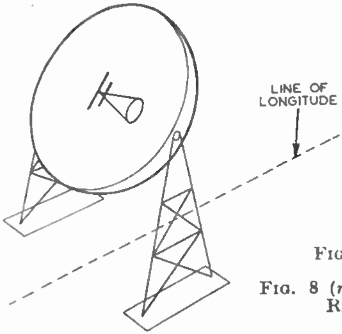
band-width is not unnecessarily limited by phase changes over a wide band.

Mountings

There are three types of mountings:

(i) *Meridian*. In this (Fig. 7) the aerial can be directed to different positions along the meridian but not away from it. A strip of sky is then scanned once in a sidereal day ($23^h\ 56^m$) by means of the rotation of the earth. The altitude of the beam may be altered by: (*a*) physically tilting the array about an East–West axis, or (*b*) altering the electrical path between the elements of the array (Fig. 8). There are examples of such aerials at Cambridge [6] and Sydney [7] among other places.

(ii) *Equatorial*. In this type the aerial can be turned to any direction, one of the rotation axes being parallel to the earth's axis.



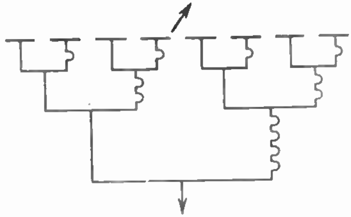


Fig. 7 (*left*).—Meridian Mounting.

Fig. 8 (*right*).—Beam Swinging by Altering Relative Phases of Elements.

It is then particularly easy to track astronomical bodies. All optical telescopes are mounted in this way. Fig. 9 shows a Yagi aerial with an equatorial mount. Many large (up to 140-ft.-diameter) paraboloidal reflectors are so mounted[8]; but for the very largest sizes mechanical difficulties become prohibitive.

(iii) *Altazimuth.* As the name indicates, this mount enables both the altitude and azimuth of the aerial to be varied (see Fig. 10). The 250-ft.-diameter reflector at Jodrell Bank is of this type. A co-ordinate converter is required with this type of mount.

Receivers

Since the function of radio-astronomy receivers is to measure temperatures, they are often called " radiometers ". If a source produces an equivalent input temperature T_e and the receiver noise temperature

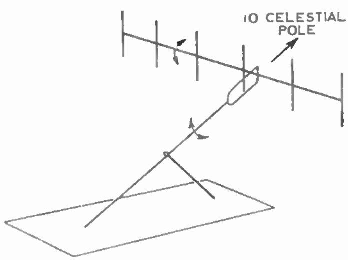


Fig. 9.—Yagi Array in Equatorial Mounting.

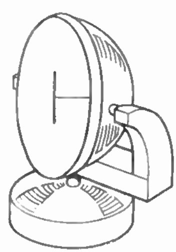


Fig. 10.—Altazimuth Mounting.

is $T_R = (N - 1)T_o$, where N is the noise factor and T_o is commonly 290° K., then it may be shown that the source is detectable if

$$T_s > T_R(\tau\Delta f)^{-\frac{1}{2}}$$

where τ is the time-constant of the output circuits, and Δf is the receiver band-width.

It is therefore advantageous:

(i) To increase Δf. Such increase may be limited by the desire to study the variation of S with f or by the necessity of avoiding interfering signals.

(ii) To increase τ. Such increase is usually limited by the need to follow time variations of the signal. Some visual averaging of the records is feasible, and there is also the possibility of integrating numbers of independent records. Typical values are $\Delta f = 1$ Mc/s, $\tau = 1$ second, i.e.,

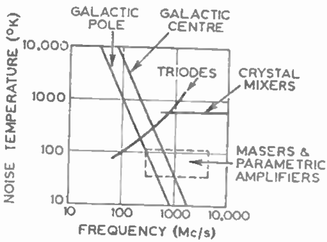
$$(\tau\Delta f)^{-\frac{1}{2}} = 10^{-3}$$

(iii) To reduce T_R. T_R is mainly determined by the input stages. Until recently the best noise factors have been obtained by the use of thermionic triodes used in grounded-grid or cascode circuits at the lower frequencies and crystal mixers at higher frequencies. The best noise temperature for triodes increases approximately linearly with frequency from 30° K. at 10 Mc/s to 1,000° K. at 1,000 Mc/s. For crystal mixers the noise temperature remains at about 600° K. from 400 Mc/s to 10 kMc/s and then increases to 10,000° K. at 100 kMc/s.

New types of low-noise receivers have now been devised. One is the MASER (Microwave Amplification by Stimulated Emission of Radiation), a solid-state amplifier working on quantum principles. In this a crystal of a rare-earth salt absorbs energy from an oscillator at a high frequency and is stimulated by the signal to emit the energy at the signal frequency. The frequency is controlled by an external magnetic field. The crystal works best when cooled to liquid-helium temperatures and the noise temperature is correspondingly low, e.g., 50° K. at 1,000 Mc/s.

The other is the parametric amplifier, which achieves amplification by the periodic variation of a circuit parameter. This may be, for example, a crystal diode forming part of the capacity in a tuned circuit. This is biased in the reverse direction by a variable voltage provided by

FIG. 11.—NOISE TEMPERATURES IN RELATION TO FREQUENCY.



an oscillator known as the "pump". The net effect is that energy is transferred from the pump oscillator to the signal. An alternative type utilizes a modulated electron beam, and has a band-width of 80 Mc/s and a gain of 30 db. For frequencies up to 1,000 Mc/s noise temperatures of less than 100° K. have been obtained. With such low noise temperatures the losses in feeders and the contribution to the aerial temperature of those side lobes of the reception pattern which are directed towards the ground become of paramount importance.

Further information on masers and parametric amplifiers is given in Section 23.

These various noise temperatures are illustrated in Fig. 11 as a function of frequency. Also shown are brightness temperatures of the galactic background which are greater than the noise temperatures of triodes at frequencies below about 300 Mc/s.

Types of Radiometer

In addition to the random fluctuations of receiver noise there are variations in gain and T_R due to such causes as change of mains voltage, etc. A 0·1 per cent change in gain would produce a change in output equal to the signal in the example previously considered. Even if voltage stabilization is provided, present techniques are not capable of maintaining the gain and T_R sufficiently constant. Radiometers are therefore designed to reduce or eliminate the effect of such changes:

(i) *Dicke Radiometer.*[9] See Fig. 12. In this the receiver, a superhet, as are the great majority of receivers used for radio

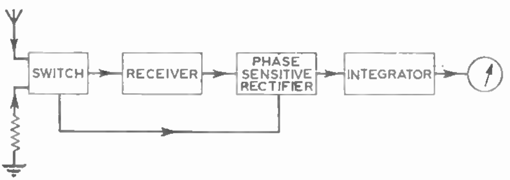


Fig. 12.—The Dicke Receiver.

astronomy, is switched (at say 30 c/s) from the aerial to a load of the same impedance and at a known temperature. The amplitude of the square wave output is nearly independent of T_R, though still linearly proportional to the gain.

(ii) *Noise-compensated Comparison Radiometer.* The gain dependence of the Dicke radiometer may be largely removed by altering the temperature of the load until it approximates to the aerial temperature. In the case of large aerial temperatures it is more convenient to use as a reference the shot noise produced by a temperature-limited diode. If the diode current is I, then the effective temperature of a resistance R connected in series with the diode is:

$$T_o + \frac{eI}{2k} \cdot R$$

where e is the electronic charge, and k is Boltzmann's constant.

Such a noise diode (e.g., CV2171) is also used for calibrating radiometers. Diodes may be used up to frequencies of about 1,000 Mc/s and can produce effective temperature of about 10,000° K. They should not be treated as absolute standards but calibrated against thermal loads.

At high frequencies an argon-filled discharge tube (CV1881) may be inserted into a waveguide and used as a noise source with an effective temperature of about 11,000° K.[10] For high frequencies at which the aerial temperature is less than ambient it may be convenient to maintain the comparison load at room temperature and to feed noise in to the aerial line by means of a directional coupler. This will result in an increase of the total noise, but it is not likely to be important. A radiometer utilizing this principle at 8,000 Mc/s has been described by Drake and Ewen (*Proc. I.R.E.*, Vol. 46, 53, 1958) and is illustrated in Fig. 13. A notable feature about this receiver is the use of three

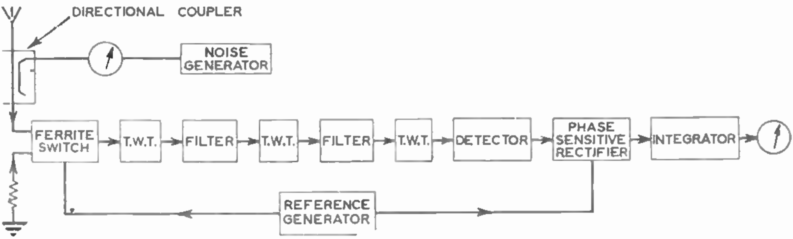


FIG. 13.—DRAKE AND EWEN TRAVELLING-WAVE TUBE RECEIVER WITH "NOISE COMPENSATION".

travelling-wave tubes in series so that a band-width of 1,000 Mc/s is possible. With an output time-constant of 100 seconds the minimum T_e detectable is 0·01° K.

(iii) *Servo-controlled comparison radiometer.*[11] In this radiometer (see Fig. 14) a servo-control adjusts the noise diode current so as to maintain the noise output equal to that from the aerial. The diode current is then a measure of the aerial temperature. The receiver acts only as a null detector, so the output is independent of changes of gain or receiver noise.

Interferometers

One of the chief difficulties in radio astronomy is that of attaining sufficient resolving power to be able to measure positions of radio sources with accuracy. For a paraboloidal reflector of diameter a, the beam-width β is approximately λ/a radians and if $\lambda = 1$ m. and $a = 60$ metres, then $\beta \simeq 1°$. With $\beta = 1°$ it may be possible to locate a discrete source to within about 10′arc, but in order to make identifications with optical objects certain, the positions must be measured to within about 1′arc. A reflector with diameter of 600 m. is impracticable. The resolving power increases at higher frequencies, but the flux

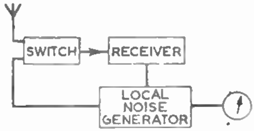


FIG. 14.—MACHIN, RYLE AND VONBERG SERVO-CONTROLLED RADIOMETER.

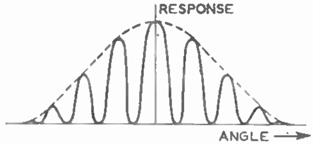


FIG. 15.—THE DIRECTIVITY OF A TWIN-AERIAL INTERFEROMETER.

density of sources falls and the receiver noise rises, so that there is an optimum frequency for the observation of radio sources, in the range 400–1,000 Mc/s, depending on the techniques used.

Radio interferometers consisting of two or more spaced aerials connected to the same receiver offer a solution to this problem. The directivity of such an arrangement is shown in Fig. 15. Consider in Fig. 16 two aerials each of effective area $A(\theta)$ and separated by a distance l. If there is a source such that the flux density from a strip bounded by directions θ to $\theta + d\theta$ is $P(\theta)d\theta$, then the total power from the source intercepted by the aerials is:

$$\int P(\theta) \,.\, A(\theta)[1 + \cos(2\pi l \sin \theta/\lambda)]d\theta$$

where the integration is taken over the angular size of the source. This becomes:

$$\int P(\theta) \,.\, A(\theta)d\theta + \int P(\theta) \,.\, A(\theta) \cos(2\pi l \sin \theta/\lambda)d\theta$$

The first term is the power from one aerial alone. As the earth rotates and θ varies the second term varies sinusoidally so that the radiometer output, which is the algebraic sum of the two terms, changes with time as shown in Fig. 17. If $A(\theta)$ can be considered as constant

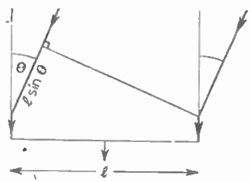


FIG. 16.—SPACED AERIALS INTERFEROMETER.

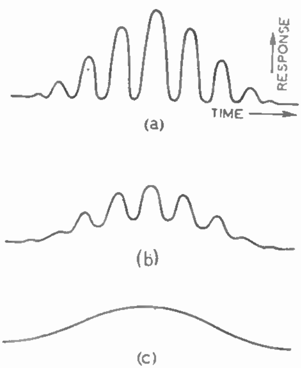


FIG. 17 (*right*).—THE RESPONSE OF A SIMPLE INTERFEROMETER TO SOURCES OF DIFFERENT ANGULAR SIZES.

over the source, then it may be placed outside the integral sign and for small θ:

$$\int P(\theta)\cos(2\pi l \sin\theta/\lambda)d\theta = \int P(\theta)\cos(2\pi\theta l/\lambda)d\theta$$

which is the component of angular frequency l/λ in the cosine Fourier Transform of the strip distribution of intensity over the source. For a point source there is an equal component for any spacing l, the second term equals the first, and the output is shown in Fig. 17 (*a*). For a source of angular size comparable with l/λ, the high-frequency components are reduced and the output may be like that shown in Fig. 17 (*b*). For a source large compared with l/λ (but small compared with the beam-width λ/a) the component of angular frequency l/λ approaches zero and the output is shown in Fig. 17 (*c*). The exact shape of the output clearly gives a measure of the angular diameter of the source. The quasi-sinusoidal traces produced by an interferometer are often described as interference fringes by analogy with Young's double-slit interferometer in optics. The angular size of the radio source corresponds exactly with the size of the source slit in Young's interferometer.

It is evident that a source of small angular size may be located more accurately with respect to the interference fringes than it may with respect to the envelope of the reception pattern. It is not possible to increase the accuracy without limit by increasing l, because of uncertainty as to which of the interference fringes is the central one.

The total strip distribution of intensity across the source may be determined by measuring the amplitudes of the sinusoidal term above as l, the aerial separation, is altered and then carrying out a Fourier synthesis.

The Phase-switching Receiver

For the detection of radio sources the principle of "phase-switching"[12] has considerable advantages over the straight radiometer. Two aerials are used as an interferometer, but a switch is arranged so that the aerials are connected in phase and in anti-phase at a frequency of say 400 c/s. A block diagram of the circuit is shown in Fig. 18. The reception pattern when the aerials are connected in phase is shown by the full line in Fig. 19 (*a*) and by the dotted line when connected in anti-phase. The receiver measures the difference between the signals received in these two cases, Fig. 19 (*b*). The record is seen to have a central zero. As in the case above, the amplitude of the quasi-sinusoidal fluctuation is proportional to the component of angular frequency l/λ in the cosine Fourier Transform of the strip distribution of intensity over the source. An alternative way of thinking of this is that the output measures the cross-correlation between the signals arriving at the two aerials. By comparison of the amplitudes at different spacings the angular distribution of intensity over the source may be obtained. For the measurement of very small angular sizes, large spacings, e.g., 20–40 km., may be necessary, and this entails the use of radio links.

Advantages of the phase-switching technique are:

(1) Angular variations in intensity having large sizes (such as the galactic background) cause a steady signal in the receiver and are eliminated from the record. This enables a high sensitivity to be used.

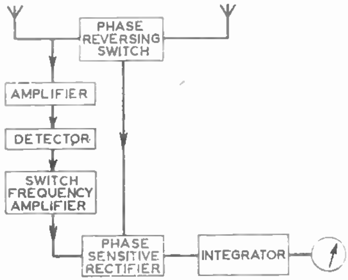


FIG. 18.—THE RYLE PHASE-SWITCHING RECEIVER.

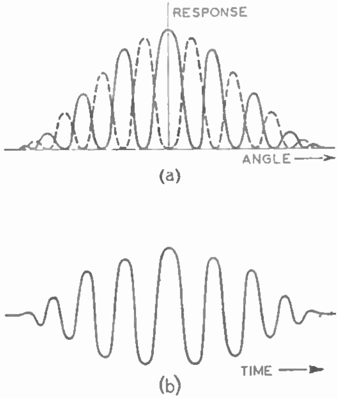


FIG. 19 (*right*).—(*a*) RECEPTION PATTERNS OF AN INTERFEROMETER WITH THE AERIALS CONNECTED IN AND OUT OF PHASE. (*b*) THE RESPONSE OF A PHASE-SWITCHING RECEIVER.

(2) Effects due to changes of gain may be reduced by the employment of automatic gain control. The gain is then determined by the galactic background noise at low frequencies and the input circuit noise at higher frequencies.

(3) The times at which the trace crosses the zero line may be very accurately determined and used to give an accurate position.

(4) Interference arriving at one aerial alone is not recorded.

If the directivities of the two aerials are D_1 and D_2, then the effective directivity of the system is: $2\sqrt{D_1D_2}$.

By continuously changing the relative phase of the two aerials the interference fringes of the reception pattern may be swept across the sky and the periodicity of the fringes in Fig. 19 (*b*) increased or reduced. This technique is known as swept-lobe interferometry.

The Mills Cross

Two aerials (Fig. 20), one extended East–West and one North–South, form the two elements of an interferometer.[7] The first receives radiation from a narrow strip of sky running North–South and the other from a strip running East–West. When they are connected to a phase-switching receiver, the output measures the intensity of the region of sky common to the two beams, i.e., the small patch at the point of intersection of the two strips (Fig. 21). The effective directivity therefore corresponds to that of a conventional paraboloidal reflector with a diameter approximately equal to the total length of the arms. A practical case is the " Mills Cross " at Sydney, in which the arms are 1,500 ft. long and at 85 Mc/s a beam-width of 50′arc is obtained. The signal strength is not as great as that of the equivalent reflector, since the area of ground covered is less, i.e., it is an " unfilled " aperture.

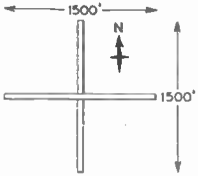


FIG. 20.—PLAN OF 85 Mc/s "MILLS CROSS". EACH ARM CONSISTS OF 250 PAIRS OF DIPOLES ALIGNED EAST–WEST ABOVE A REFLECTOR.

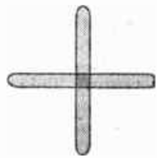


FIG. 21 (*right*).—THE TWO FAN BEAMS OF A "MILLS CROSS" AERIAL. THE PHASE SWITCHING RECEIVER RESPONDS ONLY TO THE DIAGONALLY-HATCHED REGION OF INTERSECTION.

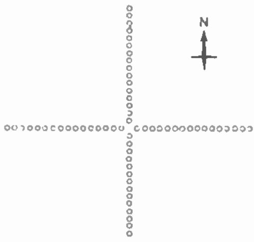
At low frequencies this may be no handicap. The altitude of the beam is altered by introducing progressive phase shifts as previously described.

Multiple-aerial Interferometry

Many of the methods used in optical interferometry are directly applicable to radio astronomy. An analogue of the diffraction grating has been designed by Christiansen (*Proc. I.R.E.*, Vol. 46, 127, 1958). He has constructed an array of thirty-two paraboloids, each of diameter 19 ft. and separated by 40 ft., along an East–West line crossed by a similar array along a North–South line (Fig. 22). The directivity of the East–West line is illustrated in Fig. 23, and when this is combined with the North–South line in a phase-switching receiver a directivity as in Fig. 24 is obtained. This instrument is designed for a frequency of 1,421 Mc/s and is used for observations of the Sun.

Aperture Synthesis

In this technique [13, 14] a series of observations, made with a pair of aerials in different positions, is combined to give results equivalent to those obtainable from a much larger aerial. It is therefore known as "aperture synthesis". Consider the operation of an aerial such as a large broadside array of dipoles or a paraboloidal reflector. The signal produced by the aerial may be regarded as the vector sum of the currents induced in each part of the aerial. In the case of a paraboloid this vector addition is achieved, automatically, at the focal point, while for an array the same result is reached by connecting all the



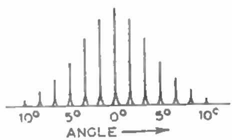


FIG. 22 (*left*).—PLAN OF A CROSSED MULTI-ELEMENT INTERFEROMETER (*after* Christiansen, *Proc. I.R.E.*, Vol. 46, 127, 1958.)

FIG. 23 (*right*).—DIRECTIVITY OF ONE ARM OF THE CHRISTIANSEN MULTI-ELEMENT INTERFEROMETER.

elements to the receiver in the same phase. If it were possible to measure the current induced at each portion of the aperture by moving a small aerial across it, vector addition of the currents would give, for a constant source, the same result as obtained by using the complete aperture. This process is not possible using a single receiving element, since it is not possible to determine the phase. It is possible, however, using two small aerials as an interferometer.

Consider a rectangular array as shown in Fig. 25. The current

FIG. 24 (*left*).—THE LOBES OF THE CHRISTIANSEN INTERFEROMETER PROJECTED ON TO THE CELESTIAL SPHERE. THE RECEIVER RESPONDS ONLY TO SOURCES AT THE POINTS OF INTERSECTION.

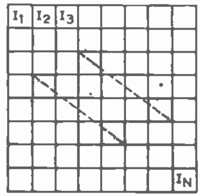


FIG. 25 (*right*).—PRINCIPLE OF APERTURE SYNTHESIS. TWO PAIRS OF SMALL AERIALS IN THE POSITIONS SHOWN ARE EXACTLY EQUIVALENT.

induced in the nth element of the array is $I_n e^{i\phi_n}$ (say), and the power P delivered to a receiver is:

$$P \propto \Sigma I_n e^{i\phi_n} \,.\, \Sigma I_n e^{-i\phi_n}$$
$$\propto \Sigma I_n^2 + \Sigma I_m I_n \cos(\phi_m = \phi_n)$$

If the N elements are all of the same size the first term is simply N times the power derived from a single element. The second term is the output obtained from a phase-switching receiver connected to the elements m and n. A single measurement of the power induced in one element suitably combined with the outputs from a phase-switching receiver connected to a pair of elements arranged successively to cover all possible combinations of positions then gives a result exactly equivalent to the use of the whole aperture. Many of the combinations of positions are equivalent, so it is not necessary to repeat them. By performing the vector addition with suitable phase changes it is possible to produce the effect of swinging the synthesized beam.

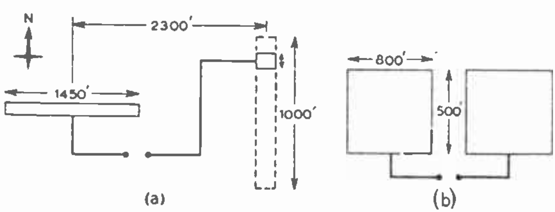


FIG. 26.—(*a*) PLAN OF THE 178 Mc/s RADIO STAR INTERFEROMETER EMPLOYING APERTURE SYNTHESIS (*after* Ryle. *Nature*, Vol. 180, 110, 1957).

(*b*) EQUIVALENT APERTURE.

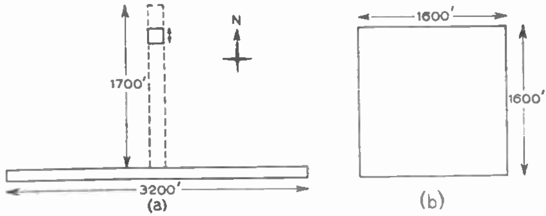


FIG. 27.—PLAN OF 38 Mc/s AERIAL PRODUCING ·1° "PENCIL BEAM" BY APERTURE SYNTHESIS (*after* Ryle, *Nature*, Vol. 180, 110, 1957).

(*b*) EQUIVALENT APERTURE.

The advantages of this technique are increased flexibility and economy, since the aerials are smaller than would otherwise be required. The time required to survey a given area of sky is of the same order of magnitude as that using the complete aperture, but the signal-to-noise ratio is, of course, worse. It is convenient to perform the vector additions by means of an electronic computor.

The principle of aperture synthesis has been applied at Cambridge to the construction of an interferometer (Fig. 26) with a beam 25′arc × 35′arc at 178 Mc/s, and an aerial (Fig. 27) giving a 1-degree pencil beam at 38 Mc/s. This is equivalent to a paraboloidal reflector of diameter 1,500 ft.

Polarimeters

Linear polarizations may, in principle, be detected by rotating an array about an axis directed towards the source and noting the variation of output.[15] A more sensitive method is to use two mutually perpendicular dipoles at the focus of a paraboloidal reflector which are connected to a phase-switching receiver. Any plane-polarized radiation incident on the aerial will, in general, give an output, whereas randomly polarized radiation gives none. The direction of polarization may be found by rotation of the dipole assembly. It should be noted that spurious signals may be produced in this method by intense sources off the centre of the beam.

Circularly polarized radiation may be detected by two mutually perpendicular dipoles with a quarter-wavelength of path between them, which are connected to a phase-switching receiver.

Spectrometers for V.H.F.

The bursts of radio radiation from the Sun have been found to have a limited band-width, the centre frequency of which varies with time. To investigate this phenomenon metre-wavelength spectrometers are required. Fig. 28 illustrates a circuit that has been described by Goodman and Lebenbaum (*Proc. I.R.E.*, Vol. 46, 132, 1958). Three such receivers are used to cover the range of frequencies from 90 to 580 Mc/s three times a second.

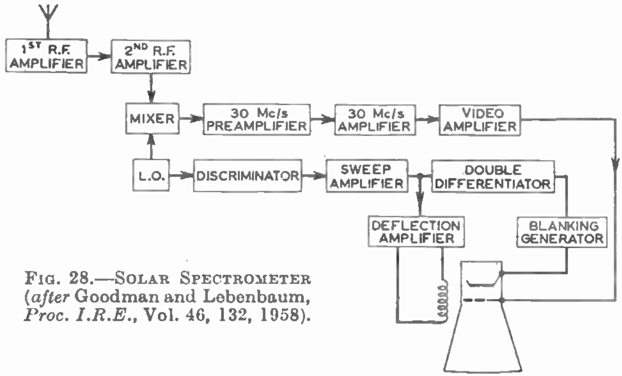


FIG. 28.—SOLAR SPECTROMETER (*after* Goodman and Lebenbaum, *Proc. I.R.E.*, Vol. 46, 132, 1958).

Hydrogen-line Radiometer

The requirement of hydrogen-line radiometers is the measurement of the intensity as a function of frequency in a band about 1 Mc/s wide centred on 1,421 Mc/s. The hydrogen may appear either in absorption or in emission, depending on its temperature compared with the brightness temperature of the radiation from beyond the hydrogen (Kirchhoff's principle). The block-diagram of a suitable receiver that has been described by Lilley and McClain (*Proc. I.R.E.*, Vol. 46, 221, 1958) is illustrated in Fig. 29. Two pass bands are arranged so that one having a width of 2 Mc/s is used as a reference against which to compare a second narrow band which is tuned across the hydrogen line.

Observational Techniques

(i) *Measurement of Position.* This depends, in general, on knowledge of the exact position of the aerial beam when the source is producing

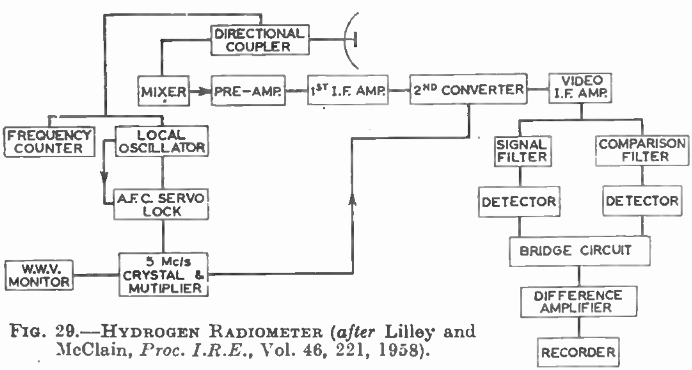


FIG. 29.—HYDROGEN RADIOMETER (*after* Lilley and McClain, *Proc. I.R.E.*, Vol. 46, 221, 1958).

a maximum output. Any collimation errors of the aerial may be determined by observation of intense sources whose positions are already known from optical studies.

The right ascension of sources is often found from the time of transit across the meridian through the aerial. In the case of the twin-aerial interferometer in which the aerials are lined up in an East–West direction the periodicity, t, of the fluctuations of the output gives a measure of the declination δ of the source in question by means of the relationship

$$t = t_0 \sec \delta$$

where t_0 is the periodicity for a source at declination 0°. This may be calculated from the spacing of the aerials and the wavelength.

(ii) *Measurement of Flux Density and Brightness.* This depends on: (*a*) knowledge of the effective area of the aerial which can be calculated in the case of horns or broadside arrays,[16] and (*b*) calibration of the radiometer by means of a thermal load.

(iii) *Measurement of Brightness Distribution.* This has already been considered in the section on interferometers and the phase-switching receiver.

J. R. S.

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SPACE-PROBE COMMUNICATION PROBLEMS

The radio signals to be received from a space probe are of very low power, and a telemetry system can function only if this signal exceeds the radio noise providing spurious signals in the receiver. Fig. 30 shows how the desired signal from a space probe, indicated by the

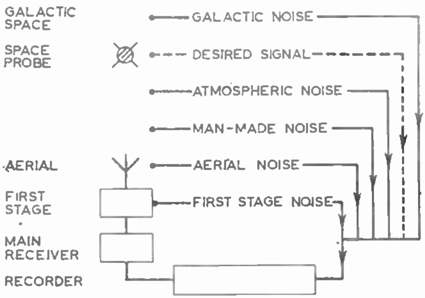


FIG. 30.—SOURCES OF NOISE COMPETING WITH THE DESIRED SIGNAL.

dotted line entering the recorder, competes with noise from galactic space, from atmospherics, man-made noise and that arising in the aerial and the first stage of the receiver. To communicate with a space probe, it is necessary to study these forms of noise, to assess their importance in any system and to decide what is the best working frequency.

Galactic Noise

If a directive aerial is turned towards the sky, then a random signal will be received, and this noise will be great in the direction of the sun and various " radio stars " which stand out from a general background of noise. This background, which cannot be assigned to specific point

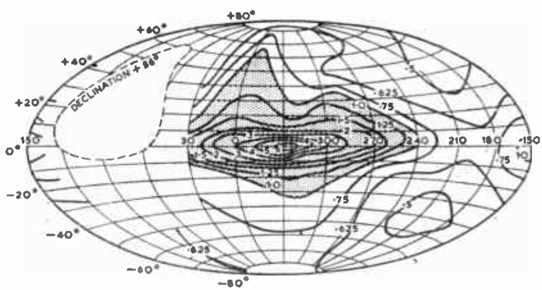


FIG. 31.—DISTRIBUTION OF RADIO BRIGHTNESS OVER THE SKY AT 100 Mc/s. Galactic Co-ordinates. Observations corrected for aerial smoothing. Unit 1000 degrees.

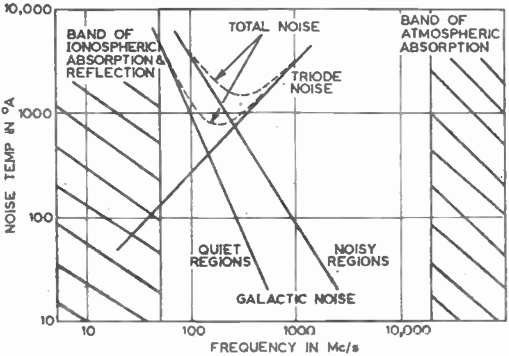


FIG. 32.—DIAGRAM ILLUSTRATING THE CHOICE OF OPTIMUM FREQUENCY.

sources, is known as galactic noise. The distribution across the sky is rather complex, and its amplitude decreases rapidly with increase of frequency. The distribution and amplitude have been studied in detail by H. C. Ko.[1] Fig. 31 shows the distribution of galactic noise plotted in galactic co-ordinates, as measured by Bolton & Westfold.[2] From this it will be seen that the noise is greatest looking towards the centre of the galaxy, while in directions well away from the centre there are quiet galactic regions. The variation of galactic noise with frequency is shown in Fig. 32, where two curves based on data published by Ko indicate the approximate limits for noisy and quiet directions. These are drawn as straight lines, but towards the higher frequencies they are somewhat unreliable because there the galactic noise tends to be too small to measure accurately in the presence of other sources of noise.

Atmospheric Noise

Atmospheric noise arises from lightning flashes and other electric discharges in the atmosphere. At low frequencies the "atmospherics" travel long distances around the earth trapped between the earth's surface and the ionosphere. At higher frequencies the ionosphere does not reflect the atmospherics, and so the noise is not transmitted over great distances. Since, if signals from space are to enter the atmosphere, the system must operate at frequencies above ionospheric cut-off; then only atmospherics from local storms will be received, and provided that the receiver is sited well away from the tropic storm regions, atmospheric noise will be negligible.

Man-made Noise

This noise arises from sparking brushes in electrical machinery, from car ignition systems, and the malfunction of diathermy or other radio

[1] *Proc. I.R.E.*, Vol. 46, No. 1 pp. 208–15 (January 1958).
[2] "Radio Astronomy", Pawsey and Bracewell.

equipment. If a receiving site is chosen well away from centres of industry, the man-made noise will not be important compared with that from other sources. In order to reduce man-made noise to a minimum the site can be chosen in a secluded valley screened from outside interference by high hills. Such siting has been chosen in the past by radio astronomers.

Aerial Noise

Under this heading we consider the contributions to the noise input to the receiver due to current flow through the resistance of the aerial and transmission line. At V.H.F. and U.H.F. this noise is so small compared with triode noise that it can be neglected. In the future, when micro-wave "maser" or parametric receivers are developed having noise temperatures very much lower than present-day triodes, this noise will probably be of importance.

First-stage Noise

At the present time, triodes give the lowest noise level, although it is likely that masers and other similar amplifiers will be developed having much lower noise levels. However, considering current practice, we must base our assessment on the performance of present-day triodes.

It is usual to express a noise level as an equivalent noise temperature, T_n. In the case of the first-stage noise this is the temperature at which a resistance R equal to the input resistance of the triode would develop a thermal noise equal to that of the first stage. In terms of V_n, the noise voltage developed across R, this noise is given by the equation

$$V_n^2 = 4\,k\,T_n\,R\,df \qquad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

in which k = Boltzman's constant, $1{\cdot}37\ 10^{-23}$ watt sec./° C., and

df = the frequency band-width of the receiver measured in c/s.

The variation of T_n with frequency for a good-quality triode is shown in Fig. 32, from which it will be seen that the noise level increases rapidly with increase in frequency.

It may be noted that the levels expressed as noise temperature are easily related to the "noise factors" which are sometimes used to assess the performance of a first stage of a receiver. This is done by taking the noise power for $T_n = 300°$ A. as unit noise power and expressing the noise power in terms of this, using a db scale; thus a noise factor of 3 db is equivalent to a noise temperature of 600° A.

Ground Aerials

The performance of the receiving aerial depends upon its effective area, and not on its gain. The larger the solid angle subtended by this area at the probe, the larger the fraction of the total transmitted power which will be collected. This fraction depends only upon the effective collecting area, and will be the same at any operating frequency. If the aerial is large and operating at the higher radio-frequencies, it will have a high gain and directivity. Because of this it will be necessary to use steerable aerials; of these the type having a parabola reflector is most convenient.

The limits on the effective collecting area are mechanical in origin. First there is a limit to the size of parabola which can be turned easily. A dish of 20 m. diameter is as big as can be handled without specially elaborate structures. One of 80 m. diameter (as at Jodrell Bank) is about the largest that is mechanically practicable. For such parabolic reflector aerials the effective collecting area is normally about 60 per cent of the circular aperture. However, for very short wavelengths, where the deviations of the reflecting surface from a true paraboloid may be significant fractions of a wavelength, the efficiency may fall much below 60 per cent. This accuracy of reflector shape, which may vary as the aerial axis changes in azimuth or elevation, provides an upper limit to the working frequency for very large aerials. However, for a well-designed structure, this frequency limit is likely to be of the order of 1,000 Mc/s or greater. It will therefore not be a determining factor in the choice of frequency for a system employing a triode first stage, for which it will be seen from Fig. 32 that a frequency of about 200 Mc/s is optimum. On the other hand, the introduction of masers or other low-noise devices may at some future date result in the use of microwave frequencies, in which case this limitation could be significant.

Space-Probe Aerial

If the space probe is free to tumble, having no axis stabilized in direction, then its transmitting aerial must be designed so as to give as near an omnidirectional cover as possible. This can be achieved reasonably well without a large aerial installation. Such a system is wasteful because the power generated in the probe at great expense is spread over all directions. If this could be concentrated in the direction of the earth the efficiency of the overall system would be improved greatly. The extent to which this concentration is practical depends upon the angular limit, ϕ, within which the direction of the axis can be controlled. Fig. 33 shows a section through the radiation pattern of a directional aerial giving a pencil beam. The points θQ^1 represent the directions at an angle θ to the axis at which the power level is half that on the axis at P. At all points within the cone of half angle θ, the gain is at least half that at P, and so this nalf gain can be assured if the axis can be stabilized to an accuracy of $\phi = \theta$. (It may seem that to take Q at the half power level is rather arbitrary—however, as the level

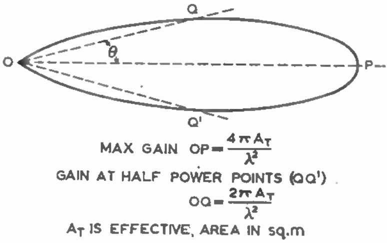


FIG. 33.—THE GAIN OF A PENCIL BEAM AERIAL.

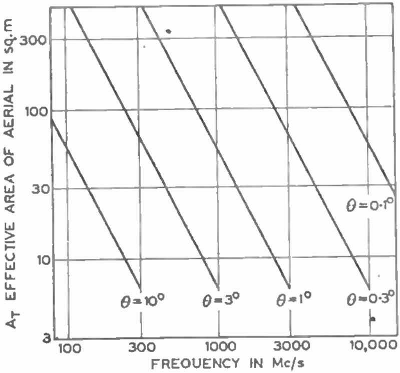


FIG. 34.—RELATIONSHIP BETWEEN AT, θ AND FREQUENCY FOR A PENCIL BEAM AERIAL.

falls off fairly rapidly near the half-power point, the value of θ is not very much affected by the ratio chosen for $OQ(OP)$.

The effective area A_t and the gain G of the aerial are related to θ and the wavelength λ by the approximate expressions

$$A_t\theta^2 = 640\lambda^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

$$G = \frac{4\,\pi\,A_t}{\lambda^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

Fig. 34 shows the necessary effective area as a function of frequency for constant values of θ. From the value of θ specified for the space probe, **Fig. 34** will give the aerial area required at any frequency in order to take full advantage of the stabilization of the probe axis. If there is a limit to the permissible value of A_t for a given probe, the figure can be used in order to determine the minimum frequency at which the desired θ can be achieved from this value of A_t.

Because the weight of an aerial increases more rapidly than its area, A_t will be strictly limited, and in making the best use of the total weight available for the space-borne radio installation any design based upon a large aerial and small transmitter would be improved by trading aerial gain for extra transmitter power.

Choice of Optimum Frequency

Based upon the above discussion, we can decide upon the optimum frequency for a space-probe telemetry system. From Fig. 32 it is noted that the operating frequency must be above about 50 Mc/s in order to be free from ionospheric effects, and below about 20,000 Mc/s in order that there should be negligible atmospheric absorption.

Because the galactic and triode sources of noise are dominant we base our choice of frequency upon these sources alone. The two dotted curves show the total noise for quiet and noisy regions of the galaxy. Both of these curves have broad minima, that for quiet regions being about 200 Mc/s, while for noisy regions the minimum is at about 300 Mc/s. If it is planned to send a probe far into space it might be assumed that a quiet direction would be chosen, and therefore we take 200 Mc/s as the optimum frequency. In this case the performance for a noisy direction would be very little worse than that to be expected at 300 Mc/s.

It is interesting to note that the frequency chosen for the Russian moon probe of 2 January 1959 was 188 Mc/s, which agrees well with the above conclusions.

Where there is a limit to the size of aerial which can be carried by a probe with its axis closely stabilized, there may be an advantage in working at a somewhat higher frequency than 200 Mc/s. This follows from Fig. 34. For instance, if the axis could be stabilized to $\phi = 3°$, then to take full advantage of this stabilization we should have $\theta = 3°$. Now if the permissible aerial area were 10 sq. m., then this value of θ could not be achieved at a frequency lower than 800 Mc/s. The question then arises whether the advantage to be gained by increasing the frequency in order to decrease θ offsets the degradation due to the consequent increase of the total noise. The increase in received power due to the decrease in θ and consequent improvement in aerial gain increases with the square of the frequency, while from Fig. 32 it is seen that the increase in noise power with frequency increases with approximately the first power of frequency. Hence the improvement in gain due to the increase in frequency outweighs the handicap of a higher noise temperature.

In such a system, then, we should deduce a minimum value of frequency from Fig. 34 using design data A_t and Q. The optimum frequency is then either this minimum or 200 Mc/s, whichever is the greater.

In this analysis the variation of transmitter efficiency with frequency has not been considered. This variation would tend to favour lower frequencies, but as the effect is not great, this is not likely to be dominant in a system. For a given type of transmitter the weight can be made less the greater the frequency. In a system which is required to operate over long periods, the weight of the transmitter may be swamped by the weight of the power supplies, in which case electrical efficiency may be more important than lightness.

Predicted Performance of a Space Probe Telemetry System

The results of this section are given in Table 1.

TABLE I.—CALCULATED PERFORMANCE FIGURES

	Moon Probe	*Mars Probe*
Range	400,000 km.	80,000,000 km.
Probe aerial	Isotropic	Power gain 25
Ground aerial	20 m. diameter dish	80 m. diameter dish
Frequency	200 Mc/s	200 Mc/s
Transmitted power	$940 \cdot 10^{-18}$ watt	$940 \cdot 10^{-18}$ watt
Bandwidth	10 c/s	1 c/s
Signal-to-noise voltage ratio	46 : 1	46 : 1

Consider now the performance to be expected from a space-probe telemetry system on 200 Mc/s using a 10-watt transmitter radiating equally in all directions. Suppose that this is in the lower region—say 400,000 km. away, and seen against a quiet region of the galaxy.

The power flux at the ground would be:

$$\frac{P}{4\,\pi\,r^2} = \frac{10}{4\,\pi\,4^2\,10^{16}} = 5{\cdot}0\,.\,10^{-18}\ \text{watt/sq. m.}$$

where P is the transmitted power in watts and r the transmission distance in metres. If the receiving aerial were a parabolic dish of 20 m. diameter and a 60 per cent area efficiency were assumed, its effective area would be 188 sq. m. This would give a received signal of 940 . 10^{-18} watts, which fed into a 50-ohm receiver input would give a voltage of

$$V_s = 0{\cdot}216\,.\,10^{-6}\ \text{volt}$$

We now determine the noise signal with which this must compete. From Fig. 32 it is seen that the total noise temperature for a quiet region at 200 Mc/s is about 800° A. If it is assumed that the system has a band-width of 10 c/s, then from Equation (1) the noise voltage:

$$V_n = 0{\cdot}00468\,.\,10^{-6}\ \text{volt}$$

which would result in a signal-to-noise voltage ratio of 46: 1. Such a signal-to-noise ratio would be acceptable for many purposes.

If now the space probe were in the region of Mars, at say 80,000,000 km., then V_s would be reduced by a factor of 200, and it would be swamped by noise. This factor of 200 could be made good by an increase in receiving-aerial area, by using a directive transmitting aerial, reducing the band-width or increasing the power. To increase the receiving dish to 80-m. diameter would give an increased voltage gain of only 4 and would represent about the limit of what could be done through modification of the receiving aerial. A directive transmitting aerial having a power gain of 25 would give a voltage gain of 5. This could be used only if the direction of the aerial could be held pointed towards the earth within limits of $\pm 7\frac{1}{2}$ degrees. For a long-term transmission an increase of power to 100 watts would be about the limit. If these three changes were made and the band-width decreased from 10 to 1 c/s, the 200: 1 voltage factor would be made good. It may be doubted whether or not all this effort would be justified for a 1-c/s band-width. If the probe were travelling through space, the parameters to be telemetered would vary so slowly that this narrow band-width would suffice. However, were it desired to send any information as detailed as a television picture of the surface of Mars, the band-width would be inconveniently small. In such circumstances it would be much better if the probe were to move in a trajectory which would return to regions near to the earth, so that data recorded on tape in the region of Mars could be telemetered by radio when the probe approached the earth. The establishment of such an orbit would present difficulties to the engineers responsible for the rocket-control system, but such difficulties could in principle be overcome, while the radio engineer's enemy—the inverse square law—is invincible.

21. RECEIVING AERIALS

21. RECEIVING AERIALS

BROADCAST-RECEIVING AERIALS

Advances in broadcast-receiving aerial design over the past three decades have not been spectacular, and because of the high sensitivity of modern receivers and the high field strength prevailing in many areas from B.B.C. transmitters, the importance of an effective aerial is frequently overlooked.

The principal benefits to be expected from a properly engineered aerial installation are :

(1) improved signal-to-noise ratio ;
(2) reduced local electrical interference ;
(3) increased sensitivity (better distant station reception on all wavelengths) ;
(4) greater freedom from the effects of night-fading on distant stations.

Before discussing in detail the properties and applications of different types of broadcast receiving aerials, it will be helpful to review some of the fundamental principles on which they depend.

Aerial System Fundamentals

A radio receiving aerial is essentially an electrical conductor exposed to the field created in the vicinity of a receiver by energy radiated from a distant transmitter as an electromagnetic wave.

Electric currents, oscillating at the frequency of the incident wave, are induced in the surface of the aerial conductor, and are taken to the receiver either through feeder conductors or by an extension of the aerial itself.

The input circuits of the receiver are energized by minute high-frequency alternating currents flowing in the conductors which connect them to the aerial, and which set up amplified oscillations in resonant circuits which are tuned to the required incoming signal frequency.

It is frequently possible to produce an output signal from a broadcast receiver even when no aerial is connected to its input circuits.

If this occurs it indicates that some of the tuned circuits in the receiver are being influenced directly by the electromagnetic-wave field, either because they are not effectively screened or because there is stray coupling between the reactive components of these circuits and other unscreened sections of the receiver or its supply leads.

A properly screened, filtered receiver is not affected by field radiation, nor by signal or noise potentials in its supply or output leads, and no trace of a signal will be found when the aerial is disconnected and the input socket is shielded.

If now, a few inches of wire be attached to its aerial socket, sufficient e.m.f. will be developed in it, in the presence of a strong electromagnetic field, to produce an output signal of programme value, if the receiver is reasonably sensitive; this should not be regarded as an indication that nothing more elaborate, in the form of an aerial, need be provided.

Effective Height

The field strength of an electromagnetic wave is diminished by absorption in the walls of buildings, in metal-framed structures and by the presence of electric wiring, plumbing, etc., and is consequently lower in the immediate vicinity of a receiver than in the space above, and surrounding, the building. It is therefore advantageous to raise the aerial conductor above the level of local absorbing media and to support it in unobstructed space where the radiated field strength is higher, and where the fields set up by local electrical interference sources are correspondingly weaker.

FIG. 1.—VERTICAL RESPONSE PATTERN OF A SHORT VERTICAL AERIAL, LOWER END NEAR EARTH.

There is a more cogent reason for raising the height of an aerial where reception of medium- and long-wave broadcasts is required; at these wavelengths (200–2,000 metres) a good receiving aerial may be only from 0·1 to 0·01 wavelengths long, and rather less than this in height, and must be regarded as being, electrically, very short and very near to the ground.

In these circumstances a horizontal aerial arranged for the reception of horizontally-polarized waves would develop an extremely small terminal potential in the wave field, because the currents induced in the conductor by the direct wave will be almost completely cancelled by induced currents set up by the ground-reflected wave, which undergoes a 180° change of phase at the point of reflection.

Vertically polarized waves are reflected from the ground without change of phase, and because of this, a vertical aerial will radiate or intercept a horizontally directed surface wave even when the conductor is electrically very short and the lower end is near to the ground (see Fig. 1). Practically all medium- and long-wave aerials are constructed on this principle, and consist essentially of one or more vertical conducting elements, insulated at their lower ends and connected to a feeder system at that point.

The defined plane of polarization of an electromagnetic wave is that which lies in the direction of the electric vector associated with the wave; in a transmitting aerial this corresponds to the direction of the major current-carrying elements.

A receiving-aerial element will develop the maximum induced e.m.f. when its length is in the plane of the electric vector, and the smallest potentials will be induced in conductors which are at right angles to this direction.

A vertically polarized wave will produce the greatest response in the vertical or near vertical parts of an aerial, and the horizontal parts will contribute little to the direct signal pick-up.

Aerial–Earth Systems

Short vertical aerials with their lower ends at, or near ground level, behave towards the incoming wave as though a reflected extension to the aerial existed in the ground beneath the aerial. Connecting the input circuits of the receiver between the lower end of the aerial and an earth plate buried in the ground will allow the aerial–earth system to develop the greatest signal potential.

It is important to recognize that in such an aerial–earth system the earth corresponds to the " other half " of a centre-terminated doublet or short aerial. If the earth connection is omitted, the aerial can only function through the effect of the capacitance between the receiver chassis and the ground; and this, being relatively small, will offer an appreciable reactive impedance to the flow of signal-frequency currents. It will thus be less effective than if a low impedance, direct return were used.

It will be seen that the receiver is virtually connected to the centre of a short aerial in all vertical aerial–earth systems and therefore at a point of low impedance. It is necessary to keep not only the reactance but also the resistance of the earth lead and earth plate to the lowest possible value to avoid loss of signal. Short direct connection to a cold-water pipe is generally effective, if it can be made near to the point where the pipe enters the ground. Hot-water-system piping and electric wiring earth returns are less effective, and are to be avoided.

If a water pipe cannot be used, a copper earth rod or plate driven deeply into ground which can be kept reasonably moist, will prove to be very satisfactory. Further information on this subject is given later in this section under " Earthing ".

Vertical Rod Aerials

The most effective use of the principles outlined above may be obtained by combining a rigid or semi-rigid vertical aerial element mounted on the top of a mast or attached to a chimney or to some convenient high point, with an external insulated wire downlead which forms an electrical extension to the aerial and should be taken directly to the receiver.

A typical installation of this type is the rod type. Such an aerial may consist of sections of taper-jointed high-tensile light-alloy tubing supported in resilient insulators on a steel angle bracket adapted to chimney corner mounting. The vertical height of the aerial is 18 ft., and, as normally installed, its top section will be approximately 50 ft. above ground, giving an effective height only slightly less than half this figure.

This type of aerial will not only give excellent results within the recognized service area of medium- and long-wave broadcast stations, but will perform very satisfactorily at much greater distances, where signal fading from ionospheric reflections may be expected at night. Its merit lies in the fact that its response is greatest in the horizontal plane, and is a minimum in the vertical plane (see Fig. 1); in this way, a vertical aerial discriminates against the high-angle reflected signals which give rise to fading, and responds most effectively to the low-angle, near horizontal, signals propagated by the surface wave, and which are responsible for non-fading reception in the extended service area.

Even at greater distances (from 200 miles upwards) where normal,

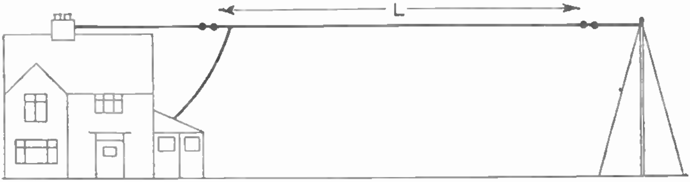


FIG. 2.—INVERTED-L AERIAL. THE HEIGHT IS IMPORTANT, BUT THE DIRECTION AND LENGTH OF THE TOP ARE NOT CRITICAL. AERIAL IS DIRECTIONAL IF L IS LESS THAN 0·3 λ.

reception is by the indirect or reflected wave, the low-angle properties of the vertical aerial help to minimize fading by rejection of high-angle multi-path signals which give rise to rapid fluctuations in strength.

Vertical aerials, although particularly useful for medium- and long-wave reception, are also quite effective on short waves; although the poor high-angle response may reduce the strength of European broadcasts, these are normally very powerful transmissions and need relatively little assistance from the aerial, while the consistent, low-angle performance is helpful in receiving long-distance transmissions from all directions.

Inverted "L" Aerials

The inverted-L aerial, Fig. 2, is probably the most widely used of all types. It consists essentially of a horizontal length of wire, insulated at both ends and continued downwards at the end nearer the receiver, and taken as near vertically as possible to the aerial socket on the receiver. The aerial should be erected as high as possible, and may be anything from 30 to 150 ft. in length.

On medium and long waves the effective or signal responsive part of the aerial is the vertical downlead. The function of the horizontal part of the aerial is to add capacitive reactance to the top of the vertical section, which increases its electrical length and also raises the electrical impedance at the terminals of the aerial. Since the normal impedance of the aerial is extremely low, the losses arising from the radio-frequency resistance of the aerial conductors represent a very large proportion of the energy collected by the aerial, and the improvement effected by raising its characteristic impedance, together with the increase in effective height, is appreciable.

The top of an inverted-L aerial, while quite unresponsive to the vertically-polarized surface wave from a transmitter, is affected by downward reflected radiation from the ionosphere, and this will increase the amount of signal-fading at medium distances (50–500 miles) during the hours of darkness.

Vertical aerials are not subject to this phenomenon, and will in general be superior unless the height of the inverted L can be raised to equal that of the top of the vertical element.

The inverted-L aerial will give a satisfactory performance on short wavebands, but may exhibit resonances which reduce the strength of signals in certain directions, particularly off the ends of the horizontal section. Directivity will change with frequency, and will also depend on the total length of the aerial and on the ratio of top length to downlead; and will be very difficult to predict with accuracy.

On medium and long waves the aerial is non-directional with respect to the downlead, and it is quite immaterial which way the top section is run in relation to the direction of the transmitter.

Loop Aerials

The vertical loop or frame aerial has its principal application in portable receivers, where an external aerial is an unwanted auxiliary (although provision is frequently made for the addition of one where the receiver is used in a fixed location). In this application the loop usually takes the place of the first tuned circuit in the receiver, and is resonated to the desired frequency by the normal tuning controls.

The efficiency of a loop aerial is controlled by its dimensions, increasing with its effective diameter; in consequence the small loop accommodated within the cabinet of a typical portable receiver is a poor collector of signal energy. Furthermore, it is invariably worsened by the close proximity of the metal chassis and loudspeaker.

A loop of substantial dimensions and mounted externally to the receiver may in some situations—for example, where outdoor aerials are impossible to erect—provide better reception than the more commonly used " wire round the picture rail ".

Its advantages lie in its ability to discriminate against local electrical interference and the fact that it is bi-directional, with sharp minima or nulls on either side of the plane of the loop. To be effective in reducing interference, the loop must be balanced to earth, i.e., its electrical mid-point is earthed and its terminals coupled symmetrically to a second tuned circuit which is connected to the input valve in the receiver. This is difficult to achieve except by screening the loop, although unscreened loops have been used for long-wave reception with some success. A balanced or screened loop (Fig. 3) responds only to the horizontal magnetic component of a radiation field and similarly to the magnetic component only in the induction field surrounding an interference source. Since the energy in the magnetic component of an induction field attenuates in proportion to the square of its distance from source, compared with the electrostatic field, which diminishes as the third power of distance, the loop is better able to separate local interference from the desired signal-radiation field (in which both electric and magnetic components obey an inverse distance law of attenuation and are consequently equal in value at all points).

The magnitudes of the induction and radiation fields are equal at a distance of $\lambda/(2\pi)$ from the source of radiation. In this formula both the wavelength (λ) and the distance are measured in metres. At a distance of 16 wavelengths the magnitude of the induction field has dropped to 1 per cent of the value of the radiation field at the same point. It is apparent from these relationships that the induction fields predominate at short distances from a radiating source, but are negligible at great distances.

A screen for a loop aerial should completely surround the loop conductors, radially, but should not form a complete circuit in the loop plane. A tubular screen should have a break in its circumference at the top or bottom (but not at the sides) in order to preserve symmetry of response to interference and to balance out any stray signal transfer to

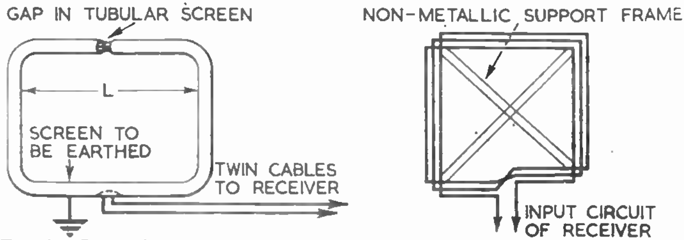


FIG. 3.—LOOP AERIAL. LENGTH L SHOULD BE LARGE, AT LEAST 24 IN. FOR EFFECTIVE INTERFERENCE DISCRIMINATION.

the loop windings. The adjacent ends should be separated by as short a distance as possible to preserve the maximum coefficient of screening.

A vertical loop aerial is not responsive to horizontally polarized radiation fields but, unlike the linear vertical aerial, it is equally responsive to energy arriving from all angles in the vertical plane. Because of this, a loop will receive signals arriving from reflected waves at all angles. At night this effect will give rise to fading to a much greater extent than with the aerials previously described.

Noise-reducing Aerial Systems

In this context noise is considered to be any random periodic electrical disturbance which is transmitted to the receiver input terminals together with the desired programme signal, to be amplified, detected and reproduced as an audible phenomenon which interferes with the quality and intelligibility of the transmission.

In the absence of a connected aerial, a sensitive receiver at full gain will develop a smooth continuous noise in the form of a hiss which is generated by the discrete random motion of electrons in the input circuits and in the cathode electron stream of the first thermionic valve in the equipment. At frequencies below 30 Mc/s the amplitude of this noise is normally below the atmospheric and cosmic noise levels set up by external radiation of a similar random character, which is intercepted by the aerial and passed on to the receiver.

Cosmic noise is of extra-galactic origin: its precise source is still uncertain, and it predominates in the 15–30-Mc/s spectrum. Below this frequency, atmospheric noise is the major source and arises from natural lightning discharges associated with thunderstorms, the effects of which are propagated for distances of many thousands of miles from their source.

There have been many attempts to devise aerial systems which will reduce the effects of atmospheric disturbances; the only effective arrangements are those employing widely spaced aerials which will respond differently to the relatively local desired signal, but which respond equally to the more distant interference. By careful phasing in the receiver, the noise component can be reduced and the desired signal separated, but the arrays are too complex and require too much space for domestic use.

Another major source of electrical noise is to be found in domestic and industrial electrical equipment in use in the vicinity of the receiver.

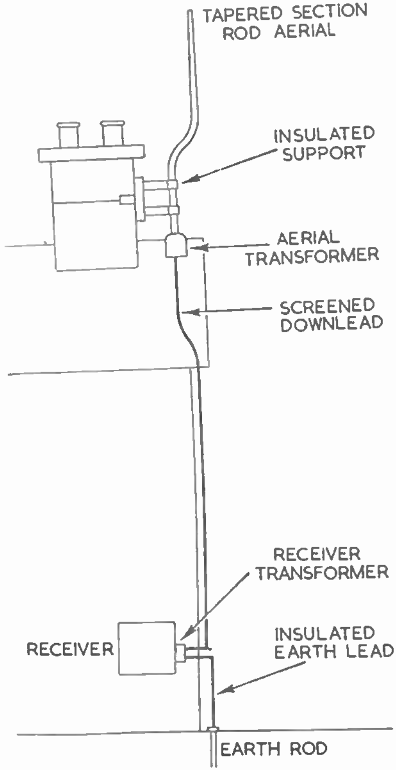


Fig. 4.—An Aerial Installation Designed to Reduce Noise.

This "man-made static" interferes very seriously with radio reception in heavily built-up areas. Indoor receiving aerials and the downleads of external aerials must pass into this noise belt, and will pick up energy from the induction and radiation fields surrounding the interference sources.

Screened Feeders

The effects can be very largely reduced by using screened and balanced downleads from aerials which are mounted sufficiently high to be above the more intense fields found inside the building. The screen of the downlead is earthed to prevent the transmission of signals or interference to the inner conductors, allowing these to pass the aerial signal, unaffected by interference, to the receiver.

Both co-axial and screened twin downleads have been employed in aerials of this type, but the latter are to be preferred, since they permit a more complete separation of signal and screening circuits, especially on medium and long waves, where the function of the aerial depends upon its ground connection.

Matching Transformer Systems

A typical installation is illustrated in Fig. 4, with the corresponding circuit arrangement shown in Fig. 5. At low frequencies signals are normally picked up by that portion of the downlead screen which is between the earth-connecting point and the aerial: the remainder of the downlead, which is normally within the building, is responsive neither to signals nor to interference. The aerial in this example is a vertical rod mounted as previously described. It is connected at its lower end to the twin conductors of a screened downlead through an impedance-matching, balanced transformer, which must be electrically screened. This is usually effected by enclosing it in a metallic weather-proof housing, connected either to the aerial element or to the aerial

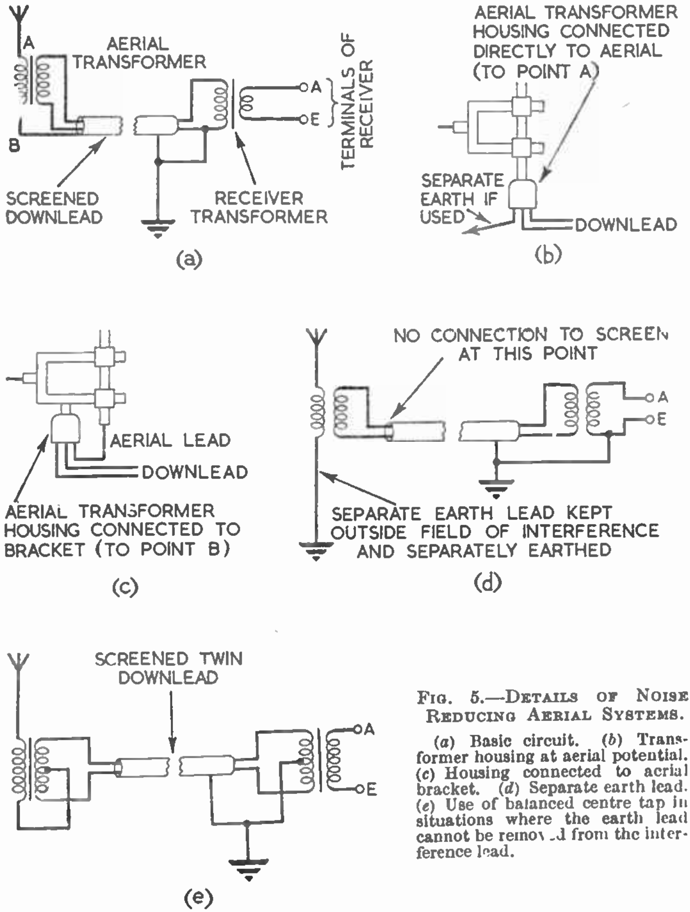


FIG. 5.—DETAILS OF NOISE REDUCING AERIAL SYSTEMS.

(*a*) Basic circuit. (*b*) Transformer housing at aerial potential. (*c*) Housing connected to aerial bracket. (*d*) Separate earth lead. (*e*) Use of balanced centre tap in situations where the earth lead cannot be removed from the interference lead.

support bracket. In the first instance (Fig. 5 (*b*)) the transformer housing will be at aerial potential, but, since there can be no electrical charge within a closed metal surface, the transformer circuits are still effectively screened from the wave field. If the housing is attached to the aerial bracket (Fig. 5 (*c*)) (which is insulated from the aerial), it is connected to the braided screen of the downlead and continues directly to earth.

The sensitivity of a vertical aerial used in this way for reception of medium- and long-wave signals is not significantly reduced, except by the losses introduced in the windings of the transformer, and in the downlead conductors, and these are very small at the frequencies in question (< 2 db).

The screen of the downlead in this type of aerial is responsive to signals in its upper section, since it forms part of the aerial–earth interceptor; because of this action the reduction in interference level may prove to be inadequate in some locations, e.g., in multi-storey flats where electrical noise may be created at quite high levels (both structurally and intensively). If the upper part of the downlead cannot be removed from the interference field, the screen should be disconnected from the aerial transformer circuit and a separate earth lead brought down by a route which avoids the interference field (Fig. 5 (*d*)). The screen should remain earthed at its lower end, and will now act solely as a screen, not being part of the aerial and not therefore carrying any signal energy. The sensitivity of the aerial will remain unchanged by this modification.

In situations where it is impracticable to remove the downlead or earth return from the interference field, the circuit may be completed through the twin conductors connected to the impedance-matching aerial transformer, either by a balanced centre tap, as shown in Fig. 5 (*e*) or by direct connection to one limb of the circuit. The screen of the downlead is isolated at the aerial transformer as in the previous example. This system will ensure freedom from interference, but the sensitivity will fall because the whole of the lower section of the downlead aerial return is screened, not only from interference, but also from the signal field. Since the most effective part of the aerial is the upper vertical element, the loss introduced by the screened lower section will not generally be greater than 3–4 db, while the reduction in extraneous noise may considerably exceed this figure, giving a significant overall improvement.

Various transformer downlead systems have been proposed using unscreened twin feeders, which rely upon the maintenance of accurately balanced currents in each conductor to minimize interference; they are not effective at normal broadcast frequencies, where unbalanced signal currents must flow to earth through the downlead, and correspondingly, an interference field will set up similar currents which will be transmitted to the balanced circuit, and thence to the receiver. It is also very difficult to maintain such a system in a state of balance without resort to screening.

The performance of noise-reducing aerial systems is usually maintained throughout the short-wave spectrum, although some variation in response will occur in the impedance-matching transformer system, as a result of the wide changes in aerial impedance which arise as the received frequencies approach the fundamental or one of the harmonic resonant frequencies of the aerial and downlead.

By careful attention to design, the transformers may be constructed to have a comparatively uniform response over the range 150 kc/s to 20 Mc/s. For this purpose they are usually wound in sections arranged so that their distributed inductances are appropriately coupled to their corresponding secondary winding sections, in such a way that the self-capacitance of the low-frequency windings acts as a by-pass to high-frequency signals, these being effectively applied to the low-inductance sections in series with them.

Short-wave Aerials

Short-wave broadcasting takes place throughout the world, mainly on frequencies in the 6-, 7-, 9-, 12 , 15-, 17- and 21-Mc/s bands. Reception of programmes from all continents is readily possible at different hours of the day, depending on ionospheric and sun-spot cycle conditions.

Although the previously discussed vertical-rod and long-wire aerials will receive signals on high frequencies, it becomes possible to construct more effective arrays specifically designed for short-wave reception and of reasonable dimensions for domestic installations.

Short-wave aerials may be raised to a height of between a quarter- and half-wavelength, and can easily be made to be a half-wave in length at any of the commonly used frequencies.

The velocity of propagation of electromagnetic radiation in air is approximately $2{\cdot}997 \times 10^{10}$ cm./second, which corresponds to 11,800 inches/microsecond. The relationship between frequency and half-wavelength is therefore :

$$\frac{\lambda}{2} = \frac{5900}{f(\text{Mc/s})}\ \text{in.} = \frac{491{\cdot}6}{f(\text{Mc/s})}\ \text{ft.}$$

and at 10 Mc/s is equal to 590 in. or approximately 50 ft.

Half-wave Dipoles

An aerial conductor insulated at its ends will exhibit a resonance when its electrical length is exactly one half-wavelength. In this condition an incoming electromagnetic wave will induce currents which set up standing waves of current and voltage, by successive reflections from the insulated ends of the conductor, as the induced oscillating currents travel along it. The distribution of current is a maximum at the centre of the conductor length and zero at its ends, where reflection occurs; correspondingly, the voltage is a minimum at the centre and rises to a maximum at the ends of the conductor.

Such a conductor is known as a half-wave dipole, and since it is (broadly) tuned to a specific frequency at which its response is a maximum, it forms a basis for most high-frequency transmitting and receiving aerial systems.

The physical length is slightly lower than the electrical length because of end effects, in particular the concentration of fields at the high potential points of the aerial. A convenient formula for calculating the resonant length of a half-wave dipole is $\frac{5705}{f(\text{Mc/s})}$ in.; this is applicable to all wire aerials having a ratio of length/diameter exceeding 1,000/1.

The centre impedance of a half-wave dipole in free space is 73·12 ohms, and if the conductors are separated by a short distance at their centre, an impedance of this value is developed across them. A downlead of 70–75 ohms characteristic impedance, e.g., flat twin or co-axial cable, may be connected to the aerial at its open centre to effect a maximum transfer of energy from it (Fig. 6 (*a*)).

Conversely, the impedance at the ends of the dipole is very high, being approximately 3,000 ohms. It is impracticable to match this value to a non-resonant line but a coupling can be effected by using a

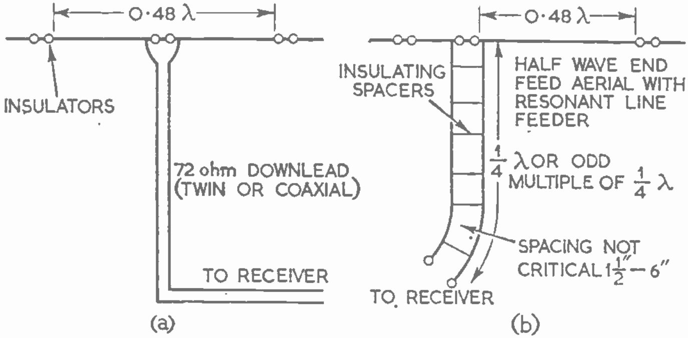


FIG. 6.—HORIZONTAL HALF-WAVE DIPOLE WITH SUITABLE MATCHING.

resonant quarter-wave section preferably of open-wire parallel line construction (Fig. 6 (*b*)) which transforms the high terminal impedance to a low value suitable for circuit coupling.

This arrangement is selectively tuned to the frequencies at which the line is a quarter-wave long or an odd multiple of this length, and so has greater application to transmission than reception.

Folded Dipoles

A centre-fed dipole with its non-resonant, low-impedance balanced feeder is better suited for wide-band short-wave reception; a modified form of construction known as a folded dipole possesses the same radiation properties, but has an increased band-width both at fundamental and harmonic frequencies, and presents a more uniform centre impedance at these frequencies. In this arrangement the aerial proper comprises two parallel conductors, of which one only is broken at its centre to connect to the downlead (Fig. 7). Since only half the total aerial current is sampled by the downlead, and since the power extracted

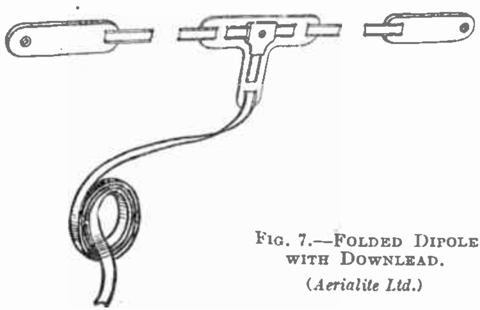
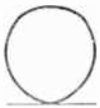
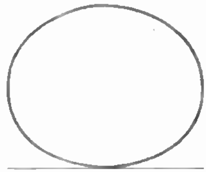


FIG. 7.—FOLDED DIPOLE WITH DOWNLEAD.

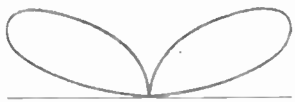
(*Aerialite Ltd.*)



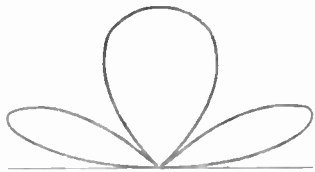
(a)



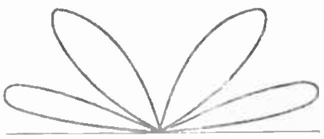
(b)



(c)



(d)



(e)

FIG. 8.—POLAR DIAGRAMS.

Vertical pattern of horizontal dipole: (*a*) 0·1 λ above ground; (*b*) 0·25 λ above ground; (*c*) 0·5 λ above ground: (*d*) 0·75 λ above ground; and (*e*) 1·0 λ above ground.

from the aerial is the same as in the single case, the centre impedance is multiplied by four, and so is approximately 300 ohms, if the aerial conductors are of the same size. Such an aerial, for short-wave reception, is very conveniently made from polythene-insulated ribbon feeder having a characteristic impedance of 300 ohms. A length of this ribbon is short-circuited at each end, and at the centre is tee-ed into a downlead of similar cable which, being non-resonant, may be as long or short as required. The ends of the folded-dipole should be insulated, since, as with the single dipole, they are at maximum potential.

Angle of Radiation and Reception

Unlike medium- and long-wave broadcasting, short-wave transmissions are generally horizontally-polarized and are propagated in quite a different way from the lower frequencies. Surface waves are rapidly attenuated, and most received short-wave signals originate in space waves radiated from the transmitting aerials at various angles to the horizontal (Fig. 8), and returned to earth by refraction and reflection from the ionosphere.

As the height of a horizontal aerial is increased, the amount of energy radiated or received by it at low angles of elevation increases steadily (see Fig. 8 (*a*)–(*d*)) and at short-wave lengths the normal height of the aerial is sufficient to ensure that incoming energy at angles greater than 20° with the horizontal will be intercepted by the first major lobe of the response pattern.

Successful long-distance short-wave broadcasting depends on the degree to which the radiated energy can be directed, both in the horizontal and vertical planes, into narrow-angle beams. These pencils of radiation not only ensure that energy is directed to those parts of the world where receiving conditions are likely to be favourable, but also, by reducing stray high-angle radiation, prevent destructive phase interference from multi-path reflected signals. Signals radiated at different angles and travelling by different paths to the receiver would cause severe signal fading. The required directivity can best be achieved with horizontally polarized arrays.

Horizontal dipoles, as used for short-wave reception, have a good low-angle response, but this is a variable factor, depending on the height of the aerial in quarter-wavelengths above the ground (see Fig. 8).

High-angle response is also generally good, and this is of value in receiving short distance (up to 2,000 miles) transmissions.

A horizontal dipole has horizontal directivity, which is a minimum off the ends of the element and a maximum at right angles to its length.

This directivity does not invariably produce a marked effect in response to short-wave signals because these signals do not all arrive horizontally; at increasing vertical angles the aerial will pick up signals off the ends as effectively as from other directions, and, for this reason, it is only necessary to consider the best direction for receiving long-distance transmissions in siting the aerial.

It is not usually of advantage to make short-wave receiving aerials highly directional; combinations of dipoles at selected spacings and appropriately phased will perform as designed only at specific frequencies, and are not suitable for general coverage. For similar reasons

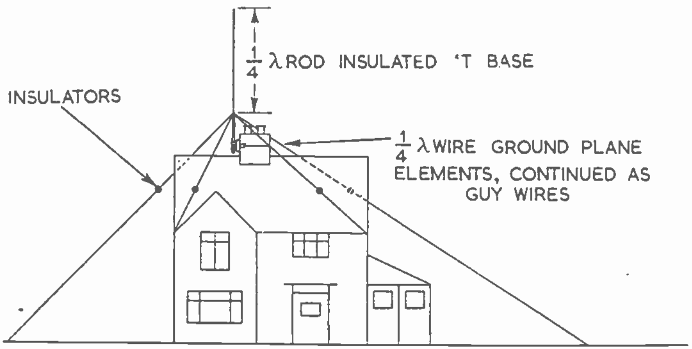
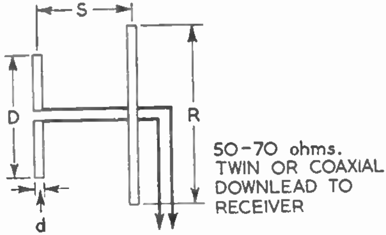


Fig. 9.—Vertical Ground Plane Short-wave Aerial.

FIG. 10.—LENGTH OF DIPOLE AND REFLECTOR ELEMENTS FOR USE IN BAND II, 87·5–100 Mc/s.



Diameter of tubing (in.)	*Dipole length* D (*in.*)	*Reflector length* R (*in.*)	*Spacing* S (*in.*)
1/2	$\frac{5270}{F\ (Mc/s)}$	$\frac{5666}{F\ (Mc/s)}$	$\frac{1770}{F\ (Mc/s)}$
3/8	$\frac{5400}{F\ (Mc/s)}$	$\frac{5783}{F\ (Mc/s}$	$\frac{1770}{F\ (Mc/s)}$
1/4	$\frac{5453}{F\ (Mc/s)}$	$\frac{5902}{F\ (Mc/s)}$	$\frac{1770}{F\ (Mc/s)}$

parasitic arrays incorporating reflectors are of little practical value, except in isolated situations where stations in a single frequency band from one direction only are to be received. Details of such aerials will be found in standard works on transmitting-aerial design.

Ground-plane Aerials

A very effective aerial for general short-wave coverage is the vertical ground-plane (Fig. 9). This consists of a vertical rod, insulated at its base and cut to be a quarter-wave long at whatever is to be the preferred centre frequency. Radiating from its base at equal angles, and as nearly horizontal as possible, are four connected wires, each a quarter-wave in length and insulated from the vertical element. The vertical element and the four " ground plane " radials are taken to the inner conductor and screening, respectively, of a 50–70 ohm co-axial down-lead, which may be of any length. The base of the aerial should be as high as possible, and the construction lends itself well to chimney mounting. In this case the radial wires may be taken downwards at an angle of about 45°, if necessary, and may serve as guy wires by including low-capacitance insulators at the quarter-wave points.

V.H.F. Aerials

The introduction of frequency-modulated V.H.F. broadcasting has stimulated interest in aerials suitable for reception in the frequency range 87·5–100 Mc/s.

In general, simple horizontal dipoles constructed from high-tensile, light-alloy tubing are sufficient, and where necessary, these may be supplemented by a reflector element. The electrical length of a dipole at these frequencies is reduced by a larger factor than that given in an earlier section because of the decrease in the length/diameter ratio. Formulæ for the calculation of dipole and reflector-element lengths are given in Fig. 10.

Car Aerials

It is difficult to mount an efficient low-frequency aerial on the body of a car, and, in practice, some compromise must be made between performance, mechanical stability and appearance.

The roof-mounting type is usually spring loaded to avoid damage from obstructions, and may be adjustable in angle or in lateral position from within the vehicle. Although the roof of a car is the most effective site for an aerial, being least subject to body screening, there are severe limitations in height which make it impossible to take full advantage of this exposed position. Telescopic elements are frequently used to increase efficiency without erecting a permanent obstruction hazard.

A rear-mounting aerial is less conveniently placed as regards screening, but may be increased in height to overcome this limitation. Its position at the rear of the car helps to minimize engine interference, but this advantage is partly offset by the need for a long feeder to the receiver, which is usually mounted at the front.

The performance of this type of aerial may be greatly improved by inductance loading at its base, but, mainly for economic reasons, this is rarely included in commercially manufactured aerials.

With the advent of modern wide-valanced front panels, a retractable multi-section telescopic aerial has become increasingly popular, and is today the most effective car aerial for all wavebands. A typical model of this type extends to 8 ft., and when retracted projects less than 14 in. from the surface of the car. The aerial housing, beneath the car panel, is an electrical screen for the aerial and its feeder connections, and despite the close proximity of the engine, ignition and other electrical interference is kept very low.

An aerial type which has had some applications where no projecting elements can be allowed is the under-car aerial. This type of aerial is not well placed for optimum reception, but the screening of the car body is not complete enough to prevent signal pick-up, and in dry weather good results are obtainable. Wet road surfaces inevitably reduce its efficiency, which must be made up by increased sensitivity in the receiver, and the aerial also suffers from interference produced by the generation of electric charges in the tyres, and other moving parts, which are electrically near.

Car aerials must be connected to their receivers through screened feeders, and since the lengths are generally short, it is practicable to use a low-loss, low-capacitance co-axial cable for this purpose.

Suppression of electrical interference from ignition systems, windscreen wipers, generators and the like is essential for satisfactory car radio working. This subject is discussed in Section 33, "Electrical Interference".

P. J.

TELEVISION RECEIVER AERIALS

The requirements of an aerial system for television reception are considerably more exacting than those for normal broadcasting purposes.

(1) the lower power, and correspondingly lower field-strength of television stations;
(2) the greater band-width required, and consequent lower gain per stage in the receiver;
(3) the higher circuit and insulation losses, and greater valve noise on V.H.F.;
(4) the greater susceptibility of V.H.F. signals to electrical and ignition interference;
(5) the necessity to avoid receiving transmissions by multiple paths in order to reduce "ghost" images;
(6) the greater susceptibility of the eye, as compared with the ear, to interference and signal variations.

On the other hand, it should be recognized that the use of a directional aerial system, tuned for optimum pick-up on a limited range of frequencies, and coupled to the receiver by a matched transmission line, represents basically a much more efficient type of arrangement than is customarily employed for broadcast reception.

The choice of an aerial and transmission line for any particular installation will depend upon the distance from the transmitter; its height and freedom from screening; the level of local interference; the length of feeder cable necessary; the sensitivity and input impedance of the receiver; and any restrictions imposed by the landlord or local authorities. It should be emphasized, however, that when planning an installation it is better to provide too much rather than too little signal input; for while it is a simple matter to incorporate an attenuator pad in the feeder to reduce an excessive signal, the raising of the signal level by even a few decibels may require the complete re-planning of the installation. Unlike valve amplification, additional gain in the aerial does not introduce extra "noise".

Dipole

If a vertical wire, whose length is short compared with the wavelength to be received, is placed in a vertically polarized radiation field from a transmitter, the current is distributed along the wire as shown in the

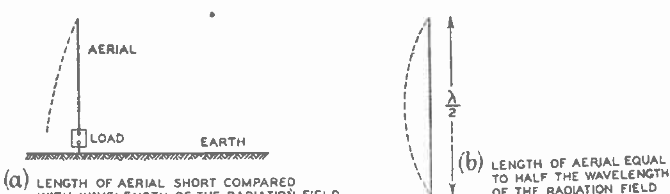


FIG. 11.—CURRENT DISTRIBUTION IN ROD AERIALS.

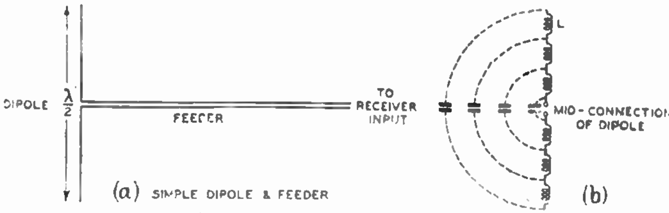


FIG. 12.—HALF-WAVE DIPOLE.

As shown in (b) this may be considered as a number of distributed inductances and capacitances acting as a tuned circuit.

dotted curve of Fig. 11 (c). This would be typical of a normal rod-type broadcast receiving aerial.

If the aerial has its physical length increased so that it is approximately half the wavelength of the radiation field, the current will be zero at each end and maximum at the centre, Fig. 11 (b). Due to the high impedance at either tip of such an aerial, it is somewhat difficult to connect a suitable downlead to it without the use of matching stubs, which are really transformers constructed from parallel wires.

The usual method is to break the aerial in the centre and connect it to the receiver by a double downlead, more correctly described as a feeder or a transmission line.

Such an aerial (see Fig. 12 (a)) is referred to as a dipole, and represents the most efficient single-element aerial which can be devised. Nothing is to be gained by adding length to either conductor, for the aerial is, in effect, a tuned circuit whose distributed self-inductance and self-capacitance cause it to resonate at the particular wavelength at which its length is half. Fig. 12 (b) shows diagrammatically the manner in which the inductances and capacitances are distributed, but, of course, it must be remembered in practice the number of inductances and capacitances involved is infinite.

A typical dipole aerial for use at the London television frequency of 45 Mc/s comprises two ½-in.-diameter metal rods, each approximately 5 ft. 2 in. long, and separated at the centre by ½ in. This aerial behaves like an unselective tuned circuit (highly damped) and has a band-width of the order of 5 Mc/s on either side of resonance before the response drops to 0·7 (3 db) of the maximum.

Over this band of frequencies the effective resistance of the dipole, as it would appear across the centre gap, is of the order of 80 ohms, and to avoid losses in transmitting the collected signal to the receiver it must be connected via a feeder whose characteristic impedance is 80 ohms or nearly so.

The formula below will enable the length to be determined

$$\text{Length (in.)} = \frac{5{,}905 \times k}{\text{Frequency (Mc/s)}}$$

where k is a factor depending upon the ratio of half-wavelength to the diameter of the aerial element, usually varying between 0·92 and 0·98. Typical values for k are 0·97 (ratio 1,000 : 1) and 0·96 (ratio 75 : 1). To ensure that the band-width of the aerial is sufficient, the length/

diameter ratio should be less than 400. The impedance at the centre is approximately 80 ohms, but this may be affected by the presence of nearby objects or additional elements.

Directional Effects and Relative Gain

The signal e.m.f. developed across the gap of a dipole when vertically disposed in a uniform, vertically polarized, radiation field is given by the formula

$$e = \frac{E\lambda}{\pi}$$

where e = developed signal voltage;
E = field strength in volts/metre;
λ = wavelength in metres.

Now the above formula gives the *open-ciruit* voltage from the dipole. When the receiver input is matched with, and connected to, the dipole, the voltage is reduced by one-half. From this must be taken the additional loss due to the feeder.

In operation the vertical dipole has no directional properties. This is rather obvious, because it is symmetrical about its axis of revolution.

The Yagi Aerial Array

Most directional television receiving aerials are based on the Yagi array, which is also widely used in H.F. and V.H.F. communications work. To understand the operation of this aerial, it is helpful to consider it as being connected to a transmitter.

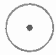
If an aerial slightly longer than one half-wave and not connected to a power source is placed parallel to but slightly less than one quarter-wave directly behind a driven half-wave dipole it will act as a parasitic reflector. It absorbs power and re-radiates it with such a phase relation to the original radiation that the fields of the two aerials add in one direction and subtract in the other, see Table 2.

TABLE 1.—TYPICAL DIMENSIONS OF TELEVISION AERIALS

Channel	*Mean Freq. (Mc/s)*	*Dipole* ft.	*Dipole* in.	*Director* ft.	*Director* in.	*Reflector* ft.	*Reflector* in.	*Spacing* $\frac{1}{8}\lambda$ ft.	*Spacing* $\frac{1}{8}\lambda$ in.
1	43·5	10	7	10	1	11	1	2	9
2	50	9	3	8	9½	9	8½	2	4½
3	55	8	5	8	0	8	10	2	2½
4	60	7	8½	7	4	8	1	2	0
5	65	7	1	6	9	7	5	1	10
6	178	2	7½	2	6	2	9		8½
7	183	2	6½	2	5	2	8		8
8	188	2	5½	2	4	2	7		8
9	193	2	4½	2	3	2	6		7½
10	198	2	3½	2	2½	2	5		7½
11	203	2	3	2	2	2	4½		7
12	208	2	2½	2	1½	2	4		7
13	213	2	2	2	1	2	3		7

Note. The exact dimensions are affected by the ratio of diameter to overall length, and so will vary slightly according to the tubing used.

TABLE 2.—PROPERTIES OF VARIOUS TYPES OF BAND I TELEVISION AERIALS

TYPE OF AERIAL	RELATIVE GAIN TO A DIPOLE	POLAR DIAGRAM (RELATIVE)	PROBABLE MAXIMUM RANGE (MILES)	OCCASIONAL RANGE (MILES)
 DIPOLE	0db (i.e. THERE IS NO GAIN SIGNAL PICK UP IS x1)		18	50
 V DIPOLE	−6·0db		14	35
 DIPOLE WITH REFLECTOR	4·0 to 5·0db		35	70
 ARRAY	7·0 to 8·0db		35+	70+

A parasitic element shorter than one half-wave, placed parallel to and slightly less than a quarter-wave ahead of a half-wave dipole, will act as a director. It absorbs power and re-radiates it with such a phase relationship that the fields of the two aerials will add in the direction of the director element.

In practice, for reasons of economy of space and because of the relatively low interaction between aerials a quarter-wave apart, it is usual to place the reflector elements considerably closer to the driven dipole, spacings between $0{\cdot}1\lambda$ and $0{\cdot}2\lambda$ being common. Since this reduction of spacing means that the necessary phase change to give approximate cancellation in the backward direction would not be obtained, it becomes necessary to obtain the required delay electrically. This is done by making the reflector slightly longer than one electrical half-wave so that, being excited above its own resonant frequency, it has inductive reactance, thus introducing the necessary lag. Similar reasoning applies also to the director; this is made slightly shorter than the resonant length.

The calculation of the exact lengths of reflector and directors is a matter of some complexity, and in practice a certain degree of experiment is often preferred, though the following formulæ will give approximate lengths:

$$\text{Reflector Length (in.)} = \frac{5{,}904}{\text{Frequency (Mc/s)}}$$

Director lengths can be determined by using the following formula:

$$\text{Length (in.)} = \frac{5{,}400,}{\text{Frequency (Mc/s)}}$$

Typical dimensions and spacing are: dipole 0·43λ–0·46λ, reflector 0·5λ, spaced 0·15λ–0·25λ from dipole. Adjustment of these dimensions will affect front-to-back ratio of the array and the terminal impedance of the dipole. Directors, 0·406λ–0·44λ. Adjustment of these dimensions may affect front-to-back ratio and also the extent of side lobes. The band-width of Yagi arrays for relatively small changes in the standing-wave ratio on the feeders is usually of the order of 4–5 per cent. Typical gains, relative to a half-wave dipole, are: three-element 4–5 dB; four-element 6–7 dB; six-element 10 dB.

In practice, the dipole plus reflector (commonly known as the "H aerial") is widely used on Band 1, whilst for fringe conditions up to four elements are in common use. To facilitate impedance matching, the dipole is often folded.

Slot Aerials

This type of aerial may be used as an alternative to the simple dipole, or dipole and reflector, to which it has roughly similar performance. For installation in the loft of a house, it has the advantage that, for vertically polarized transmissions, it is long rather than high. It consists of a vertical sheet of conducting material, such as wire netting, with a slot running horizontally in the centre and the feeder connected to the mid-points of the long sides of the slot. Typical dimensions for Channel 1 are: netting, 15 ft. long × 5 ft. high (these dimensions are not critical); slot 10 ft. long × 1 ft. high.

Aerials for Band III

On the higher frequencies of Band III the voltage induced in a dipole element is appreciably less than induced in a Band I dipole in an area of similar signal strength, while the losses in feeder cables will be approximately doubled; furthermore, the sensitivity of a receiver will not be so good on Band III as on Band I. For all these reasons, greater care has to be taken if good signals are to be presented to the receiver on Band III. On the credit side, however, the shorter elements make multi-element arrays relatively simple to construct and, in fact, arrays of ten or so elements mounted on a single cross-arm are available. These are highly directional, and great care should be taken to orientate them correctly in order to obtain the optimum results.

When planning aerial installations for both bands, the following questions need to be answered before any decision can be made as to the best type for a given location:

(1) Are the transmitters co-sited or located at different directions from the receiver?

(2) Is the location within the primary service area (i.e., high signal strengths) of one or both stations?

(3) Is a Band I aerial already installed? This may enable a Band III aerial adapter kit, as available from most aerial manufacturers, to be used.

(4) Is the feed run comparatively short (not more than about 50 ft.), or will a long cable run be required? Loss of signal in short runs is usually unimportant, but will be appreciable with long runs on Band III.

Undoubtedly, one of the major sources of difficulty with Band III aerials is the greater number of "ghost" images that occur on these transmissions. Hills, gasholders, spires, steel structures and many other reflecting surfaces may give rise to strong signals, arriving slightly out of phase with the direct signals, and thus producing ghost images slightly displaced from the main ones. These can often be eliminated by very careful orientation of the aerial, sometimes suffering some reduction in strength of the main signal.

On Band III, poor aerial siting or installation can seriously impair picture quality even within a comparatively short distance from the transmitter. Manufacturers have drawn attention to the advantages of an aerial sited above chimney stack and as far as possible from obstructions; or failing this, erected so that the chimney is to the side of, or behind, the aerial. Loft aerials for Band III are often rendered useless by the proximity of the water tank.

The following are among the suggestions put forward by a prominent manufacturer:

Install an outdoor aerial.

Have it on a high chimney.

Use a good-quality low-loss feeder cable.

Use the sensitivity control on the receiver to obtain a good picture on the weaker signal.

Use an attenuator on the stronger signal to equate the two signals.

Remember that multi-element aerials are very directional, and even a few degrees off beam will make a big difference.

Skeleton Slot Aerials

A particular form of the slot aerial that has been used recently for Band III reception—and for V.H.F. work generally—is the "skeleton slot". As it name implies, this type of aerial was developed from the normal slot by gradually reducing the metal surround. It has been found that results substantially the same as those of the normal slot may be obtained even when the surround is reduced to a mere rim of metal, which in practice may take the form of metal tubing (about ½ in. diameter) enclosing the "slot". This provides an aerial which is comparatively simple mechanically and which does not offer the wind resistance of conventional slot assemblies. Directive arrays may be formed by adding directors or reflectors, though these will be mounted at 90° to the major axis of the slot. The feeder is normally matched to the mid-points of the slot (where the impedance may be as high as 600 ohms) by means of a quarter-wave stub or a linear transformer section.

Band I/Band III Tuned Filters

Filter units for separating or combining Band I and Band III signals —known variously as "cross-over filters", "diplexers", "splitters", etc.—have a number of uses. For example, such a unit is necessary where separate Band I and Band III aerials are used with a receiver having a single input socket, or alternatively where a combined Band I/Band III aerial is used with a receiver having separate input sockets. Two filter units may be useful where separate aerials and separate input sockets exist in those locations requiring long low-attenuation feeder cables; in these circumstances the cost of two filter units may be less than that of the length of cable which is rendered unnecessary by the use of one filter mounted close to the aerials and a second filter at the receiver end of the feeder cable.

Aerials for Bands IV and V

In the U.S.A., where transmitting stations are already in operation on the ultra-high frequencies, the bow-tie type dipole has become generally popular: it is so named after its physical shape. Because, for broad-band operation, parasitic elements can operate only within a restricted frequency range, the parasitic elements in Band IV and V arrays are surfaces rather than rods: the bow-tie dipole is usually mounted in the corner formed by a surface reflector. Bowl- and mat-reflectors can also be employed in order to achieve the highly directional properties necessary in order to pick-up the maximum signal with the increased interference, reflection and propagational difficulties in the ultra-high frequencies.

Transmission Lines

There are two basic types of feeder line in general use: the co-axial and twin lines.

The co-axial feeder comprises a pair of co-axially related conductors insulated by low-loss material such as polythene. The twin feeder consists of two equi-diameter conductors laid side by side, separated and held in position by polythene or a similar flexible low-loss insulating material.

The characteristic impedance of these feeders is a property determined by the distributed capacitance and inductance of the conductors which form them. It is that value of resistance which, when connected to the end of the feeder, will absorb all the energy it receives and not permit any of it to be reflected to the sending end. If this state of affairs is permitted by mismatching, loss of energy (in the case of television, loss of signal) will occur.

The characteristic impedance of any two-conductor feeder line at high frequencies is given by the formula:

$$Z_0 = \sqrt{\frac{L}{C}}$$

where Z_0 is the characteristic impedance of the line;
L is the inductance per unit return length;
C is the capacitance per unit length.

Obviously, the closer the spacing of the wires, the lower the inductance, the higher the capacitance and, therefore, the lower the impedance.

Even if the dimensions of the feeder line are chosen so as to provide a characteristic impedance which matches exactly the aerial and receiver impedance, there will still be some loss of signal. This is due to losses in the conductors by virtue of their resistance, and losses due to the finite power factor of the intervening insulating material.

Co-axial and twin feeders in general commercial use for television reception will reduce the signal by about half (6 db) if 200 ft. is used. As the average domestic installation usually requires less than 100 ft. of feeder, the loss is fairly small.

For long-distance reception, however, where every loss must be cut to a minimum, specially designed (and more expensive) feeder possessing approximately half the loss should be used. This is important, because it must be recognized that the aerial is usually placed at a considerable height to increase the signal pick-up, and this increase can be offset if the feeder is lossy, because a greater length must be employed.

Further information on transmission lines is given in Section 12.

Matching Television Aerials to Feeder Cables

The addition of parasitic elements reduces the impedance at the centre of a half-wave dipole. This mis-match which occurs when an H type aerial is connected to, say, an 80-ohm cable is generally regarded as insignificant, but with multi-element arrays some means must be found of overcoming this difficulty.

Two popular methods of raising the effective impedance are: (1) use of folded dipoles, and (2) insertion of a quarter-wave transformer.

The folded-dipole consists of two or more half-wave aerials with their ends connected together, running parallel and closely spaced to one another, with the feeder cable connected to the centre of only one of the elements. With two similar elements such an arrangement provides an impedance step-up of 4 to 1, and three elements 9 to 1. A wide range of step-up ratios can be obtained by using elements of dissimilar-diameter tubing.

The quarter-wave transformer consists of the appropriate length of transmission line inserted between the aerial and the feeder cable. The characteristic impedance of the transformer is given by the formula:

$$Z_t = \sqrt{Z_1 \times Z_2}$$

where Z_t is the characteristic impedance of transformer;
Z_1 is the characteristic impedance of the feeder cable;
Z_2 is the impedance of the aerial.

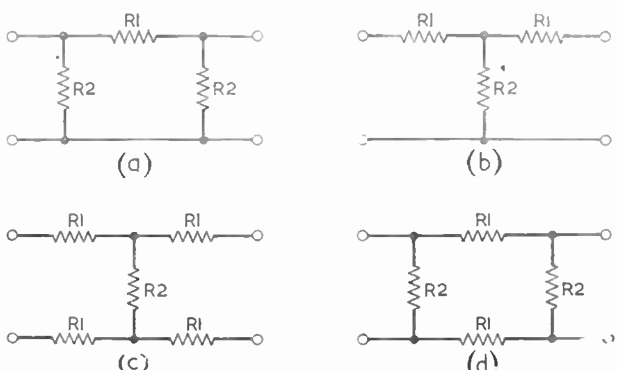


FIG. 13.—ATTENUATORS: (a) "pi" TYPE; (b) "T" TYPE; (c) "T" TYPE FOR BALANCED TWIN FEEDERS; (d) "pi" TYPE FOR BALANCED TWIN FEEDERS

Two other methods occasionally used are: (1) to make use of the relatively high impedance at the end of the dipole, and (2) T matching the cable to the aerial element. In the case of (1), which is practical only with arrays using four or more elements, a step-down quarter-wave matching stub is usually required. With (2) the feeder wires are connected not to the centre of the dipole but to points a few inches above and below the centre. Both of these methods are inclined to reduce the band-width and to make the dimensions rather critical.

For the use of cables as transformers see Section 12.

Attenuators

The two types commonly used are the "pi" and the "T". The "pi" is more suitable for test purposes, as it uses higher resistance values. Fig. 13 (*a*) shows the circuit, and resistance values for different attenuation requirements are given in Table 4.

TABLE 3.—RESISTOR VALUES FOR "pi" ATTENTUATORS

Required Approximate Attenuation (db)	*R1 (Ohms)*	*R2 (Ohms)*
10	150	100
20	470	100
30	1,500	82
40	3,900	82
50	10,000	82
60	39,000	82

Resistors should be of normal ½-watt rating, with tolerance of ±10 per cent, non-inductive.

The "T" type circuit is shown in Fig. 13 (*b*). Above 20 db attenuation this type is not suitable, as the resistance value of R2 becomes too small. Resistance values are given in Table 4.

TABLE 4.—RESISTOR VALUES FOR "T" ATTENTUATORS

Required Approximate Attenuation (db)	*R1 (Ohms)*	*R2 (Ohms)*
10	39	56
20	68	16

The circuits given are for use with co-axial feeders. Where balanced twin feeders are used the value of R1 should be halved and the circuits are as shown in Figs. 13 (*c*) and (*d*).

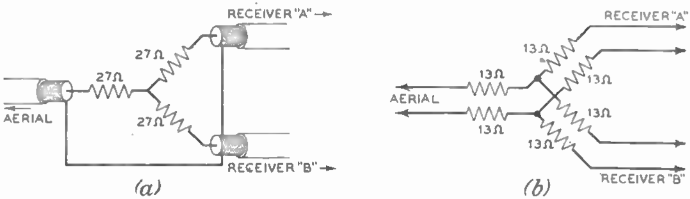


FIG. 14.—STAR NETWORKS: (a) CO-AXIAL FEEDER; (b) TWIN FEEDER.

Multi-receiver Installations

Wherever possible, aerials should be erected well clear of all neighbouring aerials. Where only one chimney stack, or other convenient mounting point, is available for two houses or flats, it may prove more satisfactory—and cheaper—to feed the two receivers from a single aerial, rather than to place two aerials in close proximity. This can be done without difficulty, provided that the signal strength in the area, for the type of aerial system employed, is sufficient to permit a loss of 6 db from the signal which would be fed to a single receiver. Star networks that enable two receivers to be fed from a single co-axial or twin-feeder line are shown in Fig. 14.

Where it is desired to operate a large number of receivers from one aerial without loss of signal, as may be the case in a large block of flats or a hotel, it is essential to ensure that there is no interaction between receivers, and that all outlets receive a satisfactory signal. This will normally entail the use of a distribution amplifier in conjunction with an efficient aerial, suitable cable runs and correct socket outlets.

Where a distribution amplifier is used, it should be fitted as near to the aerial as possible, and suitably accommodated in a dry, weather-proof room or enclosure, properly ventilated and reasonably free from dust. It is essential that flue gases or smoke do not enter the amplifier enclosure, as otherwise rapid corrosion of metalwork may be experienced. Where exceptionally long cable runs are necessary, it may be advisable to use semi-air-spaced co-axial cable.

Fringe-area Equipment

To minimize the effect of valve noise and random noise in fringe areas, it is essential to provide the receiver with the maximum possible signal; otherwise the picture will be marred by " grain ". Mention has already been made of multi-element arrays.

Since valve noise becomes increasingly important as the input to the first radio-frequency stage falls, the losses in the feeder line are often the decisive factor in determining whether or not a satisfactory picture can be obtained in a fringe area, and for this reason special low-loss cables have been developed with an attenuation, even on Channel 5, of less than 2 db per 100 ft.

When calculating feeder losses, it should be remembered that manufacturers' figures for impedance and attenuation of unshielded lines are based on the assumption that the feeder is perfectly dry and completely isolated from earth. In practice, neither condition may be fulfilled, and losses may thus be materially increased. It is possible to minimize this additional loss by keeping the feeder clear of metallic objects and by reducing to a minimum horizontal stretches of cable, particularly along the top of roofs, where there will be a greatly increased risk of water seeping into the cable. With co-axial feeder, make certain that water cannot enter the cable at the junction to the dipole, and that the cable is looped before entering the building so that external water will run off.

Since special feeder cables are considerably more expensive than the standard types, it will be necessary, in many border-line cases, to carefully weigh the relative costs of the various types of aerial systems, high masts, pre-amplifiers, cables and the like. For example, it may be found cheaper to use a more complicated array with standard cable than a simple aerial with special cable, or vice versa. The individual circumstances of each installation will govern the best course to be adopted.

For areas of low signal strength, where every decibel of loss is important, the attenuation in the feeder may reach an unwelcomely large figure, even where semi air-spaced cables, with correct impedance matching, are employed. In such cases a mast-head pre-amplifier may provide a useful gain with less valve noise than would be experienced with a normal type of pre-amplifier located beside the receiver.

" Ghost Images "

Reflection of signals from natural and man-made objects may cause a double or multiple image to appear upon the screen of a receiver; and in certain districts such conditions may prove most troublesome and difficult to overcome. Hill faces are probably the most frequent cause of " echo " signals, but almost any reflective surface, such as trees, buildings, gasometers or factory chimneys, may give rise to " ghosts ". By measuring the displacement of the spurious image, a rough estimate may be made of the distance of the offending objects.

The approximate image-displacement values and distances of the reflecting structure from the aerial for a 15-in. cathode-ray tube are given in Table 5.

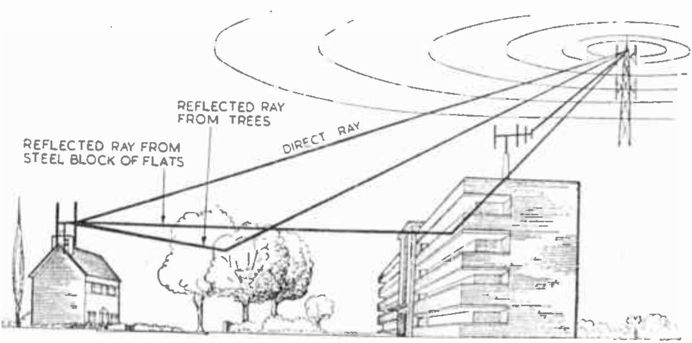


Fig. 15.—The Reception of Ghost Images from Reflected Waves.

TABLE 5.—DISPLACEMENT VALUES OF GHOST IMAGES

Displacement of Image	*Object to Right or Left*	*Object Immediately Behind*
$\frac{1}{16}$ in.	140 yards	70 yards
$\frac{1}{12}$ in.	190 yards	95 yards
$\frac{1}{10}$ in.	220 yards	110 yards
$\frac{1}{8}$ in.	280 yards	140 yards
$\frac{1}{4}$ in.	560 yards	280 yards
1 in.	1¼ miles	1100 yards
2 ins.	2½ miles	1¼ miles

Intermediate angles between the rear and side-on positions will give intermediate values between those shown in the middle and right-hand columns. Objects slightly in front of the side-on view would be at a distance greater than that given by the middle column.

The cure of double images is still largely a matter of trial and error in the positioning and orientation of the aerial, the principle being the adjustment of the system to give minimum pick-up of the "ghost" reflection. It is, for example, useless in areas prone to "ghosts" merely to point the aerial towards the transmitter. Instead, the aerial should be carefully adjusted when a programme, preferably Test Card "C", is being received. In some cases it may prove easier to reject the direct signal and concentrate on receiving the reflected signal, for example by turning the aerial on its side so as to receive horizontally rather than vertically polarized waves. With indoor aerials, a change of position of only a foot or two may make a considerable difference to results.

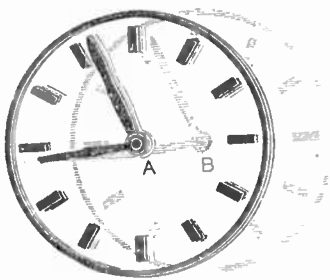


FIG. 16.—GHOST IMAGE.

TABLE 6.—TYPICAL ATTENUATION LOSSES IN FEEDERS

Type of Cable	*Impedance (Ohms)*	*Attenuation loss (db/100 ft.)*	
		50 Mc/s	*200 Mc/s*
Solid polythene co-axial . . .	75	2·9	6·4
Cellular polythene co-axial . . .	75	2·3	4·9
Semi-air-spaced co-axial . . .	75	1·6	3·5

EARTHING

Earth conductivity varies considerably according to the nature of the soil, and a system which would be satisfactory under conditions of good conductivity may be totally inadequate in rocky or sandy districts. A rough appraisal of conductivity in any particular area can be made from the following table :

Salt water	Extremely good
Rich farm soil	Very good
Average country soil	Good
Fresh water ponds	Fair
Town residential areas, rocky or sandy soil	Poor
Industrial areas	Very poor

Where more accurate figures are required, it is necessary to take into account the soil and subsoil, and the limits within which the resistivity may vary are :

	Ohm-metre
Chalk	60–400
Clay	3–160
Clay with varying proportions of sand and gravel	10–1,350
Marsh	2–3
Rock (normal crystalline)	500–10,000
Sand	90–800
Sand gravel	300–5,000
Slates and shales	100–500

For resistivity in ohm-ft., multiply figures by 3.

In general, the essential requirements for telecommunication earth connections may be summarized as follows :

(1) Must be complete, permanent and as short in length as possible.

(2) Must be of low resistance, i.e., not more than 8 ohms for telecommunication apparatus, and not more than the supply voltage divided by the circuit-breaker tripping, or fusing, current for associated power equipment.

(3) Must be of adequate current-carrying capacity.

(4) Must be mechanically strong and must not deteriorate due to corrosion.

(5) Must provide convenient access for maintenance testing purposes.

Earth connections via water-supply mains are usually of lower resistance than any form of buried earth-electrode system, and normal radio apparatus may be conveniently and satisfactorily earthed by means of connections to water mains.

For electrode systems, the graph in Fig. 17 shows the earth resistance of ⅝-in. rod(s) when installed in a soil resistivity of 100 ohm-metre. The corresponding values when installed in soils of different resistivities may be ascertained by multiplying by the actual soil resistivity and dividing by 100.

The main earth lead should preferably be of stranded copper wires, and the cross-sectional area should be proportional to the maximum current to be carried.

For power connections, the earth lead should have a nominal cross-sectional area of at least 0·0045 sq. in. (e.g., 7/0·029 in.), and be not less than one-half that of the largest of the conductors it protects, but no earth conductor larger than 0·1 sq. in. cross-section need be used.

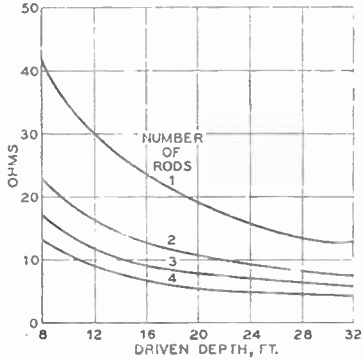


Fig. 17.—Earth Resistance of a ⅝-in. Rod.

Transmitters

There are two distinct earthing systems which may be associated with radio transmitting equipment:

(*a*) earthing system for the aerial;

(*b*) earthing system for the apparatus itself—modulator, oscillator, rectifiers, control boards, etc.

That of (*a*), which, when required, usually consists of a counterpoise or buried-wire earth to assist in forming a good reflecting surface for down coming waves, is more fully covered in Section 11.

The purpose of system (*b*) is to prevent radio-frequency currents in the frames and screening panels, by providing an electrical anchor, to obviate unwanted coupling between circuits which would cause instability, and to prevent the radio-frequency field affecting the power supply. The requirements for the earthing of radio-frequency equipment are generally the same as those for power equipment; a very important point, however, is that, with radio-frequency equipment, the earthing connection between the electrode and the equipment must be kept as short as possible. This applies at all frequencies, but is particularly vital at the higher frequencies, when the earth connection may be useless—due to its high radio-frequency impedance—particularly where its electrical length is an approximate multiple of a quarter of the wavelength concerned. For V.H.F. earthing it is often advantageous to use a copper sheet, placed under the equipment. This provides a considerable capacitance to earth, whose impedance decreases with increasing frequency. A D.C. earth connection is necessary as well.

Where the earth connection of a high-power transmitter passes through combustible material it is advisable to fit protective bushing to prevent any possibility of fire from radio-frequency sparking at joints. Any metal strips or fittings which might resonate at the transmission frequency should be bonded to earth.

Receivers

Most of the considerations for transmitters apply also to receivers, but the lower power involved cuts out the risk of fire from radio-frequency

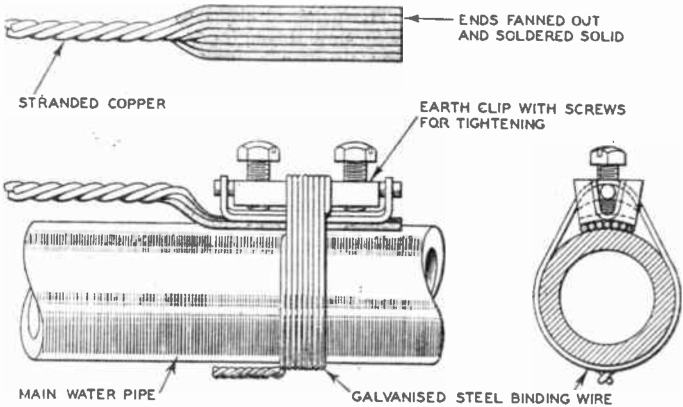


FIG. 18.—METHOD OF BONDING EARTH WIRE TO MAIN WATER PIPE.

sparking. It is often practicable to use cold-water pipes as the earth electrode.

With broadcast receivers fitted in domestic premises, the earth-continuity conductor of the electric supply can be used as the earth connection, but to obtain the best results a rising water main or a separate earth electrode should be used. The separate electrode is probably best, and it should be sited well away from any other earth electrode near the premises. It should be noted that the earth connection of a mains domestic receiver has to act as both a power and a radio-frequency earth, and its impedance must therefore be low enough to blow the fuses protecting the receiver in the event of a short-circuit to the chassis or other metalwork in the receiver, or, in the case of receivers using a double-wound mains transformer, a leakage path from chassis to supply mains.

Efficient receiver earthing systems can be made by sinking a 3–4-ft. copper rod, or several shorter rods spaced a foot or two apart, or a large copper plate as deeply as possible in moist soil; a large biscuit tin is also efficient. If the soil is sandy or particularly dry, the efficiency of the earth can be increased by placing the earth electrode in a trench filled with about 40 lb. of rock salt, magnesium sulphate or copper sulphate.

The earth lead should be of substantial gauge wire as short and as direct as possible, preferably with an insulated covering. The lead should always be soldered to the earth rod or plate and at any joints in its length, and protected where it enters the ground by a short length of suitable piping.

British Standard Code of Practice C.P.327.201 makes useful recommendations about the installation of earth connections for domestic sound and television receivers.

Radio-frequency Instruments

Where measuring instruments used at radio frequencies are concerned, the chief consideration is usually not so much the safety factor as the accuracy of the instrument. The primary object is to ensure that all the stray currents that inevitably flow through all the insulation across which a voltage exists in a radio-frequency system are directed into paths that have no unwanted effect on the indication of the instruments. The currents in question are the so-called capacitance currents. Any two metallic components separated by insulation of any sort are equivalent to a capacitor, and the insulation therefore always carries an alternating charge-and-discharge current whenever there is an alternating voltage between the components. This current is proportional to the frequency, and thus, although probably negligible at a frequency of, say, 50 c/s, is 100,000 times as great for a given voltage at 5 Mc/s, and possibly therefore quite considerable. Such stray currents, unless diverted away from the sensitive elements in a circuit, may cause instrument readings to be entirely false. The standard method of preventing any such current from reaching any particular circuit element is to enclose that element in an electrostatic screen—an enclosure the walls of which have the same voltage at every point. Ideally, the screen should be a continuous perfect conductor, but sometimes a wire mesh is a reasonably good substitute and may be necessary, if, for example, an instrument within must be visible from outside. Capacitance current from all external components must then remain outside the screen, and similarly capacitance currents from all inside components must remain inside the screen. However, it is not always a simple matter to secure that the screen of a radio-frequency instrument shall be in all its parts at earth potential. At low frequencies an earth-connection in the form of a cable of low resistance solidly connected to a water-main or an earth plate of large area well buried in moist subsoil is reliable; but at radio frequencies low resistance is not sufficient, for it is impedance that counts, and impedance arising from inductance is proportional to frequency, so that an earth cable of negligible impedance at low frequencies may behave as a choke of high impedance at high radio frequencies.

Any screen when completely insulated from earth behaves as one plate of a capacitor of which "earth" or surrounding objects form the other plate. The capacitance of this capacitor may be large relative to those in the radio circuits within it, and therefore of low impedance at the radio frequency. Thus a large electrostatic screen is automatically connected to earth by its own earth-capacitance, so far as radio-frequency voltages are concerned. Thus it often happens that, provided a radio-frequency instrument is mounted within a metallic screen, no connection to earth need be deliberately made. It may indeed be better without one, for the earth-capacitance of the screen in parallel with the inductance of an earth-lead may constitute a tuned circuit, the impedance of which is very much greater than that of the capacitance alone. At lower radio frequencies, however, when the circuit in question is far from resonance, an earth lead, if short and thick, may be very beneficial. The operator should therefore adopt whichever arrangement gives stable readings independent of movement of neighbouring objects and consistent with the theory of the instrument.

If the radio-frequency system also includes components such as power supplies of low frequency and high voltage, then the normal earth-connections required for safety must, of course, always be applied quite independently of the special considerations concerning the radio-frequency voltages and currents.

22. COMMUNICATION RECEIVERS

22. COMMUNICATION RECEIVERS

At one time a communication receiver could be described as a relatively cheap receiver, more expensive than a broadcast receiver, but very much cheaper than those designed for commercial organizations or for the Fighting Services. It normally covered the high-frequency band, 3–30 Mc/s and the medium-frequency broadcast band, and was developed largely to meet the demand by amateur radio operators.

During the last war, communication receivers were pressed into service for a great many purposes for which they were not originally designed, but nevertheless proved very successful, since they could be operated and maintained by comparatively unskilled staff and could be manufactured easily in large quantities.

Today, whilst the comparatively simple communication receiver is still being produced, there is a tendency to produce designs which give a very high degree of performance (chiefly in the matter of setting accuracy) but which are also very complicated both electrically and mechanically. Such receivers call for a very high standard of maintenance skill and, of course, are very expensive. It is proposed in this section to describe the more important design features of a receiver which may be said to fall between the very simple and cheap and the very complicated and expensive.

REQUIREMENTS FOR A COMMUNICATION RECEIVER

(1) It must be adaptable so that it can be used for almost any kind of service, telephone or telegraph.

(a) *Telephone.*—Sensitivity should be high so that a good signal-to-noise ratio is obtained from weak signals. The signal-to-noise ratio should improve linearly with increase in signal strength. The overall frequency response should be wide enough to avoid attenuating the higher-modulation frequ cies, but must be sharp enough to reject interfering adjacent signals. The frequency stability and automatic gain performance should be sufficiently high to enable the receiver to be used for rebroadcast purposes.

(b) *Telegraph.*—The sensitivity should be high and distortion of the signal pulse shape should not occur. The overall frequency response should not accept frequencies higher than the fifth harmonic of the keying speed. The frequency stability and overload characteristics should be sufficiently high to enable the receiver to be used for the operation of a teleprinter or similar apparatus.

(2) Operation should be simple, so that highly skilled personnel are not required. From this it follows that the number of controls should be minimized, and their use should be obvious without the need for elaborate instructions.

(3) Several degrees of selectivity should be provided so as to cater for telephone and telegraph reception under various conditions of noise and interference.

(4) The frequency change oscillators should be stable so as to minimize the amount of re-tuning necessary. The facility should also exist whereby the first oscillator can be crystal controlled.

(5) Spurious signals of all kinds should be negligible.

(6) The component parts should be reliable under extreme conditions of temperature and humidity, and access should be good for maintenance purposes.

(7) It should be possible to connect two similar receivers together for diversity working.

(8) The receiver should be comparatively cheap to produce. From this it follows that the preceding requirements are largely a matter of compromise, and must always be considered with the cost aspect in mind.

RECEIVER SENSITIVITY

This is usually expressed as the input necessary to produce a certain signal-to-noise ratio at the output. It is not sufficient to quote the input necessary to produce an output of say 100 mW, as this is a measure only of the overall receiver gain and takes no account of the amount of noise present in the output, which, of course, decides the readability of the signal.

Noise

Receiver noise can originate from the following sources :

(a) *Route Noise.*—This consists of bursts of atmospherics originating from certain known storm centres, and is, of course, particularly noticeable when an aerial beamed in the direction of a storm centre is used. The amount of noise is very variable, and it is not easy to express it in terms of a definite number of microvolts.

(b) *Site Noise.*—This is noise generated locally by power lines, ignition systems, electrical machinery of various kinds, medical apparatus and other sources. On an isolated country site such noise may be very small, aerials two or three miles from a town can pick up site noise of the order of 0·5–2 μV according to frequency, whilst in a town the noise may be so bad that only the strongest signals can be received effectively.

Ships are usually very bad from a noise point of view, due to the large amount of electrical machinery and the imperfect earthing of mast stays and derricks.

(c) *Thermal Agitation Noise.*—This occurs in all impedances, but is particularly important in the first tuned circuit of a receiver.

The value of such noise is given by the expression

$$E_p = 12{\cdot}6 \times 10^{-11} \times \sqrt{Z \times \delta f}$$

where E_p = noise equivalent in volts;
Z = circuit impedance at resonance;
δf = pass band of receiver in c/s.

(d) *Valve or Shot Noise.*—For convenience this is usually considered as occurring at the grid of the valve. It is usually expressed as the equivalent noise resistance, typical values are 1,000 ohms for a good radio-frequency pentode and 200,000 ohms for a pentagrid mixer.

(e) *Mains Hum.*—This is usually due to insufficient H.T. smoothing, and its removal does not represent a difficult problem.

A practical and detailed treatment of receiver noise has been covered by Zepler.[1]

Noise Factor

The actual signal voltage input necessary to produce a given output depends upon the input impedance: the noise output depends upon the pass band. These two facts make it difficult to compare readily the performance of different types of receivers on the basis of signal-to-noise ratio. To meet this problem the noise factor is sometimes used; it is defined as the ratio of the available signal-to-noise ratio at the input to the actual signal-to-noise ratio at the output of the receiver.

The determination of the noise factor of a receiver is a laboratory measurement, and great care has to be taken to prevent pick up of local noise. The method of measurement has been described by Bray and Lowry [2] and is also discussed in Section 42.

An ideal receiver, i.e., one which introduces no noise, has a noise factor of 0 db; a receiver in which the first-circuit noise completely predominates can have a noise factor of 3 db if the input impedance is matched to the first-circuit impedance. It is possible, by overmatching the input impedance, to obtain a noise factor less than 3 db, but this is of value only on ideal noiseless sites. The noise factor for a good commercial point-to-point high-frequency receiver varies from 3 to 6 db over the range 3–30 Mc/s: for a good communication-type receiver the corresponding figures are 4–10 db.

SPURIOUS SIGNALS

Oscillator Harmonics

Harmonics of the second oscillator in double-conversion superheterodyne receivers can be particularly troublesome. It is advantageous to use an oscillator circuit which is relatively free from harmonics, but it must be remembered that the mixer generates harmonics even if fed with a pure sine-wave input. Great attention should be paid therefore to the screening and decoupling of both the oscillator and mixer circuits.

Image Signal

If the first oscillator is set to frequency f_1 and the first intermediate frequency is f_2, then signals of frequency $(f_1 + f_2)$ and $(f_1 - f_2)$ will both be accepted by the intermediate-frequency circuits. Normally, in a ganged receiver, the signal-frequency circuits will be tuned to $(f_1 - f_2)$, in which case $(f_1 + f_2)$ is the image signal. The protection against the image signal depends upon the frequency response and ganging accuracy of the high-frequency circuits.

A spurious signal, which is occasionally confused with the true image signal, is one which differs from the first oscillator frequency by half the first intermediate frequency; this is sometimes known as the "half I.F. signal". A very strong signal of frequency $(f_1 \pm \frac{1}{2} \times f_2)$ may cause overloading, so that second harmonics are produced at the mixer of frequency $(2 \times f_1 \pm f_2)$, and these beat with the second harmonic, $2 \times f_1$, of the first oscillator to produce signals at frequency f_2.

Intermediate Frequency

Normally, the signal-frequency circuits offer sufficient protection against break-through of signals at the intermediate frequency, but it does occasionally happen that the frequency of a strong local transmitter coincides with one of the receiver intermediate frequencies; this case is usually a matter for local treatment.

Blocking

A very strong signal can cause overloading of the mixer stage or even the second signal-frequency amplifier, but due to the main intermediate-frequency selectivity the signal is not audible at the output. The effect of the overloading is to reduce the gain of the stage concerned, so that if the interfering signal is a keyed continuous-wave transmission, the level of the wanted signal follows the keying speed.

This reduction of the level of the wanted signal is known as blocking, and the usual method of measurement is to determine the input necessary 10 kc/s from the wanted signal to reduce the level of the latter by 3 db.

Cross Modulation

This effect is similar to blocking, but occurs when the interfering signal is modulated. The gain of the overloaded stage is varied at the modulation frequency, which is then audible at the receiver output.

Intermodulation

Various forms of intermodulation can arise, and are sometimes difficult to recognize. A typical instance is where a strong carrier acts as the first-frequency change source, so that the first-oscillator tuning has no effect on the audible output.

CALCULATION OF SIGNAL-TO-NOISE RATIO

Consider first the case where the first-circuit noise is very much larger than the valve noise, and the input is accurately matched.

If R is the source (aerial) impedance, and n is the step-up ratio, then the first-circuit impedance is n^2R, and if E is the signal voltage generated in the aerial, then the signal voltage across the first tuned circuit is $\frac{nE}{2}$.

$$\text{The source noise in volts} = 12{\cdot}6 \times 10^{-11} \times \sqrt{\delta f} \times \sqrt{R}$$
$$= k\sqrt{R}$$

where k is a constant and E = source signal.

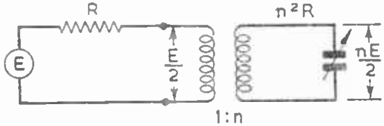


Fig. 1.—Calculation of the Signal Voltage across the First Tuned Circuit.

$$\text{Source signal-to-noise ratio} = \frac{E}{k\sqrt{R}}$$

The signal voltage across the first tuned circuit is $\frac{nE}{2}$, and the impedance, with the aerial connected, is $\frac{n^2R}{2}$.

The circuit noise is $k\sqrt{\frac{n^2R}{2}}$, and the signal-to-noise ratio is $\frac{1}{\sqrt{2}}\cdot\frac{E}{k\sqrt{R}}$.

The noise factor then, which is the ratio of the source or input signal-to-noise ratio to the receiver signal-to-noise ratio, is $\sqrt{2}$ or 3 db.

Now c[illegible] the practical case shown in Fig. 2.

In this [illegible] the following typical figures are assumed:

Frequency = 20 Mc/s
Tuning capacitance = 100 pF
Circuit Q = 50 (aerial disconnected)
Equivalent noise resistance for radio-frequency valves = 1,000 ohms = 0·3 μV for 6-kc/s band-width
Equivalent noise resistance for mixer stage = 200,000 ohms = 4·2 μV for 6-kc/s band-width
Radio-frequency stage gain = 5

$$\text{First-circuit impedance} = Q\frac{1}{\omega C}$$
$$\simeq 4{,}000 \text{ ohms}$$

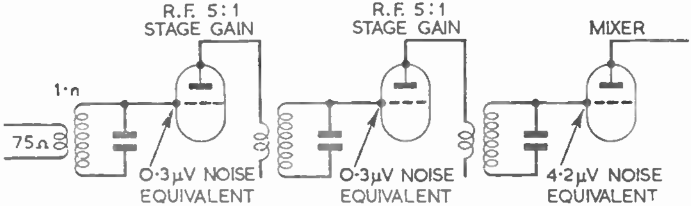


Fig. 2.—The Signal-to-Noise Ratio through the First Stages of the Receiver.

and the noise voltage (6-kc/s band-width) $\simeq$ 0·6 μV. Wh n the aerial is connected, the impedance falls to 2,000 ohms, and the noise voltage is then 0·42 μV.

The noise of the second radio-frequency amplifier and mixer stages can be reflected back to the first radio-frequency amplifier grid, so that the total noise is the square root of the sum of the squares of the individual noise equivalents.

$$
\begin{aligned}
(\text{Total noise})^2 &= (\text{1st circuit})^2 + (\text{1st radio-frequency})^2 \\
&\quad + \left(\frac{\text{2nd radio-frequency}}{\text{gain of one R.F. stage}}\right)^2 + \left(\frac{\text{mixer}}{\text{gain of two R.F. stages}}\right)^2 \\
&= (0{\cdot}42)^2 + (0{\cdot}3)^2 + \left(\frac{0{\cdot}3}{5}\right)^2 + \left(\frac{4{\cdot}2}{25}\right)^2 \\
&= 0{\cdot}18 + 0{\cdot}09 + 0{\cdot}004 + 0{\cdot}028 \\
&\doteq 0{\cdot}302
\end{aligned}
$$

$$\therefore \quad \text{Total noise} = 0{\cdot}55\ \mu\text{V}.$$

It will be noticed that with the stage gain considered, the noise of the mixer has a second-order effect on the total noise. For the circuit impedance considered, the influence of the first radio-frequency amplifier valve noise is not a major one, but it will be seen that if the circuit impedance is lowered, then the valve noise has a bigger effect and the step-up ratio will be lowered.

The step-up ratio of the input circuit is $\sqrt{\frac{4000}{75}} = 7{\cdot}3$, so that for a signal of 1 μV across the receiver-aerial input terminals (or 2 μV generated in the aerial) the signal-to-noise ratio will be $\frac{7{\cdot}3}{0{\cdot}55} = 13{\cdot}3$, which is equal to 23 db.

Now consider that the signal strength has been increased by 56 db and that the action of the automatic-gain-control system has reduced the gains of the radio-frequency and intermediate-frequency amplifiers by 28 db each, so that the receiver output has remained constant. In practice, of course, there will be a slight increase in the receiver output.

The gain of each of the radio-frequency stages will now be unity, so that $(\text{Total noise})^2 = (0{\cdot}42)^2 + (0{\cdot}3)^2 + (0{\cdot}3)^2 + (4{\cdot}2)^2$.

Thus the total noise = 4·24 μV at 1st R.F. valve grid.

$$
\begin{aligned}
\text{Input signal} = 56 \text{ db above } 1\ \mu\text{V} &= 625\ \mu\text{V at receiver input} \\
&= 625 \times \text{step-up ratio at grid.}
\end{aligned}
$$

$$
\begin{aligned}
\text{Signal-to-noise ratio} &= \frac{625 \times 7{\cdot}3}{4{\cdot}24} = 1076 \\
&= 60 \text{ db}
\end{aligned}
$$

It will be noted that the noise of the mixer stage now predominates, and that an increase in signal strength of 56 db has produced an increase in signal-to-noise ratio of only 37 db.

It will be appreciated from the foregoing example that the smallest possible proportion of the total automatic-gain-control action should be applied to the radio-frequency circuits, the exact amount being decided by the overload point of the mixer and the required overall automatic-gain-control characteristic.

Calculation of Image Signal Protection

The response of a tuned circuit is given by

$$\frac{I}{I_{max.}} = \frac{1}{\sqrt{1 + (yQ)^2}}$$

where $$y = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}$$

for a small percentage mistune

$$y = \frac{2(\omega - \omega_0)}{\omega_0} = \frac{2 \times \text{percentage mistune}}{100}$$

for large percentage mistune

$$\frac{I}{I_{max.}} = \frac{1}{yQ}$$

Consider now the case given by Fig. 2 and an intermediate frequency of 1·6 Mc/s.

Then for the input circuit, with aerial connected,

$$Q = 25$$
$$\omega = 20 \text{ Mc/s}$$
$$\omega_0 = 23{\cdot}2 \text{ Mc/s}$$
$$y = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} = 0{\cdot}3$$
$$yQ = 7{\cdot}5$$

$$\frac{I}{I_{max.}} = \frac{1}{\sqrt{1 + (7{\cdot}5)^2}} = \frac{1}{\sqrt{1 + 56{\cdot}25}} = \frac{1}{7{\cdot}57}$$

The protection then is 7·57 to 1 or 18 db.

For the other circuits $Q = 50$, so that they each offer a protection of 24 db.

Total image protection = 18 + 24 + 24 = 66 db.

In practice, this figure may be modified by ganging errors and possibly a small amount of reaction. The effect of the latter will be to improve the image protection at the ganging points, but may degrade the protection at other points.

CALCULATION OF OSCILLATOR TRACKING

The ganged variable capacitor used in a high-frequency receiver is usually fitted with identical sections for the signal-frequency and oscillator circuits, and it is necessary to modify the law of the oscillator

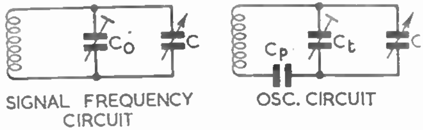


FIG. 3.—BASIC ARRANGEMENT FOR OSCILLATOR TRACKING.

section so that the resonant frequency is higher than the signal frequency by the value of the first intermediate frequency. This is achieved by using different values of inductance and parallel capacitance and adding a series capacitance as shown in Fig. 3.

In Fig. 3, C_t is the parallel capacitor, usually referred to as the trimmer. C_p is the series capacitor, usually referred to as the tracker or padder. C is the variable capacitor, identical with the signal-frequency capacitor.

It is not possible to obtain accurate tracking over the whole of the variable capacitor sweep, but the method gives accurate tracking at three points. Probably the best tracking points are those which give the minimum value of peak-tracking error between the tracking points. If the lowest signal frequency is f_4 and the highest is f_0, then a good empirical method is to take the geometric mean $\sqrt{f_0 f_4}$ for the centre point (f_2) and $f_0 - \dfrac{f_0 - f_2}{4}$ and $f_4 + \dfrac{f_2 - f_4}{4}$ for the others (f_1 and f_3 respectively).

Knowing the sweep of the variable capacitor, the settings of the latter for f_1, f_2 and f_3 can be calculated and three equations obtained, the unknowns being C_t, C_p and L. The solution of these equations is not easy, and a comparatively easy series of formulæ has been quoted by Zepler.

$$\text{Then trimmer capacitance} = C_t = \frac{C_3(\alpha - 1) - C_1(n - 1)}{n - \alpha}$$

$$\text{Padder capacitance} = C_p = \frac{\dfrac{C_1 + C_t}{n - \alpha}}{\dfrac{\;}{(n-1)(1-p_1)}} - 1$$

where
$$p_1 = \left(\frac{f_3'}{f_1'}\right)^2$$
$$p_2 = \left(\frac{f_2'}{f_1'}\right)^2$$
$$\alpha = \frac{1 - p_1}{1 - p_2}$$
$$n = \frac{C_3 - C_1}{C_2 - C_1}$$

Take as an example a signal-frequency range of 1·425–2·89 Mc/s, an intermediate frequency of 1·2 Mc/s and a variable capacitor sweep of 163·8 pF. Tracking points to be 1·5, 2·02 and 2·75 Mc/s.

Then $f_0 = 2{\cdot}89$ Mc/s, $f_1 = 2{\cdot}75$ Mc/s, $f_2 = 2{\cdot}02$ Mc/s, $f_3 = 1{\cdot}50$ Mc/s, $f_4 = 1{\cdot}425$ Mc/s.

Intermediate frequency $= 1{\cdot}2$ Mc/s, $f_1' = 3{\cdot}95$ Mc/s, $f_2' = 3{\cdot}22$ Mc/s, $f_3' = 2{\cdot}70$ Mc/s

TABLE 1.—THREE-POINT TRACKING CALCULATIONS

Given	*Highest Frequency*	*Tracking Frequencies*			*Lowest Frequency*
Signal frequencies	f_0	f_1	f_2	f_3	f_4
Signal tuning capacitance	C_0	$C_0 + C_1$	$C_0 + C_2$	$C_0 + C_3$	$C_0 + C_4$
Intermediate frequency	f	—	—	—	—
Oscillator frequency	—	$f_1 + f$	$f_2 + f$	$f_3 + f$	—
Oscillator frequency	—	f_1^1	f_2^1	f_3^1	—

$$C_0 = \frac{C_4}{\left(\frac{f_0}{f_4}\right)^2 - 1} = \frac{163\cdot8}{3\cdot11308} = 52\cdot617 \text{ pF}$$

$$C_1 = C_0\left(\left(\frac{f_0}{f_1}\right)^2 - 1\right) = \cdot494 \text{ pF}$$

$$C_2 = C_0\left(\left(\frac{f_0}{f_2}\right)^2 - 1\right) = 55\cdot084 \text{ pF}$$

$$C_3 = C_0\left(\left(\frac{f_0}{f_3}\right)^2 - 1\right) = 142\cdot699 \text{ pF}$$

$$1 - p_1 = 1 - \left(\frac{2\cdot70}{3\cdot95}\right)^2 = 0\cdot5328$$

$$1 - p_2 = 1 - \left(\frac{3\cdot22}{3\cdot95}\right)^2 = 0\cdot3355$$

$$\alpha = \frac{1 - p_1}{1 - p_2} = 1\cdot5881$$

$$n = \frac{C_3 - C_1}{C_2 - C_1} = 2\cdot7668$$

$$C_t = \frac{C_3(\alpha - 1) - C_1(n - 1)}{n - \alpha}$$

$$\underline{C_t = 62\cdot97 \text{ pF}}$$

$$C_p = \frac{C_1 + C_t}{\frac{n - \alpha}{(n - 1)(1 - p_1)} - 1}$$

$$\underline{C_p = 271\cdot5 \text{ pF}}$$

The misganging, in terms of frequency error, has been calculated and the results shown in Fig. 4. The corresponding loss in gain obviously depends upon the number and the Q of the tuned circuits employed.

The foregoing calculations are based on the assumption that ideal circuit conditions exist, and in practice it is probable that some experimental adjustment of the tracking and trimming values may be necessary.

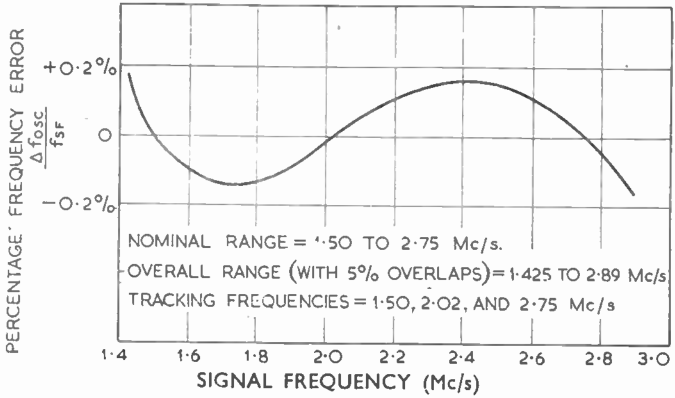


FIG. 4.—OSCILLATOR TRACKING MISGANGING.

In the example quoted, the tracking capacitor is comparable in value with the variable capacitor, and a value of tolerance of the order of 5 per cent should be specified. For a signal-frequency range of 16–30 Mc/s, for the same intermediate frequency the tracking capacitor is 2,810 pF, and is not so critical in value—say within ±15 per cent.

Oscillator Frequency–Temperature Characteristic

Most components have a natural tendency to increase in value with increase in temperature. This means that the tuning of a receiver drifts during the warm-up period and again during any subsequent ambient temperature changes. The simplest method of compensation is to use a capacitor with a high negative-temperature coefficient for the whole or part of the trimmer capacitor. This gives good compensation at the middle of the band, but over-compensates at the high-frequency end and under-compensates at the low-frequency end of the band. The best form of capacitance compensation uses a bimetal strip which bends with temperature and so varies the spacing between the plates of an air dielectric condenser.

The highest temperature coefficient obtainable from a fixed capacitor is about 700 parts in 10^6 negative per degree Centigrade, whereas with a bimetal compensator a figure of 5,000 parts in 10^6 can be achieved. The compensator may, of course, be designed to give positive or negative coefficient.

Temperature compensation of the tuning inductance, whilst very desirable, is not easy to achieve cheaply in practice. A comparatively easy alternative, which gives a reasonable compromise, is to fit a high-coefficient capacitor as a part or whole of the tracking capacitor.

Many components, apart from changing their value during a heat cycle, do not return to their original values on the completion of the cycle. Obviously, components should be checked for this effect before being included in an oscillator design. It is not difficult to find designs of fixed and variable capacitors which are cyclic, but great trouble is necessary in the coil designs. It is not uncommon to find that a batch of apparently identical coils shows a wide variation in performance. This can be due, for instance, to non-uniformity of winding tension. In the case of random-wound coils, it has been found in practice that best stability figures are obtained when the minimum of winding tension is used.

INTERMEDIATE-FREQUENCY AMPLIFICATION

The choice of frequency for the intermediate-frequency amplifier is influenced by two main considerations: first, it should be as low as possible so that good selectivity can be easily obtained; and secondly, it should be as high as possible in order that the image signal response is negligible. These two requirements are in direct conflict, and probably the best solution is to cater for them independently by employing two frequency changes. The first intermediate-frequency circuits can be centred within a few hundred kilocycles of the lowest signal frequency, and need consist of only one pair of coupled circuits feeding straight to the second frequency changer. This scheme has the further advantage that the second oscillator can be made variable over say ± 5 kc/s of the centre frequency, thus providing a very effective bandspread device. The band-width of the first intermediate-frequency circuits must be wide enough to accept this bandspread in addition to the band-width of the second intermediate frequency, but this does not normally present any difficulty.

The easiest method of obtaining the necessary selectivity for the second intermediate-frequency amplifier is to use pairs of mutually coupled tuned circuits between valve stages, the degree of coupling depending upon the circuit Q and the band-width required.

Intermediate-frequency Stage Gain with Coupled Circuits

If we consider the circuit shown in Fig. 5, where

d = damping of each tuned circuit $= \dfrac{1}{Q}$

I = Anode current $= G_m E_g$

E_2 = secondary voltage

Z = impedance of each circuit at resonance $= \omega_0 LQ$

$y = \dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega} \simeq \dfrac{2\Delta F}{F_0}$

ΔF = difference between resonant frequency F_0 and any other frequency at which gain is required,

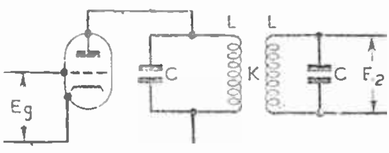


FIG. 5.—STAGE GAIN WITH COUPLED CIRCUITS.

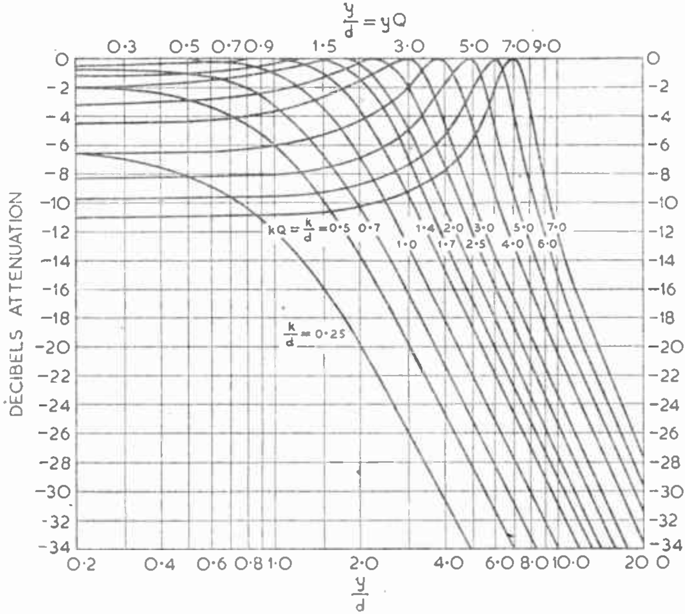


FIG. 6.—RESONANCE CURVES OF COUPLED CIRCUITS.

then $E_2 = \dfrac{IZkQ}{\sqrt{(k^2Q^2 + 1 - y^2Q^2)^2 + 4y^2Q^2}}$

At mid-band $y = 0$ and E_2 then $= IZ \cdot \dfrac{kQ}{1 + k^2Q^2}$

Resonance Curves of Coupled Circuits

The product kQ is known as the coupling factor, and for $kQ = 1$ the resonance curve of two coupled circuits is a maximum at the resonant frequency. For $kQ > 1$, then, the resonance curve has a minimum at the resonant frequency and two peaks or maxima on either side, the spacing being given by $y = \pm\sqrt{k^2 - \dfrac{1}{Q^2}}$.

The coupling factor is said to be critical when $kQ = 1$.

The curves shown in Fig. 6 give the attenuation with respect to the mid-band gain of a pair of critically coupled circuits plotted against yQ for various values of kQ.

When several degrees of selectivity are required it is necessary to vary the coupling for some, if not all, of the pairs of circuits. This can be done by mechanical movement of the coils. It has the advantage that the selectivity is continuously variable, but is sometimes difficult to

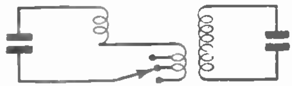


FIG. 7.—ARRANGEMENT FOR OBTAINING DEGREES OF SELECTIVITY FROM COUPLED CIRCUITS.

arrange. The scheme shown in Fig. 7 has proved satisfactory in practice, and has the merit that the response curves are symmetrical about a common centre frequency.

AUTOMATIC GAIN CONTROL

The purpose of the automatic-gain-control system is to prevent overloading due to strong signals and to maintain an output level which is reasonably constant even though the input signal may be varying in strength. Obviously, the gain of the receiver should not be affected when weak signals are being received so that the ideal automatic-gain-control curve should be as Fig. 8.

It is also important that the noise output should progressively decrease with increase of signal once the automatic-gain-control system is operative. Ideally, this relationship should be linear.

A simple automatic-gain-control circuit is shown in Fig. 9. The output from the final intermediate-frequency stage is fed to a diode, and a positive delay voltage is applied to the diode, so determining the point of operation of the system. Decoupling is provided to remove the intermediate- and low-frequency components of the signal and to provide a suitable time constant.

Supposing it is desired that an increase in signal strength of 60 db above the operational point should cause an increase of 6 db in receiver

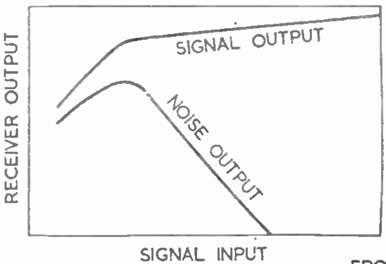


FIG. 8.—IDEAL AUTOMATIC-GAIN-CONTROL CURVE.

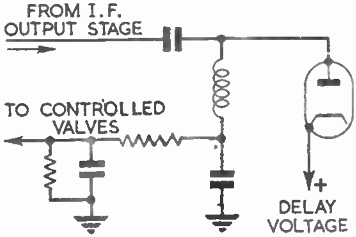


FIG. 9.—A SIMPLE AUTOMATIC-GAIN-CONTROL CIRCUIT.

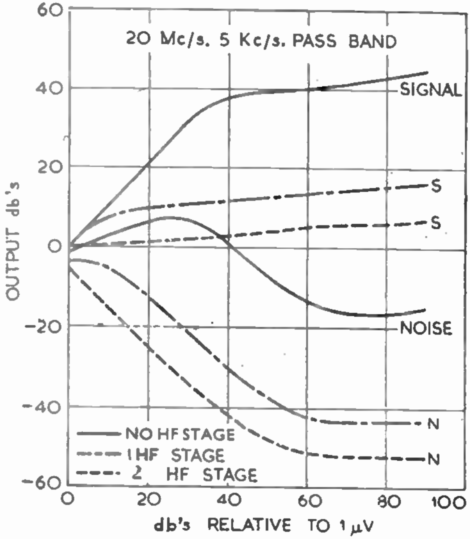


Fig. 10.—Typical Automatic-gain-Control Performance Curves.

output, then if the delay voltage is x, the peak signal voltage must also be x at the operational point. When the signal strength rises by 60 db, the signal level at the diode will rise by 6 db and the delay voltage will be

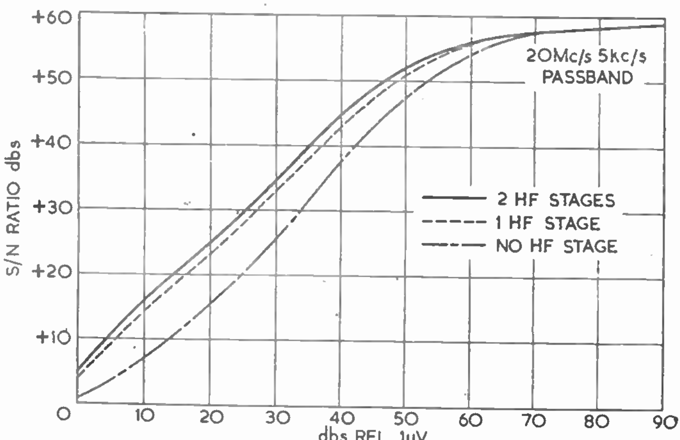


Fig. 11.—Analysis of the Curves in Fig. 10 in Terms of Signal-to-noise Ratio.

exceeded by x, and this is used to reduce the gain of the control stages by 54 db.

Now supposing the delay voltage is increased to $2x$ and additional intermediate-frequency gain introduced so that the operational point is unchanged in terms of signal input. An increase of signal level of 6 db at the diode will now produce $2x$ control volts, and the original reduction in gain of 54 db will be substantially increased.

It will be seen, then, that the performance of the automatic-gain-control system depends upon the delay voltage.

The simple scheme as described has two main disadvantages: first, it is difficult to obtain the high level necessary in the intermediate-frequency amplifier without experiencing overload troubles; and, secondly, as there is direct coupling between the signal and automatic-gain-control diodes there is a danger of the automatic-gain-control circuits being operated by the beat-frequency oscillator unless the level is kept low, in which case strong continuous-wave signals may not be fully modulated.

These disadvantages can be overcome by feeding the automatic-gain-control diode from a separate valve, whose input is in parallel with that of the last intermediate-frequency valve.

In Fig. 10 are shown automatic-gain-control performance curves for three receivers with identical circuitry from the first frequency changer onwards, but with different radio-frequency amplifiers. Fig. 11 shows the result of analysing Fig. 10 in terms of signal-to-noise ratio, and it will be noticed that one radio-frequency stage shows an appreciable improvement compared with no radio-frequency stage, but that the addition of a further radio-frequency stage shows little improvement in signal-to-noise ratio at the frequency considered.

A TYPICAL RECEIVER DESIGN

It will be of value to discuss the main features of a typical modern double-conversion communication receiver, though it should be appreciated that considerable variations may be met in practice.

A skeleton circuit is shown in Fig. 12, and it will be noted that it is a double superheterodyne. The first intermediate frequency is centred at 1·2 Mc/s to give a high value of image signal protection, and the second intermediate frequency is 100 kc/s and provides the main adjacent channel selectivity. Continuous tuning is provided by ganging the signal-frequency circuits to that of the variable first oscillator, but up to a maximum of six preset frequencies are available by substituting a crystal oscillator for the variable oscillator.

Radio-frequency Circuit Design

It is advantageous to cover the whole frequency range in as many bands as is practical, for the following reasons:

(1) The circuit impedances are subject to less variation and are higher than when fewer bands are used. This is of particular value at the higher frequencies, where it is normally difficult to make the first-circuit noise high in comparison with the first-valve noise. It is also easier to obtain the stage gain necessary so that the relatively high mixer noise does not degrade the signal-to-noise ratio.

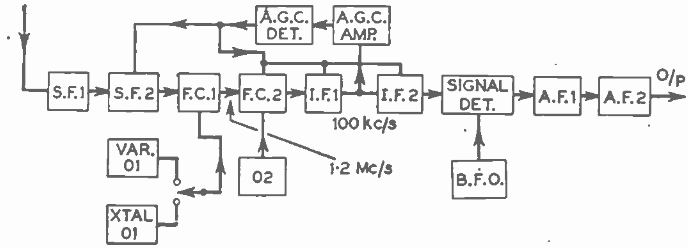


FIG. 12.—BLOCK DIAGRAM OF A TYPICAL DOUBLE SUPERHETERODYNE COMMUNICATION RECEIVER.

(2) Misganging errors are minimized.

(3) The calibration scale is more open, and can be read to a greater degree of accuracy.

It follows, of course, that the stray circuit capacitance should be kept as low as possible. This involves detailed attention to the mechanical layout and run of wiring, but it should not be difficult to attain a figure of 40 pF for the stray capacitances, so that using a variable capacitor of 164 pF sweep, a range factor of 2 to 1 can be obtained. Allowance must be made for the parallel trimming capacitors and for adequate overlaps between ranges.

The coils used have adjustable dust-iron cores to facilitate ganging, and should have a high-Q value so as to maintain a high circuit impedance and to provide the minimum response to the image signal. A free coil may have a Q value of 100, but the effect of the wiring, switches, etc., can reduce the Q to 50, and in the case of the input circuit this can be further reduced to 25 by the aerial damping. There is little point, then, in devoting effort to the design of a high-Q coil unless attention is paid also to the associated circuitry.

For best performance, two radio-frequency stages of amplification are normally used; this has the effect of ensuring that the gain is sufficient to mask the mixer noise and that the number of tuned circuits provides a good value of image-signal protection. If the circuit impedances are high, two radio-frequency stages may not show much improvement over one radio-frequency stage in the matter of sensitivity, the only gain being in image-signal protection. It is possible, of course, to couple two tuned circuits together without the aid of a valve, but it is difficult to maintain the correct coupling over a frequency band, and valve coupling is much easier.

A four-gang capacitor tunes the two radio-frequency stages, the mixer and the oscillator, circuits simultaneously. The coils not in use are shorted out, so that unwanted resonance effects are avoided. Automatic gain control is applied to the second radio-frequency amplifier stage, which is a variable-mu type valve. The first stage has a short-grid-base valve of low noise equivalent.

The variable first oscillator uses a tuned-anode circuit, and is coupled through a small capacitor (C26) to the pentagrid mixer; alternatively, a triode hexode could, of course, be used as a mixer. When using the crystal oscillator, the variable oscillator stage is used as an amplifier or as a multiplier at the higher frequencies.

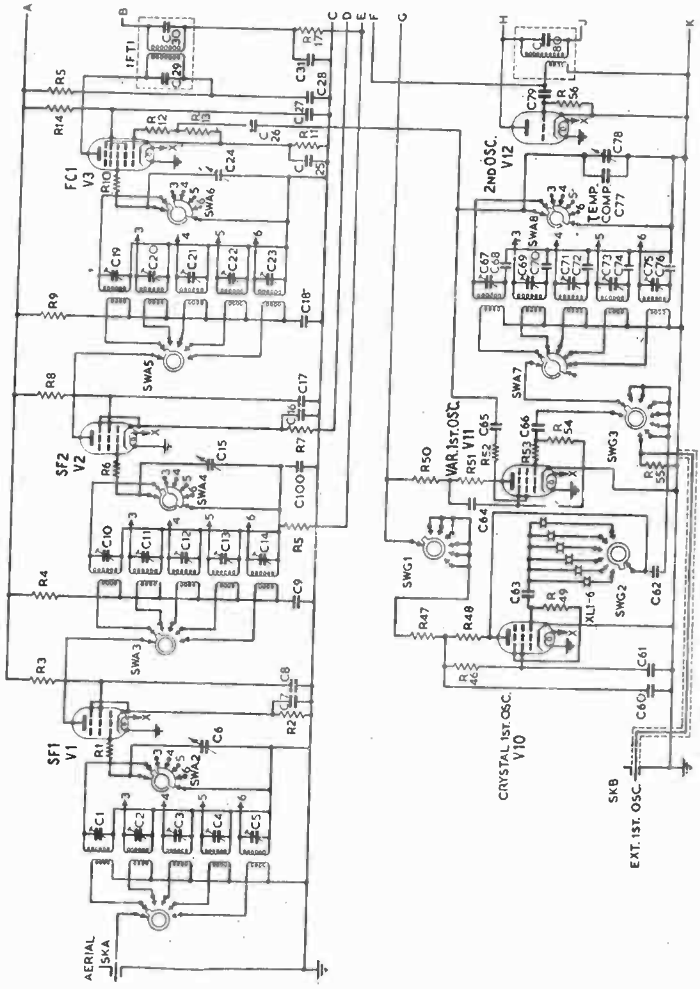


FIG. 13 (a).—R.F. AMPLIFIERS AND FREQUENCY CHANGERS OF TYPICAL DOUBLE-CONVERSION COMMUNICATION RECEIVER.

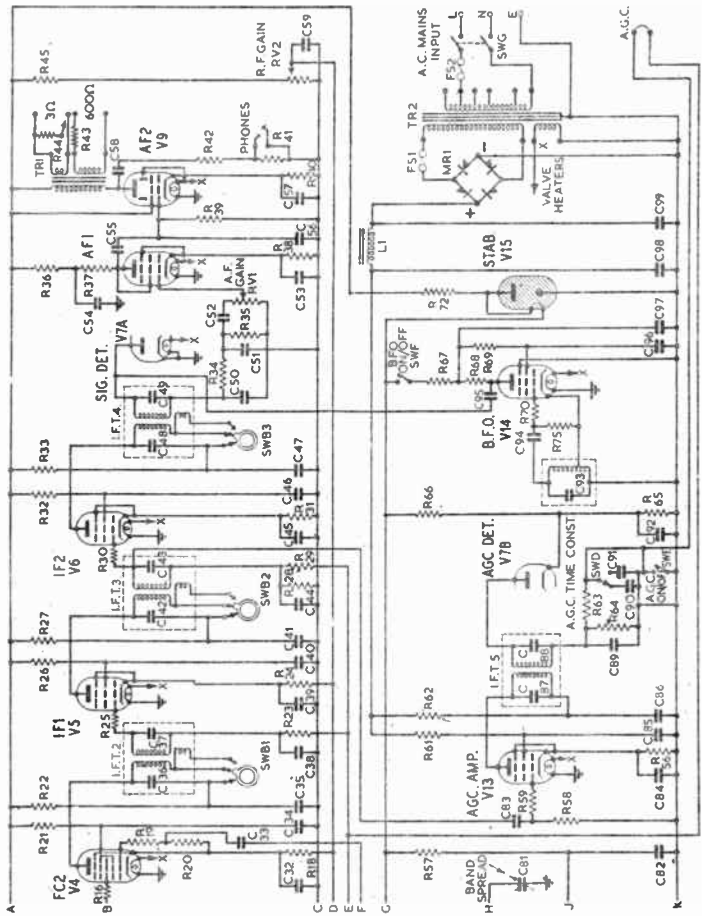


FIG. 13 (*b*).—INTERMEDIATE-FREQUENCY AND AUDIO-FREQUENCY SECTION OF TYPICAL DOUBLE-CONVERSION COMMUNICATION RECEIVER.

TABLE 2.—FREQUENCY COVERAGE OF TYPICAL RECEIVER

Band	*Nominal (Mc/s)*	*Actual including Overlaps (Mc/s)*
1	1·5–2·75	1·425–2·89
2	2·75–5·0	2·61–5·25
3	5·0–9·0	4·75–9·45
4	9·0–16·5	8·55–17·3
5	16·5–30	15·7–31·5

VALVE AND COMPONENT DATA, SEE FIG. 13.

Valve Types.

V1	CV138
V2	CV131
V3	CV453
V4	CV453
V5	CV131
V6	CV131
V7	CV140
V8	CV138
V9	CV2136
V10	CV138
V11	CV138
V12	CV133
V13	CV131
V14	CV131
V15	CV216

Resistors.

R1	22
R2	330
R3	47k
R4	10k
R5	47k
R6	22
R7	330
R8	47k
R9	10k
R10	22
R11	330
R12	22
R13	22k
R14	33k
R15	10k
R16	100
R17	47k
R18	330
R19	100
R20	22k
R21	33k
R22	10k
R23	47k
R24	1k
R25	100
R26	47k
R27	10k
R28	2·2M
R29	3·3M
R30	100
R31	330
R32	47k
R33	10k
R34	47k
R35	220k
R36	47k
R37	47k
R38	1k
R39	470k
R40	270
R41	4·7k
R42	47k
R43	560
R44	3·3
R45	68k
R46	47k
R47	68k
R48	22k
R49	47k
R50	1k
R51	10k
R52	22
R53	22
R54	10k
R55	75
R56	100k
R57	150k
R58	470k
R59	100
R60	330
R61	47k
R62	10k
R63	1M
R64	100k
R65	10k
R66	150k
R67	33k
R68	22k
R69	33k
R70	100
R71	47k
R72	3·9k

Capacitors.

C1–C5	3–30 pF
C6	164 pF sweep
C7	0·01
C8	0·01
C9	0·01
C10–C14	3–30 pF
C15	164 pF sweep
C16	0·01
C17	0·01
C18	0·01
C19–C23	3–30 pF
C24	164 pF sweep
C25	0·01
C26	10 pF
C27	0·01
C28	0·1
C29	560 pF
C30	560 pF
C31	0·01
C32	0·1
C33	100 pF
C34	0·1
C35	0·1
C36	330 pF
C37	330 pF
C38	0·01
C39	0·1
C40	0·1
C41	0·1
C42	330 pF
C43	330 pF
C44	0·01
C45	0·1
C46	0·1
C47	0·1
C48	330 pF
C49	330 pF
C50	220 pF
C51	220 pF
C52	0·01
C53	2
C54	2
C55	0·01
C56	470 pF
C57	2
C58	0·1
C59	0·1
C60	0·1
C61	0·1
C62	47 pF
C63	22 pF
C64	0·01
C65	100 pF
C66	100 pF
C67	3–30 pF
C68	2,700 pF
C69	3–30 pF
C70	1,500 pF
C71	3–30 pF
C72	876 pF
C73	3–30 pF
C74	485 pF
C75	3–30 pF
C76	270 pF
C77	Temperature compensator
C78	164 pF sweep
C79	100 pF
C80	270 pF
C81	3–10 pF
C82	0·1
C83	100 pF
C84	0·1
C85	0·1
C86	0·1
C87	330 pF
C88	330 pF
C89	470 pF
C90	0·2
C91	0·5
C92	0·1
C93	1,000 pF
C94	100 pF
C95	22 pF
C96	0·1
C97	0·1
C98	8
C99	8
C100	0·01

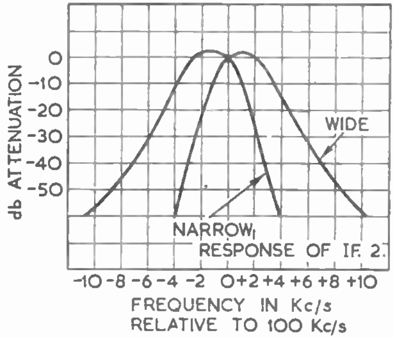


FIG. 14.—SECOND INTERMEDIATE-FREQUENCY RESPONSE CURVE.

The intermediate-frequency output from the first mixer at 1·2 Mc/s is taken to the second mixer via the double-tuned circuits (IFT1). The second oscillator is variable to the extent of ±5 kc/s, and this provides the band-spread control.

The second intermediate-frequency amplifier employs two variable-mu valve stages and a total of six tuned circuits. Two degrees of coupling provide band-widths of 1·5 kc/s for telegraphy and 6 kc/s for telephony. The mid-frequency is 100 kc/s, and is sufficiently low to enable good selectivity to be obtained without the use of a crystal-gate.

Audio-frequency Stages

Diode detection is used, the load being provided by R35, and the intermediate-frequency filter by R34, together with capacitors C50 and C51. The degree of intermediate-frequency filtering must not be so high as to affect the modulation frequencies. The output from the signal diode is taken to the first audio-frequency stage via the audio-frequency gain control (RV1) and thence to the output stage. Three audio-frequency outputs are shown, these being for loudspeaker, 600-ohm line and headphone monitoring The line output is normally limited to about 10 mW, as the maximum line level permitted by the G.P.O. is about 5 mW.

A separate beat-frequency oscillator is coupled to the signal detector for continuous-wave reception, and is pre-set to give a beat note of 1 kc/s.

An automatic-gain-control amplifier feeds the automatic-gain-control detector, the delay voltage being supplied by the potentiometer formed by R65 and R66, and the time-constant being selected by the switch SWD. The time-constants chosen are 0·2 seconds for telephony and 0·5 seconds for telegraphy. Resistance values for R63 and R64 are 1 MΩ and 100 kΩ respectively, so that there is no appreciable difference between the "on" and the "off" time-constants. If a short "on" time-constant is used, a strong noise element can operate the automatic-gain-control system, so that several telegraph characters may be rendered inaudible, even though the duration of the noise element may be so short that the noise itself cannot be heard.

The manual gain control operates on both the radio-frequency and intermediate-frequency amplifiers simultaneously. A separate radio-frequency gain control is sometimes adopted, and is particularly useful in preventing cross-modulation due to strong signals; such a control is frequently misused, however, and causes degradation of the signal-to-noise ratio, so that on balance it is probably best omitted.

The power supply is conventional; it uses a metal rectifier on the score of economy of maintenance as compared with a valve rectifier. The smoothed H.T. is about 250 volts, and is stabilized for supply to the oscillators and automatic-gain-control delay potential.

In common-frequency, spaced-aerial diversity working, it is usual to common the automatic-gain-control systems, so that the receiver with the strongest signal controls the gain of the other receiver, or receivers, and reduces the unwanted noise. It is also advantageous to use a common first-frequency-change oscillator, as this simplifies the location of the signal and subsequent monitoring.

The receiver circuit as given in Fig. 13 is intentionally simple for ease of presentation, obviously it can be elaborated by the addition of more intermediate-frequency pass-bands, a distortionless audio-frequency amplifier, valve-feed metering and the like.

Estimated Performance

The following information gives guidance as to the standard of performance that could be expected in practice with this receiver:

(1) *Noise Factor.*—Band 1, 4 db; Band 2, 4 db; Band 3, 5 db; Band 4, 6·5 db; Band 5, 9 db.

(2) *Signal-to-noise Ratio.*—(*a*) Signals modulated 30 per cent at 400 c/s to give 10 db signal-to-noise ratio with a pass-band of 6 kc/s.

(*b*) Continuous-wave signals to give 20 db signal-to-noise ratio with a pass-band of 1·5 Kc/s.

Band	*Signals as* (a)	*Signals as* (b)
1	1·45 μV	0·7 μV
2	1·45 μV	0·7 μV
3	1·6 μV	0·8 μV
4	1·9 μV	0·9 μV
5	2·6 μV	1·2 μV

(3) *Image Protection.*—Band 1, 100 db; Band 2, 100 db at 2·75 Mc/s, 85 db at 5 Mc/s; Band 3, 95 db at 5 Mc/s, 72 db at 9 Mc/s; Band 4, 80 db at 9 Mc/s, 50 db at 16·5 Mc/s; Band 5, 50 db at 16·5 Mc/s, 35 db at 30 Mc/s.

(4) *Automatic-gain-control.*—Two time constants—0·2 and 0·5 seconds. The action of the automatic-gain-control is such that if the inputs given in (2) above are increased by 60 db the output will increase by approximately 7 db.

(5) *Input Impedance.*—75-ohms, unbalanced.

(6) *Audio Output.*—1-watt into 3 ohms; 10 mW into 600 ohms.

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F. W. J. S.

23. VALVES

23. VALVES

The thermionic valve or tube in its present form owes its origin to a number of basic discoveries and much organized research; there are, however, three contributions of a fundamental character. Edison, in 1883, noticed that certain substances when heated in a vacuum emitted electrons, and that these particles could be attracted by an electrode held at a potential positive in respect to the emitter. Then, in 1904, J. A. Fleming applied this effect, for the first time, to radio communication, and thus discovered the thermionic diode detector. In 1907 Lee de Forest conceived the idea of interposing a grid between the emitter (called the cathode) and the collector (anode), which enabled the valve to be used, for the first time, as an amplifier, since by suitable arrangement a change of voltage at the grid produced a corresponding but greater change of voltage at the anode. The 1914–18 war gave considerable impetus to valve design, and marked the first steps towards achieving consistency of characteristics between similar types of valves. A steady increase in efficiency, reliability and type specialization to meet particular requirements has followed.

CONSTRUCTION

Thermionic emission is dependent upon the property exhibited by certain materials of throwing off electrons when heated to a suitable temperature. A metal unaided by chemical coating, such as tungsten, must be raised to a temperature of about 2,000° C. (approaching white heat) before worth-while emission occurs; it is usual therefore to use a wire or, in the case of an indirectly heated valve, a fine-gauge tube coated with chemicals which emit electrons at a more convenient temperature, namely at about 800° C. Pure tungsten has a maximum emission of the order of 4 mA for each watt of cathode heating. Owing to the absence of any activating agent, a pure tungsten cathode cannot be poisoned or de-activated.

In practice, a number of highly emissive coatings are used: the chemicals associated with low-temperature emission are barium and strontium oxides.

Thoriated tungsten, produced by the addition of thorium and ·rbon, is designed to be heated to approximately 1,600° C. with an emission of up to 30 mA/watt of cathode heating. It has the disadvantage that the thin layer of thorium formed on the surface of the cathode during manufacture may be destroyed by overheating or excessive emission, and for this reason the average emission should be of the order of 7·5 mA/watt.

The oxide-coated cathode has a coating of alkaline-earth metals, mostly barium, deposited by dipping or spraying. The operating temperature may be as low as 700° C., with an emission figure of the order of 200 mA/watt of cathode heating. For receiving valves a mixture of barium oxide and strontium oxide is common. The average D.C. cathode current for Class A operation should not exceed about 20 mA/watt of cathode heating.

It will thus be seen that the choice of cathode material depends upon : (1) obtaining the required emission at a working temperature that will not damage the cathode; (2) the material being easy to work; (3) it not being liable to chemical reactions during the life of the valve ("poisoning").

Filaments of directly heated valves are usually of nickel alloy, pure nickel or tungsten. For indirectly heated valves, the heaters are generally of tungsten wire, coated with magnesium oxide, which acts as insulation between heater and cathode.

Heat Dissipation

The heat generated by the impact of electrons at the surface of the anode and other electrodes must be continually removed by radiation, or conduction through the pins, as otherwise the temperature of the electrode will rise progressively until it is destroyed, or will cause excessive liberation of gas which may poison the cathode or affect the vacuum. In order to increase the radiation, the anode is made with as large a surface area as possible; where the dissipation is considerable this area may be increased by the use of projecting fins on the anode, or by the use of wire mesh, or—for large transmitting types—by air- or water-cooling. Radiation may also be improved by darkening the surface. The anodes of most low-power valves are made of nickel or

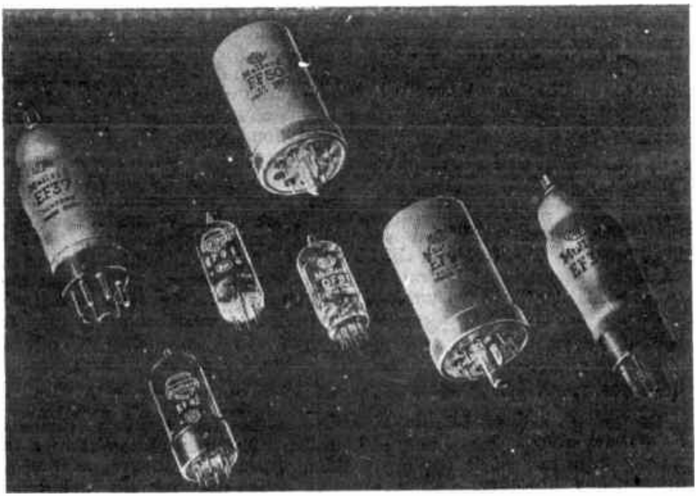


FIG. 1.—TRENDS IN VALVE DESIGN.

Trends in the design of valves for low-power applications can be seen in this illustration. The earlier pinch-type construction of the EF36 and EF37 has tended to give way to the miniature all-glass construction of the EF91, DF91 and later types

(*Mullard Ltd.*)

TABLE 1.—CLASSIFICATION OF VALVES

Number of Electrodes	*Classification*
2	Diode
3	Triode
4	Tetrode
5	Pentode
6	Hexode
7	Heptode
8	Octode

nickel-plated iron. Carbon may be used for valves requiring heavier dissipation. Transmitting valves may use tantalum, as this metal—unlike most others—tends to absorb rather than liberate gas when operating at high temperatures.

The anode dissipation under no-signal conditions is equal to the anode current multiplied by the anode voltage (and is thus equal to the power input to the anode circuit). When a signal is applied, the anode dissipation is equal to the difference between the power fed to the anode circuit and the power dissipated in the load.

Screening

To prevent stray electrons from producing unwanted coupling, particularly in multiple valves, it is necessary to fit screens between the electrode systems, or between electrode connections. Stray electrons may also strike the walls of the envelope and cause secondary emission. To minimize this the inner surface of the envelope is often coated with carbon, which has a low secondary emission factor.

To prevent external fields from affecting the valve characteristics, many valves have external coatings of a metallic material such as copper or zinc. This coating is painted on the outside of the envelope, and may be protected by a coating of lacquer. Where the valve is required to dissipate a fair amount of heat, the metallizing may be limited to the lower part of the envelope. The trend of design is to use an internal screening cage rather than external coating. This cage may take the form of a wire-mesh cylinder completely surrounding the electrode assembly, and replaces both the external coating and the internal carbonizing.

Gettering

In order to provide as near a vacuum as possible, the valve is heated by passing it through a heated chamber, and by the use of high-frequency heating, while the air is being withdrawn. Even with these precautions, however, a small amount of gas may remain occluded in the assembly, and this may later be released when the valve is put into operation.

To remove the last traces of free gas, a little barium or magnesium is usually evaporated within the envelope just before sealing. The getter, as this material is called, is fired by means of high-frequency eddy currents, and after evaporation condenses on the envelope near the getter-holder, a small metal holder welded to the electrode assembly. It is important that as little as possible of the deposit falls on the valve insulation.

Base Construction

For many years the electrode assemblies were built up around a glass "foot", with the electrode supports and electrode connecting wires pinched into this tube. In order to make air-tight seals, the lead-in wires were of composite construction, with the section forming the seal made of a material having a similar coefficient of expansion to glass.

The modern trend, however, for receiving valves is to use a pressed-glass base with sturdy wires moulded into the glass base of the valve and serving both as lead-in wires and valve pins.

VALVE FUNDAMENTALS

Diodes

The diode is a two-electrode valve having a cathode and anode. Providing that the cathode is raised to emission temperature and the anode held at a suitable positive potential, electrons will flow from cathode to anode, and back through the power supply to the cathode. The value of this current will be affected by cathode temperature and anode potential. Over a certain range, at a given cathode temperature, the change of anode current will be directly proportional to the change of anode potential. After a certain anode potential has been reached, no further increase in current will occur: this is known as saturation point.

It is, however, important to take into account "contact potential", which has the effect that voltages applied externally to the valve will not be the same as those present across the electron stream within the valve. This means that in practice electrons will flow from cathode to

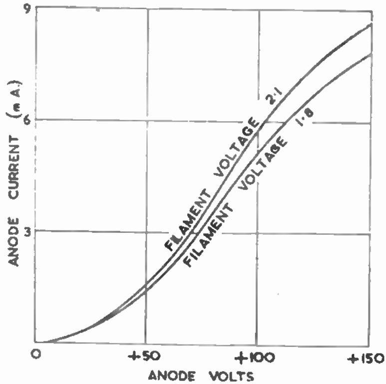


Fig. 2.—Showing How the Anode Current in a Diode Valve Depends upon the Anode Voltage and the Cathode Temperature.

(Typical 2-volt filament valve.)

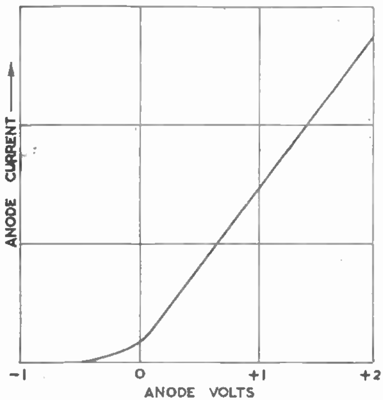


FIG. 3. CHANGE OF ANODE CURRENT WITH CHANGE OF ANODE VOLTAGE.

Showing that the anode current may start when the anode is slightly negative.

anode when the latter is apparently slightly negative to the former. This is shown in Fig. 3, where current commences to flow when the external voltage difference between cathode and anode is 0·5 volt negative.

A diode tube, because of its unidirectional conductive quality, can rectify an A.C. signal and, if the signal is modulated, act as a detector by superimposing the modulation on the D.C. resulting from the rectification.

Diode valves for power rectification are described more fully in Section 25.

Triodes

The triode is in effect a diode valve with an electrode of open construction interposed between cathode and anode; as this electrode is placed closer to the cathode than the anode it exercises greater control, thus enabling the valve to function as an amplifier. This greater control means that a change of voltage applied to the grid will cause a greater change of anode current than would be brought about by a similar change in anode voltage. Fig. 6 shows pictorially the effect of grid potential on the electron stream : (*a*) electron stream when the grid is at cathode potential; (*b*) with the grid at a normal negative bias voltage; and (*c*) with sufficient negative bias to cut-off the passage of electrons. By plotting the value of anode current for various grid

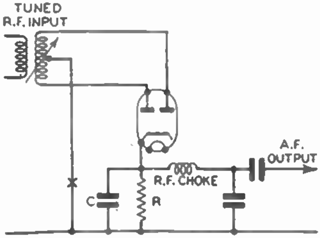


FIG. 4.—PUSH-PULL DIODE DETECTOR.

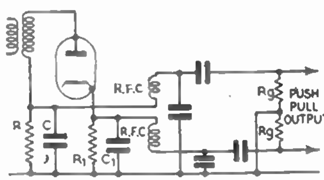


FIG. 5.—DIODE DETECTOR AS PHASE-SPLITTER.

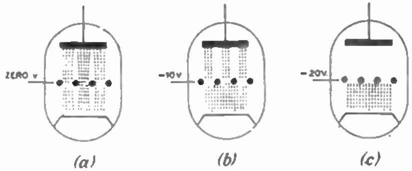


FIG. 6.—HOW THE GRID CONTROLS THE ELECTRON STREAM.

(a) No bias; (b) working bias; (c) high negative bias.

voltages at fixed anode voltages, a static characteristic curve can be obtained. The curve of a typical triode is shown in Fig. 7, which gives the value of anode current for any value of grid voltage under conditions of three selected anode voltages.

It is informative, in this example, to compare the relative influence of grid and anode voltage upon anode current. A change of anode voltage from 50 to 150, at zero bias, changes the anode current from 3 to 10 mA, which is a change of approximately 0·07 mA/volt. On the other hand, with an anode voltage of 100 volts, the valve passes 6·5 mA at zero grid bias and 3·5 mA at 1 volt negative bias; thus a change of 1 grid volt has brought about a change of 3 mA. It will be obvious that if the figure of 0·07 (the change of current due to a change of 1 anode volt) is divided into 3 (the change due to a change of 1 grid volt), the figure so obtained will indicate the superiority of the grid over the anode, and this is termed the *amplification factor* of the valve. It should be noted, however, that the amplification factor calculated in this way is not necessarily the magnification of the valve when used in an amplifying stage, as it does not take into account the associated circuit components. The amplification factor of triode valves varies from about 3 to 100: valves with an amplification factor above about 30 are usually known as high-μ types; from 8 to 30 as medium-μ types; and below 8 as low-μ types.

The figures given above may be used to ascertain the related characteristics of *impedance* and *slope*. Impedance is the equivalent A.C. resistance of the electron stream and, like any other variant of Ohm's Law, anode impedance can be ascertained by the effect of voltage on current, and is thus arrived at by dividing a selected change of anode volts by the resultant change in anode current: in this case 1 volt divided by 0·00007 amperes, or approximately 14,300 ohms.

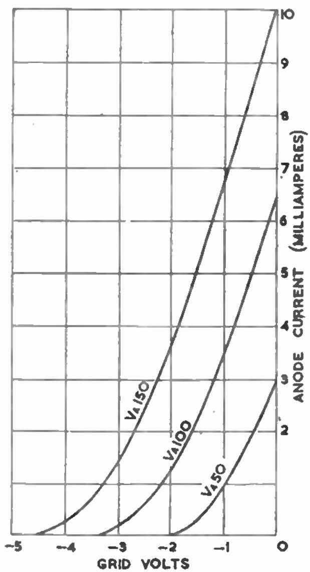


FIG. 7.—CHARACTERISTIC CURVE OF A TYPICAL SMALL TRIODE.

Slope, or mutual conductance, can be ascertained by a direct inspection of the curve, or by combining the impedance and amplification factor. Slope is a means of annotating directly the change of anode current to be expected from a stated change of grid voltage, and is expressed as so many milliamperes change for a change of 1 volt. Thus in Fig. 7, a change from zero to 1 grid volt will change the anode current (at an anode voltage of 100) from 6·5 to 3·5 mA, or a change of 3 mA/volt. It should be noted that the mutual conductance, which gives a measure of the efficiency of the valve, will vary with different anode voltages. For example, in the curves of Fig. 7 it will be seen that with an anode voltage of 50, a change from zero to 1 grid volt only brings about a change of 2·1 mA, whereas at 150 anode volts, the change is 3·3 mA.

We thus have the following relationships:

$$\text{Mutual conductance} = g_m = \frac{\mu \times 1{,}000}{r_a}\ \text{mA/volt}$$

$$\text{Amplification factor} = \mu = \frac{g_m \times r_a}{1{,}000}$$

$$\text{Anode impedance} = r_a = \frac{\mu \times 1{,}000}{g_m}\ \text{ohms.}$$

While mutual conductance is a direct indication of the efficiency of a valve under stated conditions, amplification factor means little unless impedance is taken into account. For example, a valve with an impedance of 30,000 ohms and an amplification factor of 90 still has a slope of 3 mA/volt, so that no advantage has been gained by doubling the amplification factor, unless the circuit conditions are such that a valve of higher impedance can be used to greater advantage.

American manufacturers use the term transconductance instead of mutual conductance: this gives the same indication of valve efficiency, but is derived from the ability of the electron stream to conduct, rather than—as in the case of mutual conductance—to oppose, and is thus given not in mA/volt but in micromhos. Transconductance is obtained by dividing the change of anode current by the change of grid voltage. Thus in the example given the transconductance is 3,000 micromhos. To convert transconductance to mutual conductance, divide the number of micromhos by 1,000; this gives the answer in mA/volt.

The static characteristic of a valve shows the relation between the steady voltage applied to the grid and the anode current, when the anode voltage is kept constant, or it may show the relation between anode voltage and anode current when the grid voltage is kept constant. A dynamic characteristic, however, shows the performance of a valve under working conditions when alternating voltages are being applied to the grid.

Interelectrode Capacitances

Each pair of electrodes in a valve forms a small capacitor. Although these capacitances are small—seldom more than a few picofarads—they can have a considerable effect upon the performance of the valve as an amplifier.

The input capacitance of a valve, that is the effective capacitance "seen" by a signal source connected across the grid and the cathode is—owing to the Miller effect—considerably larger than the grid-cathode capacitance; this is because the signal must supply the capacitive current which flows from grid-to-anode and which is directly proportional to the sum of the A.C. grid and A.C. anode voltages; and this, in turn, will depend upon the voltage amplification of the valve. The input capacitance of a high-mu valve may amount to several hundred picofarads.

The input capacitance of a triode is thus:

$$C_{\text{input}} = C_{gk} + C_{ga}(A + 1)$$

where C_{gk} is the grid to cathode capacitance, C_{ga} is the grid-to-anode capacitance, and A is the voltage amplification of the valve.

The reactances of the small interelectrode capacitances drop to a low figure at radio frequencies. A resistance-coupled amplifier, for example, is of comparatively little use at radio frequencies, since the input and output circuits are in effect short-circuited by these capacitances. This difficulty can be overcome by making the valve capacitances part of the tuning capacitances by substituting tuned circuits in place of the resistive loads.

A further difficulty then arises at radio frequencies in that the grid–anode capacitance forms a coupling capacitor between anode and grid circuits. Energy fed back via this path from the anode to the grid will generally be in the form of positive feedback, and will thus cause oscillation.

Limitations of Triode Amplification

The amplification factor of a triode valve is governed by the distance of the grid from the filament, and by the spacing of the grid wires: a close-mesh grid exercising greater control over the electrons than one of coarser construction having wires spaced relatively far apart. Unfortunately, however, as the mesh of the grid is increased, certain difficulties tend to arise.

Firstly, the area of the grid electrode will increase, and thus increase the inter-electrode capacitances.

Secondly, not all the electrons emitted by the cathode reach the anode, and these, together with any electrons repelled by a negatively charged grid, tend to accumulate in the space between the cathode and the grid, giving rise to what is known as the "space charge". The fresh electrons emitted by the cathode have to break through this charge in order to reach the anode. Since a closely-wound grid exercises more control over the electron flow, it will tend to drive a greater number back, thus increasing the magnitude of the space charge and impeding the cathode-anode flow.

Tetrodes

The limitations of the triode can be overcome by inserting a further grid between the control grid and the anode; this grid acts as an electrostatic shield, and thus greatly reduces the capacitative coupling between grid and anode; furthermore, by connecting a positive potential on this electrode, it will assist in attracting electrons from the cathode,

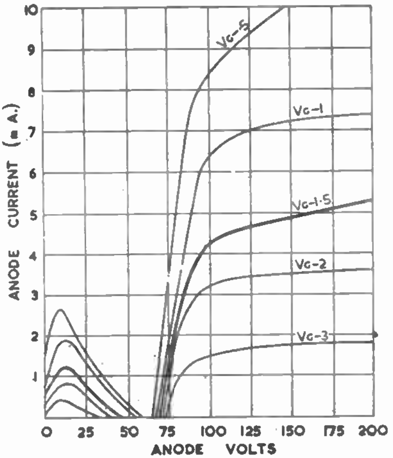


FIG. 8.—CHARACTERISTIC CURVE OF TYPICAL SCREEN-GRID VALVE.

and thus help to overcome the space charge. By the time the electrons reach the second grid (screen grid), they are travelling sufficiently fast for most of them to pass through and strike the anode. By thus overcoming the limitations of the triode, the control grid can be made of much finer mesh, so that the amplification factor may exceed 1,000.

In its turn, however, the tetrode valve has a characteristic which severely limits its use. This is shown by considering the typical characteristic curve shown in Fig. 8. If the screen voltage is kept constant whilst the anode voltage is increased from zero, the anode current rises in a normal manner until the anode voltage reaches approximately 20 volts. The anode current then falls until the anode voltage reaches approximately 80 volts. If the anode voltage is further increased, the current begins to rise again until a saturation value is reached.

This kink is caused by the fact that when an electron travelling at considerable velocity strikes the anode, the force of the impact dislodges other electrons into the inter-electrode space; this is termed secondary emission. In the triode valve these electrons are repelled by

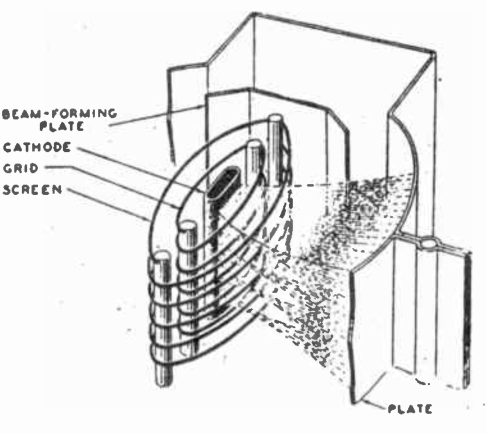


FIG. 9.—BEAM TETRODE VALVE.
(*R.C.A.*).

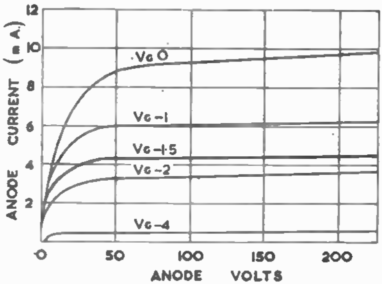


FIG. 10.—STATIC CHARACTERISTICS OF A TYPICAL PENTODE.

the negative grid and fall back on to the anode, and thus have no effect on the valve characteristics. In the tetrode, however, the positively charged screen will attract these electrons, unless the anode voltage is greater than the screen voltage, thus causing a reverse current to flow between anode and screen.

The effects of secondary emission may be overcome either by the introduction of a further grid (see below) or by the insertion of beam-forming plates. In such valves, known as "beam tetrodes", the electrons emitted by the cathode are directed to those parts of the anode which are at a critical distance from the other electrodes.

A variable-mu valve is one in which the mutual conductance varies smoothly with changes in grid bias.

The pitch of the control grid wire in an ordinary screen-grid valve is constant throughout, whereas the pitch of the wire at one end of the control grid of a variable-mu valve is different from that at the other end. The degree of control over the anode current depends upon the closeness of the mesh or pitch of the wires forming the grid. The closer the grid wires, the greater will be the control exercised by the grid, and with it the mutual conductance of the valve. It follows, therefore, that with the variable-mu valve the degree of control, and thus the mutual conductance, will be different on different portions of the cathode stream emerging from the cathode.

Pentodes

The kink in the characteristic of the tetrode valve, caused by secondary emission, may be eliminated by inserting a further grid between the screen grid and the anode. This electrode, known as the suppressor grid, is—for most applications—connected to cathode potential, thus making it negative with respect to anode, and repelling the electrons dislodged from the anode by secondary emission.

Frequency Conversion

The prime essential required before a valve can act as a frequency converter is that it should operate over a non-linear portion of its characteristic. From this it may be inferred that any valve that is capable of detection is also capable of frequency conversion. However, not all valves are equally efficient for this purpose: useful measures of their relative efficiencies are provided by the *conversion conductance* and *conversion gain.* Conversion conductance (g_c) is the ratio of the required beat-frequency component in the output current to the signal-input voltage, and is generally stated in units of μA/volt. Conversion gain is the ratio of the beat-frequency output voltage to the signal-frequency input voltage.

There are also a number of other factors that determine the suitability of valves for frequency conversion. These include: the noise contributed by the valve must be as low as possible if weak signals are involved (frequency converters contribute considerably greater noise than straight amplifying stages); the conversion efficiency over the frequency range concerned should not depend too critically upon the local oscillator injection voltage, nor should too great an injection voltage be required (otherwise the design of a stable local oscillator may be rendered difficult); cross-modulation must not occur in the valve; "pulling" of the local oscillator frequency by alteration of the signal-frequency circuits or by changes in valve potential should be as small as possible.

Valve Noise

Noise voltages which are generated in a valve amplifier include three main types: Thermal agitation ("Johnson" noise); shot effect; and induced grid noise.

All these factors include frequency components throughout the entire frequency spectrum, and the amount of noise is therefore affected by the band-width.

Thermal Agitation.—At any instant, due to the random motion of electrons in a conductor, there are likely to be more free electrons moving in one direction than in the other; this causes a voltage drop across the conductor. If the temperature of the conductor is now raised, the agitation of electrons in both directions increases, and the instantaneous currents are therefore greater.

The value of the mean square voltage can be calculated (see Section 22).

Shot Effect.—Shot effect is caused by irregular emission from the cathode. The electron flow in the anode circuit varies slightly in regard to the number of electrons reaching the anode from one instant to another, and also in the velocities of individual electrons. Although such noise is generated in the anode circuit, for convenience it is usual to consider it as an equivalent noise resistance, which, when connected in the grid circuit, would generate the same amount of noise. When a positive grid is placed in the electron path, the shot effect is magnified because the division of electrons is irregular. For this reason, pentodes and multi-grid valves are more noisy than triodes.

Induced Noise.—Strong varying electromagnetic and electrostatic fields may induce voltages and currents in resistors, leads, and in the valves. Irregularities in the flow of the electron stream may also induce current flow in the grid circuit. This is due to random variations in the transit of electrons from cathode to anode, and should be distinguished from electrons passing from cathode to grid. Stray fields external to the valve may produce noise effects by induction: long feed lines are particularly susceptible. In general, therefore, induced noise voltage is reduced by the use of low-impedance circuits, shielding, filtering and short leads.

SPECIAL TYPES OF VALVES

Valves for Television Reception

Since the opening of the B.B.C. high-definition service in 1936, the requirements of V.H.F. television have led to the introduction of a number of new types of valves. Radio-frequency pentodes with a slope of about 7·5 mA/volt were soon introduced, the relatively high input capacitance being of less consequence than for narrow-band-width applications. Post-war improvements have included the reduction of microphony, better cathode-to-heater insulation for series-heater operation, and improved line-output valves capable of withstanding peak forward anode voltages and of passing high anode currents at low anode voltages. Special E.H.T. rectifiers with heaters designed to be fed with the peaky waveform derived from the line-output transformer, and booster diodes capable of withstanding high peak cathode-heater voltages are available. Much has been achieved in the production of a restricted range of valves suitable for a wide range of applications in a normal domestic television receiver, in order to reduce the number of different types that would otherwise be necessary. The choice of valves for television is discussed further in Section 15.

Frame-grid Valves

Improved sensitivity and performance of V.H.F. radio and television receivers is possible with recent valves designed for use in cascode amplifier stages. These valves, which use what is known as a frame-grid, have a mutual conductance of about 12 mA/volt compared with the 6 mA/volt or so of conventional valves and make possible a correspondingly higher " figure of merit ". The high efficiency of grid control necessary to achieve these characteristics is obtained by winding the grids as a very close mesh of extremely thin wires and siting them much closer to the cathodes than in earlier valves. If inter-electrode short-circuits are to be avoided, this calls for precision manufacture of a high order. For example, the main rods, which are much stouter than usual, may be drawn to size within the close limits of $\pm 0{\cdot}005$ mm. This enables the grid winding to consist of two exactly parallel flat meshes of taut, very thin wires. A higher order of accuracy is also necessary in the location of the cathode and the spacing micas. The assembly and inspection of these valves are carried out under microscopes, with large images of the components projected optically on to a large drawing.

Ceramic Valves

In recent years ceramic instead of glass has been used for valve envelopes, particularly for transmitting purposes. This increases the mechanical strength, allows smaller size for a given power dissipation and enables the valve to operate at higher ambient temperature. It is also claimed that more effective de-gassing is possible during manufacture, permitting greater emission for pulse operation. Ceramics have also been used to replace the mica spacers in the valve assembly.

Typical maximum operating temperatures of low-power ceramic valves are in the region 180–250° C.

Special Service Valves

For applications where a high degree of valve reliability under adverse operating conditions is essential, a restricted range of " special service " valves has been introduced by a number of manufacturers. These valves are electrically similar to a number of standard types, but use improved mechanical structures, designed to give better electrode and getter support, and are manufactured by selected personnel to finer tolerances, under dust-free conditions; during manufacture they are subjected to more rigid inspection and testing than would be economically possible for standard types.

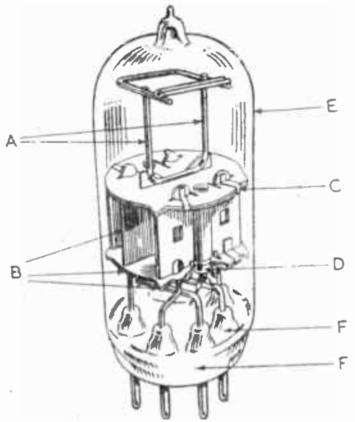


FIG. 11.—CONSTRUCTIONAL FEATURES OF A TYPICAL SPECIAL SERVICE VALVE.

A—Double getter supports to prevent resonant vibration and to reduce microphony.
B—Double micas and pre-bent points to secure accurate location of the system within the bulb and minimize inter-electrode short-circuiting.
C—Electrodes strapped to micas to firmly lock the assembly.
D—" Belled " cathode ends to improve heater-cathode insulation.
E—Internal diameter of the bulb selected to give rigid assembly.
F—Button seal selected to withstand thermal shock test and to minimize strain in the glassware.

(*G.E.C.*)

Cold-cathode Valves

Gas-filled, " cold-cathode " rectifier valves were introduced primarily for use in vibrator-type H.T. units. The valve consists of two anodes and a cathode coated with electron-emitting material, the envelope being filled with an inert gas at very low pressure. The name " cold-cathode " is to some extent a misnomer, since the cathode is heated, not by a heater wire, but by ionic bombardment. Ionization of the gas occurs between the cathode and the anode, which is at positive potential, but not between the cathode and the anode, which is at negative potential: rectification will thus occur. The voltage drop between anode and cathode is higher than with mercury-vapour rectifiers, owing to the loss of energy required to heat the cathode. Circuit arrangements must be such that sufficient current is drawn at all times to maintain the cathode at working temperature.

Heaterless Valves

The discovery that a thin layer of magnesium oxide can emit electrons under the influence of a positive charge near its surface has led to the development of a new type of valve which has no continuously operating heater. To start the electron emission a small filament is at present fitted, but this may be eliminated in later designs. To provide the necessary charge a positive potential is applied to the first grid. The future development of valves using this principle seems likely to prove of considerable importance.

Gas-discharge Triodes

The gas-discharge triode—or "Thyratron"—is different in theory and application from other forms of triode. The valve passes either unlimited anode current or no anode current whatsoever: there are no graduations between these extremes. For each electrode structure there is a particular value of negative grid voltage that will cut off the anode current completely. When, however, the grid voltage is reduced below this critical figure, anode current flows—limited only by external circuit resistance. Once anode current has commenced to flow, the grid loses all control, and will not affect the value of anode current, no matter how negative it may be made: anode current can be stopped only by breaking the anode circuit or by causing the anode to cease being positive with respect to the cathode.

Gas-discharge triodes intended for use as rectifiers or relays may contain mercury-vapour gas, and require an appreciable de-ionization time of the order of 100 microseconds to permit the grid to regain control after the anode current has ceased to flow. Valves intended for use in television or oscilloscope time-bases, however, usually contain argon gas, and have a de-ionization time of 10 microseconds or less; this can be further reduced by limiting the peak and mean anode current.

VALVE OPERATING CONDITIONS

The useful life of a valve is governed to a considerable extent by the conditions under which it is operated. The following notes are based on "The Use of Electronic Valves" (British Standard Code of Practice CP 1005 : Parts 1 and 2 : 1954), on BS : 1106 and on the recommendations of members of the British Radio Valve Manufacturers' Association.

General.—Published ratings should be carefully observed, particularly in regard to maximum dissipation in the various electrodes. Apart from the application of incorrect potentials, excessive dissipation may be caused by incorrect tuning of associated circuits, unnecessarily high no-signal currents or parasitic oscillation in audio-frequency and radio-frequency stages.

Heaters and Filaments.—Filament and heater voltages should generally be maintained within ± 7 per cent of the rated values. The heater current of valves connected in series should be maintained within ± 5 per cent of the rated values. Thoriated-tungsten and oxide-coated filaments should be maintained within closer tolerances than the above figures: 5 per cent voltage fluctuations are permissible, but permanent deviation from rated values will reduce valve life. Directly heated and indirectly heated valves having similar filament-current ratings should not be connected in series.

With 2-volt filaments a tolerance of ± 10 per cent is permissible, though this may result in some variation of the valve characteristics. 1·4-volt filaments are designed for use with dry-cell batteries having a rated terminal voltage of 1·5 volts, but when mains operation or accumulators are used, the voltage drop across each 1·4-volt section of filament should have a nominal value of 1·3 volts. It is important to note that with certain types of valves under-running the heater is as harmful as over-running. If the voltage at the filament or heater terminals rises by 7 per cent, the current will increase by considerably less than this figure, since the resistance of the filament or heater increases as its temperature rises.

Cathodes.—The maximum D.C. potential difference (i.e., A.C. peak value) between heater and cathode should be kept to a minimum. A D.C. path should exist between the cathode and each electrode and between the cathode and all internal and external screens. The heater-cathode path should not be included either in the audio-frequency or radio-frequency circuit. A valve should not be rendered inoperative by disconnecting the cathode unless there is a resistor not exceeding 250,000 ohms between heater and cathode. It is not good practice for the heater-dropping resistor to be common to both H.T. and heater supplies; this, however, is not important where the resistor is of low value.

Grids.—The resistance between the grid and cathode should be the minimum practicable. Typical figures are: radio-frequency pentodes and frequency changers—not greater than 1MΩ, or 3·5MΩ where automatic bias circuits are used; mains-output valves—0·1MΩ, or 0·5MΩ where automatic bias circuits are used; 1·4-volt battery valves are an exception, and values up to 10MΩ may be used in certain circumstances. Valves should not be run under conditions which result in appreciable grid current, unless so specified.

Screen Grids.—The voltage supply for the screen grids of frequency changers and beam tetrodes should be obtained from a potentiometer network, the resistor values being as low as possible to prevent undue variation of voltages.

Mounting.—Ventilation should be adequate to ensure that bulb temperatures are not unduly high. Valves should preferably be mounted vertically with the base downwards. Where mounted horizontally, the major axis of the first grid of indirectly heated valves of high mutual conductance and the plane of the filament of directly heated valves should be mounted vertically. Spare valve-base contacts should preferably not be used as connecting tags, and in no circumstances should pins shown as internally connected be so used. Valve-holders which incorporate contacts should not have heavy wire connections made to them. Cushioned holders should be used where the valves would otherwise be subjected to continuous vibration. To reduce microphony, the chassis should not be rigidly fixed to the cabinet containing the loudspeaker. The characteristics of valves placed in a strong magnetic field may be affected: miniature 1·4-volt valves, in particular, should not be mounted in close proximity to loudspeaker magnets.

Rectifiers.—An adequate limiting resistance should always be placed in series with a rectifying valve used in conjunction with a capacitor-input smoothing filter. The life of a rectifier operating near its maximum current rating can be seriously affected if the heater is under-run.

Mercury Vapour Rectifiers.—Before putting a mercury vapour rectifier into service, or where it has been out of service for a considerable period, it should be run for at least 30 minutes with normal filament voltage, but without any anode potential; this is to vaporize mercury deposited on the electrodes.

At all times the filament must attain full working temperature, and the condensed mercury temperature be within the prescribed limits, before the anode supply is connected. This involves a delay of at least 1 minute, and longer if the ambient temperature is appreciably less than the prescribed condensed mercury temperature. An automatic delay mechanism is recommended.

A permanent deviation of more than 2 per cent or temporary fluctuation of more than 5 per cent in the filament voltage may cause damage. With directly-heated rectifiers, the filament supply should preferably be 60–120° out of phase with the anode supply.

They must be mounted vertically, with the cathode connections at the lower end, with free circulation of air. An increase of temperature above the specified value will reduce the safe peak inverse voltage. Where screening is employed, suitable ventilation holes at the top and bottom of the box must be provided.

They should not be operated in strong radio-frequency fields unless enclosed in separate earthed screen boxes, and radio-frequency filters should be used to prevent radio-frequency current from reaching the rectifiers. Otherwise ionization, causing flash-over, may occur.

Since parasitic oscillation may occur, it is advisable to place small radio-frequency chokes or resistors in the anode leads where there is any risk of causing interference to nearby apparatus.

To limit peak anode current, a choke of specified minimum inductance should precede the first smoothing capacitor. The value of this inductance may be decreased if a surge-limiting resistor is included. To ensure good voltage regulation, the value of smoothing capacitor must be chosen to suit the maximum current, and the value of the smoothing inductance should be large enough to provide uninterrupted current at the minimum load.

Interpretation of Valve Data

The information supplied by valve manufacturers can usually be considered in three categories: (*a*) general electrical and mechanical characteristics; (*b*) maximum electrode ratings; and (*c*) typical operating conditions. Of these, (*a*) includes heater or filament ratings, base and pin connections; amplification factor and anode impedance; inter-electrode capacitances; and dimensions: the information supplied in this category is self-explanatory.

Maximum "absolute" ratings are those determined by the valve manufacturer in relation to a reasonable life expectancy for the valve. The operation of an electrode beyond the limits set by the manufacturer may not affect directly the electrode concerned, but may be considered harmful to the valve as a whole. Normally, the ratings are set independently for each electrode: however, in certain cases, the values are dependent upon the conditions of other electrodes within the valve envelope; in such cases it is common practice to assume, when determining values, the simultaneous operation at maximum values of all interdependent electrodes. It should not be assumed, however, that one rating may be exceeded by a corresponding reduction to another.

An alternative to the "absolute" method of valve rating is the "design centre" system, which takes into account normal fluctuations of mains or battery voltages. No matter which system of rating is used, however, it is necessary to consider the voltage fluctuations likely to be encountered, owing to component tolerances and load fluctuations, etc.

The typical operating conditions suggested by valve manufacturers are not, of course, the only ones under which the valve may be operated safely: they do however show the conditions in which the maximum performance of the valve is likely to be achieved with a specified anode voltage, or in a specified circuit condition.

Negative Grid Current

Under normal Class A operation, only negligible current flows in the control-grid circuit, and this current is in such direction as to produce, in conjunction with the grid-cathode resistor, a negative potential on the grid. Under certain conditions, however, this current may be reversed and negative grid current flow, lowering the bias applied to the valve and seriously affecting the performance of the stage concerned and also of the preceding stage.

The most common causes of negative grid current are:

(1) Internal leakage currents across the insulation between the grid and other electrodes, commonly as a result of the deposit of volatilized metallic material on the insulating surfaces.

(2) Positive ion current, due to the presence of gas in the envelope, commonly as a result of the liberation of gas caused by overheating the valve.

(3) Electron emission from the grid, commonly as a result of the deposit of active cathode material on the grid, and high grid temperatures.

A rise in the operating temperature of the valve will contribute to a[illegible] these factors and—since negative grid current tends to increase anod[illegible] current—once started it will increase progressively.

Careful attention to manufacturer's data, and in particular to heater voltages, electrode dissipation and the permissible upper limit of the grid-cathode resistance will reduce the likelihood of this fault arising.

Microphony

Physical vibration of the electrode assembly may cause slight variation in the characteristics of a valve, resulting in rhythmic variations in the output. In practice, this can give rise to interference known as "microphony", which, in extreme cases, may result in "feed-back" howl being set up in an audio-frequency amplifier. Microphony in a radio-frequency stage will cause modulation of the carrier, and will thus be noticeable only when signals are present: in a cathode-ray tube it may cause a fluttering and coarsening of the picture.

The vibrations may reach the valve via the chassis, valve-holder and valve pins, or as sound waves striking the valve envelope, and thence being conveyed to the electrode assembly via the mica supports. The effects of microphony will be most pronounced when considerable amplification follows the valve concerned. and where the valve is mounted near a moving or vibrating mechanism: for example, amplifiers used in conjunction with magnetic-tape recorders are particularly susceptible.

Microphony can be reduced by careful design of the electrode assembly, and by such methods as: resilient mounting of chassis and source of vibration; use of anti-microphonic valve-holders; careful positioning of the early valves in an amplifier (e.g., not close to the loud-speaker, or in the direct line of sound waves).

Series Operation of Heaters and Filaments

The series operation of valves gives rise to a number of problems, including those of heater-cathode insulation and the provision of correct voltage drops across filaments.

The potential difference between the heater and cathode should always be kept as low as possible. The maximum safe figure is generally stated by the manufacturers on valves developed for series operation, but where this is not known, a figure of 150 volts is normally assumed, although a limit of about 50 volts may be preferable. The breakdown of heater-cathode insulation most often occurs during the warming up period and, for this reason, delayed application of heater-cathode potentials tends to improve reliability. It is not advisable to include heater-cathode insulation across any tuned radio-frequency circuit, since this may give rise to modulation hum or frequency instability.

Where the filaments of 1·4-volt battery valves are connected in series, shunting resistors should be used to correct the voltage-drop across each filament, or section of filament. This is because anode and screen currents will affect the filament current, even though the total L.T. consumption of the filament chain may appear correct; the extra current contributed by the anode and screen currents will be greater across the more negative sections of the chain.

FREQUENCY LIMITATIONS OF VALVES

The main factors limiting the output and efficiency of valves with an increase of frequency are:

(*a*) the stray inductance and capacitance associated with the valve electrodes and connecting leads;

(*b*) electron transit times;

(*c*) increased radio-frequency losses in dielectric materials, and resistive losses.

Stray Inductances and Capacitances

The stray inductances and capacitances associated with a triode valve are shown in Fig. 12. At ultra-high frequencies the combined effect of the electrode capacitances on the generated frequency is equivalent to an increase in capacitance of the tank capacitor C by an amount C_1 where

$$C_1 = C_{ga} + \frac{C_{ak} \times C_{gk}}{C_{ak} + C_{gk}}$$

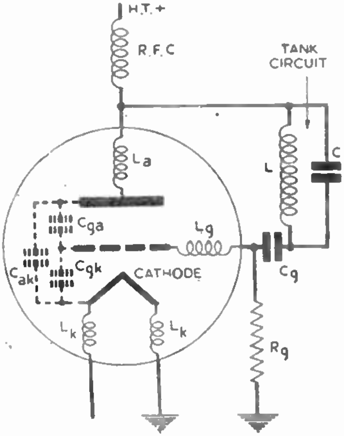


FIG. 12.—VALVE INTER-ELECTRODE CAPACITANCES AND LEAD INDUCTANCES SHOWN IN DIAGRAMMATIC FORM.

At lower frequencies when C is large compared to C_1, the effect is negligible, but as the frequency is raised the capacitance of C must normally be made smaller, and thus the effect of C_1 increases.

The lead inductances are in parallel with the inductance L of the tank circuit, and these lower the value of the upper frequency limit for the valve. For example, a wire 0·040 in. diameter and 4 in. long has inductance of approximately 0·1 μH. At 1 Mc/s the reactance would be 0·63 ohms, but at 100 Mc/s becomes 63 ohms. Furthermore, the inductance of the cathode lead is common to grid and anode circuits, and so produces negative feedback.

Reducing the size of the electrodes and spacing them farther apart would reduce inter-electrode capacitance, but would make high voltages necessary, and would increase transit time.

However, if all the dimensions of a valve are reduced by a given factor, the characteristics of the valve are unchanged, but inter-electrode capacitance, lead inductance and transit time are all reduced. This reduction in size, however, reduces the power-handling capabilities of the valve. Lead inductance may also be decreased by making the leads of large diameter. To shorten their length and to avoid the capacitance

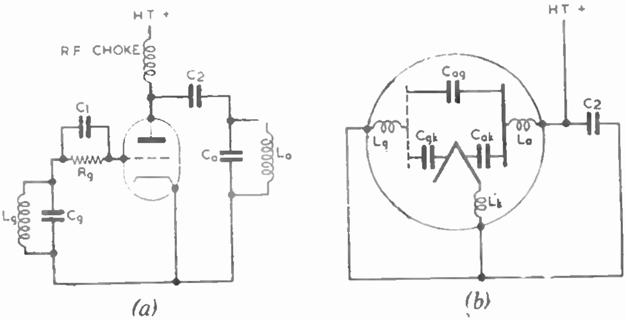


FIG. 13.—EFFECT OF INTERNAL INDUCTANCES AND CAPACITANCES AT VERY HIGH FREQUENCIES.

(a) Basic tuned-anode tuned-grid oscillator; (b) equivalent circuit at very high frequencies.

effects of a base, the leads in ultra-high-frequency valves may be brought out directly through the envelope.

The "acorn" type of valve construction represents an approach to ultra-high-frequency design on such lines, and is of use up to about 500 Mc/s.

Electron Transit Time

The passage of electrons from the cathode to anode takes a finite time. For a standard valve this is of the order of one thousandth of a microsecond. Since it is required that the flow of electrons be governed by the potential on the grid, it follows that the potential on the grid should not change appreciably during the passage of the electrons. One thousandth of a microsecond represents about one-tenth of the time for one cycle at 100 Mc/s : the change of potential represented by this amount could probably be tolerated for most applications. However, at 400 Mc/s the same period represents one quarter of a cycle, and the operation of the same valve at this frequency would be seriously affected; the practical effects termed "transit loss" are "back-heating" of the cathode, increased loading of the grid circuit and a reduction of mutual conductance. As a result the power gain of the valve becomes progressively smaller as the frequency is increased.

The reduction of spacing between the valve electrodes—and in particular that between the cathode and grid, where the electron velocity is low—will reduce "transit loss", but there will be a practical limit to how far this process can be carried in valves of a conventional type.

Radio-frequency Losses

Radio-frequency valve losses increase as the frequency is raised, owing to increasing skin effect and dielectric loss, chiefly in the glass parts of the valve. The oscillatory current flowing in the valve leads gives rise to resistive losses, which tend to cause overheating of the leads, and which, if not taken into account, may affect the seals where the leads pass through the envelope.

V.H.F. Double Tetrodes

It has already been noted that the cathode lead inductance may have appreciable reactance at very high frequencies, causing a reduction in driving voltage and negative feedback. A popular method of overcoming this difficulty is the twin-tetrode valve : two separate electrode assemblies are placed in a common envelope, and the two cathodes joined by means of a low-inductance strip. When such a valve is used in a balanced or push–pull condition there will be no flow of radio-frequency current through the external cathode lead. In practice, a number of additional constructional improvements have been introduced in such valves in order to raise the frequency limits; these include reduction of anode–grid capacitances, the use of a common screen grid, which may be extended to provide shielding of the input from the output leads, silver plating of leads required to carry radio-frequency currents, etc. By such means, valves suitable for power amplification throughout the V.H.F. range have been made available.

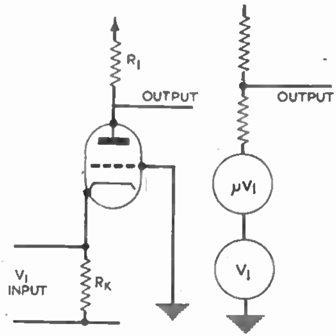


FIG. 14.—BASIC AND EQUIVALENT CIRCUITS OF GROUNDED-GRID AMPLIFIER.

Disc Seal and "Lighthouse" Valves

The cathode-input amplifier, or as it is commonly termed the grounded-grid amplifier, offers a number of advantages above about 500 Mc/s, and special triodes have been developed for use in this circuit. Since these valves have extremely low inter-electrode capacitances they may also be used in other applications requiring this property.

In the British form of construction, flat rings, sealed directly in the bulb, are used to connect the electrodes to the circuit. The discs provide not only low-loss and low-inductance connections, but materially assist in cooling the valve electrodes by conduction.

Disc-seal tetrodes have also been developed for a number of applications, although their upper frequency limit is still relatively low compared with the triodes, which may be used up to about 5,000 Mc/s.

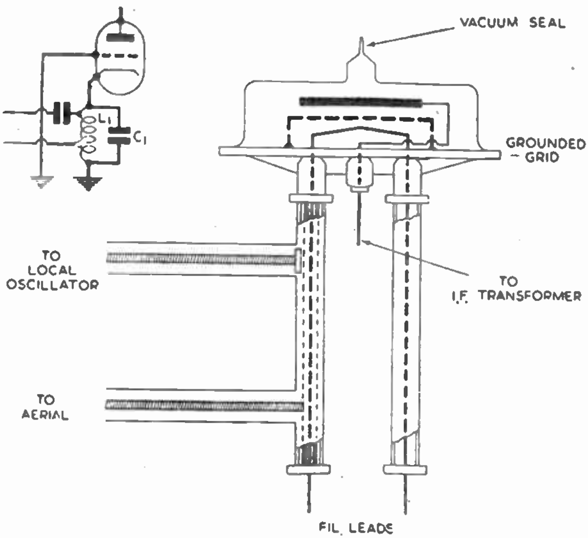


FIG. 15.—GROUNDED-GRID TRIODE EMPLOYED AS MIXER WITH EQUIVALENT CIRCUIT.

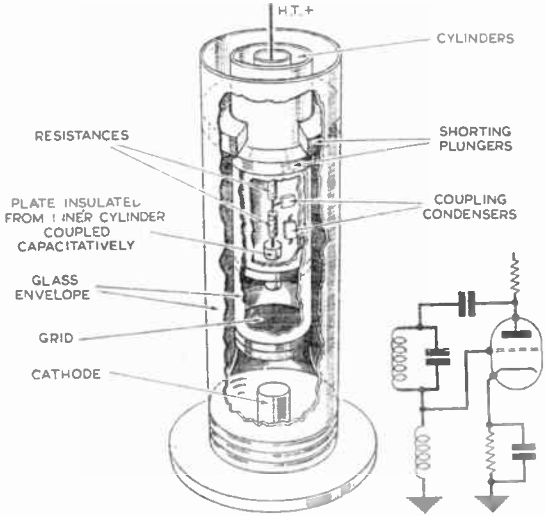


FIG. 16.—GENERAL CONSTRUCTION AND EQUIVALENT CIRCUIT ARRANGEMENT OF THE "LIGHTHOUSE" VALVE.

The shielding action of the grounded-grid is so effective that when oscillation is required, external feedback is necessary.

The general appearance of American-type disc or planar valves has given rise to the name "lighthouse" tube. The tube base and glass are so designed that the valve can be screwed into tuned concentric lines. Tuning is then effected by adjusting shorting plungers between cylinders.

Apart from the planar electrodes, the principal features of the lighthouse triode are: the small size of the anode and grid (3–6 mm. in diameter), their close spacing (0·1–0·4 mm.) and the inclusion of other components such as fixed capacitors and tuning plungers inside the envelope.

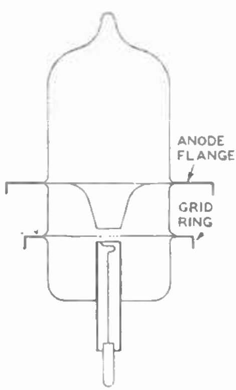


FIG. 17.—FORM OF CONSTRUCTION OF BRITISH DISC SEAL TRIODE.

MAGNETRONS

The magnetron is basically a diode valve operating in a magnetic field of which the lines of force are partly coincident with the fields of electron movement within the valve. The original magnetrons, developed in the 1920s, consisted of a filament mounted co-axially within a cylindrical anode, the assembly being enclosed in an evacuated glass envelope. The diode thus formed is positioned between the poles of a magnet so that the magnetic lines of force lie along the axis of the anode, parallel to the cathode (Fig. 18). In the absence of a magnetic field the device behaves as a normal diode. When a magnetic field is created, however, the electrons moving from the cathode to the anode are forced to follow a curved path; as the field increases this movement also increases until a point is reached when the electrons miss the anode altogether (cut-off point) and return to the cathode. During the return journey the electrons are moving against the force exerted upon them by the magnetic field, and because they are doing work on the field they must be surrendering energy to it, and therefore to the anode. If an

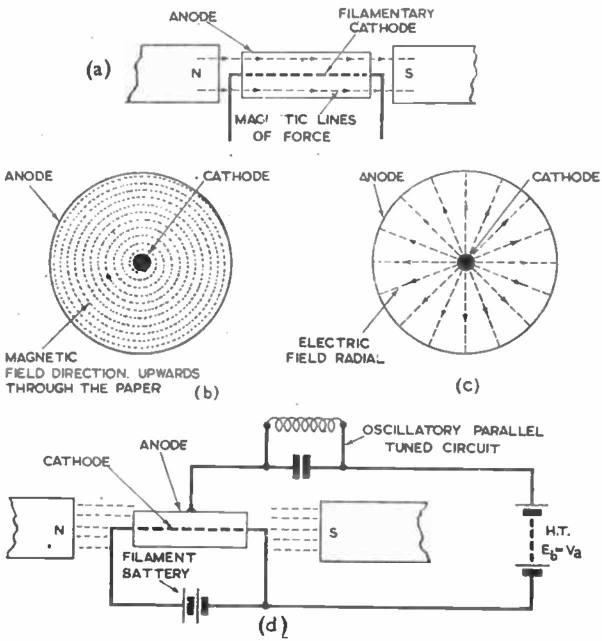


FIG. 18.—ORIGINAL CYLINDRICAL ANODE MAGNETRON.

(*a*) Disposition of cathode, anode and magnetic systems; (*b*) magnetic field; (*c*) electric field; (*d*) circuit arrangement of magnetron oscillator.

oscillatory circuit is now connected between anode and cathode, oscillations can be generated, the time period being equal to the transit time of the electron.

Improved efficiencies can be obtained by splitting the anode into two equal segments by means of two longitudinal slots and then connecting the oscillatory circuit between the two segments. As a result of splitting the anode, the radio-frequency fields due to oscillations excited in the parallel-tuned oscillatory circuit now appear across the gaps between segments and modify the electric field between cathode and anode. The effect is to change the curvature of an electron travelling from cathode to anode so that it travels in a spiral path with a general bias towards the anode segment of lower potential, which it ultimately reaches. Power output is greatly increased owing to the greater net loss of electrons from the cathode to the anode.

Extremely high frequencies (over 50,000 Mc/s) have been attained by using this mode of oscillation.

Cavity Magnetrons

A notable improvement in the magnetron was made during the Second World War by building the tuned output circuit—in the form of resonant cavities—into the valve itself. The anode of the resonant magnetron consists of a solid block of copper (or punched laminations) in which the resonant elements, formed by a combination of hole and slot, are drilled. The cathode is constructed from a nickel mesh to which an oxide coating is keyed. End shields prevent electrons from escaping to the anode without first interacting with the magnetic field. The end blocks, due to the raised rim on which they rest, do not make close contact over the whole of the anode surface, and there is, therefore, an air gap between the end block and anode, through which the fields of the respective cavities are coupled together. The output is obtained by magnetic coupling between the pick-up loop and the resonant cavity in which it is situated. The valve may operate in different modes at different frequencies: by connecting together alternate cavities frequency stability is improved.

Generally the number of cavities increases with the frequency of oscillation (e.g., 8 for a 10-cm. valve, 14 for a 3-cm. valve). It will be appreciated that such valves can be used only on the frequency for

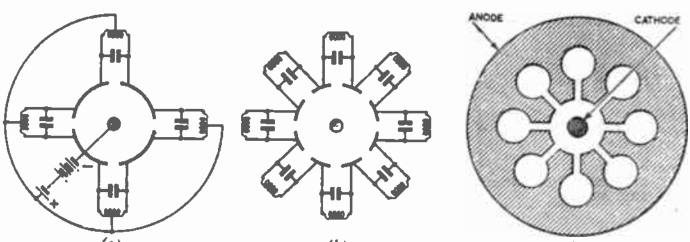


FIG. 19.—DEVELOPMENT OF THE RESONANT-CAVITY MAGNETRON.

(a) Four anode segments with separate tuned circuits across each gap; (b) eight anode segments; (c) tuned circuits replaced by resonant cavities.

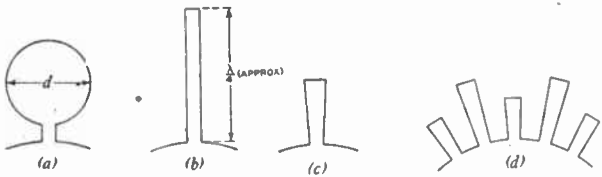


FIG. 20.—MAGNETRON CAVITIES.

(a) Hole and slot; (b) slit; (c) vane; (d) rising sun.

which they have been designed, since it would be difficult to alter the dimensions of the cavities: great care must also be taken during manufacture to ensure that all cavities in a particular valve have the same resonant frequency. Magnetrons which permit a limited range of tuning have been developed. Methods include:

(1) Use of conducting annular ring placed over the straps concentrically with the axis in a strapped magnetron. Frequency is changed by varying the distance between this ring and the straps.

(2) A ring of larger diameter placed over the resonant cavities will alter the inductance of the system.

(3) Insertion of rods into the cavities will diminish the inductance.

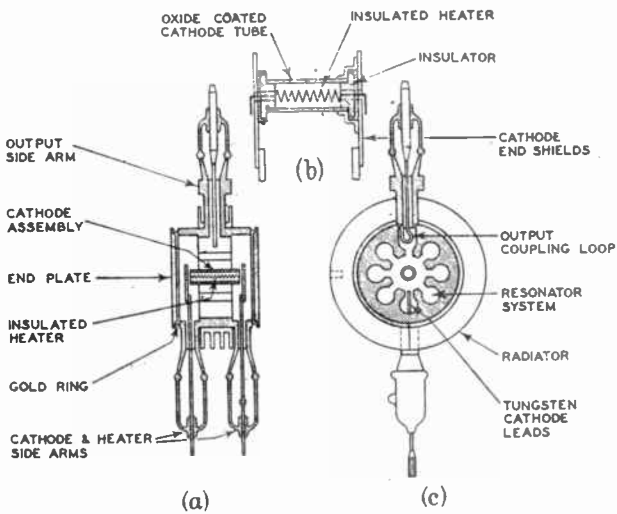


FIG. 21.—A MULTI-RESONANT MAGNETRON.

(a) Side elevation; (b) details of the cathode, heater and cathode suspension; (c) plan of magnetron looking along the axis of the cathode (end block removed).

(4) A tunable cavity resonator coupled to the resonant system of the magnetron.

The magnetic field is generally produced by placing a permanent magnet externally to the anode block so that the field is parallel to the cathode : in some designs, however, the magnet forms an integral part of the valve, and such valves are known as " packaged magnetrons ".

The efficiency of such valves may be as high as 50 to 60 per cent, and considerable power output can be obtained : peak outputs of the order of 2,000 and 250 kW at 10 and 3 cm. respectively have been obtained using pulse operation.

KLYSTRONS

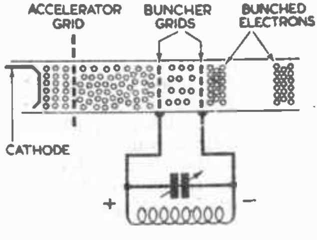
Basically, the klystron is a valve in which the velocity of the electrons in a steady stream is made to alter in such a way that the electrons form " bunches ", the velocity corresponding to an external or internal modulation. In a simple form the klystron consists of a heated cathode, an accelerating grid, two buncher electrodes, two catcher electrodes and an anode or " collector " as shown in Fig. 22

Electrons leaving the cathode are accelerated uniformly by the positive potential on the accelerator grid. As the buncher grids and catcher grids are maintained at a mean potential equal to that of the accelerating grid, any alteration in the potential will increase the velocity of the electrons with a positive change, or decrease it for a negative one. As the buncher grids are connected to opposite ends of a resonant circuit, the potentials vary according to the sine wave of the oscillating source (i.e., a resonant circuit).

When the first buncher grid is at peak positive amplitude, the electrons approaching it are accelerated, while those that have passed it are retarded. The second grid simultaneously is at peak negative potential (relative to the mean), and retards electrons approaching it while accelerating those that have passed it. These two actions, which are reversed every half-cycle, combine to cause the electrons to " bunch ", thus splitting up the electron stream into a series of bunches corresponding to the modulating source.

This stream of regularly spaced bunches of electrons then travels through two " catcher " grids. These have a similar oscillating circuit, but are spaced so that the transit time of the electrons between the grids will equal the duration of half of one cycle of the oscillations. The phase of the catcher grids in relation to the spacing of the bunches is arranged so that a bunch arrives at the first buncher grid when the

Fig. 22.—Modulation of the Velocity of an Electron Stream by the Alternating Electric Field between a Pair of Grids Connected to a Resonant Circuit.



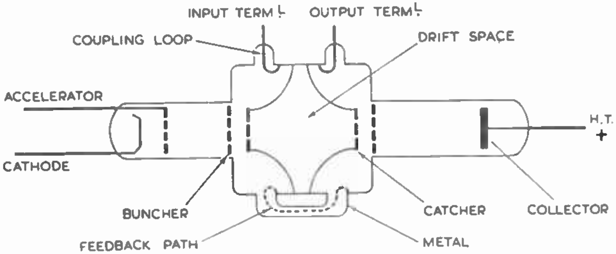


FIG. 23.—KLYSTRON WITH CAVITY RESONATORS.

latter is at peak negative amplitude and, therefore, retards the electrons. As the transit time is equal to that of one half-cycle, the second grid is also at peak negative amplitude when the bunch arrives, still further retarding it.

At each retardation the bunch gives up energy to the catcher grids. This energy may be used either for providing an amplified signal from the original oscillation or it may be fed back to the buncher grid to cause oscillations either with tuned circuits or with internal or external resonant cavities, in which case it acts as an oscillator (reflex klystron) or as a mixer.

Beyond the catcher grid is a "collector" electrode (at a positive potential) to attract electrons that have not been usefully absorbed in the operation of the valve.

The size of klystron valves varies widely according to the application, from the size of small receiving-type valves to large transmitting types with forced air or water cooling, with accelerating voltages of the order of 15 kV. The efficiency, in practice, may be of the order of 15 per cent. Klystrons are available for frequencies up to at least 10,000 Mc/s. When used at high frequencies, the resonant circuits are usually resonant cavities, the grids being connected to each side of the cavity; the cavities may be constructed internally or externally. Tuning is often adjusted by screwing plugs into the periphery. The energy is coupled into or out of the cavities by means of single-turn loops.

Various electrical methods of tuning the cavities are also used.

The Reflex Klystron

A simplified form of klystron is generally used as a local oscillator in the U.H.F. range: one set of grids, together with their associated cavities, are used to perform the double function of bunching and catching, and the positive collector is replaced by a highly negative repeller electrode (see Fig. 24).

By correct adjustment of the negative voltage applied to the repeller, electrons which have passed through the buncher field may be made to return again through the resonator grids in the correct phase to deliver energy to the circuit, thus providing the feedback required for oscillation. The frequency of oscillation can be varied slightly by adjustment of the repeller voltage, but at the cost of variation in output.

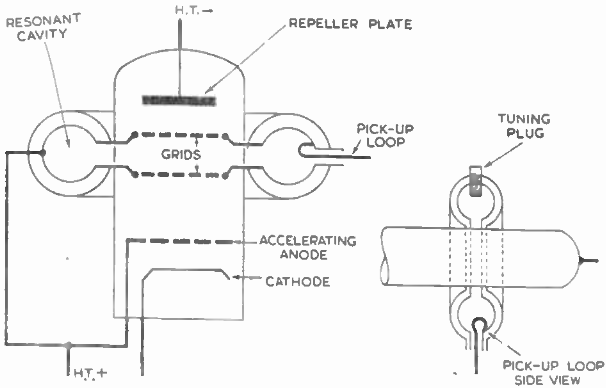


FIG. 24.—REFLEX KLYSTRON.

The resonant cavities are formed by a copper tube which encircles the envelope.

While the use of this type of klystron is limited to oscillation, the comparative simplicity of setting up, small size, cheapness of construction and ease of tuning (generally carried out by adjustment of the volume of the cavity by means of plugs) have all tended to bring such valves into general use on micro-wavelengths.

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Multi-cavity Klystrons

By adding extra cavities between the buncher and catcher cavities of a klystron amplifier, the efficiency and gain can be enormously increased. By using two buncher cavities, for example, the bunching efficiency can theoretically be increased to 80 per cent. Also, stagger-tuning of the cavities enables large band-width to be obtained, though at the expense of gain. High-power pulsed klystrons and high-gain high-efficiency valves for lower power applications using three or more cavities have been developed. An experimental three-cavity high-power valve has given 20 MW pulsed power at 30 per cent efficiency at 10 cm. wavelength. The valve operates at 300 kV beam voltage, 180 amperes beam current and gives a gain of 30–35 dB. Similar high-power valves giving 1–2 MW in the 3-cm. and 30-cm. bands have been constructed.

For lower power applications, five- and six-cavity klystrons have been used. A typical five-cavity valve at 3 cm. may have a gain of up to 70 dB compared with 15 dB for a normal two-cavity version. Magnetic fields for focusing the electron beams are usually necessary in such valves. In general, the maximum useful gain which can be obtained is determined either by the signal-to-noise ratio at the output or by the onset of oscillations.

TRAVELLING-WAVE TUBES

Valves such as the magnetron and klystron have very narrow band-widths, due mainly to their use of cavity resonators. At most a 10 per cent tuning range can be obtained. Also, interaction between the electron beam and the R.F. field occurs only in the narrow region of the cavity gaps. In travelling-wave tubes, however, no cavities are used, but instead circuits having the properties of wide-band filters. Also, interaction is made to occur over an extended region by making an electron beam and an R.F. wave travel together, interacting continuously. The band-width limitations of such tubes are thus limited mainly by the input and output matching arrangements rather than by the circuits themselves.

Various types of travelling-wave tube have been developed. The most important of these are the travelling-wave amplifier, in which interaction occurs between an electron beam and an R.F. wave travelling in the same direction; the backward-wave oscillator or O-Carcinotron, in which interaction occurs between an electron beam and an R.F. wave travelling in opposite directions; and the M-Carcinotron, a backward-wave oscillator using crossed electric and magnetic fields. i.e., fields acting at right angles to one another and to the beam.

Travelling-wave Amplifiers

In the travelling-wave amplifier an electron beam and the R.F. wave to be amplified are made to travel alongside each other at more or less the same speed. The action of the wave on the beam causes electron bunching to occur, as in the klystron, which increases along the length of the tube. The bunched beam reacts back on the wave in such a manner as to cause a certain amount of electron kinetic energy to be converted into R.F. energy of the wave. A much amplified wave thus emerges from the tube.

Since, with the normal accelerating voltages used in practice, the electron beam travels at less than one-tenth of the velocity of an electromagnetic wave, the wave must be slowed down. This is done by means of a " slow-wave structure " or delay line, which may take many forms. The one generally used is a helix of wire. The wave travels round the helix so that, although its velocity along the wire is roughly its normal velocity, its velocity along the axis of the helix is much slower. By suitable choice of helix pitch the velocity of the wave along the axis can be made as little as one-tenth its normal velocity.

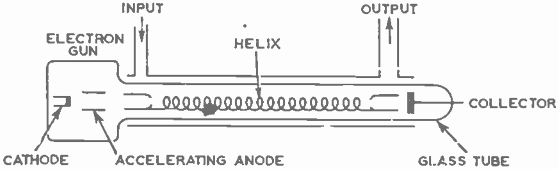


FIG. 25.—TRAVELLING-WAVE AMPLIFIER.

The helix is a non-dispersive structure, i.e., the velocity of electromagnetic waves along it is the same irrespective of frequency, within wide limits. Thus interaction can occur with R.F. waves covering a wide range of frequencies without any alteration of beam voltage or circuit parameters being necessary.

The structure of a travelling-wave amplifier is shown in Fig. 25. It consists essentially of a helix of wire about 6 to 9 inches long and about 0·1 inch internal diameter supported in an evacuated glass tube, which contains at one end an electron gun and at the other a metal plate or collector. A beam of electrons produced by the gun travels through the helix towards the collector. In order that most of the beam should reach the collector and not be intercepted by the helix, a focusing magnetic field is applied along the tube, either by a solenoid round the helix and gun or by permanent magnets forming a magnetic lens system. Power is fed into and out of the helix by means of either coaxial or waveguide couplings at the helix extremities. In order to prevent feed-back of reflected R.F. power from the output end, an attenuating layer is applied to part of the helix near the centre of the tube.

A number of different types of travelling-wave amplifiers are now available in the microwave band. At 10 cm., low-noise tubes having noise figures around 6 dB and gains of about 20 dB, operating at 500 volts and currents of under 1 mA at power levels around 1 mW, intermediate power tubes with gains of 30 dB operating at 1 to 2 kV and currents of a few milliamperes, and high-power tubes capable of handling over 100 kilowatts at 20 dB gain have been developed. The helices for such high-power tubes are generally copper tubes through which cooling water flows.

The most important property of the travelling-wave amplifier is its large band-width. Thus a typical valve operating at 4,000 Mc/s may have a band-width of over 1,000 Mc/s and could therefore handle 10,000 telephone channels or nearly fifty high-definition television programmes simultaneously. Travelling-wave amplifiers are thus extensively used in non-demodulating radio relay links.

Backward-wave Oscillators

Backward-wave oscillators (or O-Carcinotrons) are travelling-wave tubes in which the electron beam and the R.F. wave travel in opposite directions. Thus the power emerges at the gun end of the tube while the electrons travel towards the collector end. As in the travelling-wave amplifier, a slow-wave structure is used, but one having rather different properties from those of the helix. In the backward-wave oscillator the slow-wave structure is dispersive, i.e., the velocity of a wave along it depends on the frequency of the wave. Analysis of the wave travelling on such a structure shows it to consist of an infinite number of components, some travelling in the same direction as the wave and others in the opposite direction. It is one of these latter components, travelling in the opposite direction to the R.F. power flow but in the same direction as the beam, with which the electrons interact. Bunching of the electrons occurs and the travelling wave gains energy from the beam as in the travelling-wave amplifier. Interaction normally occurs with the fundamental of the infinite series of backward-wave components, the higher components being too weakly

Efficiencies of about 30 per cent are obtained in practice. M-Carcinotrons are generally curved into a circle for convenience, but this does not affect their operation.

PARAMETRIC AMPLIFIERS

Parametric or reactance amplifiers derive their name from the fact that the equations governing their operation contain one or more time-dependent reactance parameters. Recently the name Mavar (*M*ixer *A*mplification by *Va*riable *R*eactance) has been used.

The operation of the parametric amplifier depends on the fact that a time-varying reactance can exhibit negative resistance characteristics under certain conditions, or can act as a frequency converter under other conditions. Consider the system shown in Fig. 29. This consists of a resonant circuit and source of frequency f_1 coupled to a resonant circuit of frequency f_2 by a reactance, represented in this case by a capacitor C_3. If the capacitance is varied sinusoidally at a frequency $f_3 = f_1 + f_2$, analysis shows that power will be fed to the two circuits at their respective frequencies. Thus if a load is coupled to the circuit of frequency f_1 an amplified signal at this frequency will be obtained in the load. Alternatively, if the load is coupled to the other circuit, an output signal of frequency f_2 will be obtained, the system in this case acting as a frequency converter.

An arrangement such as this may be obtained in practice by using a source of frequency f_3 in conjunction with a non-linear reactance, e.g., a capacitor whose charge is a non-linear function of the voltage, or an inductance for which the flux varies non-linearly with the current. The frequency f_1 is called the signal frequency, f_2 the idler frequency and f_3 the pump frequency. The system thus converts energy at the pump frequency into energy at the signal and idler frequencies, thereby giving amplification of the signal. A schematic representation of a parametric amplifier would thus be as shown in Fig. 30.

A device such as this has the disadvantage that it is necessary to supply power at a higher frequency in order to obtain amplification at a lower frequency. However, by using two pumping sources in conjunction with a reactance having a third-order non-linearity, e.g., an inductance for which the flux varies as the cube of the current, the pump frequency can be lower than the signal frequency.

It can be shown that gain in such devices is equivalent to the intro-

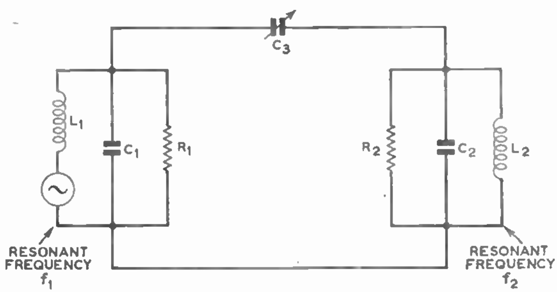
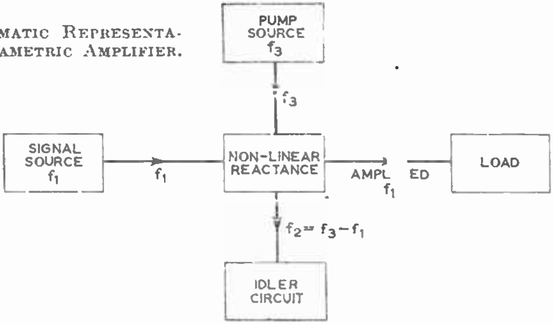


Fig. 29.—Two Circuits Coupled by a Time-varying Capacitor.

FIG. 30.—SCHEMATIC REPRESENTATION OF A PARAMETRIC AMPLIFIER.



duction of negative resistance in the signal circuit. This negative resistance is a function of the pump power, and hence is controllable. Amplifiers using these general principles may be constructed at any frequency from audio up to microwave frequencies, the essential requirement being the availability of a suitable variable reactance. It is at microwave frequencies, however, that most of the development has occurred, since the devices are inherently low-noise ones due to the absence of electron beams. At microwave frequencies successful amplifiers have been constructed using ferrites, reverse biased semiconductor diodes and modulated electron beams as the variable elements.

The performance of these devices may be illustrated by the results obtained for parametric amplifiers using semiconductor diodes. These make use of the fact that the capacitance of a p–n junction varies non-linearly with voltage when negatively biased. For a germanium junction, for example, a change of the order of 15 volts in bias produces a 3 to 1 change in capacitance. With such diodes, parametric amplifiers have been constructed using 50–500 mW of pump power at 12,000 Mc/s to give a stable gain of 45 dB at a signal frequency of 6,000 Mc/s. The gain band-width product is constant. With the gain adjusted to about 20 dB a band-width of about 5 Mc/s is possible with a noise figure of 5 dB. Noise figures as low as 1 dB have been obtained.

THE MASER

The maser is a device which amplifies microwave signals by making use of the radiation emitted when transitions occur between energy states in atoms or molecules. The name derives from "*M*icrowave *A*mplification by *S*timulated *E*mission of *R*adiation".

According to the principles of quantum physics, atoms or molecules may exist in any one of a number of discrete energy states. Most of the atoms or molecules will exist in the lowest or ground state, but they may be raised to any of the possible higher states by causing them to absorb radiation of frequency f given by

$$hf = W_2 - W_1$$

where W_2 is the energy of the higher state, W_1 that of the lower state and h is a constant called Planck's constant. Conversely, molecules or atoms in a higher state, on transferring to a lower state, will emit radiation of frequency f depending on the difference in energy between the states in the same way. Such transitions from higher to lower states may also be induced by applying a small amount of radiation of frequency f. If it were possible to prepare a material in which more molecules existed in the higher energy state than in the lower, the application of a small amount of radiation of the correct frequency would cause more downward than upward transitions. There would thus be a net emission of radiation at this frequency and consequent amplification of the applied signal. This is the essential principle of maser action.

The first maser constructed used ammonia gas. This can exist in two possible energy states, the difference between which corresponds to a frequency of 24,000 Mc/s. Molecules in the higher energy state may be separated from those in the lower energy state by placing them in an electrostatic field, since those in the higher state are repelled while those in the lower state are attracted by such a field. Thus if a beam of ammonia molecules traverses such a field in a cylinder, only those in the higher energy state will emerge, the lower-energy-state molecules being attracted to the walls of the cylinder. The emerging beam is then fed into a cavity resonator tuned to 24,000 Mc/s and subjected to a small signal of this frequency. Transitions to the lower state are thus induced and energy is given to the microwave signal. Such a device, called a two-level maser, can easily be adapted to oscillation.

The signal from the ammonia maser is stable to one part in 10^{10}, and it can thus be used as a frequency or time standard. As an amplifier, however, its disadvantage is that it has virtually no band-width, since it can only amplify a signal of 24,000 Mc/s. Masers using solid-state materials (artificial ruby being a commonly used substance) can, however, be tuned by subjecting them to a magnetic field, which alters the energy difference between states and thus gives a limited range of frequency operation.

Three-level masers have also been constructed. In these, materials having three possible energy levels W_1, W_2 and W_3 are used. The material is first subjected to radiation corresponding to the difference $W_3 - W_1$, thus transferring the substance from its ground state W_1 to the higher state W_3. By applying a small signal of frequency $(W_3 - W_2)/h$ transitions from W_3 to W_2 occur, with the emission of radiation of this frequency and consequent amplification of the applied signal. Using potassium cobalti-cyanide containing 0·5 per cent of chromium as maser material and applying 1 mW of "pumping" power at 9,000 Mc/s, 30 dB gain at 3,000-Mc/s signal has been obtained.

Such masers must be operated at extremely low temperatures to be successful and are normally surrounded with a liquid helium bath. Under such conditions they have a very low noise level; noise figures as low as 0·3 dB have been achieved. Such low-noise operation is far in advance of anything obtainable with conventional amplifiers.

R. C. G.

24. CATHODE-RAY TUBES

24. CATHODE-RAY TUBES

The cathode-ray tube, together with its associated time-bases and amplifiers, is now to be found in almost every branch of electronic and scientific work. Recent years have seen many advances in the design and construction of the electron gun, while the rapid development of television has made necessary the mass-production of tubes with large screens. Gas-focused tubes, which suffered from comparatively short life and excessive spot-swelling when the beam current was increased, have now been superseded by the high-vacuum types in which pressures lower than 10^{-6} mm. Hg are attained.

Development

The apparatus used for early experiments has little in common with the modern cathode-ray tube, but the pioneers did succeed in establishing certain basic principles which paved the way for subsequent development.

The cathode-ray tube is named after Hittorf's discovery in 1868 of "cathode rays", although since J. J. Thomson's investigations in 1879 it is now known that the once mysterious "cathode rays" consist of electrons travelling at extreme speed from a negative electrode towards parts of the tube at positive potential; a more correct name would therefore be "electron beam".

It has been known since 1868 that the cathode rays could be deflected by magnets. A. Hess was the first to propose their use for oscillographic purposes, but this was realized for the first time by Ferdinand Braun in 1897. The Braun Tube was evacuated to a pressure of a few bars. Its gas filling was usually nitrogen. At a pressure of this order, the electronic pencil left the finest trace. Two years later Wiechert proposed to make it finer by a magnetic coil, termed the "concentrating coil". Such tubes were used for laboratory work for some twenty-five years. In 1921, A. Dufour built the first high-speed cathode-ray oscillograph in which the electrons recorded directly on a photographic plate which was introduced into the vacuum. Many workers, including Rogowski, Gabor, McEachron, Knoll, Burch and Whelpton, have helped to bring this instrument to its present state, and since 1925 it has been used extensively. Direct photography, however, has now been generally superseded by the sealed-off cathode-ray tube with thermionic cathode.

After the discovery by Wehnelt in 1905 of thermionic cathodes, attempts were made to use them in high-vacuum cathode-ray tubes, these were unsuccessful until 1926, when Burch explained the action of the concentrating coil as an electron lens. These theoretical results were exploited by V. K. Zworykin, who in 1929 combined for the first time the two essential elements of the modern cathode-ray tube: a small thermionic cathode and an electron optical-lens system.

The simple electrode arrangement of Fig. 1 is now seldom found in practice, but illustrates the basic principles of electrostatic focusing.

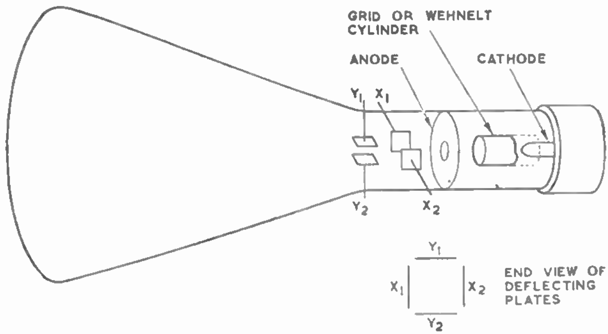


FIG. 1.—SIMPLE GAS-FOCUSED ELECTROSTATIC TUBE.

These "soft" (gas-filled) tubes generally required a heater supply of 2–4 volts, anode potential of 500–1,000 volts and a negative shield bias of 5–50 volts. Control of the grid potential was the primary means of focusing the spot, which, when correctly focused, was 0·5–1·0 mm. in diameter.

If a potential is applied across the X plates, depending upon the sign of the applied voltage, the cathode ray will be repelled by one X plate and attracted by the other. In Fig. 2, plate X_1 is positive and plate X_2 is negative, with the result that the electron beam is deflected towards X_1. The movement of the ray at these plates will be apparent as a movement of the fluorescent spot on the screen, the degree of movement on the screen being enlarged in proportion to the length of the ray. If now the connections to X_1 and X_2 are reversed, the spot will move across to the opposite side of the screen. When an alternating voltage is applied to the X plates, therefore, the spot will clearly move from side to side in step with the frequency of the alternations. Because of the afterglow in the fluorescent screen and the rapidity of the movement, the effect to the eye will be a steady line across the screen. The length of the line will be directly proportional to the voltage applied to the deflecting plates.

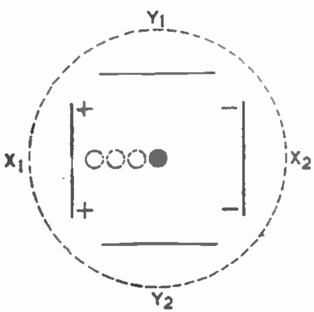


FIG. 2.—A POSITIVE POTENTIAL ON THE X_1 PLATE CAUSES THE ELECTRON BEAM TO SWING TOWARDS THAT PLATE.

Suppose that while this line is being traced out on the screen a D.C. potential is applied to the Y plates. It is obvious that the spot must tend to respond to the force thus exerted on the ray as it passes between the Y plates. This response will take the form of shifting the line bodily up or down on the screen, depending on whether Y_1 is positive or negative. But what will happen if instead of D.C. on the Y plates an alternating voltage is applied? It is assumed for the moment that both the X and Y voltages are derived from the power mains and are therefore both sinusoidal and of the same frequency and phase.

Two forces are now acting on the ray at the same time, and its movement will be proportional to the resultant of those forces at any instant. The line on the screen will, therefore, assume a new position, making an angle with the horizontal which is governed by the relative value of the deflecting forces. If the alternating voltages applied to the X and Y plates are equal, the angle of the line on the screen will be 45°. By variation of the relative values of the voltages applied to the X and Y plates the angle will correspondingly become greater or less.

The straight line on the screen is preserved only while the two applied voltages are of the same phase. Where there is a phase difference, the resultant of the forces acting on the ray differs from instant to instant, with the result that the spot traces out an elliptical path ranging from the straight line already discussed when the phase angle is 0° to a circle when the phase angle is 90° (and the alternating voltages are equal).

Hard Cathode-ray Tubes

The gas-filled cathode-ray tube makes use of ionization by collision to assist focusing of the ray. This is a simple and effective method for low-frequency work, but the gas-filled tube has largely given place to the vacuum type for modern use. These hard tubes usually operate at far higher potentials than the gas-filled types and incorporate additional anodes or accelerators to assist focusing. In the double or triple accelerator type of cathode-ray tube, the Wehnelt cylinder is still adopted but does not play the important part in focusing that it does in the gas-filled tube. The arrangement of the electrodes and the potentials applied are so designed as to produce electrostatic fields which form "electron lenses" acting on the electron beam in the same manner as optical lenses focus a ray of light.

The accelerator potentials are normally progressively increasing, although in some designs the first and third anodes are strapped internally.

Construction

Modern cathode-ray tubes consist of: (1) a gun, which, in the case of a triode arrangement, comprises a cathode, modulating electrode and first accelerator; (2) a lens, either electrostatic or magnetic, for focusing the beam; (3) means for deflecting the electron beam, again either electrostatically or electromagnetically; and (4) a fluorescent screen, which when bombarded by the beam converts some of its energy into visible light.

There are two main types of magnetically focused tubes, namely triodes and tetrodes. The tetrode has, between the modulator and the final anode, an auxiliary anode to which is applied a voltage of a few hundred volts. With identical positioning of the focusing magnet, and the same E.H.T. voltage applied to both tubes, the tetrode requires less field strength for focusing than the triode.

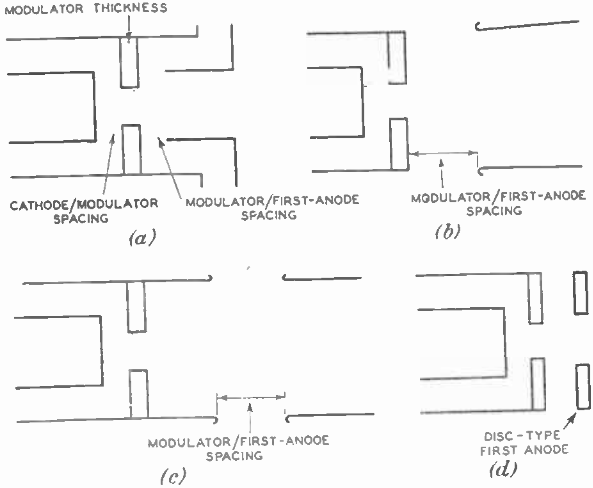


FIG. 3.—TRIODE ASSEMBLIES FOR CATHODE-RAY TUBES.

(*a*) Triode for electrostatic tube. (*b*) Triode for magnetic tube giving large beam angle. (*c*) Triode for magnetic tube giving small beam angle. (*d*) Disc-type triode forming part of tetrode gun.

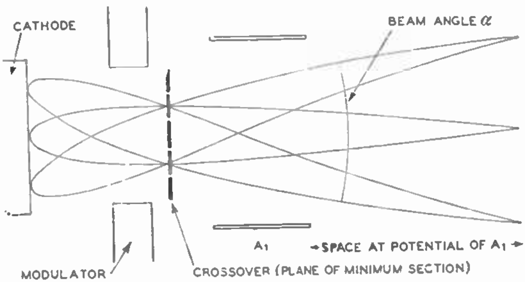


FIG. 4.—ELECTRON TRAJECTORIES.

The choice between permanent and electromagnets for focusing is usually one of economics. The stray fields of permanent magnets are usually greater than those of electromagnets. A further technique sometimes employed for focusing a television tube uses a combination of both permanent and electromagnets. The electromagnet is located inside the permanent magnet and a small variable current passed through the coil to serve as a fine adjustment of beam-focus.

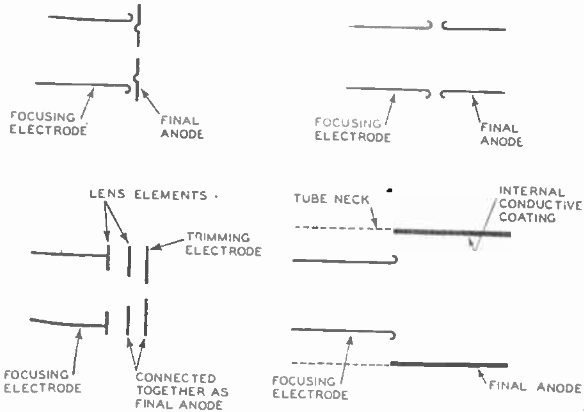


FIG. 5.—FORMS OF ELECTRODES FOR FOCUSING.

Although tubes using electrostatic means of deflection are now seldom employed for television work, they are still extensively used in oscillography and other forms of instrument work.

TELEVISION PICTURE TUBES

The desirable features of a cathode-ray tube for television reception may be summarized as follows :

(*a*) GEOMETRICAL ACCURACY. It should be possible to provide on the face of the tube a picture which appears in correct proportion not only to people sitting immediately in front of the tube, but also to those sitting some way on either side. Fundamentally this is impossible of achievement. The best approximation is given by a large flat screen and the worst by a small curved one.

(*b*) DEFINITION. The spot should be as small as possible. Theoretically there is no lower limit to its permissible size, but there is no point in making its diameter much smaller than the distance between the centres of adjacent scanning lines.

(*c*) CORRECT GRADATION AND ADEQUATE LUMINANCE. The relative luminosity observed by the viewer when looking at different parts of the received picture should be identical with that which he would experience if he were looking at the original scene. This is a complex subject which is still far from being fully under control in practice, as it relates not only to the luminance of the spot as a function of the applied modulating voltage, but also to the ambient viewing conditions and the type of transmission characteristic employed in the system.

(*d*) ACCEPTABLE COLOUR, FREEDOM FROM FLICKER AND ABSENCE OF STREAKING AND COLOUR FRINGING. All modern television picture tubes employ phosphors which give a black-and-white picture. The

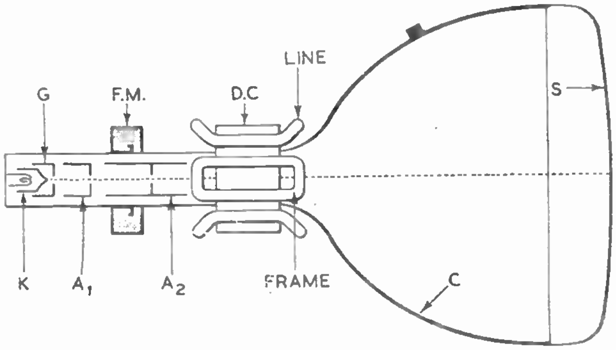


FIG. 6.—ARRANGEMENT OF A TYPICAL TETRODE PICTURE TUBE.

decay time of the phosphorescence must not be so long that moving objects become blurred (possibly with coloured fringes if the components of the phosphor have different decay times), but within this limitation it should be as long as possible, as this reduces the amount of flicker present in the picture.

(e) LONG LIFE. A cathode-ray tube is an expensive item, and it is essential that it should be a reliable one.

(f) EASY MODULATION AND DEFLECTION. The required grid drive and scanning power should be as low as possible in order to facilitate the design of the receiver.

(g) PRODUCTION. The tube should be suitable for large-scale production, so that its price can be as low as possible.

Many of these factors are mutually conflicting, and it is necessary to strike a compromise.

The majority of post-war television receiver tubes have magnetic focusing and deflection. Within the past few years, however, there has been a trend towards electrostatic focusing of the electron beam, thus dispensing with the bulky external focusing magnets. Fig. 6 shows the general arrangement of a typical tetrode tube employing magnetic focusing, while Fig. 7 shows a typical tube employing electrostatic focusing. Referring to Fig. 6, electrons are emitted by the indirectly heated cathode K and are attracted by the first anode A_1, which is usually 150–300 volts positive with respect to the cathode. A_1 consists of a skirted disc with a small central hole through which most of the electrons pass up the tube to strike the fluorescent screen S, which is composed of " phosphors " that emit light under electron impact, e.g., zinc sulphide, blue, or zinc cadmium sulphide, yellow. To obtain an approximation to white light, mixtures of phosphors are often used. Surrounding the cathode is the cup-shaped grid G, also with a small central hole. An electric field is produced between grid and cathode by virtue of their potential-difference, and this controls the number of electrons drawn through the first anode—and hence the brightness of the spot. A grid voltage 25–100 volts negative to cathode will cut off the electron flow altogether. The second anode, A_2, connected internally to

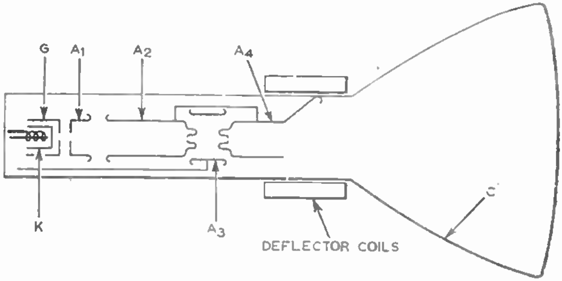


FIG. 7.—TYPICAL TELEVISION TUBE EMPLOYING ELECTROSTATIC FOCUSING.

the graphite coating C inside the wall, gives further acceleration and governs the final electron velocity: its voltage is 6–16 kV (20–50 kV in projection models), determining the ultimate possible brightness of the screen fluorescence.

Focusing

The electrons, travelling at high velocity, must be focused to a small spot, i.e., $\frac{25}{1000}$ in. in diameter on a 14-in. tube. Now a flow of electrons, whether in a wire or not, is a current; and when a current flows across a magnetic field, it experiences a force which tends to move it. In the cathode-ray tube the electron stream thus experiences the deflecting force when passing across a magnetic field. By means of a short coil, or a permanent-magnet arrangement, a field, as shown in Fig. 8 (*a*), is produced. Electrons passing along the tube axis do not cross this

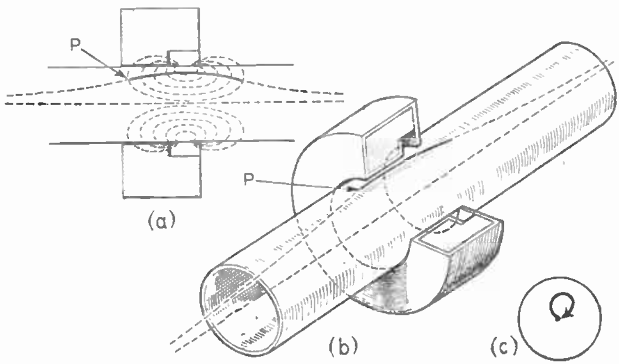


FIG. 8.—FOCUSING BY MEANS OF A SHORT COIL.
(*c*) Shows an end view of the electron path.

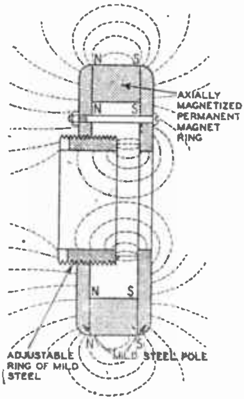


FIG. 9.—FOCUS UNIT USING PERMANENT-MAGNET RING AND POLE PIECES.

This type of focus unit is very common because of the popularity of the A.C./D.C. type of television receiver, in which it is difficult to find power for an energized focus coil. A thicker iron circuit is needed than for a solenoid in order to eliminate peripheral variations in magnet strength. Two opposing axially magnetized rings of Magnadur—a magnetic ceramic material—are often used. This has the advantage of having smaller external fields, and it requires no iron circuit because, being pressed from powder, the rings are more homogeneous than rings made from normal cast magnetic materials.

field, and so are unaffected, but those entering at P are deflected into a helical path whose radius is proportional to the strength of the field, and whose pitch is proportional to the *tangential* velocity of the electrons. Since the *axial* velocity is not affected by the magnetic field, all electrons may be brought to a focus on the screen in a spot corresponding to the origin—the cathode. It is thus important for the cathode to be small so that the focused spot is well defined.

In electrostatically focused tubes (see Fig. 7), electrodes A_2 and A_4 are connected together internally and also to the graphite coating C to which the tube E.H.T. is applied. The potential of electrode A_3 is adjusted (usually about cathode potential) to focus the beam on the screen.

Deflection

The spot must be made to traverse the screen rapidly from side to side and more slowly from top to bottom to produce the " raster ". For line scanning a varying magnetic field is required which varies linearly to deflect the spot left-to-right and then rapidly right-to-left. Simultaneously, there must be a corresponding top-to-bottom and return motion for frames. D.C. in Fig. 6, shows the deflecting coils: the deflection is at right angles to the direction of the magnetic field, so that the vertically disposed field is that of the line-deflection coils. In this way the spot will trace out the " raster ", which corresponds to the pattern traced by the camera tube.

Modulation

To reproduce the " picture " as seen by the camera, the instantaneous brightness of the spot is modulated by the instantaneous variation of the cathode-grid potential from its mean or quiescent value, determined by the " brilliance " control. This may be done either by applying a positive going signal to the modulator, which is normally biased so that the beam is suppressed, or by applying a negative-going signal to the cathode.

Ions

In addition to the electrons, positively and negatively charged ions are also produced in the cathode-ray tube. The former are attracted to the grid and cathode, but the latter travel to the fluorescent screen. They are between 5,000 and 500,000 times heavier than electrons, and though like electrons they are attracted by the first and second anodes, they are much less deflected by magnetic fields. The deflection of electric particles by a magnetic field is directly proportional to the strength and the length of the deflecting field and its distance from the screen, and inversely proportional to the particle velocity. But the latter varies as the square root of the quotient of the final anode voltage and particle mass : hence the heavy negative ions are less deflected, and they destroy the fluorescent property of the screen by bombardment over a small central area. The result is a dark spot referred to as

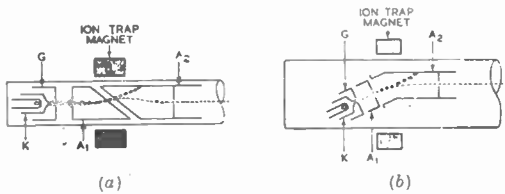


FIG. 10.—POPULAR ION-TRAP ARRANGEMENTS.
(a) With " undeflected beam." (b) " Bent-gun " type.

" ion-burn ". Ion-burn may be avoided by removing the ions from the beam, either by passing the beam through an electrical and then a magnetic field so that the ions are deflected in the first and not in the second (see Fig. 10 (a)) or by a " bent-gun " in which the beam is initially projected at about 10° to the tube axis and an " ion-trap " magnet deflects the electrons back to the tube axis. The ions, however, carry nearly straight on to the second anode (see Fig. 10 (b)).

A different treatment of the problem is to prevent the ions in the beam from reaching the fluorescent material by backing it with a very thin metallic film. Aluminium is the usual metal, and considerable success in reducing ion-burn has attended "aluminizing", which was originally introduced mainly to improve brightness and contrast. In some modern tubes both methods of reducing ion-burn are used.

Aluminizing

In a non-backed tube about half of the total emitted light is visible at the front of the phosphor. The remainder appears at the back, and even with an internal graphite coating some of this light is reflected back to the screen, where it reduces contrast by faintly illuminating the whole screen. With aluminizing, the useful output is increased, as the aluminium film reflects the backward output, improving both brightness and contrast. For maximum efficiency the thickness of the aluminizing film is closely related to the final E.H.T. voltage, so that the indiscriminate fitting of an aluminized tube may not result in any noticeable improvement in performance.

Dark- or tinted-glass screens have also been introduced to improve picture contrast when viewing in strong ambient light. Ambient light illuminates the " black " portions of the screen and reduces contrast. With tinting, the small loss of brightness (as the phosphor is seen through the tinted glass) is compensated by increasing the brilliance control. The " black-level " seen by the viewer is determined by the ambient light reflected from the screen, and as this has to pass through the tinted glass twice (once to reach the phosphor and once to reach the viewer) the result is a darker " black " and improved contrast.

Metal Tubes

At one stage in the development of television tubes difficulty was experienced in manufacturing glass bulbs of about 16 in. diameter or more. To overcome this difficulty, metal-cone tubes were introduced.

Within recent years, however, it has become possible to produce rectangular glass bulbs of any size up to 27 in. diagonal, and, as a result, very few metal-cone tubes are now used for monochrome television receivers. At present, metal-cone bulbs are used in certain designs of colour television tubes, but it is possible that even in this application, the metal-cone bulbs will be replaced by glass bulbs.

Fluorescent Screen Shape

The fluorescent screen, formerly circular, is now available in a rectangular shape approximately the size of the reproduced picture.

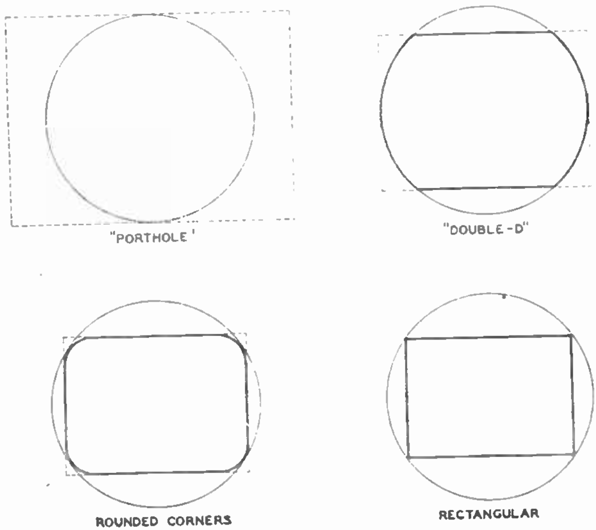


FIG. 11.—THESE ILLUSTRATIONS SHOW THE DIFFICULTY OF ACCOMMODATING A RECTANGULAR PICTURE ON A ROUND SCREEN. THE DOTTED LINES REPRESENT LOST PICTURE AREA.

The picture sizes available on circular tubes are approximately 8 × 6 in. on a 10-in. tube; 9½ × 7 in. on a 12-in. tube; 11 × 8 in. on a 14-in. tube; and 13 × 10 in. on a 16-in. tube; so that if the unused height of 4–6 in. can be saved with the rectangular face, cabinet size and cost will be reduced.

Handling Cathode-ray Tubes

Although the danger of implosion when handling tubes is relatively slight, provided that care is taken not to let the neck strike the chassis or hard surface, it *can* occur, and it is advised that precautions to guard against the effects of such an implosion should always be taken. The high vacuum of modern tubes means that the pressure on the envelope is very high, amounting to about 1½ tons on a 14-in. tube. It is therefore advisable to wear gloves at all times when handling tubes, while protective goggles give an added sense of security.

Tubes should always be lifted and carried by placing one hand beneath the face, this hand taking the weight and fulfilling the lifting action. The other hand should be placed on the flare to steady the tube. This avoids placing any strain on the junction between the cone and the neck, which is mechanically the weakest part of the tube. With metal-cone tubes, the supporting hand must be beneath the face and not merely hooked under the lip, as this would impose a strain between the glass face and the metal cone.

The tube must never be placed face downwards on hard surfaces, as this is likely to cause scratches. It is best to place the tube on a piece of felt or other thick material, or failing this a few sheets of paper.

The risk of implosion is greatly increased if the glass surface is scratched or if the rim of a metal cone is knocked. It should also be remembered that the internal and external coatings of many tubes form a capacitor, and if the tube is handled when this is in a charged condition, it is possible to receive a shock sufficiently strong to induce the recipient to drop the tube; this risk can be eliminated by always connecting the final anode to the external coating when the tube is removed from its socket. Also, to touch this coating with damp hands will impair its insulating properties.

Viewers should be protected from implosion by a strong glass or plastic screen: suitable screens include ¼-in. armour-plate glass (not ordinary plate glass) or $\frac{3}{16}$-in.-thick flat perspex sheet.

In time, a film of dust will be formed on the face of the tube by electrostatic attraction. This can be removed by wiping with a soft, slightly moistened cloth, after which it must be dried thoroughly. Anti-static preparations may be used, but the makers' instructions should be carefully followed.

When refitting, never use force when inserting the tube into the deflection yoke and focus coil, and take care not to tighten the tube straps and clamps excessively: remember that while this might not by itself cause the tube to break, a tube subjected to excessive clamp pressure is much more liable to break should it be accidentally struck.

Where it is desired to destroy a faulty cathode-ray tube, a recommended method is to place it in an enclosed box and then break it by means of a metal rod inserted through a hole in the lid of the box.

Correct Usage of Cathode-ray Tubes

A British Standards code of practice (C.P. 1005 Parts 1 & 2: 1954) on the use of cathode-ray tubes makes the following recommendations on how to secure optimum performance and life:

Manufacturers' ratings should never be exceeded.

Heater voltage should not vary more than 5 per cent from the rated value; low voltages are as much to be avoided as high. Television tubes which are designed to operate in a series heater chain should have the heater current restricted to 2·5 per cent of the rated value.

When the tube is mounted horizontally, the correct method is to support the tube near the bulb at its maximum diameter and also to clamp the neck lightly near, but not actually on, the base. The fixing should be resilient, and metal-to-glass clamps avoided. The connections to the socket should also be flexible. Provision should be made for the rotation of the tube.

There should be adequate ventilation to ensure a safe temperature under all conditions, and it should be appreciated that the heat generated in adjacent components may largely determine the final temperature of the tube.

Before applying a potential between the heater and cathode of a cathode-ray tube, the manufacturer's tube specification should be studied. If no maximum voltage is given it is better to restrict this value to the minimum possible, preferably less than 5 volts.

There must always be a D.C. connection between each electrode and the cathode. The resistance of this connection should be the minimum practicable.

Care should be taken to avoid scratching or otherwise damaging the surface of the glass.

To ensure maximum cathode life, tubes should be run at the minimum useful brightness.

To prevent damage to the screen material, tubes should not be operated with a stationary or slowly moving spot, except at low beam current density.

Stray magnetic fields may produce serious adverse effects. Interference from such fields may be minimized by the suitable spacing and orientation of neighbouring components.

New Display Tube Devices

Cathode-ray tubes for television generally consist of a glass screen, a conical glass bulb and a glass neck, which when joined together give the tube a considerable length. The length of the tube is the main factor determining the depth of the cabinet in which the receiver is located, and there has been a continued demand for television receivers to be as small as possible, without affecting the size of the picture. Considerable reduction in cabinet size has been made possible by progressively increasing the scanning angle of the cathode-ray tube to 70°, then to 90° and more recently to 110°. In this way a worthwhile reduction in the length of the bulbous part of the tube is achieved, although it leaves the neck in which the electron gun is sealed virtually unaltered.

To obtain a still greater reduction in tube length, flat tubes have been developed, both in this country, by Dr. Gabor, and in America, by

W. Ross Aiken. These tubes consist essentially of a flat glass box having a depth of about 4 in. and a rectangular cross-section, on the end of which the fluorescent screen is deposited. The electron gun is located within the glass box, and hence there is no neck attached to the tube. The display tube resembles in many respects a picture which can be hung on the wall separately from the rest of the television receiver.

The Gabor Tube

The Gabor tube is illustrated in Fig. 12.* Dividing the box into two compartments is a magnetic shield situated at the mid-plane of the box. The electron gun is situated centrally at the top of the tube behind this magnetic shield. The electron beam is directed vertically downwards, passing through a conventional focusing lens, and is then deflected from side to side across the width of the tube by a pair of deflector plates. At the bottom of the magnetic shield the electron beam comes under the influence of a special reversing lens. The effect of this lens on the beam is to bend the beam through 180° so that it then travels upwards and in front of the magnetic screen. The design of this reversing lens is very critical, and the dimensions have to be maintained to very close limits to preserve the quality of the beam. In addition to reversing the beam, the lens also increases the divergence angle of the fan by a factor of about four, which means that on leaving the lens the beam is being swept through an angle of about 120°. The beam is, therefore, being deflected across the whole width of the tube very soon after leaving the reversing lens. To ensure that the beam does not over-scan the width of the picture, the beam is influenced by a collimating lens. This lens is produced by a magnetic element. After passing through the collimating lens, the beam continues on this path vertically upwards parallel to the magnetic shield while being deflected horizontally through a distance equal to the picture width. It only now remains for the beam to be deflected through a further 90° to bombard the fluorescent material deposited on the tube face-plate.

This final deflection is achieved by a scanning array, shown in Fig. 13, consisting of a large number of horizontal linear conductors. These conductors are supported on an insulating base and are not connected. They are charged and discharged by the action of the electron beam at the ends of the scanning lines. A negative potential wave travels down the array of horizontal conductors, and, at the particular point at which the wave has reached, the electron beam is deflected through 90° to bombard the fluorescent screen. An important feature of this deflection is that the beam, in addition to being bent through 90°, is also focused to give a very fine image.

The charging action described in the previous paragraph is achieved by secondary emission, which is effective when the beam has traversed its line sweep. Fig. 13 illustrates the principle of the frame-scanning action, which is brought about by the electron beam.

* Figs. 12 and 13 are reproduced by permission of The Institution of Electrical Engineers from "A New Cathode-Ray Tube for Monochrome and Colour Television" by D. Gabor, P. E. Stuart and P. G. Kalman, *Proc. I.E.E.*, Vol. 105, Feb. 1958.

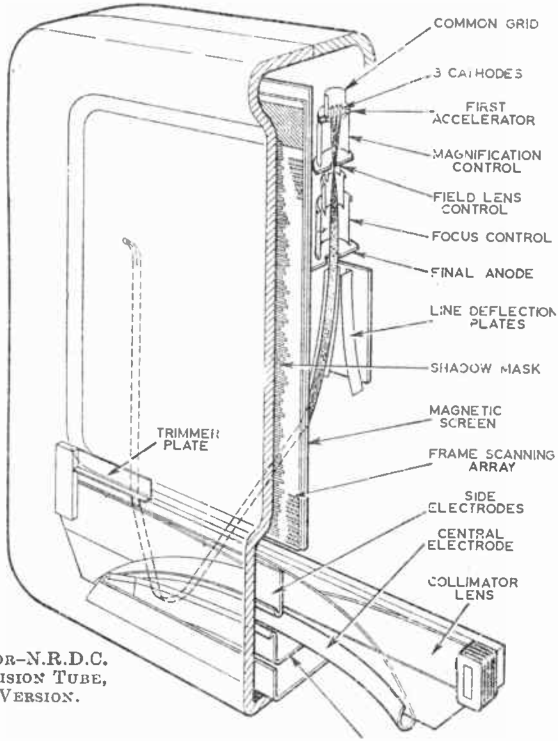


Fig. 12.—Gabor–N.R.D.C. Colour Television Tube, all Glass Version.

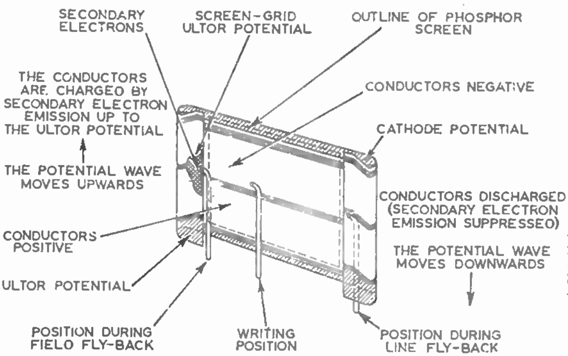


Fig. 13.—The Principle of Self-scanning by Writing Beam.

For economic reasons, it is unlikely that this type of tube will be used for monochrome displays. The main use for such a tube will probably be in colour television. To achieve colour pictures three separate cathodes are situated in the electron gun. The three beams are focused and deflected by a common electrode structure, and are then reversed in direction and collimated as previously described. Because of the slightly different relative positions of the three beams, they bombard the three colour phosphor strips, which are deposited as triplets of horizontal rows, thus giving the colour picture.

The development of the flat tube in this country has only so far reached the laboratory stage, and there are several technological difficulties still to be overcome. It can be seen, for example, that instead of simplifying the internal structure of the cathode-ray tube, the converse has taken place, and the structure is more complicated. Flat tubes have been made in America by W. Ross Aiken, and have been applied to radar applications. An example of their use is in aeroplane cockpits, where it is possible to superimpose a radar presentation on the general view seen by the pilot.

THE ELECTRON BEAM

Cathodes

Modern cathode-ray tubes use indirectly heated oxide-coated cathodes.

Mixtures of barium, strontium and calcium carbonates are extremely good emitters compared with pure metal emitters at the same temperature. The normal operating temperature for the oxide-coated cathode is approximately 1,100° K. If this is greatly exceeded, the life of the cathode will be reduced due to barium evaporation.

The carbonate mixture used for cathodes consists of nearly equal parts of barium and strontium carbonates, with a small percentage of calcium carbonate. The mixture can be applied to the core material, which is of nickel, in a liquid carrier, the most common being nitrocellulose lacquer. The coatings are required to have a uniform thickness of the order of 0·07 mm. If the coating is too thick, the surface will run cool, while if the coating is too thin, the cathode life will be shortened due to high temperature, and emission will not be uniform.

Cathodes are activated by heating the cathode in vacuum to a temperature considerably higher than that of normal operation.

Oxide-coated cathodes can be poisoned by contaminating gases or vapours, which can further completely destroy the cathode surface by ionic bombardment. The latter defect is partly overcome by means of the L-cathode, in which the oxide cathode is surrounded by a pellet of tungsten through which the free barium migrates to the surface. Bombardment of the surface by positive ions may temporarily destroy the barium layer, but this can readily be reformed, provided there is sufficient barium oxide present in the cathode mixture. A further advantage claimed for the L-cathode is that it can operate at a much higher loading, thus enabling smaller fluorescent spots to be obtained on the screen.

Deflection Coils

The three main requirements of scanning coils are that they should give: (1) linearity of scan; (2) rectangular raster; (3) freedom from

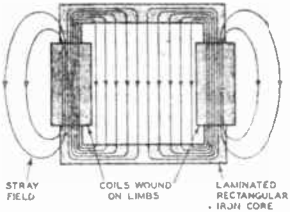


FIG. 14.—RECTANGULAR IRON-CORE COIL PRODUCING UNIFORM DEFLECTING FIELD.

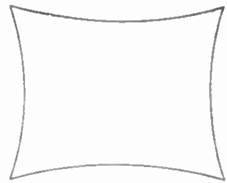


FIG. 15.—PIN-CUSHION DISTORTION INTRODUCED BY FLAT-FACED SCREEN, USING UNIFORM DEFLECTING FIELD.

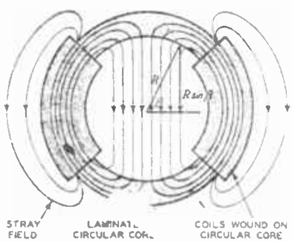


FIG. 16.—TOROIDAL-WOUND CIRCULAR IRON-CORE DEFLECTING COIL.

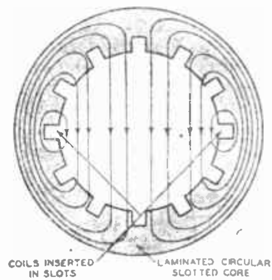


FIG. 17.—SLOTTED IRON-CORE DEFLECTING COIL.

deflection defocusing. In addition to these requirements, the coils should have a high sensitivity, whilst certain inductance and resistance values will tend to simplify the design of scanning circuits.

In practice, the tendency is for deflector coils to be developed to suit particular tubes, rather than for the adoption of a general design theory.

It is necessary to compromise between scan distortion (i.e., pin-cushion or barrel) and deflection defocusing, for a given beam diameter. Both qualities depend upon the uniformity of the deflecting field. In pin-cushion distortion the edges of the raster are curved inwards instead of being straight, while in barrel distortion the raster edges are curved outwards. To achieve a uniform field, it is usual to have coils wound on the opposite limbs of a laminated core. The coil can be made to fit tightly around the tube neck by having a toroidal wound circular iron-core deflecting coil, with the density of the windings proportional to cos β (Fig. 16). Very uniform fields can be obtained, but, in practice, it is usual to make an approximation to the cosine distribution of winding. By using slotted cores it is possible to use the slots to obtain uniform positioning of the windings (Fig. 17). A true cosine distribution, however, could cause scan distortion giving a pin-cushion shape, and this can be overcome by increasing the number of turns at low values of β or decreasing the turns for high values of β. This is usually determined by experiment.

For coils used with low-scan-angle television tubes economic considerations usually enforce the use of air-core coils provided with a low-reluctance-return path around the outside, reducing the stray external magnetic field, and improving the field inside the tube. Grading is usually done by using differing wire gauges; a coil may consist of three different gauges.

Care is needed with the end turns of the coils, for these contribute nothing to the required beam deflection, and can cause harmful results, such as defocusing and non-linearity. Improved performance is usually obtained when the end turns are taken as far away from the tube neck as possible (see Fig. 19).

The modern trend in television practice is towards shorter tubes and larger screens, thus causing an increase in the maximum angle of deflection up to 110°. This increase in scan angle, combined with higher E.H.T., calls for considerably increased power in the deflection circuits. Some compensation in the scanning-power requirements may, however, be obtained by reducing the neck diameter, thereby enabling the coils to be situated much nearer the beam.

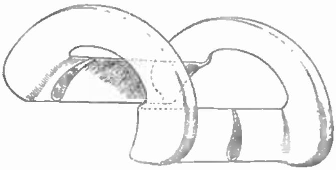


FIG. 18.—ONE SADDLE-SHAPED COIL FROM A DEFLECTING-YOKE.

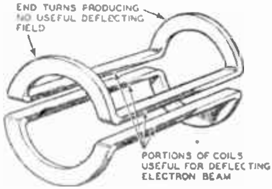


FIG. 19.—AIR-CORE COILS FOR TELEVISION. A FURTHER SET OF COILS IS SUPERIMPOSED AT RIGHT-ANGLES, AND THE ASSEMBLY SURROUNDED BY IRON WIRE.

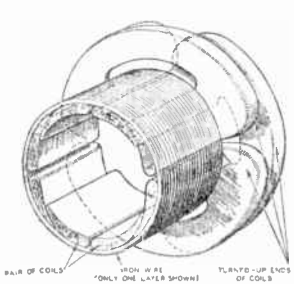


FIG. 20.—TYPICAL YOKE ASSEMBLY CUT IN HALF TO SHOW SECTIONS OF WINDINGS.

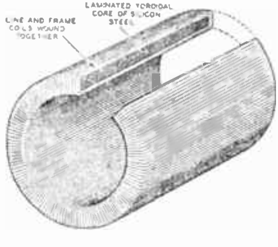


FIG. 21.—TOROIDAL DEFLECTING-YOKE CUT AWAY TO SHOW LAMINATED STEEL CORE.

PHOSPHORS

A fluorescent material (or "phosphor") consists of a crystalline, usually inorganic, powder which is applied to the inside of the glass, and has the property of absorbing invisible radiation and re-emitting it as visible light. The chief materials used are sulphides, silicates, tungstates and phosphates; but their fluorescence nearly always depends upon the presence of a very small quantity of a second substance, known as an activator, which intrudes into the crystal lattice, and is usually metallic, such as silver, copper or manganese. An important requirement is a high degree of purity for both substances.

Phosphors differ among themselves in three respects, namely, (*a*) colour of luminescence, (*b*) afterglow or persistence and (*c*) luminous efficiency; the effects of which are as follows:

(*a*) Colour. The colour of the fluorescence obeys Stokes' law, i.e., the wavelength of the emitted light must be longer than that of the exciting radiation. For example, a screen is used which receives blue light from a second screen in contact with it, and re-emits it as a yellow light. For cathode-ray-tube purposes, however, the exciting radiation is in general the electron stream, which has a wavelength of the order of 10^{-5} that of visible light.

As regards the various available colours, green is somewhat at an advantage for visual tests and observations, since it is that to which the human eye is most sensitive. Blue is preferred for photography. For television, white is naturally required; but a slight tinge of blue is frequently added to counteract the effect of room lighting.

(*b*) Afterglow. With some materials, the luminescence ceases almost immediately the exciting cause has been removed; e.g., in about 20 microseconds. In other cases, the trace takes an appreciable time to die out, such as 0·1–180 seconds, and this persisting effect is known as phosphorescence. It may be of a different colour from the fluorescence proper, and in some cases at least it can be eliminated by the addition of an inhibitor or "killer", such as nickel. Afterglow is a considerable advantage for some applications and a serious drawback for others. For general test observations, a slight afterglow obviates flicker, and the trace is hence less tiring for the eyesight. For most radar purposes, such as plan position indicators, a long afterglow, of up to 40 seconds or more, is desired, in order that the display may remain visible between successive "paintings". But for photographic recording on a moving film, even the shortest afterglow causes blurring. For television, any persistence from one frame to the next (i.e., for $\frac{1}{25}$ second in Great Britain) would give double images, and must be rigorously avoided.

In some radar applications a very long afterglow—up to 15–20 minutes—may be required. Conventional long-after-glow phosphors, such as fluoride screens, are not really suitable for these applications, as the afterglow trace can be seen only as a very faint signal when viewed in complete darkness following a long period of dark adaptation. Such viewing conditions are very fatiguing, and therefore not highly favoured.

There are solutions now available to this problem of long afterglow. The first is to use storage tubes (see "electronic storage tubes" later in this section); the second is by the application of heat to the fluorescent screen.

When a phosphor is bombarded by an electron beam, part of the beam energy is converted into visible light, which is emitted instantaneously, while a further part of the energy is used to provide the afterglow. In the phosphor material there are empty electron traps which become filled when the phosphor is excited by the scanning beam. During the natural return of the excited electrons from the traps to their steady state, light is emitted and is visible in the form of afterglow. This decay of light output usually follows an exponential law. It is known that some electrons are held in the traps for a very long time, and therefore if these electrons could be released in a controlled manner the afterglow of the phosphor could be modified. A possible way by which this can be achieved is by the application of heat. Thus, in the operation of the tube, the signal is applied in the normal manner, and the natural decay law is observed. After the required interval of time, which for some phosphor materials can be several days, the stored information is released as a visual display by heating the screen.

(c) LUMINOUS EFFICIENCY. The effectiveness of a phosphor depends to some extent on the wavelength of the exciting radiation. In the case of low-pressure fluorescent lamps, most of the radiation has a wavelength of about 0·25 micron in the ultra-violet, for which silicates, tungstates and halo-phosphates are the most suitable; but for cathode rays, sulphides give the maximum efficiency, which may amount to as much as 7–8 candles per watt at 15,000 volts from cathode to screen, and they are tending to replace all other types.

Fluorescent Materials

The principal fluorescent materials used for cathode-ray tubes are as follows:

(1) Zinc orthosilicate, with 1 per cent by weight manganese activator, giving a green colour with slight afterglow. This substance is met with as the ore willemite, some varieties of which fluoresce in the natural state. It is, however, best prepared artificially. It has for long been the standard "general-purpose" screen material, its robustness especially as regards an over-bright spot, being a special advantage.

(2) Calcium tungstate, blue-violet, very suitable for photography, practically no afterglow. This is also found in nature as the ore scheelite, but as a screen material is made synthetically. It is now generally replaced by (3)(*b*) below, which has about six times the actinic effect on all photographic films.

(3) Zinc sulphide and cadmium sulphide are considered together, as they crystallize in the same system and form solid solutions in any proportions. The former fluoresces blue or green (depending on the activator), but by adding varying proportions of the latter, the colour may be varied even as far as red. They are the most efficient phosphors of all. Examples are as follows:

(*a*) Zinc sulphide activated with 0·01 per cent of copper, green with very long afterglow. A variant is also made having a blue colour with green afterglow.

(*b*) Zinc sulphide with 0·01 per cent of silver, blue, very sensitive, no afterglow.

(*c*) Zinc sulphide with 2 per cent of manganese, orange.

(*d*) Zinc–cadmium sulphide (90 : 10), activated with copper, greenish-yellow, very long afterglow.

(*e*) Double-layer screen giving blue flash and yellow afterglow of up to 40 seconds, consisting of (*d*) next the glass and (*b*) superposed upon it.

(*f*) "White" screen for television tubes, very short afterglow, consisting of a mixture of (*b*) powder mixed with a 50 : 50 zinc–cadmium sulphide activated by silver and giving a yellow colour. Direct-vision screens have a momentary brilliance of about 1,000 c.p./sq. cm., and screens for projection even more.

Application of Screen Material

The fluorescent powder, which has a fineness ranging between about 1 and 20 microns, is applied to the glass in several ways. A typical method is to coat the glass with a binder consisting of water-glass (sodium or potassium silicate) or of phosphoric acid dissolved in acetone, and to dry-spray the powder on to it.

Another method which is used extensively on television tubes is the settling technique. This technique is considerably more economical on powder than the dry-spray method. In the settling operation a suspension of the fine powder in water or other liquid is placed in the bulb, and the powder is allowed to settle on to the glass. The water is decanted and the screen is then dried. Whatever method of screening is used, the aim is to form a very uniform, continuous layer of phosphor particles on the tube face.

A recent development in the deposition of fluorescent screens is the evaporation technique. Conventional crystalline screens suffer from two defects. First, due to the thickness of the screen, internal light reflections in the crystal layer, as well as possible electron scattering, cause an increase in the spot size; and secondly, as the body colour of the material is white, the contrast is reduced when the screen is viewed under ambient illumination. Both of these defects are overcome by using very thin transparent evaporated screens.

The fluorescent material is evaporated from a boat on to the glass face plate. After evaporation, the screen is given a high-temperature bake to ensure that the phosphor and the activator are intimately mixed. Depending upon the phosphor material used, the post-evaporation bake is performed either in air or vacuum.

Screens prepared by this technique have shown a worthwhile improvement in resolution, although at the present moment the luminous efficiency is not as high as is obtainable from crystalline screens. On the other hand, however, evaporated screens are transparent, and therefore any external light which falls on the tube face is not reflected, but passes through both the glass face-plate and the evaporated phosphor layer, and is absorbed by the black coating of graphite on the internal surface of the bulb. The contrast is thereby increased, and such screens can be easily viewed even under exceptionally high levels of ambient illumination.

Although these screens are, at the present, in the development stage, it is highly probable that within a few years evaporated screens will be used on radar tubes and also possibly on monochrome television tubes. A further possible application is on colour television tubes. Separate layers of material capable of emitting light of different colours can be evaporated on to the tube face. By modulating the energy of the electron beam, the depth of penetration of the beam into the evaporated screen layer can be controlled, thus altering the colour of the emitted light.

Cathode-ray-tube Screens

There is no general classification of luminescent screens in Great Britain. Table 1 uses American R.M.A. notation. The description of purpose is sufficient for the identification of most normal phosphors from catalogue data. Persistence, or afterglow, is the period of continued visibility after cessation of electron bombardment.

TABLE 1.—NOTATION OF AMERICAN PHOSPHORS

Phosphor Number	*Colour*	*Persistence*	*Purpose*
P1	Green	10–100 millisec.	General repetitive work in which observation is visual and frequency greater than about 25 c/s.
P4	White	Short	Most common use in television tubes; rarely used for oscillography.
P7 (double layer)	Blue-green	Up to several minutes	Used for a low-speed work, observation of transients and radar. Persistence depends upon voltage, current and duration of excitation.
P11	Blue	Very short	Photography, including moving-film recording.

ELIMINATION OF ION BURN

If a tube has not been thoroughly evacuated, any residual gas will become ionized when the tube is operated. Negative ions, together with those emitted from the cathode, will travel with the electrons and bombard the screen, while positive ions will proceed back down the tube and bombard the cathode. Negative-ion bombardment causes the screen to show a dark discoloration when scanned. With electrostatically deflected tubes the deflection of ions is the same as for electrons, and the whole raster becomes uniformly dark. With magnetically deflected tubes, however, the negative ions are not deflected by the magnetic fields, and a dark central circular patch, which is a shadow of the tube neck, appears on the screen.

There are at present three methods of overcoming this trouble.

First, to ensure that there is no residual gas present in the tube; second, if there are ions present, to prevent them reaching the screen; and third, if the ions do reach the screen, to ensure that no damage is done.

Considering these three methods in detail:

(1) To ensure that there is no gas present in a cathode-ray tube is an almost impossible task. Furthermore, to attempt to achieve this state would be a lengthy and costly process.

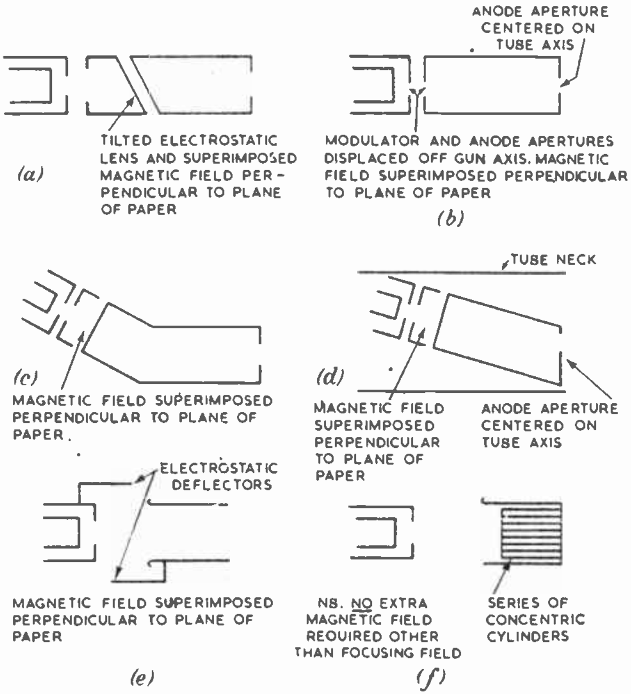


FIG. 22.—ION-TRAP DESIGNS.

(2) Several ingenious designs of gun have been introduced to prevent ions from reaching the screen. Either these employ superimposed magnetic and electrostatic fields in the triode region of the gun; or the axis of the beam, which is initially off the tube axis, is deflected on to the tube axis by an external magnetic field, which acts only on the electrons. The negative ions proceed in their off-axis path and are trapped on the anode.

It can be seen that the successful operation of such ion traps depends upon the different paths followed by the electrons and negative ions under the action of magnetic fields. This path difference is already existent in all magnetically focused tubes, and has led to the design of the following ion trap which needs no extra electrostatic or magnetic field. The negative ions proceed from the cathode-modulator region as a divergent cone, while the electrons upon entering the focusing field have their paths altered. Over a portion of their path, through the focusing field, the electrons travel parallel to the tube axis. Therefore by arranging a series of coaxial cylinders at the appropriate position, the electrons will continue towards the screen, while the negative ions will be trapped on the cylinders.

Fig. 22 indicates various forms of ion traps. In traps that employ combined electrostatic and magnetic fields, the positive ions which travel back towards the cathode will be deflected away from the cathode and will harmlessly bombard the modulator. Thus, in addition to preventing negative ion burn of the screen, these traps save the cathode from being destroyed by the back-streaming positive ions.

(3) The third technique of preventing ion burn is by protecting the screen with a layer of aluminium. The layer permits the passage of electrons, absorbing only a small fraction of their energy, but it acts as a barrier to the negative ions. The aluminium film is obtained by evaporation, and to make the film continuous, a nitro-cellulose layer is initially applied to the phosphor. This can be done by allowing a drop of nitro-cellulose to spread over a water surface which covers the screen. By tipping the water off, the nitro-cellulose blanket rests on the phosphor crystals and serves as a foundation on which to evaporate the aluminium. After the evaporation process, the nitro-cellulose blanket can be decomposed by a high-temperature bake, leaving a continuous aluminium film. The aluminizing of screens introduces advantages additional to the prevention of ion burn. The aluminium layer reflects in a forward direction, outward from the screen, light which would otherwise have been lost in the interior of the tube. Thus, other conditions being equal, aluminized screens will have a higher brightness. The contrast is also improved since the aluminium layer prevents light from the bright parts of the picture illuminating the dark parts. Finally, the difficulty of the screen potential "sticking" is overcome, enabling the tubes to operate at higher potentials. This sticking is explained in the next paragraph.

Normally the potential of a non-aluminized screen is maintained at anode potential by virtue of secondary emission. If, however, the anode voltage is increased to exceed the value for which the secondary emission coefficient becomes less than unity, the screen potential will be unable to follow the anode potential and will "stick" at the potential value corresponding to a secondary emission coefficient of unity. With aluminized screens, however, no such limitation exists, as the aluminium layer is connected to the final anode, and thus maintains the screen at anode potential.

RADAR TUBES

Cathode-ray tubes used for radar do not differ basically from those used for other purposes. Cathode-ray tubes using electrostatic deflection and focusing are used for A- and B-scan displays, because the necessary high writing speeds and accurately linear traces are more easily obtained from voltage- than from current-generating circuits, such as are required for tubes using magnetic deflection.

In displays where range measurement is liable to be made at any point along the trace, it is important that astigmatism and deflection de-focusing should be small.

Brightness requirements do not usually present much difficulty, because room lighting can be kept reasonably low: however, some difficulty may arise where the indicator is viewed in daylight and also with some very short-range displays. The problem in the latter case arises from the fact that a short range means a high writing speed and a correspondingly short period of illumination of the trace. This cannot be compensated by increasing the pulse-repetition frequency beyond a certain point, owing to limitations of the magnetron, now almost universally used for radar transmitters. For these reasons, it may be necessary to compromise between brightness, focus quality and tube life.

It is best to increase brightness by increasing the final anode voltage, but difficulties arise due to the necessity for increased deflecting voltages to obtain the same deflection; and also in the tube construction, it is often necessary to provide separate side-arm connections for the final anode and the four deflector plates, since the increased voltages are too great to permit these connections being brought out through the base.

For P.P.I. displays, the magnetically focused and deflected tube is mainly used. The duration of the sweep is relatively long, and the inductance of the deflecting coils is not objectionable, in fact it may be used to assist in forming the saw-tooth waveform.

Advantages of magnetic tubes are that higher beam currents are obtainable for a given focus quality, and that a higher operating anode voltage necessitates only one side connection to isolate the anode potential from the other connections in the base.

Higher beam power is almost essential in P.P.I. displays to obtain adequate brightness of the long afterglow screen that must be employed, owing to the slow repetition rate of the scan of the aerial, which is of the order of six to thirty per minute.

Against the advantages of the magnetic tube, it should be noted that tubes using magnetic focusing require individual adjustment of the alignment of the magnetic lens with respect to the electron beam to obtain optimum focus. A badly aligned magnetic lens can produce worse spot distortion than the best electrostatic lens (of necessarily smaller aperture) for equal beam currents. For this reason P.P.I. tubes using electrostatic focusing but with magnetic deflection are sometimes used in cases where only semi-skilled maintenance is possible.

Screens

The choice of screen largely depends on the periodicity with which the information displayed is written on the screen.

In the majority of A-scan displays, for example, the trace repeats over the same path on the screen at least twenty-five times per second, and persistence of vision makes the trace appear continuous. For this type of display, therefore, a screen having high visual efficiency and negligible afterglow is used. As in ordinary oscilloscope practice, willemite (zinc orthosilicate) or green response sulphide screens are suitable.

In a P.P.I. display, since the trace may scan the same point on the screen only once every 2–10 seconds, it is necessary to use a screen having a long afterglow in order that the picture may persist for several revolutions. The three main types of long-afterglow screens are: (1) double-layer sulphide screen; (2) single-layer fluoride screen; (3) potassium chloride screen for use in the "Skiatron". Of these three, the single-layer fluoride screen is now used most extensively in tubes operating at 10–15 kV.

The "Skiatron" or dark-trace screen is formed by evaporating potassium chloride *in vacuo* on to the glass bulb of the tube. The resulting screen is white, but turns magenta where bombarded by the electron stream. This coloration persists for a time, depending on the intensity of the electron excitation.

Compared with a bright trace tube, the skiatron is of low efficiency, and an electron beam power of about 10 kV at 300 μA is required for good contrast.

Large-scale Displays

For case of plotting, or where several observers are concerned, a large screen is desirable.

Although optical projection is common in television, this method cannot be used with a double-layer afterglow tube. One method is to separate the two layers of a double-layer screen, leaving only the first layer within the tube. The blue light from this layer can then be projected by optical means on to a large screen coated with the afterglow powder normally used for the second screen.

Another method of obtaining a large-scale display is by the use of the skiatron, which has been developed especially for this purpose. The tube has a standard magnetic deflection and focusing system and a screen which darkens under electron bombardment. As the trace is dark, it is possible to illuminate the tube screen by external light and project an image of it on to a large viewing screen by using ordinary optical methods. The viewing screen can be flashed opal glass, on which chinagraph pencil marks can be made for plotting.

Since powerful illumination is required for satisfactory projection, it is usual to illuminate the screen by the cold light of mercury vapour lamps, and in practice it is usually desirable to keep down the screen temperature by means of an air blast.

Although the skiatron provides an easy method of obtaining a large-scale display, it suffers from the disadvantage that permanent echoes give stains which take a long time to decay. If the equipment is used on a moving ship this results in the picture being smeared.

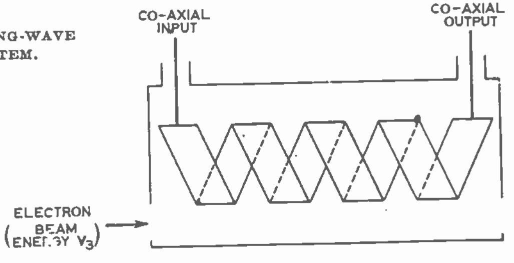
Post-deflection Acceleration

Not all radar applications necessarily display the information in P.P.I. form. There are several instances in which the incoming signal is used to deflect the electron beam, such as, for example, in A-scan displays, rather than to intensity-modulate the beam as in P.P.I. displays. In the A-scan displays tubes are used employing electrostatic focus and deflection of the electron beam, and the highest possible deflection sensitivity is required from the tube. There are several ways of achieving this increase of sensitivity. One method is to deflect the beam at low energy, and then to accelerate the beam to high energy during its travel from the deflecting region to the screen. This technique, which is known as post-deflection acceleration (P.D.A.), has been done in the past by applying successively increasing potentials to adjacent conducting bands on the internal surface of the wall of the bulb. Usually there are only two such bands, but in very high-voltage tubes there may be as many as four or five. The reason for having a greater number of bands is that the ratio of the potentials of adjacent bands has to be kept as low as possible, otherwise the trace becomes distorted, due to the lens action, which is always present when two adjacent cylindrical conductors are maintained at different potentials. Although this form of P.D.A. has been used quite extensively, the limitations of scan distortion have been known for a long time, and hence the ratio of the screen voltage to gun voltage has been limited to about two. Recently a proposal first made in the early days of cathode-ray-tube development has been revived. This is to paint a spiral of graphite on the internal surface of the bulb extending from the region of the deflector plates up to the screen. The resistance of this graphite spiral is very high, and across the spiral the post-deflection accelerating potential is applied. The action of the spiral is to apply the accelerating field in a more uniform manner, thereby minimizing any scan-distortion effects. By this technique it has been possible to increase the ratio of screen voltage to gun voltage up to six or more.

Travelling-wave Deflection Tubes

There are also applications where it is necessary to display waveforms containing frequency components up to 1,000 Mc/s. Conventional A-scan display tubes cannot be used in these applications for two reasons. Firstly, the time taken for an electron to traverse the deflector plates is long compared with the time of one cycle of the highest frequency to be displayed, and hence the resultant deflection of the beam will not be a faithful reproduction of the applied signal. Secondly, the deflection sensitivity of the tubes is too low, and hence several tens or hundreds of volts are necessary to produce an appreciable

Fig. 23.—Travelling-wave Deflecting System.



deflection on the screen. This limitation imposes difficulties on the design of suitable amplifiers to provide sufficient volts for deflecting the beam.

To overcome these difficulties a cathode-ray tube employing a travelling-wave deflection system has been developed. The deflecting system, which is shown in Fig. 23, consists of an outer metallic cylinder and an inner helix of metal tape wound on an insulating former. The signal is applied to the helix. The electron beam travels in the space between the inner helix and the outer conductor. The velocity of the beam is adjusted to equal the axial velocity of the signal along the helix. By this means the electron beam experiences the same deflecting force each time the beam passes the separate turns of the helix. Transit-time distortion can be eliminated by reducing the width of the tape, while sensitivity can be increased by increasing the number of turns of the helix. The deflecting system shown in the diagram is for non-push–pull deflection voltages. By having two helices side by side inside the outer cylinder and passing the electron beam between the two helices it is possible to apply push–pull voltages to deflect the beam. Deflecting systems of the type shown in Fig. 23 have been developed having a frequency response extending up to 1,300 Mc/s, and a deflection sensitivity such that only 0·5 volt is required to deflect the spot through a distance equal to its diameter.

Electronic Storage Tubes

Electronic storage tubes are becoming more popular and are being used to a greater extent. The main uses to which these tubes are put include computors, where it is necessary to store information required in the arithmetical processes, oscillography, where information can be recorded and studied at leisure instead of making photographic records, and radar applications. In this last application it is often necessary either to obtain a degree of control on the afterglow of the display or to convert the radial P.P.I. display into a television-type raster display suitable for relaying to various display panels.

All storage tubes depend for their action upon the charging and discharging of an insulated surface. The charging action can be effected either by secondary emission or else by bombardment-induced conductivity. The secondary emission ratio of an insulator decreases with increasing electron velocity after the ratio has passed through a maximum, usually at about 1 kV. The voltage of the writing beam is adjusted so that the secondary emission ratio is less than unity, and there is, therefore, a net gain of electrons on the insulated surface. This is the action of writing on the information. For reading, the potential of the beam is adjusted to operate at a value where the ratio is greater than unity and there is, therefore, a loss of electrons as the beam scans the surface. While this explanation is somewhat over-simplified, it serves to explain the basic principle. In brief, the writing beam effectively deposits on the storage surface a charge pattern representing the radar-plan view, and the reading beam reads this picture and reproduces it by television scanning methods.

In applying these principles to a radar-scan conversion storage tube, two electron guns are normally used.

The first beam, known as the writing beam, creates a charge on the surface of the insulator, the charge varying from point to point as the

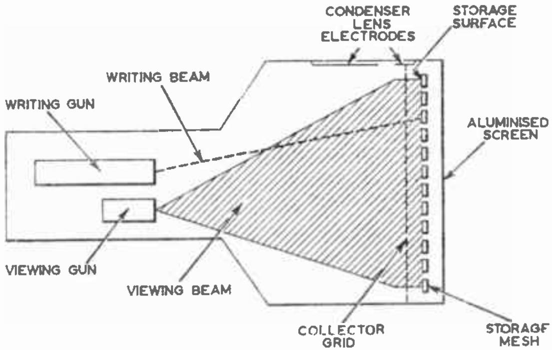


FIG. 24.—DIRECT-VIEWING STORAGE TUBE.

intensity of the scanning electron beam is modulated. The second beam, known as the reading beam, has a very low current and can be made to discharge a charged point by a fractional amount per scan. The storage target consists of a metal plate, for example aluminium, which has an insulated surface a few microns thick. This insulated surface can be aluminium oxide, which may be deposited by anodizing the aluminium plate. In such tubes, the two guns are on the same side of the target surface. Tubes are made, however, with the guns on the opposite sides of the target, as indicated in Fig. 24. The target in these tubes consists of a very thin aluminium film as the signal electrode through which the writing beam penetrates, and on the side of the target facing the reading gun is deposited a thin layer of a suitable insulator, for example, zinc sulphide. With these tubes it is possible to retain the information for several minutes, although by increasing the current in the reading beam the decay time can be shortened to a second or so. This technique, therefore, affords the possibility of controlling the after-glow of the radar display.

The reading-beam signals are taken from the metal backplate of the target surface and, after amplification, applied to a television tube for direct viewing or projection. The television-tube scan must be synchronized with the reading-beam scan. The number of scans for full discharge of any area on the screen depends on the reading-beam current, and this can be adjusted, within fairly wide limits, to give any desired decay rate.

Since the scanning rate of the reading beam, and therefore of the display tube, is high, the screen of this tube need not have any afterglow properties. High beam power can be used, and therefore high brightness obtained with this method.

Although the storage tube has considerable advantages over earlier methods, it should be remembered that the system is complicated by the additional television circuitry required. Another drawback is that the definition of the storage tube is barely good enough to show signals of low-signal-to-noise ratio.

There are radar applications where it is necessary to view the information under high levels of ambient illumination, and for these applications directly viewed storage tubes have been developed, as indicated in Fig. 25. Such tubes are also now being employed in oscillographic applications, where information can be retained on the tube face and inspected over an indefinite period. Again, two electron guns are normally used, although sometimes a third gun is incorporated to act as a replenishing gun to retain the charge pattern on the target.

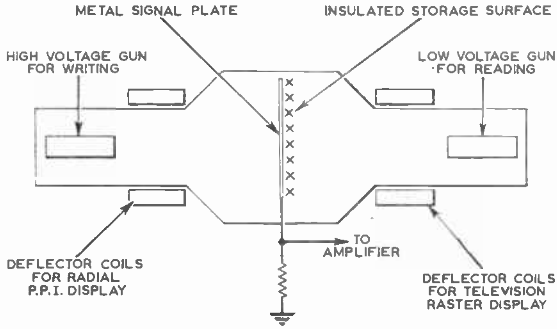


FIG. 25.—TWO GUN STORAGE TUBE FOR RADAR SCAN CONVERSION.

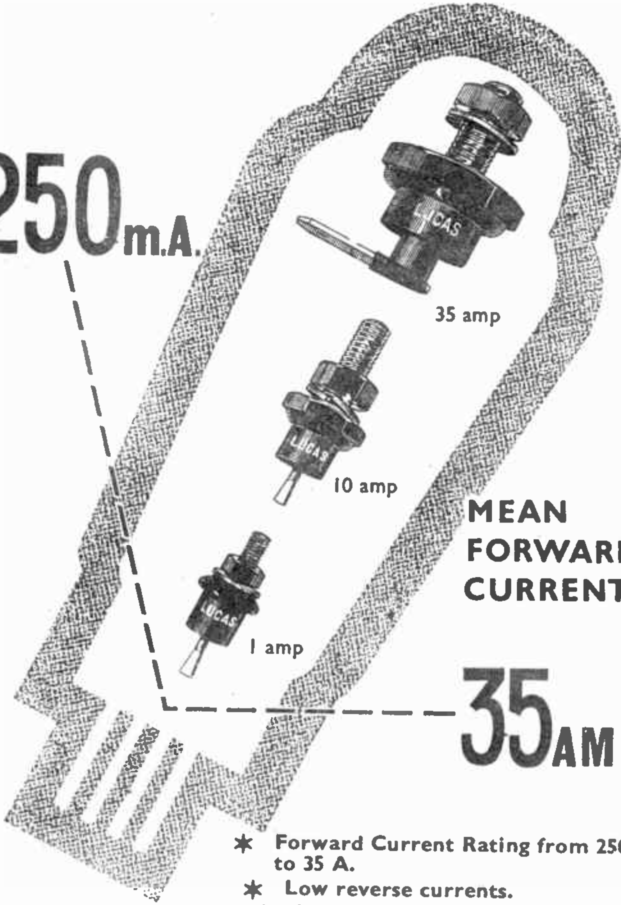
The storage surface in these tubes consists of a metallic mesh, on one side of which is deposited an insulated layer. This mesh is spaced a short distance away from the fluorescent screen, which is maintained at a very high positive potential.

The reading gun produces a flood beam of electrons which bombard the storage mesh normally. The number of electrons from the flood gun which penetrate the storage mesh is governed by the potential to which the discreet parts of the insulated layer have been charged by the writing beam.

The effect, when the reading beam floods the storage mesh continuously rather than sequentially, as in a television system, means that a considerable gain in brightness is obtained. This gain in brightness allows the trace to be viewed in high levels of surrounding illumination, such as are encountered, for example, in aircraft flying at very high altitudes. In the two gun versions of the directly viewed storage tube, one of the limitations is the natural decay of the stored information from the surface. The main reason for this decay is the action of the positive ions, which are produced in the tube and are attracted towards the negatively charged storage mesh. It is essential therefore that in the production of these tubes the best possible vacuum is obtained, and this usually means that the tube has to be given a prolonged pumping schedule.

L. S. A.

250 m.A.



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THERMIONIC VALVE RECTIFIERS

An ordinary metallic conductor presents the same resistance to current flowing in either direction, but there are many types of circuit element in which the resistance to current flow in one direction is much greater than the resistance to current flow in the reverse direction. The introduction of a non-linear device of this nature into an A.C. circuit will result in the average current on one half-cycle being greater than on the other half-cycle. Rectification takes place and a D.C. component of current appears in the circuit.

Principle of the Thermionic Diode

There are many types of non-linear element suitable for use as rectifiers, but the thermionic diode is convenient and efficient: it is small and cheap, and the resistance in the forward direction is low, while the resistance in the reverse direction is extremely high. In its simplest form the thermionic diode is a two-electrode device, a heated filament —the cathode—forming the electron emitter, and an anode electrode surrounding the cathode to receive the electrons. During the half-cycle of applied voltage when the anode is positive with respect to the cathode, current will flow, convention requiring that it be considered to flow from anode to cathode, though the electron movement is from cathode to anode, but during the reverse half-cycle the current flow is negligible.

The unidirectional flow is the result of the basic mechanism of current conduction in the diode, the current to the anode consisting of a stream of electrons, negatively charged particles emitted by the heated cathode and attracted to the anode only when that electrode is positive with respect to the emitter. Childs and Langmuir have shown that the anode current in a thermionic diode is given by an equation of the form

$$I = kV^n \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

where I is the anode current; k a constant; V applied voltage; n a constant, 1·5 approx. This equation is strictly accurate only when the two electrodes consist of infinitely large parallel plates, but it is very nearly accurate for practical valves, Fig. 1 showing the measured and computed values of current for a particular sample of the 5U4G rectifier valve.

The anode current computed from equation (1) holds good only at a specified value of heater voltage, the anode current for a fixed value of anode/cathode voltage being a function of cathode temperature. Present designs of rectifier valve have permissible maximum peak currents at least five to ten times the rated D.C. load current, and in consequence only a small fraction of the cathode emission arrives at the anode.

At low values of anode current the current is not greatly affected by heater temperature, for there is a more than adequate supply of electrons, but at high values of anode current the emitter temperature is the controlling factor. Fig. 2 indicates the result of measurements on a UU5 at three values of heater voltage, and anode currents up to 100 mA. It will be apparent that the condition of a valve will not be indicated by a check on the anode current made at a low anode voltage.

Indirectly-heated Cathodes

The majority of the commercial rectifier valves now available do not employ a filamentary cathode structure but have indirectly heated cathodes consisting of a circular or rectangular tube having a coating of the oxides of thorium and barium as the emitting surface. The cathode tube or sleeve is maintained at the working temperature (800–1,000° C.) by a heater or filament supported inside the tube but insulated from it by a spacer of Alumina. Typical constructional details are shown in Figs. 3 and 4.

Indirectly heated cathodes of the type described have the advantage that the whole of the cathode is at the same potential with respect to the anode, and therefore shares the total emission uniformly over the surface. A filamentary cathode will have one end of its filament more

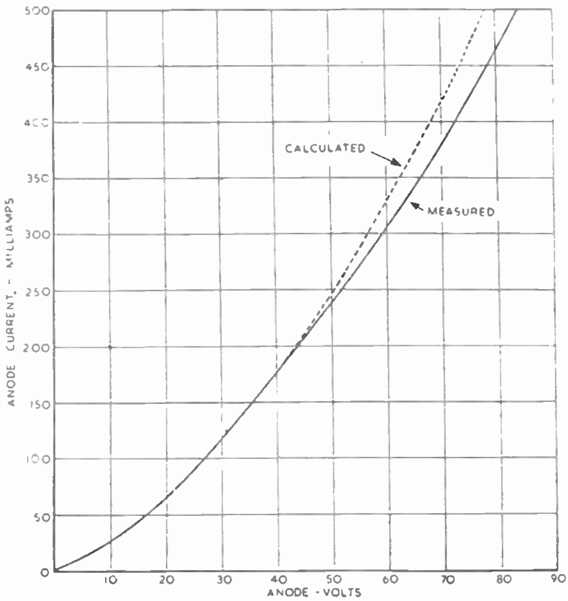


Fig. 1.—Comparison of Calculated and Measured Anode Currents for Single Anode of 5U4G.

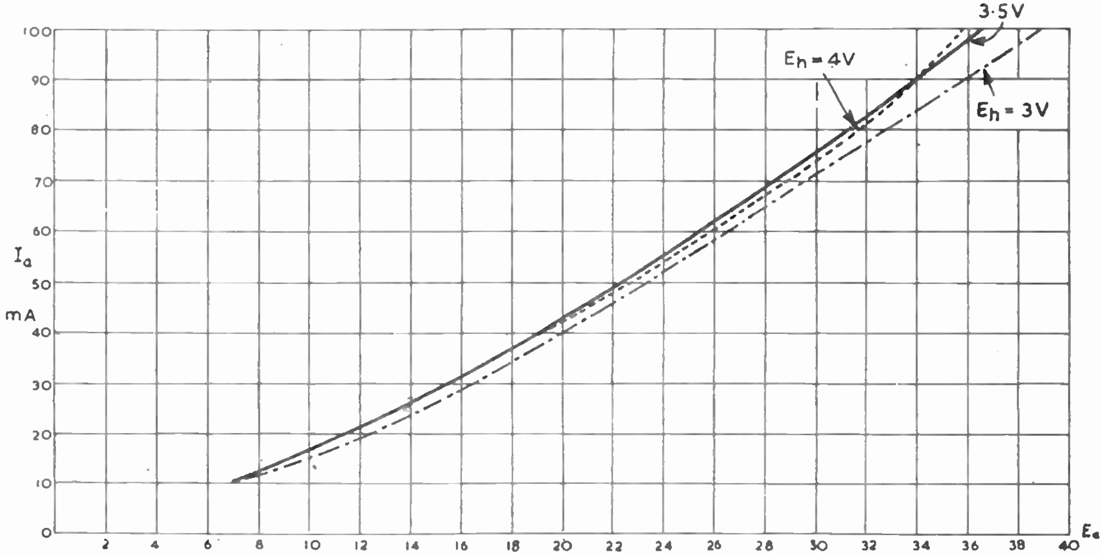


FIG. 2.—TYPICAL ANODE CURRENT/ANODE VOLTAGE CURVES AT LOW VALUES OF HEATER VOLTAGE = E_h.

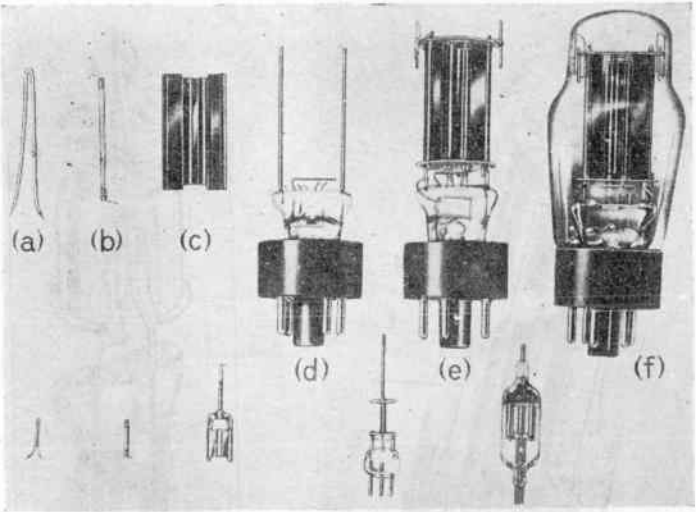


FIG. 3.—CONSTRUCTIONAL DETAILS.

(*a*) Heater. (*b*) Cathode sleeve. (*c*) Anode. (*d*) Glass press and anode support. (*e*) Assembly ready for insertion in the bulb. (*f*) Completed Mazda U903.

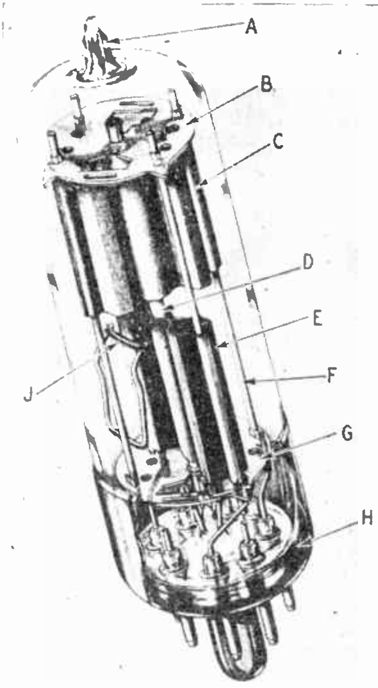
positive with respect to the anode by the amount of the voltage drop in the filament. This leads to a non-uniform distribution of the emission along the filament, and hence a shorter life.

Anode Construction

The anode electrode is basically a tubular or rectangular tube, generally of nickel or nickel-plated iron, supported concentrically round the cathode; but in practice it is generally provided with ribs or fins to stiffen the structure against distortion in order that the anode/cathode clearance be maintained. The ribs and flanges serve to increase the heat dissipation, but it is also common practice to coat the anode with carbon for the same purpose. It will be obvious that the anode temperature must be kept well below the temperature at which it emits electrons, or current conduction would take place in both directions through the valve.

E.H.T. and High-current Types

For valves intended for high-voltage rectification, such as are necessary in cathode-ray tube E.H.T. supplies, it is necessary to take special precautions in the mechanical and electrical design of the valve to avoid corona and flashover. Anode and cathode leads are taken out at opposite ends of the bulb, the anode being supported clear of the pinch to increase creepage clearances, and having rolled edges to reduce corona. A typical valve of this type is shown in Fig. 5.



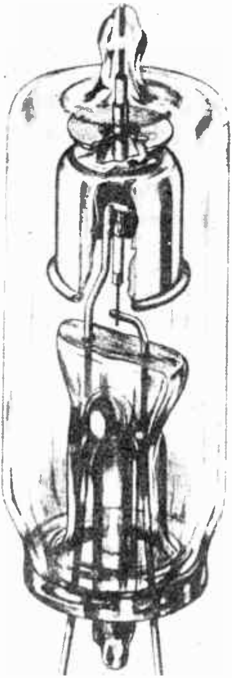


FIG. 4 (*left*).—CONSTRUCTIONAL DETAILS OF THE MAZDA UU9.

(*a*) Exhaust Seal. (*b*) Mica spacer. (*c*) Anode No. 1. (*d*) Common cathode. (*e*) Anode No. 2. (*f*) Support rods and connections. (*g*) Mica spacer. (*h*) Pressed glass base. (*j*) Getter carrier.

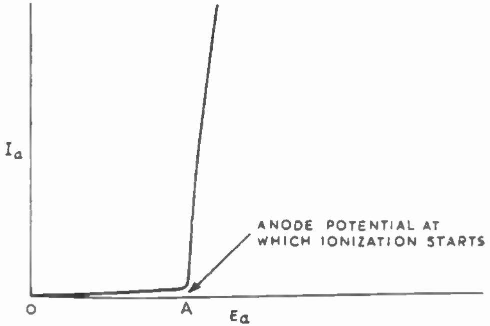
FIG. 5 (*right*).—ENLARGED VIEW OF TYPICAL E.H.T. RECTIFIER.

Many types of valve such as the UU5, UU9, 5U4G, 6X5, etc., employ two separate anode/cathode assemblies mounted inside one envelope to give full-wave rectification. Some manufacturers use the same idea to produce a half-wave rectifier of very high current rating, as many as four separate structures being mounted in one bulb and connected in parallel to share the load.

Gas-filled Types

The internal resistance of a valve of the high-vacuum type is basically due to the mutual repulsion of the negatively charged electrons in the space around the cathode. This may be reduced by operating with a supply of positive ions in the cathode/anode space to neutralize the

FIG. 6.—TYPICAL ANODE CURRENT/ANODE VOLTAGE CURVES FOR A GAS-FILLED RECTIFIER VALVE.



negatively charged electrons. This result can be achieved by introducing a trace of gas into the bulb, the positive ions being produced by the collision of the fast-moving electrons on their way to the anode with the molecules of the included gas.

The effect on the anode volts/anode current characteristic is shown in Fig. 6. At low anode voltages, the velocity of the electrons is insufficient to produce positive ions by collision, and the current to the anode is entirely due to the normal electron emission from the cathode. The emission increases in the usual way, following the relation of equation (1), until the point A is reached, but at this point the electron velocity is sufficient to produce ions by collision, the space charge is neutralized and the valve resistance decreases rapidly to the point at which the current is limited only by the external circuit resistance.

The potential at which electrons have sufficient velocity to ionize the gas is a characteristic of the gas employed, typical figures for the gases usually employed being given in Table 1. The onset of ionization is easily detected by the appearance of a glow discharge in the anode/cathode space, the colour of the glow being a characteristic of the particular gas.

From the point of view of prolonging the life of the valve, the maximum current through the valve is restricted by the necessity of limiting the velocity with which the heavy positive ions are returned to the cathode by the voltage existing across the anode/cathode space. If the potential across the valve is allowed to rise above this limiting value, the ions are returned with sufficient velocity to produce

TABLE 1.—IONIZING POTENTIALS OF GAS-FILLED RECTIFIER VALVES

Gas	*Ionizing Potential*
Argon	15·7
Hydrogen	13·6
Helium	24·5
Mercury	10·4
Xenon	11·5

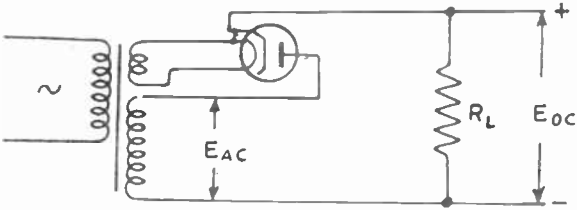


FIG. 7.—HALF-WAVE CIRCUIT WITH RESISTANCE LOAD.

mechanical disintegration of the oxide coating, an effect that is clearly evidenced by violent scintillation at the cathode surface.

In the small sizes commonly employed in radio equipment the construction of gas-filled valves is generally similar to that of high-vacuum valves, except that greater anode/cathode clearances are usually employed.

Rectifier Circuits

The simplest circuit is that shown in Fig. 7. It is not usually employed in radio circuits because of the high-ripple-voltage existing in the D.C. output, but it is an excellent point at which to commence the study of the more usual circuits. In the circuit of Fig. 7, current from the transformer secondary flows through the rectifier valve into the load only when the anode is at a positive potential with respect to the cathode, the load current and the load voltage having the form of half-sinusoids with substantially the same waveform as the positive half-cycle of the transformer secondary voltage.

Effect of Reservoir Capacitor

The addition of a capacitor across the load resistance as in Fig. 8 is a simple modification that profoundly modifies the mode of action. Assuming that steady conditions are established in the circuit there will be a D.C. voltage across the load resistance R_L and the parallel capacitor C_L. Current will not flow through the rectifier valve into the load until the instantaneous value of the transformer voltage on the positive half-cycle exceeds the D.C. voltage on the capacitor, that is :

$$E \sin \theta > E_{DC} \quad \ldots \ldots \quad (2)$$

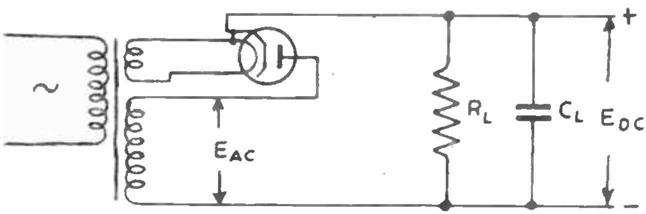


FIG. 8.—HALF-WAVE CIRCUIT RESERVOIR CAPACITOR.

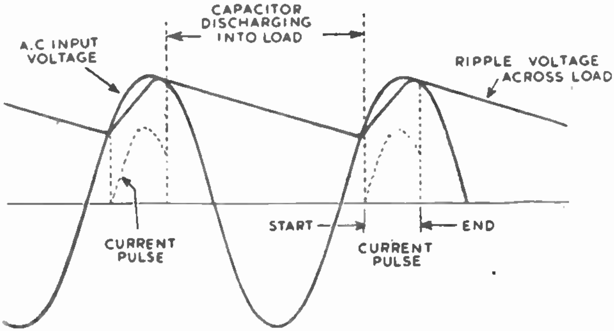


Fig. 9.—Approximate Waveforms in Half-wave Capacitor Input Circuit.

Current then flows through the valve into the load resistance and capacitor in parallel, and the voltage across these rise with the rising 50-c/s voltage up to the peak of the A.C. cycle. It then commences to fall with the supply voltage, but at this point the capacitor commences to discharge into the load resistance, and the rate at which the capacitor voltage falls is controlled only by the exponential relationship governing the discharge of a capacitor into a resistor. In practical circuits, the capacitor is sufficiently large to maintain the voltage across the resistor, and as the decreasing A.C. supply voltage falls below the voltage on the capacitor, the voltage across the rectifier valve decreases to zero and the current through the valve stops, generally just after the peak of the A.C. voltage wave is passed. Typical waveforms of the voltage and current in the circuit are shown in Fig. 9.

In the half-wave circuit without a shunt capacitor across the load, the voltage across the valve on the non-conducting half-cycle is equal to the A.C. supply winding voltage E_{ac}, but when a capacitor is shunted across the load the valve will have to withstand an inverse voltage equal to the D.C. output voltage plus the A.C. supply voltage, the peak value of the inverse voltage being roughly $2\sqrt{2} \times E_{ac}$.

The fundamental difference in the mode of action of a circuit with a reservoir capacitor will now be clear. Without the capacitor the current flows through the valve for practically the whole positive half-cycle of the supply voltage, but with a reservoir capacitor the current passes in the form of a short pulse lasting for only a small fraction of the positive half-cycle. During the portion of the half-cycle when the valve is not conducting, the load current is supplied by the capacitor, the voltage falling exponentially until it is picked up again on the following half-cycle by the rising voltage across the transformer secondary winding. As the whole of the D.C. load power must be supplied by the transformer, the average value of the current pulse over the whole supply voltage cycle must equal the mean D.C. current; and thus the peak value of the current pulse must exceed the mean value by a large factor. In practical circuits of the half-wave type, the peak value of current pulse may be ten times the mean D.C. current in the

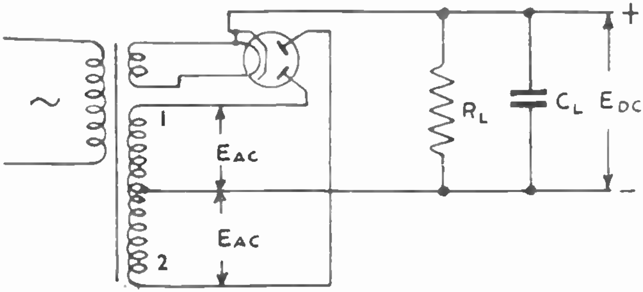


Fig. 10.—Full-wave Circuit with Reservoir Capacitor.

load. Increase in the size of the reservoir capacitor will reduce the rate at which the voltage on the capacitor falls as it discharges into the load resistance, and will therefore decrease the time during which the supply voltage exceeds the voltage on the capacitor. For a given D.C. load current, the shorter current pulse must have a higher peak value. As the ripple voltage that exists across the reservoir capacitor is equal to the amount the capacitor voltage falls during the discharge period, an increase in the value of capacitor must decrease the ripple voltage. The process has been explained in some detail because it is fundamental to practically all the rectifier circuits used in radio equipment.

The half-wave circuit is widely used in radio and television equipment for loads requiring low values of current, particularly at high voltages. It has also recently become popular in A.C./D.C. equipment, where its use eliminates the mains transformer and allows the receiver to be used on either A.C. or D.C. circuits without change.

Full-wave Circuits

In the full-wave circuit shown in Fig. 10 a single, centre-tapped transformer secondary winding supplies the D.C. load through two rectifier valves on alternate half-cycles. On the half-cycle during which terminal 1 is positive, terminal 2 is going negative with respect to the earthed centre-tap. The two halves of the secondary winding and the two rectifier paths are used alternately. As charging current flows into the reservoir capacitor on both negative-going and positive-going half-cycles of the supply voltage, the voltage swing across the

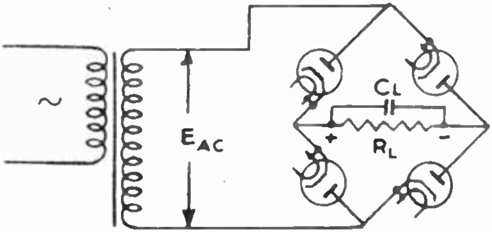


Fig. 11.—Bridge Circuit.

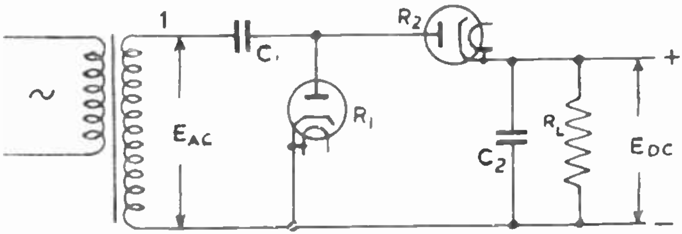


FIG. 12.—HALF-WAVE VOLTAGE DOUBLER CIRCUIT.
(*Note:* The connections to the anode and cathode of R_1 should be reversed.)

capacitor will be nearly halved, as will the ratio of peak A.C. to mean D.C. current.

The full-wave circuit is the most generally used circuit in radio and television equipment intended for A.C. main supplies only.

Full-wave Bridge Circuit

In high-power circuits, such as those used in large audio amplifiers, etc., it is occasionally found that the inverse voltage across the rectifier valve is outside the maximum permissible figure. In the full-wave bridge-circuit shown in Fig. 11, it will be noted that there are two valves in series at any particular instant, and that the maximum inverse voltage across either rectifier is equal to half the D.C. load voltage. In the half- and full-wave circuits previously described, the voltage across the rectifier on the reverse half-cycle will rise to a value that is roughly equal to the D.C. voltage plus the peak value of the A.C. voltage.

Voltage-multiplying Rectifiers

The circuit shown in Fig. 12 results in a D.C. voltage that is approximately $2\sqrt{2} \times E_{ac}$, about twice the voltage derived from either the full- or half-wave circuits. The circuit operation is as follows. On the half-cycle during which terminal 1 is positive going, current flows into capacitor C_1 through rectifier R_1, leaving it charged to a voltage of roughly $\sqrt{2} \times E_{ac}$ at the end of the half-cycle. On the next half-cycle, the voltage applied to rectifier R_2 will be the sum of the transformer output voltage and the voltage on the capacitor, roughly $2\sqrt{2} \times E_{ac}$, and this voltage will be applied to the rectifier, charging capacitor C_2 to a peak voltage of approximately $2\sqrt{2} \times E_{ac}$.

This device may be applied to produce greater voltage multiplications, but because of the number of separate heater windings required, is little used with thermionic rectifiers.

Filters

With any finite value of capacitor across the D.C. load, additional smoothing circuits will be required to remove the residual ripple voltage. These generally take the form of series inductance and shunt capacitance across the D.C. load; but the development of small electrolytic capacitors of large capacitance is tending to eliminate the inductance in favour of a resistor which is appreciably smaller and

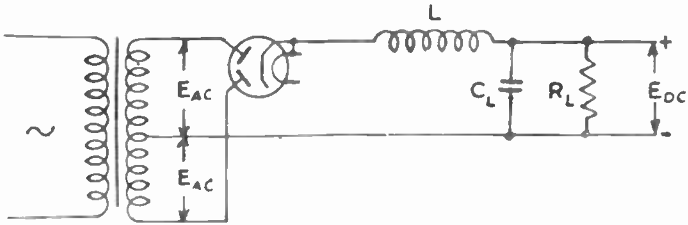


FIG. 13.—FULL-WAVE CHOKE-INPUT CIRCUIT.

cheaper and has no external field, a matter of considerable importance in high-gain audio amplifiers.

The design of these smoothing filters is dealt with under "Ripple Filters", which should be consulted for further details.

Smoothing Circuits for Gas-filled Rectifiers

Gas-filled rectifier valves may be used in any of the circuits described above, though the advisability of operating valves of this type with a low ratio of peak to mean D.C. current suggests that the half-wave rectifier circuit be avoided. In the full-wave and bridge circuits current conduction during the half-cycle can be prolonged by the use of a choke between the reservoir capacitor and the rectifier valve, the so-called choke-input circuit, shown in Fig. 13.

It can be shown that the output voltage waveshape is given by the Fourier series

$$E = \frac{2E}{\pi}(1 - 0{\cdot}67 \cos 2t - 0{\cdot}134 \cos 4t \; . \; . \; .) \qquad (3)$$

the value of the D.C. component E_{DC} being $\frac{2E}{\pi}$ and that of the ripple component at twice supply frequency by $0{\cdot}67 \times E_{DC}$. If the inductance of the first choke is sufficiently high the ripple current flowing in the circuit as a result of presence of the ripple e.m.f. $0{\cdot}67 \times E_{DC}$ will be reduced to a value less than that of the D.C. component $\frac{2E}{\pi R_{DC}}$. Under these conditions, the output current will consist of substantially square pulses, each approximately half a cycle in length; and the ratio of peak to D.C. current will approach unity. To achieve this, the D.C. component must be larger than the A.C. component at twice supply frequency.

The D.C. component of output-circuit current will be

$$\frac{E_{DC}}{R_{DC}} = \frac{2E}{\pi R_{DC}} \quad . \; . \; . \; . \; . \; . \; . \; . \qquad (4)$$

and the A.C. component $\frac{E_{ac}}{Z} = \frac{0{\cdot}67 \times E_{DC}}{2\pi f L}$, which at a supply frequency of 50 c/s becomes:

$$\frac{0{\cdot}67 \times E_{DC}}{628L} \quad . \; . \; . \; . \; . \; . \; . \qquad (5)$$

D.C. and A.C. current components will be equal when :

$$\frac{E_{DC}}{RL} = \frac{0{\cdot}67 \times E_{DC}}{628L} \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

This requires a value of inductance of :

$$L \simeq \frac{RL}{1{,}000} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

This fixes the minimum value of choke inductance that should be used, but if current conduction is to be continuous under conditions of a variable D.C. load, it is necessary that the value of R used in calculating L should be the value corresponding to minimum D.C. load current, i.e., the maximum value of D.C. load-circuit resistance. By the use of an iron-cored choke having an air gap proportioned to give the required inductance at the minimum load current, it is possible to meet the choke requirements with a choke appreciably smaller and cheaper than would be obtained by the standard design procedure.

It will be obvious that the choke cannot be designed to meet equation (7) down to zero D.C. currents, for this would require an infinite value of inductance. If the voltage soaring that accompanies discontinuous conduction is to be avoided, it is necessary to shunt some load resistance across the output of the rectifier circuit to limit the range over which the D.C. load current swings. The use of a gas-filled rectifier valve in conjunction with a choke-input circuit results in a rectifier circuit having a very good output regulation, as is shown in Fig. 14. Discontinuous conduction sets in at the point marked A, and it will be noted that the change to capacitor-input-circuit conditions at that point results in the output voltage rising very rapidly, with further decrease in load current.

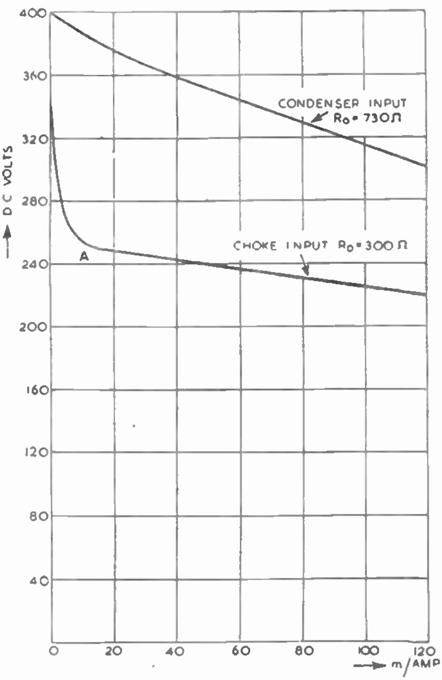


FIG. 14.—PERFORMANCE OF CHOKE INPUT AND CAPACITOR INPUT CIRCUIT USING IDENTICAL COMPONENTS.

Rectifier Circuit Design

A direct mathematical analysis of the capacitor–input circuit is very complicated and of little direct value to a designer, but it does serve to show the parameters that are of importance. The analysis shows that for any given A.C. input voltage, the D.C. load current and voltage are functions of two parameters

$$\frac{\text{D.C. load resistance}}{\text{Rectifier circuit resistance}} = \frac{R_L}{R_s}$$

and

$$\frac{\text{D.C. load resistance}}{\text{Reservoir capacitor reactance}} = \frac{R_L}{X_c}$$

The D.C. load resistance is E_{DC}/I_{DC}, and the reservoir capacitance is computed at the frequency of the main ripple component, that is, at supply frequency for the half-wave circuits, and at twice supply frequency for the full-wave circuits. The source resistance, R_s, is the sum of all the resistances that are effectively in series between the supply terminals and the reservoir capacitor. These are :

(Transformer primary resistance) $\times\ n^2$ +
(Transformer secondary resistance) + (valve resistance)

i.e., $$R_p \times n^2 + R_s + R_v$$

where n is the turns ratio, very approximately equal to the open-circuit voltage ratio E_2/E_1.

The valve resistance is a somewhat ambiguous term to apply to a non-linear element, but satisfactory accuracy for all design purposes can be achieved if this is taken as :

$$R_v = 1{\cdot}16 \times \frac{E_{\text{peak}}}{I_{\text{peak}}}$$

E_{peak} and I_{peak} being the D.C. values taken from the valve characteristic at the peak value of the current pulse. The ratio of D.C. load voltage to A.C. supply voltage, in the full-wave circuit of Fig. 10, is

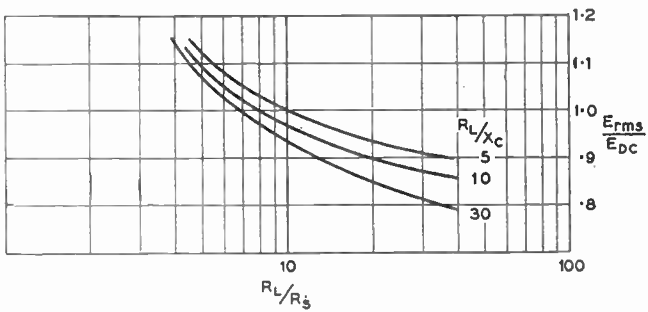


FIG. 15.—E_{rms}/E_{DC} AS A FUNCTION OF R_L/X_C AND R_L/R_S FOR THE FULL-WAVE RECTIFIER CIRCUIT.

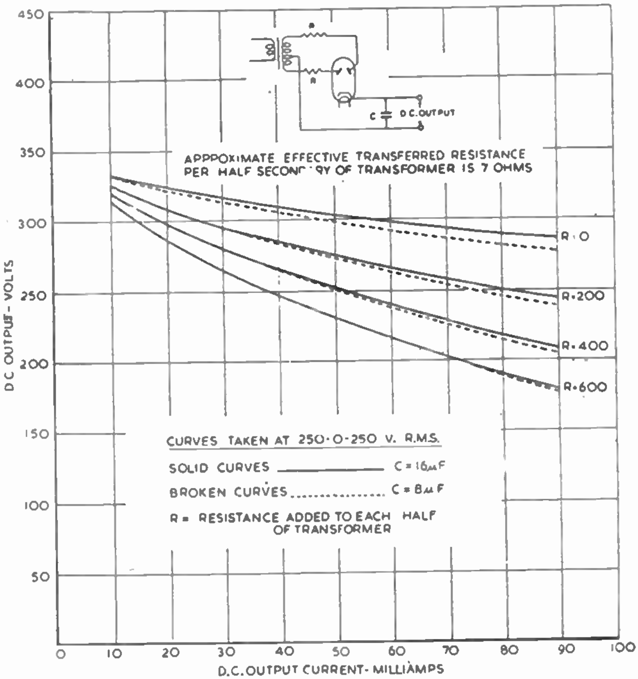


Fig. 16.—Characteristic Curves of Average Mazda Valve UU6.

plotted as a function of these two parameters in Fig. 15. On a given core size, the secondary winding voltage and the winding resistances are interdependent, and this necessitates a trial calculation with assumed values of winding resistances, followed by the final check made with the more accurate values of winding resistances computed from the approximate values of voltage determined by the first calculation.

Most of the valve manufacturers supply data of the type shown in Fig. 16 enabling the A.C. voltage required for any load to be approximated by interpolation. If in any case a preliminary design of transformer is to be available, these curves may be used to indicate the required value of secondary voltage, or if the first design is to be as accurate as possible, the curves may be used to indicate the probable secondary voltage for use in calculating the transformer winding resistances, required when the basis curves of Fig. 15 are used.

Operating Precautions

With all types of rectifier valve it is advisable to operate with low values of peak-current/mean-current, and with the gas-filled types it is imperative to do so. Practically all manufacturers of the high-vacuum types suggest a maximum value of reservoir capacitor that should be employed, and in addition specify a minimum value of the supply-circuit resistance that should be realized. Thus, if the source resistance—as discussed in the section on circuit design—is not reached by the normal design procedure, it is necessary to add external resistance, either as a single resistance in the centre-tap, or as resistance in both legs of the supply transformer, to bring the source resistance up to the specified minimum value.

If the rectifier valve is operating at or near maximum ratings, it is essential that the heater voltage at the valve pins should reach the rated value, a point that is of particular importance when using gas-filled valves. With gas-filled valves, it is good practice to use some form of delayed switching of the anode voltage in order to allow the cathode to reach full operating temperature before load is applied. This is sometimes applied to hard valve rectifier circuits, but it is of doubtful advantage, since it results in a very large current rush into the reservoir capacitor when the anode circuit is closed on to the discharged capacitor. Under ordinary conditions, the resistance of the valve during its warming up period is sufficient to limit the current to a safe value, but it should be noted that this comment does not apply to gas-filled rectifiers.

An anode/cathode short-circuit or an overload on the D.C. side may not result in an increase in transformer primary current sufficient to blow primary side fuses, particularly when the transformer supplies several heater circuits or other loads. For complete protection against a transformer burn-out due to an overload of this type it is necessary to fit fuses of suitable rating into the anode leads of the rectifier valve, and it may also be necessary to add a fuse in the D.C. circuit beyond the smoothing choke. This complication has led to the development of thermal circuit-breakers consisting of a bi-metallic strip in contact with the transformer winding and arranged to open the supply circuit if the winding temperature rises above normal.

J. M.

METAL RECTIFIERS

Great advances have been made in the metal rectifier, and, for example, modern "Westalite" selenium rectifiers are available which will operate at twice or even four times the reverse voltage of the original selenium rectifiers (20 volts). These elements are termed Westalite "double-voltage" and "quadruple-voltage" respectively. Typical D.C. characteristics are shown in Fig. 17.

Note that the double-voltage (D.V.) type has only a very slightly greater forward resistance than the single-voltage (S.V.) type. This greatly reduces rectifier size and weight, since a rectifier unit built of D.V. elements will have only half the number of discs and half the losses of a unit built of S.V. elements, D.V. elements are therefore used for

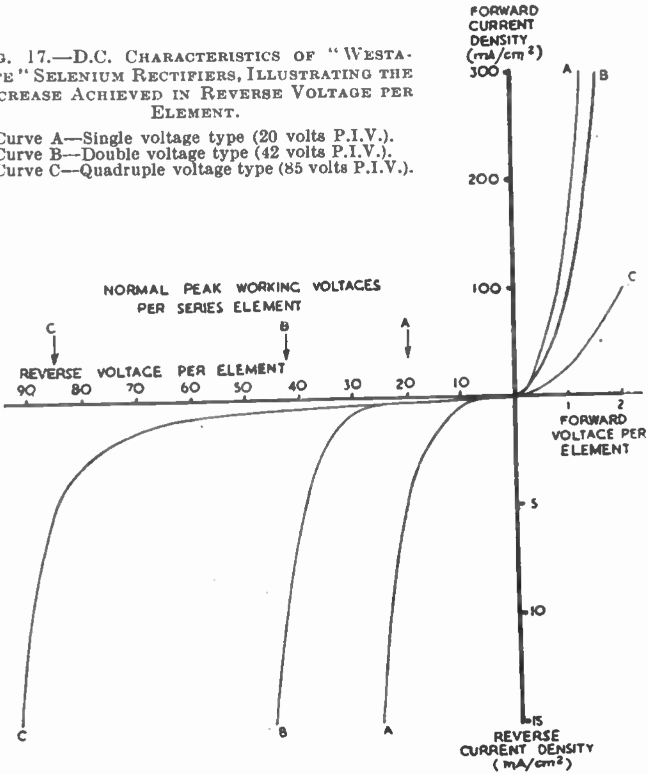


FIG. 17.—D.C. CHARACTERISTICS OF "WESTALITE" SELENIUM RECTIFIERS, ILLUSTRATING THE INCREASE ACHIEVED IN REVERSE VOLTAGE PER ELEMENT.

Curve A—Single voltage type (20 volts P.I.V.).
Curve B—Double voltage type (42 volts P.I.V.).
Curve C—Quadruple voltage type (85 volts P.I.V.).

nearly all radio rectifiers. Quadruple-voltage (Q.V.) elements are used for miniature tubular E.H.T. rectifiers, and have made it possible to achieve a peak reverse voltage rating of 1,500 volts/in. length. Modern copper oxide types have also been improved, and are preferable to selenium or germanium rectifiers where high and stable reverse resistance is necessary.

Rating, Heat Dissipation and Life

The working life of metal rectifiers is determined almost entirely by the operating temperature of the elements themselves. Apart from sustained voltage overload, which can cause breakdown, the life is not directly affected by the electrical loading, but only by the temperature rise which the electrical loading causes under given cooling conditions. The radio designer using metal rectifiers should therefore forget his conventional "valve" approach, and instead of concentrating primarily on a fixed current limit (as with a valve) he should pay prime attention to heat dissipation, suiting this to the load he requires to impose on the rectifier and the life he desires. Since the nominal current rating of a metal rectifier is based on an assumption of average ventilation conditions, a designer can often safely obtain a greater current output by improving the cooling of the rectifier. This is often a better policy than attempting to economize in total H.T. current demand, as it produces a more reliable receiver design and, if a lower rectifier temperature results, a longer rectifier life.

The maximum voltage rating, however, is not similarly flexible, as will be apparent from the curves in Fig. 17. Owing to the sharp bend in the reverse characteristic, the reverse current increases rapidly at excess voltages, so causing heating and damage. Here again, the damage is caused by the heat and not primarily by the electrical conditions, so that transient over-voltages, which do not produce serious heating, are permissible. The sudden decrease in resistance at the reverse bend can in fact be used as a surge absorber of very high index ($I = kV^8$ approximately), or as a means of improving E.H.T. regulation, since in both these cases the average energy is limited and no undue temperature rise is produced.

Cooling

Metal rectifiers operate at a much lower surface temperature than valve rectifiers, and therefore must dissipate their losses by convection and conduction rather than by radiation. Convection-cooled units are either provided with cooling fins or with spaced rectifying elements, and it is essential to ensure an unrestricted flow of cool air through the unit. Since the volume of cooling air required is considerable, and the temperature rise imparted to it is small, it is good practice to cause the main air intake of the receiver to flow through the power rectifier. In this way, with good receiver cabinet design, it is actually possible to obtain a greater safe output from the metal rectifier when in the receiver cabinet than would be possible if it were freely mounted in still air outside the receiver. This careful attention to air flow through the power rectifier and cabinet is usually further rewarded by lower operating temperatures through the receiver.

Element Temperature and Life: Practical Test for Radio Power Rectifiers

The electrical loading, the cooling and the ambient temperature together determine the temperature of the rectifying elements. The element temperature then determines the rectifier useful life. Thus the total element temperature permissible depends on the useful life expected. The rate of increase of forward resistance (ageing) of the rectifier increases with temperature, becoming rapid at element temperatures above 70° C.

A compromise between rating, cooling and average ambient temperature for different applications is therefore necessary. For radio applications a practical test recommended for Westalite selenium rectifiers is as follows: The receiver is run on the maximum mains voltage (i.e., nominal voltage plus 6 per cent), with maximum current demand (e.g., maximum contrast setting, etc.), and with the mains tapping at the most unfavourable setting (e.g., the setting which results in the highest voltage at the rectifier within the range of voltages covered by that tapping). All covers should be in position, and the receiver should be carefully shielded from draughts. The temperature *rise* of the rectifier as measured by a thermocouple pressed into close contact with the rectifier elements themselves should not exceed 30° *C. above the ambient room temperature,* using a rectifier unit having characteristics as near to the lower pass limit as is practicable.

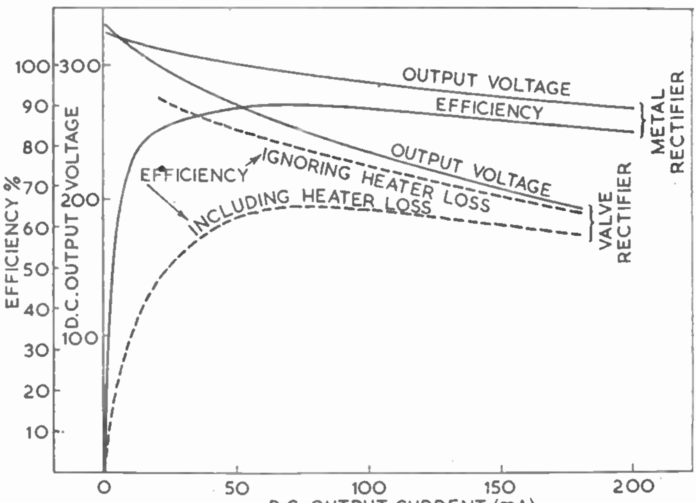


FIG. 18.—COMPARISON OF METAL AND VALVE RECTIFIERS IN A HALF-WAVE CIRCUIT FOR A "TRANSFORMERLESS" RECEIVER.

Mains supply 230-volt A.C. Reservoir capacitor 60 μF.

This arbitrary test is designed to allow for typical variations in conditions of use, and is applicable to power rectifiers in radio receivers intended for use in temperate climates. For tropical applications special advice should be obtained from the manufacturers.

H.T. by Rectification Directly from the Mains

Transformerless Television Receivers

A typical television receiver requires a power supply of about 60 watts for the H.T. supply alone, and in order to save the cost and weight of a transformer of comparable rating it is common practice to obtain this H.T. power by half-wave rectification directly from the mains. In some cases a transformer is used to supply the heaters, and the primary winding can then be used as an auto-transformer to maintain the rectifier input at 250 volts A.C. for all supply voltages, but if a series heater chain is used, the A.C. input to the rectifier may fall as low as 200 volts. In order to obtain an adequate H.T. voltage with low ripple, a large value of reservoir capacitance is used, under which condition the rectifier carries short pulses of current which may reach a very high peak value (e.g., 3 amperes peak for 220 mA mean D.C. output). These high peak

TABLE 2.—TYPICAL TELEVISION H.T. SUPPLY RECTIFIERS

Type	*Maximum A.C. Voltage (r.m.s.)*	*Nominal D.C. Output*		*C1 (μF)*	*C2 (μF)*	*C3 (μF)*
		Voltage	*Current (mA)*			
14A86	240	270	220 *	60–100	100–150	30–60
14A100	270	305	220 *	60–100	100–150	30–60
LW7	240	270	300 *	60–100	150	60
LW9	270	305	300 *	60–100	150	60

N.B. Two units in circuit (c) will give double output voltage at the same current (thus two units LW7 will develop 540 volts at 300 mA from a 240-volt supply).

* Safe current rating depends entirely on the degree of cooling provided (see "Rating, Heat Dissipation and Life").

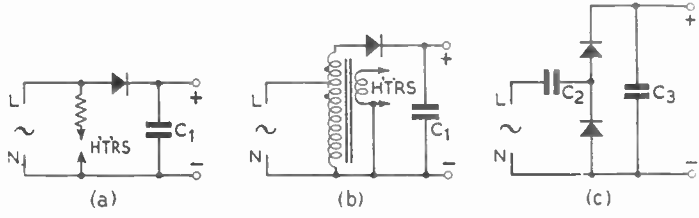


FIG. 19.—TELEVISION H.T. SUPPLY RECTIFIERS (SEE TABLE 2).

values do not damage the metal rectifier, as they would a valve, no limiting resistor being necessary, and the circuit accordingly gives a high D.C. voltage with high efficiency (see Fig. 18). Design precautions are :

(a) The rectifier must be rated for a peak inverse voltage $2\sqrt{2} \times V$ (where V = maximum r.m.s. value of supply voltage).

(b) The rectifier operating temperature should not exceed recommended value (see " Rating, Heat Dissipation and Life ").

A.C./D.C. Radio Receivers

The same half-wave arrangement is commonly used for A.C./D.C. radio receivers. The principles are the same as for the heavy-duty television rectifier, but in general less vigorous " peaking " is employed, since by reducing the value of the reservoir capacitance the H.T.

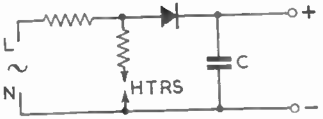


FIG. 20.—A.C./D.C. RADIO RECEIVER RECTIFIER CIRCUIT (SEE TABLE 3).

voltage can be made substantially the same when the receiver is fed from either A.C. or D.C. mains, without change of connections. In miniature portable receivers employing series resistors, with 110 volts input to the rectifier, it is necessary to ensure that the rectifier and reservoir capacitor will be safe in the event of a valve heater failure, which can result in 250 volts A.C. being applied to the rectifier.

TABLE 3.—TYPICAL A.C./D.C. RADIO RECEIVER RECTIFIERS

Type	*A.C. Voltage*	*Output Current (mA)* *	*C (μF)*
15B35	240	60	16
14B35	240	80	16
HT46	250	120	16
HT47	250	60	16
HT48	250	30	8
HT54	110	60	16

* See " Rating, Heat Dissipation and Life ".

Valve Replacement Rectifiers (Centre-tap Circuit)

Metal rectifiers are available for use in the conventional centre-tap rectifier circuit for use either as valve replacements or for use in conjunction with centre-tapped transformers which are often readily available.

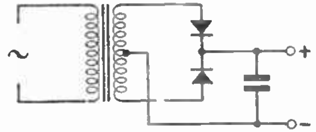


FIG. 21.—VALVE REPLACEMENT (CENTRE-TAP) CIRCUIT.

TABLE 4.—TYPICAL VALVE-REPLACEMENT RECTIFIERS

Type	*Input Voltage*	*Mean Output Current (mA)* *
15B39	120–0–120	110
HT50	300–0–300	40
HT51	350–0–350	100
HT52	350–0–350	200
HT53	500–0–500	200

* See " Rating, Heat Dissipation and Life ".

E.H.T. Rectifiers

The Westalite tubular selenium E.H.T. rectifiers employ quadruple-voltage elements operating at 85 peak inverse volts per element. The rectifier discs are assembled in insulating tubes under light spring pressure, provided with connecting tags or wires at each end, by which all but the largest units can be supported directly in the wiring. The overall length is about 0·6 in./1,000 volts, and the maximum voltage which can be handled by a single tubular rectifier is over 20,000 volts. The self-capacitance is small, and is given approximately by the following expression :

Self-capacitance of type 39E rectifiers $= \frac{55}{N}$ pF + strays (rating 0·1 mA).

Self-capacitance of type 36 EHT rectifiers $= \frac{350}{N}$ pF + strays (rating 2 mA), where N is the number of elements.

The self-capacitance is dependent on the reverse voltage, as can be seen from the typical curve in Fig. 22.

These tubular E.H.T. rectifiers are particularly suitable for use in voltage multipliers owing to the simplification which results from the elimination of multiple high-voltage heater windings. Only a few typical circuits can be described in this brief review.

E.H.T. by Flyback Pulse Rectification

An E.H.T. supply may be obtained by rectification of flyback pulses from the line-scanning output transformer, using either a simple half-wave circuit or a pulse doubler or tripler, depending on the degree of overwinding provided on the transformer and the E.H.T. voltage

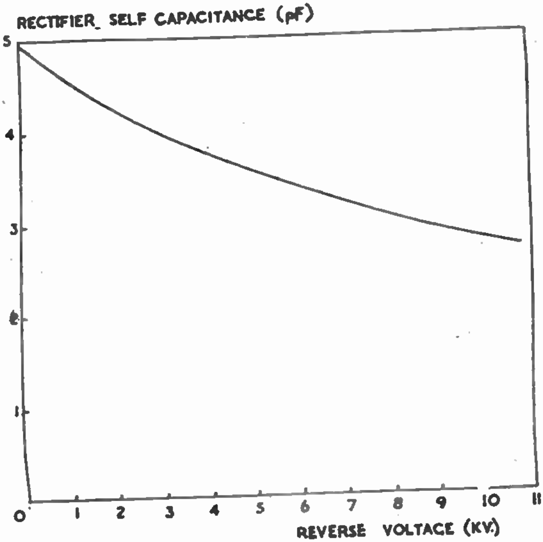


Fig. 22.—Variation of Apparent Self-capacitance of an 11-kV Tubular E.H.T. Rectifier (36EHT130) with Reverse Voltage.

(Additional stray capacitance approx. 2 pF.)

required. The use of a metal rectifier doubler or tripler provides a ready means of increasing E.H.T. without modification to the transformer, but the use of the tripler is generally necessary only with transformers having no overwindings. Typical arrangements are given in Fig. 23.

Table 5.—Typical E.H.T. Rectifiers

Rectifier	*Peak Pulse Input Voltage*	*Approximate D.C. Output Voltage at 100 μA*		
		Half-wave	*Doubler*	*Tripler*
36EHT30	2,180	1,960	3,520	5,000
36EHT45	3,470	2,950	5,280	7,520
36EHT60	4,350	3,930	7,020	10,000
36EHT100	7,250	6,550	11,700	16,700
36EHT160	11,600	10,400	18,700	26,700
36EHT200	14,500	13,000	23,400	33,400
36EHT240	17,400	15,700	28,000	40,000

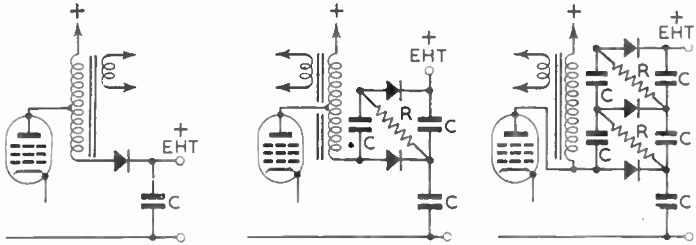


FIG. 23.—HALF-WAVE, DOUBLER, AND TRIPLER PULSE-DRIVEN RECTIFIER MULTIPLIERS.

E.H.T. by Half-wave Rectification from Mains Transformer

The main advantage of this arrangement is the good voltage regulation which can be obtained. The disadvantages are the size and cost of the mains transformer and the reservoir capacitor.

TABLE 6.—E.H.T. HALF-WAVE RECTIFIERS (SINE-WAVE INPUT)

Rectifier	*Approximate E.H.T. Output Voltage*		*R.M.S. Input Voltage*	*C * (μF)*
	At 100 μA Load	*At 2 mA Load*		
36EHT30	1,050	1,000	810	0·5
36EHT45	1,575	1,500	1,220	0·25
36EHT60	2,100	2,000	1,620	0·25
36EHT100	3,500	3,250	2,700	0·1
36EHT160	5,600	5,300	4,320	0·1
36EHT200	7,000	6,600	5,400	0·1
36EHT240	8,400	7,900	6,480	0·05

* The values given for C are for 50-c/s supplies. For other frequencies C should be varied in inverse proportion to the frequency. Capacitance may be reduced for 100 μA load.

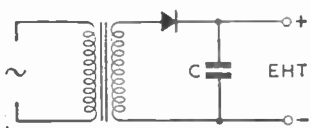


FIG. 24.—E.H.T. HALF-WAVE RECTIFIER (SINE-WAVE INPUT).

E.H.T. by Modified Voltage Doubling from Mains Transformer

The doubler shown below is usually to be preferred to the conventional doubler, as it permits one side of the transformer to be earthed, a condition for which the majority of E.H.T. transformers are designed.

FIG. 25.—MODIFIED VOLTAGE DOUBLER PERMITTING COMMON CONNECTION OF ONE SIDE OF A.C. AND D.C. CIRCUIT.

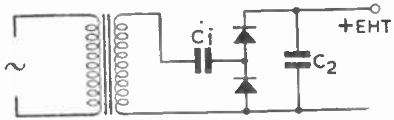


TABLE 7.—E.H.T. VOLTAGE DOUBLER RECTIFIERS (SINE-WAVE INPUT)

Rectifiers (Two Required)	*Approximate E.H.T. Voltage*		*R.M.S. Input Voltage*	*C1* *		*C2* *	
	At 100 μA	*At 2 mA*		*μF*	*Voltage Rating (kV)*	*μF*	*Voltage Rating (kV)*
36EHT30	2,100	1,800	810	0·5	1·2	0·25	2·5
36EHT45	3,150	2,850	1,220	0·5	1·75	0·25	3·5
36EHT60	4,200	3,600	1,620	0·25	2·5	0·1	5
36EHT100	7,000	5,800	2,700	0·25	4	0·1	8
36EHT160	11,200	8,500	4,320	0·1	6	0·05	12
36EHT200	14,000	12,000	5,400	0·1	8	0·05	16
36EHT240	16,800	14,300	6,480	0·05	10	0·02	20

* The values given for C1 and C2 are for 50 c/s supplies, for other frequencies C should be varied in inverse proportion to the frequency. Capacitance values are not critical, and can be reduced for loads under 2 mA.

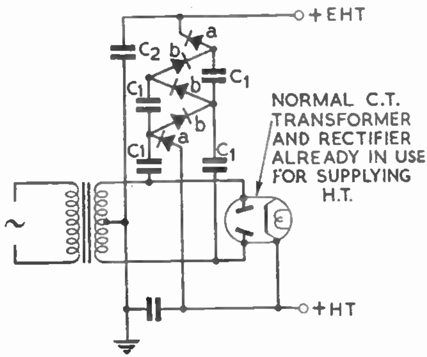


FIG. 26.—WALKER MULTIPLIER.

All capacitors are 0·1 μF for 150 μA load. Pro-rata for different loads up to 2 mA maximum.

C_1 Voltage Rating

2 kV for 350–0–350,

3 kV for 500–0–500.

C_2 must be rated for full E.H.T. developed.

"Walker Multiplier" Producing E.H.T. from Existing Centre-tapped E.T. Transformer

This circuit is suitable for developing any potential up to about 8 kV for oscillographs, etc., from the existing centre-tapped H.T. transformer. Various values of E.H.T. can be developed by suitable choice of the number of stages. "Half-section" rectifiers (*a*) are used at each end of the multiplier, and must be retained. Any required number of intermediate full sections (*b*) may be used up to about five. Typical designs for 350–0–350-volt and 500–0–500-volt transformers are given below.

TABLE 8.—"WALKER MULTIPLIER" RECTIFIERS

Rectifier	350–0–350	500–0–500
Rects (*a*)	36K14	36EHT20
Rects (*b*)	36EHT25	36EHT40
Approximate Output Voltage		
2 (*a*) + 5 (*b*)	5·0 kV	7·8 kV
2 (*a*) + 4 (*b*)	4·2 kV	6·0 kV
2 (*a*) + 3 (*b*)	3·4 kV	4·8 kV
2 (*a*) + 2 (*b*)	2·6 kV	3·7 kV
2 (*a*) + (*b*)	1·7 kV	2·4 kV

E.H.T. Stabilizer and Limiter

The sharp bend which occurs in the reverse characteristic of Westalite selenium rectifiers has already been shown in Fig. 17. A large-scale graph of the reverse bend for quadruple-voltage elements (Westalite type 36) is given in Fig. 27, plotted for one series element only. (Thus for 36EHT100, which contains 100 elements, the voltage scale of this graph requires multiplying by 100.)

This sharp bend can be used to provide a limiting action in low-energy circuits. The critical part of the characteristic for this application is that portion lying between reverse voltages of 75 and 90 volts, and within this range the curve can be represented to a first approximation by the expression

$$I = kV^{8\cdot5} \times 10^{-20}$$

where I = reverse current (amperes);
V = reverse voltage per series element;
K = 2·13 for Westalite E.H.T. rectifiers type 36.

It is clear from this eighth-power law that at the bend the rectifier presents a shunt resistance whose value is highly sensitive to the applied voltage, and as such it can form a useful voltage regulator in any circuit where the energy is limited or the source impedance is comparatively high.

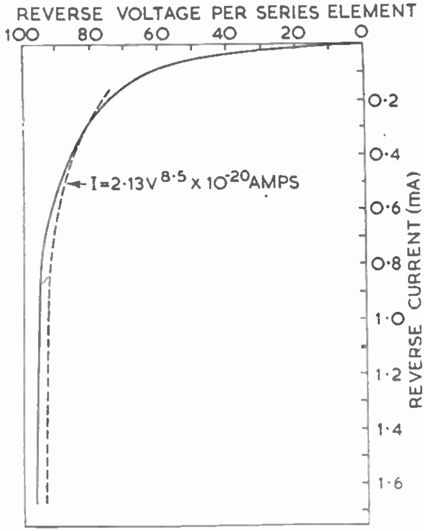


FIG. 27.—REVERSE CHARACTERISTIC OF A SINGLE QUADRUPLE-VOLTAGE (36EHT) DISC, ILLUSTRATING THE SHARP TURNOVER.

In all forms of flyback E.H.T. circuits the rectifier can be used to stabilize and improve the regulation by direct connection across the E.H.T. output. However, if a suitable metal rectifier is itself used as the rectifying means in any of the pulse circuits illustrated above it will provide this regulating action automatically at the same time, and will provide improved voltage regulation with load changes. The limiting action also prevents undue E.H.T. rise in the event of faulty line time base operation, loss of synchronizing, etc., and so protects capacitors and tube against severe overvoltage.

Conduction Cooled Rectifiers

Although metal rectifiers are conventionally assembled with cooling fins for air convection cooling, a recent development which results in a considerable saving in space is the conduction cooled rectifier. In this type of unit the rectifying elements are assembled in a number of small stacks in a flat metal case which is mounted on the receiver chassis. The lower end of each stack of rectifying elements is insulated from the bottom of the case by a thin sheet of insulating material having good thermal conductivity, and since this surface of the case is effectively cooled by conduction to the chassis, the heat generated in each short stack of elements is effectively dissipated. The case is constructed with a flat under-surface which is bolted down into good contact with the chassis, and is provided with raised lugs on the upper surface to ensure that it cannot be incorrectly mounted.

A typical example of such a conduction-cooled unit (18RA1-1-8-1) is

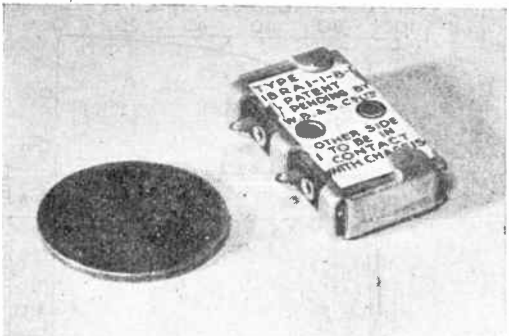


FIG. 28.—CONDUCTION COOLED RECTIFIER. OUTPUT 140 VOLTS, 60 mA.

shown in Fig. 28. This is a half-wave unit for miniature A.C./D.C. receivers, and is rated at 120 volts A.C. input and 140 volts, 60 mA, D.C. output when mounted on a chassis operating at 40° C. It measures only $1\frac{1}{2} \times \frac{7}{8} \times \frac{3}{8}$ in. thick.

Westectors

Westectors are miniature copper oxide rectifiers assembled in insulating tubes with wire ends, and are approximately the same size as $\frac{1}{4}$-watt resistors. They are suitable for straightforward audio-frequency and high-frequency rectification up to about 1·5 Mc/s, and are very suitable for pulse operation, e.g., clipping, clamping, etc., at video frequencies. Two types of element (WX = high impedance, and W = lower impedance) are made, and each type is available in assemblies containing from one to fifteen elements. A very wide range of voltage ratings and impedance is therefore available. The code number indicates the type of element and the number in series; thus Westector WX6 comprises six type WX elements.

Westector Characteristics

Typical D.C. characteristics for both W and WX types are given in Fig. 29, plotted for one series element only.

From these curves the best Westector for a given application can be selected by choice of a suitable number of series elements (see preferred list below).

Westector Self-capacitance

The self-capacitance of Westectors varies with the reverse voltage over a range of about 2 : 1 for a change of reverse voltage from zero to 10 volts. The following table enables the approximate self-capacitance of any Westector to be found at various reverse voltages :

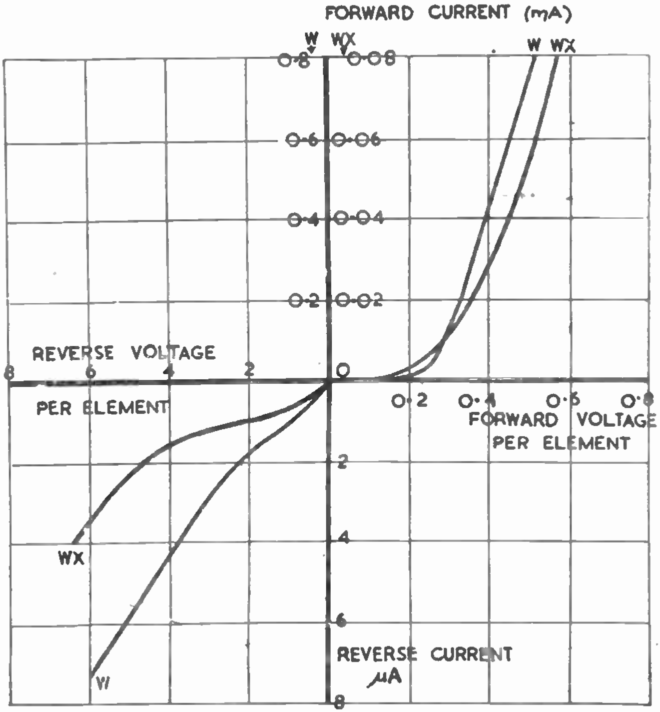


FIG. 29.—D.C. CHARACTERISTICS OF WESTECTORS TYPES W1 AND WX1 (i.e., FOR ONE SERIES ELEMENT ONLY).

For other types multiply voltage scales by the number of elements in series (e.g. WX6 has 6 elements).

TABLE 9.—SELF-CAPACITANCE OF WESTECTOR

Westector	*At Less than 1 V (pF)*	*At 6 V (pF)*	*At 10 V (pF)*
W type . . .	$\frac{150}{N}$	$\frac{100}{N}$	$\frac{70}{N}$
WX type . . .	$\frac{50}{N}$	$\frac{35}{N}$	$\frac{25}{}$

Where N = the number of series elements used in the Westector.

(Thus the self-capacitance of a WX6 Westector at 6 V reverse is $\frac{35}{6}$ approximately 6 pF.)

Pulse Operation of Westectors (Clippers, Limiters, etc.)

There are many applications in television circuits where a miniature diode is required for such purposes as an amplitude-biased clipper (e.g., frame-pulse separators), or as a clamping rectifier (e.g., D.C. restorer). In these applications it is often essential that the reverse resistance of the rectifier should be high and stable, or the circuit operation may become impaired. In such cases a copper oxide rectifier offers a higher and more stable reverse resistance than either the selenium or germanium rectifier.

The pulse ratings in such circuits may be allowed to exceed the continuous ratings, the transient forward current is unlimited and the transient reverse voltage may be greatly increased (see table below). For example, a WX6 Westector has a reverse resistance of over 10 MΩ at 35 volts, a self-capacitance of only 6 pF, and a transient inverse voltage rating of 90 volts.

TABLE 10.—WESTECTOR RATINGS

Westector Type *	*Continuous Operation*				*Transient Operation*	
	Maximum R.M.S. Input Voltage (half-wave with capacitor load)	*Maximum Reverse Voltage*	*Maximum Mean Forward Current (mA)*		*Transient Peak Reverse Voltage*	*Transient Forward Current*
			Type W	*Type WX*		
W1 and WX1	2·3	6	0·25	0·1	15	Unlimited
W2 and WX2	4·6	12	0·25	0·1	30	Unlimited
W3 and WX3	6·9	18	0·25	0·1	45	Unlimited
W4 and WX4	9·2	24	0·25	0·1	60	Unlimited
W6 and WX6	13·8	36	0·25	0·1	90	Unlimited
W9 and WX9	20·7	54	0·25	0·1	135	Unlimited
W12 and WX12	27·6	72	0·25	0·1	180	Unlimited
W15 and WX15	34·5	90	0·25	0·1	225	Unlimited

* Preferred types only are listed, but intermediate types having any required number of elements up to fifteen are available.

A. H. B. W.

GERMANIUM AND SILICON POWER RECTIFIERS

An ideal rectifier would offer zero impedance to current in one direction and infinite impedance in the other. A germanium or silicon power rectifier approximates to this definition; its forward resistance is very low and its reverse resistance extremely high. A single germanium cell may drop only 0·5 volt to a forward current of 100 amps., yet pass only a few milliamperes when blocking a reverse voltage of 100 volts; in consequence the power loss is small and the rectifying efficiency high.

As with selenium and copper oxide rectifiers, rectification takes place at a junction between dissimilar materials, but the junction of a germanium or silicon rectifier exists inside a single crystal of the semi-conductor. This difference results in current densities of 100 amps./sq. cm. for germanium, compared with 0·25 amps./sq. cm. for selenium.

Because of their small size, high efficiency and low heat dissipation, semi-conductor power rectifiers offer considerable advantages over the equivalent thermionic-valve or dry-plate rectifiers. They do not require heater supplies, and the forward resistance does not deteriorate with age. The voltage drop across the rectifier is usually less than 10 per cent of that across a valve or dry-plate rectifier.

A wide range of germanium and silicon power rectifiers is now available. Generally, germanium junction rectifiers which have a lower forward resistance are used for high currents, and silicon rectifiers for applications of under about 1 amp. A typical silicon rectifier for 200 mA might be the size of a quarter-watt resistor.

Semi-conductor rectifier failures are generally caused by voltage surges, which are most likely to occur when the circuit is switched on or off. It is therefore essential to make sure that any voltage surge does not exceed the maximum permissible peak inverse voltage. Manufacturers usually recommend that if the values of the voltage surges are not known, an attempt should be made to measure them, using measuring equipment with input impedance high compared to that of the rectifier, as otherwise the surge voltage may be damped. Suitable capacitance or resistance should then be connected across the rectifier to ensure that the surges never exceed the maximum permissible peak inverse voltage. The values of these components will depend on the nature of the circuit and on the loss of efficiency that can be tolerated.

Series operation of several rectifiers of the same type may usually be arranged by reducing the peak inverse voltage and shunting each rectifier with an equalizing resistance. Parallel operation of rectifiers is not recommended where a single rectifier of higher current rating can be used: should it be necessary to use rectifiers in parallel, it will generally be desirable to reduce the current rating of each rectifier by about 30 per cent to avoid overloading an individual rectifier.

Miniature silicon rectifiers are hermetically sealed and often are wire ended for direct connection into circuit; care should be taken not to overheat the body when soldering nor to bend the wires too close to the rectifier. Germanium power rectifiers are often intended to be used with cooling fins, which will often allow an increase in permissible mean forward current of the order of 100 per cent provided a good thermal contact is made between the base of the rectifier and the cooling fin.

RIPPLE FILTERS

Rectification of alternating currents is one of the most important methods of producing direct currents, whether it is by mechanical means (rotating commutators and, to a much lesser extent, by vibrators) or electronic ones (metal and valve rectifiers).

Unfortunately, for many of the usual communication applications the rhythmic voltage variations during commutation can rarely be accepted directly from the D.C. source. Consequently, some means of correcting them must be adopted. The most important of these is the ripple filter. It is, of course, possible to correct for ripple voltages using valve circuits, but usually such devices are likely to be expensive in power and components compared with the normal resistance/inductance/capacitance types of filter, although they also have their uses in modern techniques.

In general, "ripple" may be classified under two major headings:

(i) that due to repetitive cycles of frequencies related to the fundamental;

(ii) that due to transient damped waves.

The first is of primary importance because it usually needs most attention from a power viewpoint: the second is usually much less significant (except in the case of wide-band circuits), and is generally caused by commutation deficiencies primarily due to circuit capacitance and inductance. These are impossible to remove because self-capacitance and leakage inductance arise naturally in most power circuits. Their inherent oscillatory tendencies are usually eliminated or, at least,

TABLE 10.—RIPPLE FREQUENCIES AND MAGNITUDES

Circuit	*Fundamental Ripple Frequency (f = supply frequency)*	*Percentage Peak Ripple Magnitude in Terms of D.C. Component*
Single-phase, half-wave	f	157
Single-phase, bi-phase full-wave	2f	68
Single-phase full-wave bridge	2f	68
Three-phase, half-wave: (i) star; (ii) inter-connected star or zig-zag	3f	26
Two-phase, double bi-phase	4f	14
Double two-phase with balancing coil	4f	14
Three-phase, full-wave bridge	6f	6
Double three-phase with balancing coil	6f	6
Three-phase interconnected star or zig-zag	6f	6
Six-phase, half-wave	6f	6
Three-phase T/eight phase	8f	3·4
Triple three-phase, half-wave with balancing coil	9f	2·6
Twelve-phase, half-wave	12f	1·5

minimized to a negligible degree by the use of suitable capacitors and resistances.

It should be mentioned that non-linear devices such as valve and metal rectifiers, assymetric crystals, etc., are often used successfully for these purposes.

Initial Ripple

The initial ripple of a rectifier is that produced at its output.

This ripple voltage contains a number of harmonics which, in general, contribute little to the total magnitude and are readily attenuated by the use of a suitable ripple filter. Before examining the method of filtering the ripple, it is worth while noting their magnitudes immediately after rectification. Table 10 gives this information for the more common rectifier circuits used in transmitter practice.

The Low-pass Filter

This is a network which passes freely only frequencies below the cut-off value. An example of a filter of this type is shown in Fig. 30. This may be split into equal sections, separated by dotted lines as in Fig. 31, forming a T-section filter.

For such a circuit the radian cut-off frequency ($2\pi f_c$) for resistanceless elements is given by $\omega_c = \frac{2}{\sqrt{LC}}$. When ω ($2\pi f$) exceeds ω_c, the higher frequencies become rapidly attenuated.

In practice, as the resistance of the inductor elements is finite, its effect on the circuit performance shows itself in two ways :

(i) The higher the effective resistance value of the choke, the greater the attenuation in the pass range. This is undesirable, hence the resistance value is kept as low as possible to maintain a high Q-value, $\left(\frac{\omega L}{R}\right)$.

(ii) The higher the effective value of resistance, the more rounded becomes the sharp corner at ω_c, as indicated in Fig. 32.

To ensure complete absence of reflections on relatively short lines, it is essential to close the line in its characteristic impedance Z_0, so making it behave as a line of infinite length.

From a simple ripple filter viewpoint, however, it is rarely possible to

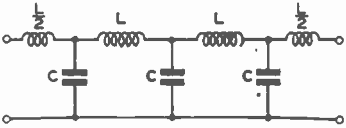


Fig. 30.—Basic Low-pass Filter Circuit.

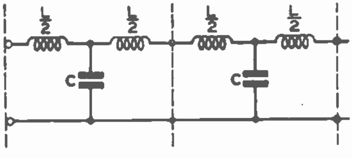


Fig. 31.—T-Section Low-pass Filter Circuit.

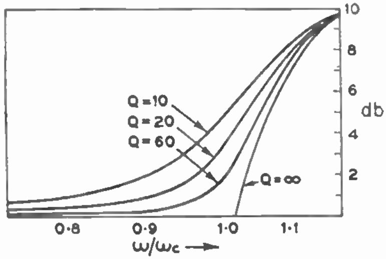


FIG. 32.—ATTENUATION CURVE OF A LOW-PASS FILTER.

close the circuit on its characteristic impedance. To simplify the circuit, most ripple filters omit the first half-section, without serious detriment to its ripple-reduction characteristics.

Divider Networks

From a general consideration of ripple filters, the circuit is basically a divider network in which a high impedance is connected in series with the low-impedance section, as shown in Fig. 33. Thus the relative reduction of the input signal, or Ripple Reduction Ratio is :

$$\frac{Z_1}{Z_1 + Z_2} = \frac{1}{k}$$

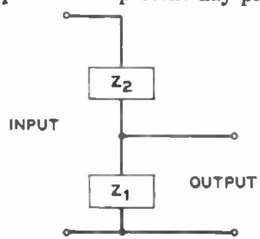
where k is the ripple reduction factor.

For two such systems in tandem (Fig. 34) the Ripple Reduction Ratio becomes :

$$\frac{Z_1}{Z_1 + Z_2} \times \frac{Z_3}{Z_3 + Z_4} = \frac{1}{k_1} \times \frac{1}{k_2}$$

For two identical systems the ratio is $1/k^2$, hence for n such systems it is $1/k^n$. It would appear that many small systems of L–C " links " would be more satisfactory than one or a few ; in practice this is not necessarily so, as will be seen later under " Inductance/Capacitance Filters ".

In the above expressions the impedance may be resistive, reactive or capacitative, or a combination of all three, thus giving the three basic types used in present-day practice.



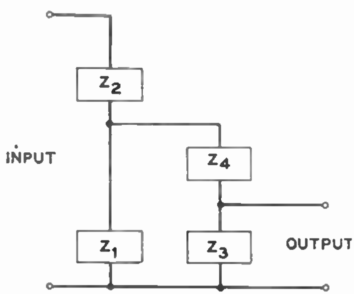


FIG. 33 (*above*).—DIVIDER NETWORK, FIRST STAGE.

FIG. 34 (*right*).—SECOND STAGE.

Ripple Filter Types

The three main types are as follows:

(i) *Resistance/Inductance Filters.*—These are used when the capacitative section is prohibitively large, from a mechanical, electrical or economic aspect.

(ii) *Capacitative/Resistance Filters.*—Used for low-current conditions where the D.C. voltage drop and power loss are unimportant.

(iii) *Inductance/Capacitance Filters.*—Used for all power requirements where overall efficiency and load regulation are important.

Resistance/Inductance Filters

This class of filter utilizes the resistance of the load as the low-impedance element, with the inductance of a series choke as the high-impedance arm. This arrangement fills an important role in connection with valve-filament heating, where the valve current is of a high order and the voltage relatively low. In these cases the ripple requirement is usually of the order of 1 per cent of the D.C. voltage value.

For currents exceeding 20 amperes, three-phase supplies are usually adopted to give the advantage of low fundamental ripple as illustrated in the following typical example:

EXAMPLE.—*Filament rating:* 6 *volts* 200 *amperes. Supply: three-phase* 50 *c/s. Rectifier: selenium in three-phase bridge circuit. Basic ripple:* 6 *per cent at* 300 *c/s. Smoothing choke:* 0·18 *mH.*

From the above data the "hot" resistance of the valve is 6 volts 200 amperes, which is 0·03 ohms.

Impedance of choke $\omega L = 2\pi \times 300 \times 0{\cdot}18 \times 10^{-3} = 0{\cdot}337$ ohms; hence the ripple reduction ratio is $\dfrac{0{\cdot}03}{0{\cdot}337} = \dfrac{1}{11{\cdot}2}$ and the ripple percentage $= \dfrac{6}{11{\cdot}2}$ or 0·54 per cent.

To make any appreciable difference using capacitor filtering, the impedance of the capacitor element must be of the order of one-fifth that of the valve "hot" filament resistance.

Thus in the above example, the capacitor impedance would be of the order of 0·006 ohm, i.e., $\dfrac{1}{\omega C} = 0{\cdot}006$ ohm.

Hence
$$C = \frac{10^6}{\omega \times 0{\cdot}006}$$
$$= \frac{10^6}{1{\cdot}885 \times 10^3 \times 6 \times 10^{-3}}$$
$$= \frac{10^6}{11{\cdot}31} = 89{,}000\ \mu\text{F}$$

which is too large from a mechanical and cost aspect for normal filament filtering. In many instances the cost of the capacitors would be much greater than that of the choke.

In general, the addition of capacitors under the conditions outlined above is unusual in practice.

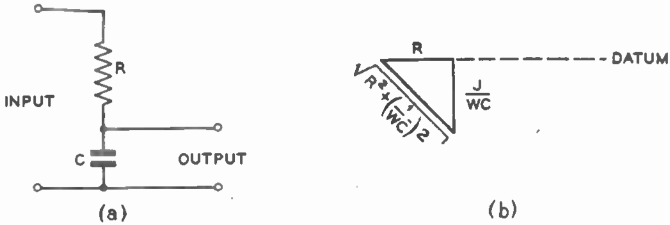


FIG. 35.—RESISTANCE/CAPACITANCE FILTER: (*a*) CIRCUIT; (*b*) VECTOR.

Resistance/Capacitance Filters

This system utilizes resistance as the high-impedance arm and capacitance as the low-impedance one, as shown in Fig. 35 (*a*).

Since the capacitative reactance is 90° out of phase with the resistive element (see Fig. 35 (*b*)), the estimation of ripple ratio must involve the correct vector relationship. Hence for this circuit, the magnitude of the ripple reduction ratio will be :

$$\frac{\frac{1}{\omega C}}{\sqrt{R^2+\left(\frac{1}{\omega C}\right)^2}}=\frac{1}{k}$$

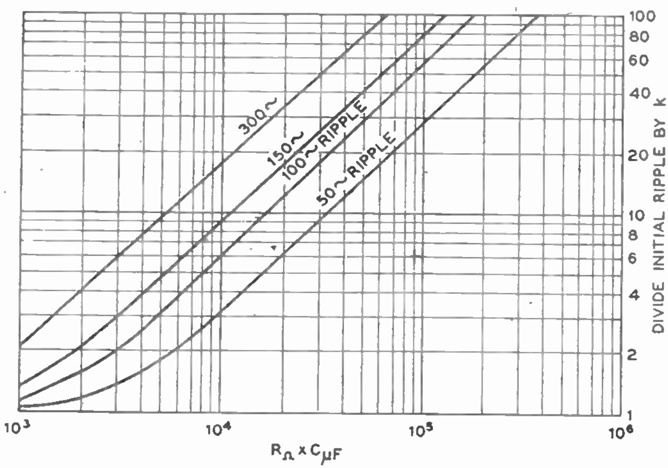
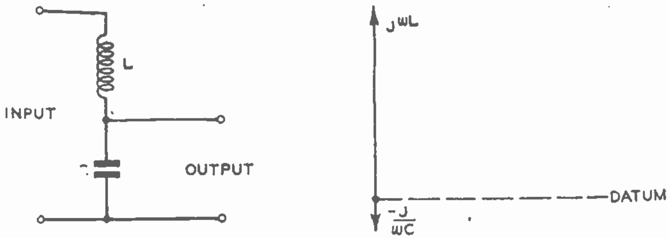


FIG. 36.—RIPPLE REDUCTION FACTOR FOR A SINGLE RESISTANCE/CAPACITANCE FILTER.



(a) (b)

FIG. 37.—INDUCTANCE CAPACITANCE FILTER: a) CIRCUIT; (b) VECTOR.

For ratios less than $\frac{1}{3}$, the approximation $\dfrac{\frac{1}{\omega C}}{R} = \dfrac{1}{\omega CR}$ may be used without serious error. For values greater than $\frac{1}{3}$, however, the full expression must be used. Fig. 36 shows the ripple-reduction factor, k, for a single-stage resistance/capacitance filter.

Inductance/Capacitance Filters

For this type of filter the inductance is the high-impedance arm, with the capacitance as the low-impedance member, as shown in Fig. 37 (*a*).

From the above vector relationship (Fig. 37 (*b*)) it is clear that the net impedance to A.C. is given by $\omega L - \frac{1}{\omega C}$, and since $\frac{1}{\omega C}$ is across the output circuit, the magnitude of the effective ripple reduction ratio is given by :

$$\frac{\frac{1}{\omega C}}{\omega L - \frac{1}{\omega C}} = \frac{1}{\omega^2 LC - 1} = \frac{1}{\left(\frac{X_L}{X_C} - 1\right)}$$

which indicates that $\frac{X_L}{X_C}$ must exceed 2 to make any contribution to the ripple-reduction ratio. So that if $\frac{X_L}{X_C} = 10$, then the ripple reduction ratio $= \frac{1}{9}$, i.e., $k = 9$.

"Link Resonance"

As indicated under "Divider Networks", it appears that the larger the number of sections for a given electrical specification, the greater the ripple-reduction ratio. In practice, other limiting factors come into play, the major one being the "link resonance", which is a governing factor in many applications. Fig. 38 shows the "link" diagram:

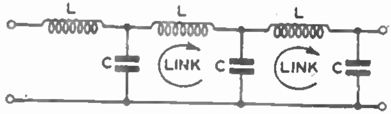


FIG. 38.—"LINK" DIAGRAM.

It is preferable for this resonance to be outside the working frequency of the system. For instance, if the minimum working frequency of the system is 50 c/s, the filter should not resonate above this figure. In general, the "link" resonant frequency should be 15–20 per cent less, i.e., 42·5–40 c/s. It should be remembered that effective capacitance is that due to the series value round the link, i.e., if both capacitances are each C, then the effective capacitance C', is $C/2$. Fig. 39 shows the relationship between resonant frequency f and the product of L (henrys) and C (microfarads).

Although it is possible to give precise figures for inductance (L) and capacitance (C), it is worth while bearing in mind that the value of L may vary between ± 10 per cent and C vary between ± 20 per cent of the nominal values, which would thus give the following excursions of "link" resonant frequency from the nominal value.

EXAMPLE.—*A ripple filter to operate over the range of 50–3,500 c/s comprises two stages of* L = 6 *henrys and* C = 4 μF. *Assuming* L *to be*

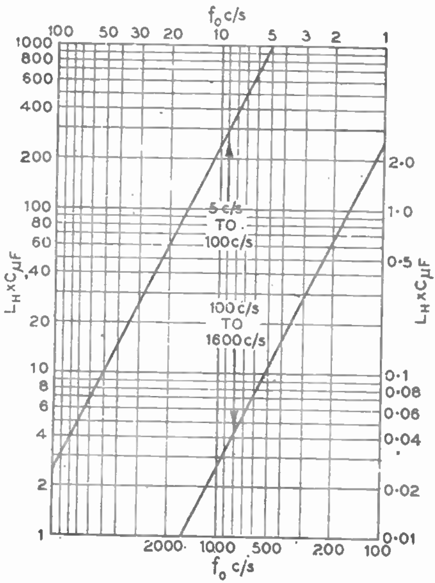


FIG. 39.—RESONANCE CHART LC/f.

within ±10 per cent and C to be within ±20 per cent, then three link resonances, as given in Table 11, could occur on top, normal and low element values.

TABLE 11.—EFFECT OF ELEMENT VARIATIONS ON LINK RESONANCES

	(1) *Top Element Values*	(2) *Normal Element Values*	(3) *Low Element Values*
L (henrys)	6·6	6·0	5·4
C (μF)	4·8	4·0	3·2
C_1 (μF)	2·4	2·0	1·6
LC_1 (μFH)	15·85	12	8·62
$\sqrt{LC_1}$	3·98	3·465	2·94
$\omega_0 = \dfrac{10^3}{\sqrt{LC_1}}$ (radians/sec)	252	289	340
$f_0 = \dfrac{\omega_0}{2\pi}$ (c/s)	40	46	54·2

This shows that cases (1) and (2) are well outside the working frequency range, but case (3) indicates that, under suitable conditions, resonance could occur, and consequently the values of L or C would have to be increased to avoid this condition.

Fig. 40 gives the ripple reduction factors for one stage of L–C filtering.

Filters for D.C. Generators

In some instances it is necessary to use D.C. generators for power supplies for communication purposes, particularly on telephone networks, where the ripple applied to the circuit must be of a very low order.

Before considering this problem in detail, it is worth while examining the fundamental ripples inherently produced in D.C. machines. There are three main sources :

(i) the number of poles, resulting in pole frequencies varying between 25 and 100 c/s, dependent upon the speed of the machine :

(ii) the slot arrangement, producing slot frequencies of the order of 500–1,000 c/s, depending upon the number of armature slots and the machine speed ;

(iii) the commutator, producing commutator ripple, depending upon the number of commutator segments and the machine speed, producing frequencies varying from 1,000 to 3,000 c/s. for a well-maintained commutator.

For a badly maintained commutator, the frequency spectrum may extend well into the ultra-high-frequency region, with consequent interference with telephone, radio and television reception.

The slot frequency is usually of greatest importance because of its medium-frequency value and relative magnitude.

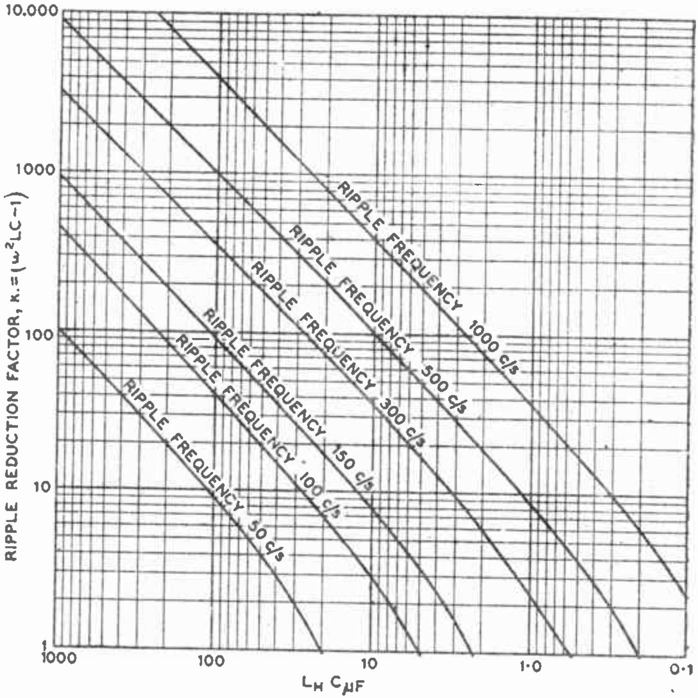


FIG. 40.—RIPPLE REDUCTION FACTORS FOR ONE STAGE OF INDUCTANCE/ CAPACITANCE FILTERING.

In many cases the commutator-segment frequency is much higher and relatively smaller in amplitude; hence its effect is negligible and may be handled by the filter for slot frequencies. If the commutator-segment frequency approximates that of the slot frequency, then it, too, would be dealt with in the same manner.

Should the double-pole frequency fall in an important part of the frequency spectrum, from the communication aspect, then it would necessitate larger component values to attenuate the lower frequencies subject to the "weighting" correction given in Figs. 43 or 44. In view of the foregoing remarks, it is necessary to filter the output of the generator to attenuate the ripple to a level which will not cause interference within the operating frequency range of the system.

For large machines the upper limit of ripple is of the order of 1 mV at 800 c/s for a 50-volt machine, i.e., a ripple-reduction ratio of 1/50,000 or 94 db below 1 mV. To obtain such low values of ripple-reduction ratio, it becomes necessary to use at least two filter sections. Fig. 41

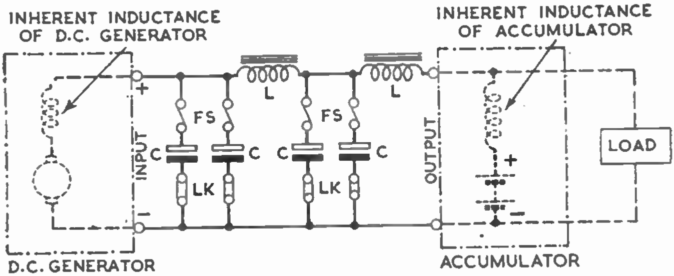


Fig. 41.—Machine Filter.

shows the basic diagram, in which it will be noted that the following precautions are taken :

(i) the total capacitance is divided into convenient blocks, usually not exceeding 1,000 μF (where the total capacitance is 1,000 μF or less the blocks are divided into two or more sections) ;

(ii) each capacitor is suitably fused to limit the power dissipated under fault conditions ;

(iii) isolating links are fitted so that the capacitor element may be completely disconnected for testing or removal.

Effect of Audio-frequency Disturbances

As indicated previously, the ripples generated range from very low to high audio frequencies, their magnitude having a large bearing on the degree of attenuation necessary. Before examining the problem in

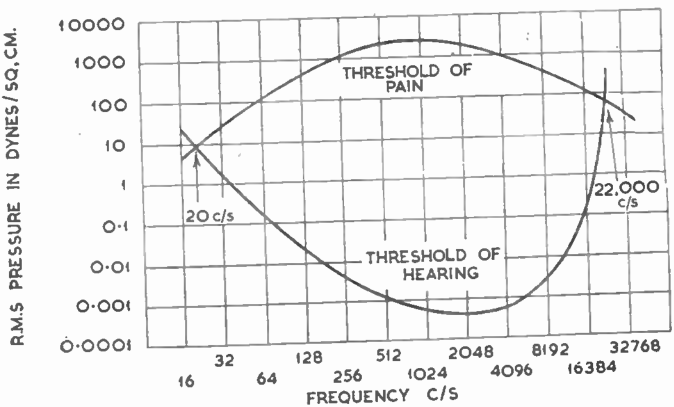


Fig. 42.—Threshold Curves.

greater detail, a brief outline of the behaviour of the human ear is desirable.

Frequency Characteristics of the Human Ear

The human ear, although quite a sensitive organ, does not respond in a linear manner to sound pressure: it is least sensitive at each end of its frequency range, where the threshold of hearing meets the threshold of pain, which normally ccurs in the region of 20 c/s and 22,000 c/s respectively. Fig. 42 shows a curve of this relationship.

There are, of course, many people who have defective hearing due to illness or occupational deafness, the latter including boilermakers, riveters and kindred trades, in which case the curves of Fig. 42 become modified.

From a broadcasting-service aspect the frequency range transmitted is between 30 and 12,000 c/s, which is well within the range given in Fig. 42. For commercial speech—utilizing telephone lines, line apparatus and radio equipment—the frequency band ranges from 200 to 3,500 c/s. From Fig. 42 it may be seen, for example, that at a frequency of 200 c/s the ear requires a sound pressure approximately eleven times greater than that at 800 c/s to be equally effective at the threshold of hearing. Consequently, the level of the disturbance at 200 c/s may be eleven times greater for an equal interference level. This large variation led to the investigations which resulted in the adoption of a recognized noise meter.

C.C.I.F. Noise Meter

In 1934 the C.C.I.F. recognized the desirability of having an interference or noise-meter, which took into account the great variation of pressure sensitivity of the human ear over the audio-frequency range, and decided that some attempt must be made to rationalize the position regarding its effect on international communication systems.

The result of the 1934 meeting was the definition of a new meter to give permissible variations of pressure sensitivity in terms of an 800-c/s current into a non-inductive resistance of 600 ohms. Thus it became possible to measure the effect of the interfering signal in terms of the 800-c/s standard. (At a subsequent meeting it was decided to include another network to cater for broadcast systems in which the datum was 1,000 c/s.)

The Psophometer

The instrument, originated by the British Post Office and accepted by the C.C.I.F., is called a Psophometer, and contains a network associated with an indicating instrument.

The function of this network is to "weight" every frequency in accordance with its interference value (related to a frequency of 800 c/s).

Weighting Curves

Fig. 43 shows a curve of the "weighting" constant, k, based on the C.C.I.F. recommendations for a broadcast system. Fig. 44 gives the relationship for a commercial telephone system.

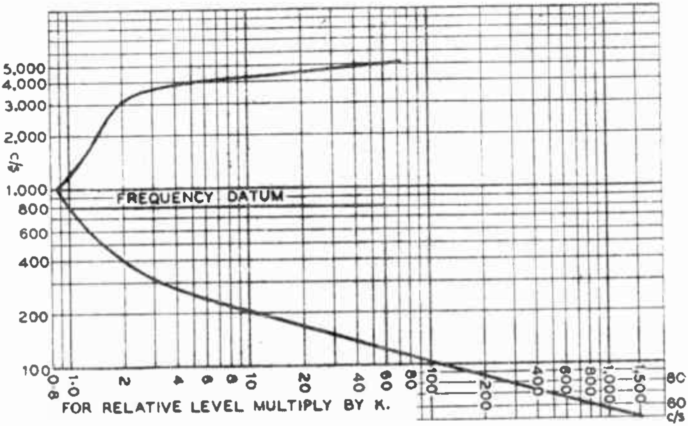


Fig. 43.—C.C.I.F. "Weighting" Curve for Broadcasting System.

For a commercial telephone circuit, the peak response frequencies are in the neighbourhood of 1,000–1,200 c/s. Consequently, any signal within this range will show greatest response for a given input level, therefore any interfering signal would also have greatest effect. Thus it will be noted that both the ear and the system have greatest effect in

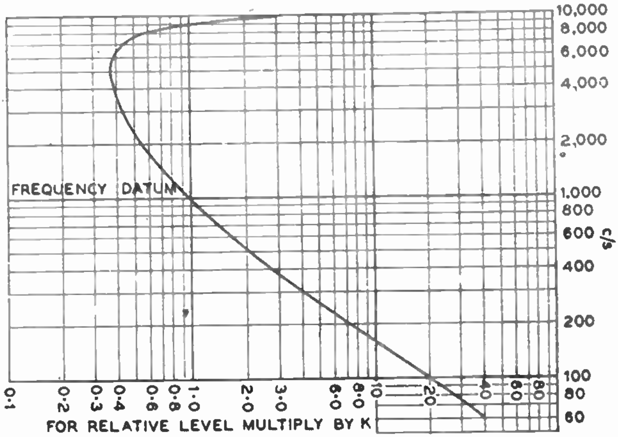


Fig. 44.—C.C.I.F. "Weighting" Curve for Communication System.

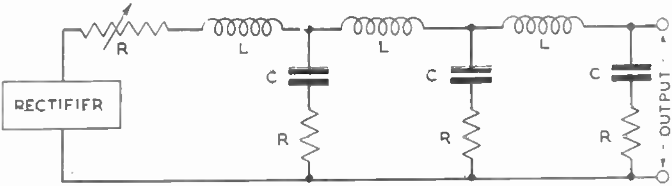


FIG. 45.—" BUILT-OUT " FILTER.

the same frequency region. Hence the necessity of the weighting assessment of the whole transmission.

Assessment of Audio-frequency Noise

Considering this problem in terms of an interfering frequency of 100 c/s, and supposing the limiting signal for the standard is 1 mV at 800 c/s, then from the " weighting curve ", the appropriate value of k (for 100 c/s) is chosen and the effective level of the interference signal determined from the relationship, 2 mV $\times$ k, where $k = 112{\cdot}2$; hence the equivalent audio-frequency interference level would be $2 \times 112{\cdot}2 = 224{\cdot}2$ mV. Thus the correct limitation of interfering signal may be applied via the filter system.

Smoothing Circuits for Television Service

In considering the smoothing of H.T. supplies for circuits passing a series of rapidly changing pulses, it is worth while noting two major conditions under which such supplies may be operated :

(i) To present a very low impedance to the load over a wide frequency band so that the load regulation is negligible. This may be achieved by using a series type stabilized power supply in which feedback conditions are applied to control the circuit impedance over a wide range of load and frequency.

(ii) To make the circuit behave as a constant resistance, whatever the load. This may be done by (*a*) using a shunt valve stabilizer and (*b*) by " building out " the smoothing circuit.

" Built-out " Filters

For television modulator H.T. circuits the load varies with the picture signal, and, as it is essential to avoid reflections in them, constant-resistance networks are used.

Fig. 45 shows one type commonly used in practice.

For this circuit $R^2 = \dfrac{L_{(\text{henrys})}}{C_{(\text{farads})}}$ or $R_{(\text{ohms})} = \sqrt{\dfrac{L_{(\text{henrys})}}{C_{(\text{farads})}}}$.

Table 12 gives relative values of R and C for an inductance of 0·5 henry where $C_{(\mu F)} = \sqrt{\dfrac{L_H}{R_\Omega^2}} \times 10^6$.

TABLE 12.—RELATIVE VALUES OF R AND C

R	20 ohms	50 ohms	70·7 ohms	100 ohms
C	1,250 μF	200 μF	100 μF	50 μF

It is clear from Table 12 that the lower the value of "built-out" resistance R, the greater is the capacitance, size and cost of the capacitor. Thus the resistance becomes a major factor in determining the value of capacitance to be used in circuits handling high powers. From the input-power aspect, however, the larger the value of R, the greater the voltage input for a given current rating, and consequently, the lower the overall efficiency of the circuit. A compromise between power efficiency and electrical expediency is usually necessary to obtain a satisfactory arrangement: hence R = 100 ohms and C = 50 μF is a reasonable solution for circuits handling power of the order of 100 kW D.C. (From the ripple-reduction ratio viewpoint, the value of $1/k$ for a three-stage circuit of this type is of the order of 1/800 for the 300-c/s omponent, and $\frac{1}{8}$ for the 50-c/s one.)

Since the value of R chosen is usually above 50 ohms, the rectifier circuit resistance, including the transformer primary (referred to the secondary) choke and other resistances must be "padded out" to approach the value of R. Therefore, if soft rectifiers are used, additional resistance must be included in series with the rectifier, either lumped in the D.C. output or in each valve anode, the latter being preferable, as it minimizes serious damage should the valve backfire.

Hard valves may be used in these circuits, providing their effective resistance can be held within the R value.

From the viewpoint of general convenience and expediency, it is obvious that if the "built-out" resistance can be incorporated within the rectifier, a much improved arrangement will result. One such form is the selenium metal rectifier, which is very suitable for this service, although it is important to ensure that the inherent rectifier resistance is slightly below the "built-out" value after the rectifier has aged.

The difference between the rectifier resistance and the built-out resistance value (some 10–20 per cent of R) is made up by using a variable non-inductive unit.

To allow for the variation of the resistance of individual rectifier arms, it may be necessary to include a balancing resistance in each. This will

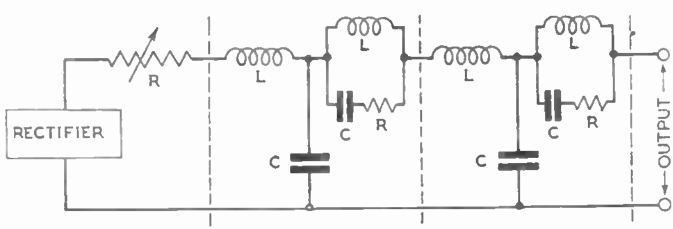


FIG. 46.—IMPROVED FORM OF "BUILT-OUT" FILTER.

minimize the increase of fundamental ripple occasioned by the resistance being out of balance. The "built-out" filter, shown in Fig. 46 (for which $R^2 = \frac{2L}{C}$), is an improvement on that shown in Fig. 45. This is because the series-resistance elements are only included in the series arm, i.e., the capacitors in shunt across the supply are of low impedance, and therefore are of greater use from a pure ripple-reduction aspect.

Typical Ripple Figures for Radio Transmitter Equipment

One of the guiding requirements on the ripple-reduction figures for these filters is the subsequent degree of amplification the signal receives, and whether any cathode-follower assistance is obtained from the circuit (this effect is of great value, because it results in a reduction proportional to $1/(1 + \mu)$). In general, the earlier the stage in a given group of circuits, the greater must be the attenuation to the unwanted signals applied via the smoothing circuit. Thus the ripple permissible for a microphone pre-amplifier would be extremely low, because of the high gain necessary before the final power stage is reached.

On the higher-powered stages the amplification is usually less, and consequently the degree of ripple reduction for a general level of unwanted ripple noise is less. Table 13 gives ripple figures for circuits used in radio and telephone-line service.

TABLE 13.—TYPICAL PERMISSIBLE RIPPLE VALUES

Ref.	*Circuit*	*Ripple Value Relative to D.C. Level*	
		Percentage	*Decibels*
1	Microphone pre-amplifier	0·0005–0·001	106–100
2	Microphone		
3	Studio amplifier H.T.		
4	Telephone speech circuits		
5	1st audio-frequency amplifier	0·002–0·005	94–86
6	Modulator and sub-modulator G.B.		
7	H.T. for modulator and all broadcast service amplifiers, including final stage television transmitter H.T.	0·01–0·005	80–86
8	H.T. for C.W. transmitters	0·1–1	60–40
9	Low-voltage high-current heater circuits		

Chokes for Smoothing Filters

Chokes for this service may be classified under two headings :

(i) Those for nominal "constant" inductance applications, where the variation of inductance with D.C. is relatively small, of the order of ± 15 per cent.

(ii) Those for "swinging" choke service, where the required variation of inductance is very wide. This is brought about by the use of partial magnetic saturation at high D.C. values. The range of inductance variation may be as high as ten to one. The object of the high inductance at relatively low values of D.C. is to prevent the output voltage "soaring" to the peak of the A.C. voltage wave as the D.C. tends to zero. By this scheme good load-regulation characteristics are maintained over a much wider range of load current than would otherwise be possible.

"Swinging" chokes are used mostly on low-power circuits.

Testing of Chokes

Unless the choke performance is within its specified tolerances, the ripple-reduction ratio of the smoothing filter cannot be attained; consequently choke testing is of considerable importance to ensure this condition. The main tests applied to chokes are as follows:

(i) D.C. resistance;
(ii) inductance;
(iii) voltage proof tests;
(iv) heat run.

These items will now be considered in greater detail under the above item references.

(i) *Direct Current Resistance.*—It is important to ensure that the D.C. voltage drop is within the specified value. Variations in the size of wire for a given nominal gauge, the interleaving insulation and the tension of the wire during winding all contribute to the variations about the nominal value. The usual manufacturing tolerance is of the order of ± 15 per cent, although in some cases ± 10 per cent is specified.

(ii) *Inductance.*—The inductance of chokes for ripple-filter applications is a function of the incremental permeability of the iron forming the core, hence the A.C. voltage under test conditions is of fundamental importance, as shown in Fig. 47. If the choke is first in the circuit, then the majority of the ripple voltage will appear across it. In the case of a single-phase full-wave rectifier the peak voltage across the choke is approximately 68 per cent of the D.C. value, thus producing a large cyclic change of flux density, which tends to give a larger value of effective inductance than when a very low ripple voltage is applied.

To ensure the correct performance of the choke, the specified A.C. ripple voltage must be applied along with the correct D.C. value.

(iii) *Voltage Proof Tests.*—Under this heading the voltage across the choke must be specified to enable the correct test voltage to be applied. For instance, if a choke is connected to the output of a 10,000-volt rectifier on a full-wave bi-phase circuit, this winding must be insulated to operate at 6,800 volts peak across the coil: a very different condition for a circuit operating at, say, 500 volts D.C., where the peak ripple voltage would be 340 volts. In the first case the creepage voltage per inch would be some twenty times greater than for the second case (for windings of equal

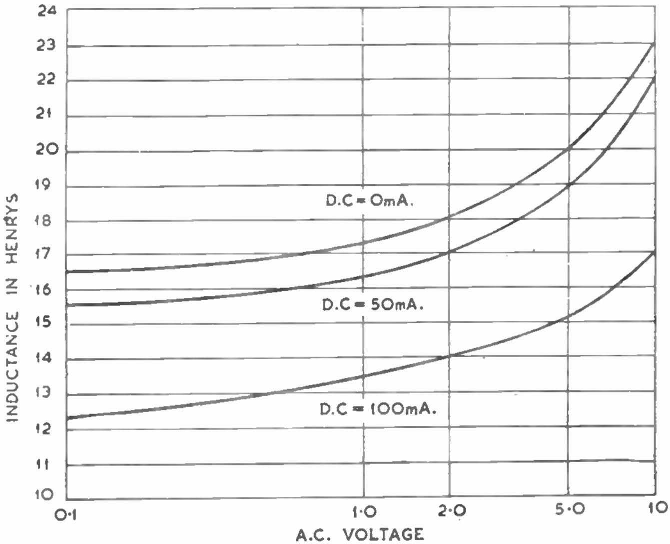


FIG. 47.—EFFECT OF INPUT VOLTAGE ON INDUCTANCE.

mechanical size), thus necessitating greater care in design to eliminate ionization hazards which might occur under these conditions.

In some applications, providing the insulation over the winding is adequate, the coil could be centre-tapped and connected to the core to minimize the voltage stress between the winding and the core. In this case the whole of the choke structure should be mounted on insulators suitable for the required test voltage. It is essential that the ionization level is not exceeded under any test or working condition.

There are a number of proprietary designs of A.C. and D.C. H.T. insulation testers with audio-ionization indicators which simplify tests of this type.

(iv) *Heat Run.*—The temperature rise of the choke is of great importance, and must be safely inside the specified temperature tolerance. Where temperature rise exceeds 50° C., the insulation on the coil and general insulation overall should be reviewed and changed to glass fabric, or the newer plastics, nylon and terylene, and high-temperature varnishes, as circumstances dictate.

Measurement of Inductance

There are a number of circuits suitable for the measurement of inductance, but the Owen, or one of its variants, is often used. This

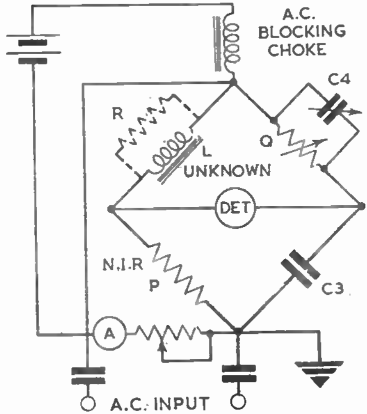


FIG. 48.—OWEN BRIDGE.

bridge may be arranged to give the elements as series or parallel components of inductance and resistance.

To eliminate the human element as much as possible, the balance of the bridge may be facilitated by the use of an amplifier (which may be tuned or not, as required) and an indicator in the form of a tuneon or microammeter. To obtain a fair degree of accuracy, good magnetic and electric screening is essential on the bridge components and wiring. Should the inductive and capacitative residuals of these not be small, large errors are likely to arise on low- and high-inductance determinations.

Apart from the above causes of error, the bridge balance is often difficult if: (i) the A.C. source contains harmonics; (ii) the iron-cored component itself produces them, particularly when using telephones.

In general, the flux densities used for chokes are such that the initial straight part of the magnetization curve is utilized. Fig. 48 shows the usual form of bridge for relatively low-voltage test conditions. When the A.C. voltage across the choke is large the circuit shown in Fig. 49 is used. This circuit avoids the difficulties in obtaining decade resistances and capacitances which can be operated at high voltages. The arms of the bridge are separated, and the difference of voltages between the two parts is compensated for by suitable tappings along the ratio arm.

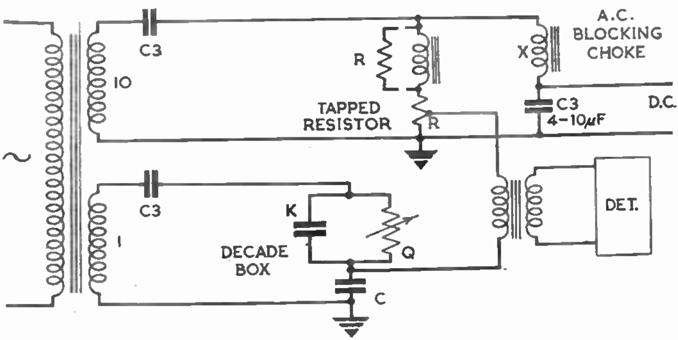


FIG. 49.—OWEN BRIDGE FOR HIGHER VOLTAGES.

These bridges have the advantage that they are independent of frequency. The bridge will balance when $R = \frac{PC_3}{C_4}$ and $L = PQC_3$. With fully screened components, accuracies of the order of 1 per cent may be obtained on inductances up to 1 henry, whilst 2½ per cent may be obtained for inductances from 1 to 150 henrys.

The value of the A.C. blocking-choke inductance must be at least ten times that of the choke to be tested, to minimize its shunting effect on the bridge. To avoid heating troubles on the blocking choke, the current through it should be kept on only for a time sufficient to make a balance.

R. H. B.

MERCURY ARC RECTIFIERS

Prior to 1932 the high voltage D.C. power required by large radio broadcast transmitters was obtained from motor generator sets. However, some three years before this time the Marconi's Wireless Telegraph Company Ltd. had carried out experiments with high-voltage pumped steel-tank rectifiers. In 1932 the British Broadcasting Corporation decided to incorporate 20-kV 600-kW steel-tank rectifiers in the new 1,500-metre Droitwich station. Whilst at this time the relatively untried steel-tank rectifier could hardly be said to be as reliable as the motor generator set, its approximate efficiency of 96 per cent, compared to the 85 per cent of the motor generator set, was no doubt the deciding factor.

With its advantages of voltage control, high-speed circuit protection and low internal impedance (particularly suiting it to supply Class B modulated transmitters) further recommending its use, the steel-tank rectifier, once established, soon superseded the motor generator set. During the past decade, however, the steel-tank rectifier has in its turn been almost completely supplanted for this duty by the high-voltage, single-anode rectifier called the Excitron.

Steel-tank Rectifiers

It is a sign of the rapid progress that has recently been made in the field of high-voltage rectification that in 1945 the steel-tank rectifier, having since 1932 given ample proof of its reliability, was looked upon as the type of plant that would be used for many years to come. The latest type tanks, which had gradually evolved from the water-cooled, pumped units first used, were then—and still are—the sealed air-cooled units, and it was anticipated that any future developments would take place along these lines. Today, as mentioned above, the high-voltage steel tank is virtually obsolete for radio applications.

A typical high-voltage sealed steel-tank rectifier is shown in Fig. 1. The bulb is of the air-cooled, side-arm type, fitted with two grids, the upper a control grid and the lower an energized grid, which, in view of the long narrow arms, is required to assist the natural commutative action of the rectifier. The photograph also illustrates the normal method of installing the tank and its associated equipment. On the right-hand side can be seen the control cubicle, in which are mounted the grid control, arc suppression, excitation and ignition auxiliary circuits. The whole equipment is contained in an enclosure measuring about 15 ft. by 13 ft., and having access doors interlocked with the A.C. breaker and rectifier earthing switch. Behind the enclosure in a cell

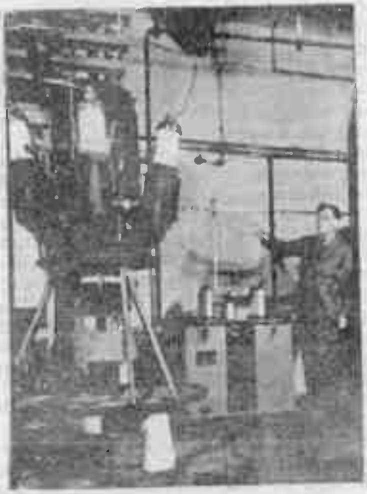


FIG. 1 (*above*).—A 500 KW, 12,500-VOLT HIGH VOLTAGE SEALED STEEL TANK RECTIFIER ENCLOSURE IN A TRANSMITTING STATION.

FIG. 2 (*right*).—TYPE AR.64 SINGLE ANODE RECTIFIER.

(*Photographs reproduced by kind permission of the English Electric Co. Ltd.*)

is mounted the rectifier transformer, the latter being normally connected either in six-phase fork or six-phase double star.

High-voltage Single-anode Rectifiers—Excitrons

While the steel-tank rectifier had proved its worth for large equipments, it was, in fact, expensive for all but the larger-powered transmitters. In view of this fact, work was started around 1947 to develop a mercury-pool rectifier that would provide an alternative at the smaller ratings to the hitherto normally used hot-cathode tubes. The outcome of this work was the single-anode rectifier, named the Excitron. The first excitron equipments, rated at 260 kW, 12 kV, to be commissioned went into service early in 1951. Since that date further development has shown that equipments of several times this rating can be constructed by connecting numbers of tubes in parallel per phase, equipments having many advantages over the steel-tank rectifier, not the least being a much-reduced capital cost.

The excitron is a continuously excited, mercury-pool cathode, glass-envelope rectifier. Fig. 2 illustrates a high-voltage excitron of recent design. It is constructed from a heavy glass tube of approximately $3\frac{1}{2}$ in. in diameter, it is $17\frac{1}{2}$ in. in length and weighs $5\frac{3}{4}$ lb. The anode, cathode, grid and excitation-anode seals are of the glass-to-metal type. Graphite is used for the solid cylindrical anode, and the tube has a mercury-pool cathode. Anchoring of the cathode spot is achieved by means of a semicircular molybdenum spot-fixing ring. Two grids fully screen the anode, and are supported inside the tube by the intermediate disc seals through which they are energized. The upper grid is the H.T. or potential dividing grid, and the lower the control grid.

The excitron is ignited by a simple gravity-controlled dipper device, which can be clearly seen in the photograph. The pivoted arm carries at one end an armature, and at the other a small anode, performing the dual function of an ignition and excitation anode. The energizing of a gapped choke, mounted approximately ¼ in. from the tube, attracts the armature and lifts the anode from the mercury pool, drawing an arc, which is subsequently maintained as an excitation arc continuously exciting the tube. Extinction of this arc results in the dipper falling back into the mercury and automatic re-ignition. With the excitation established, the tube is immediately ready to carry full load. The tube is mounted vertically and air-cooled, the air being blown on to the cathode radiator from insulated nozzles at the rate of 30–40 cu. ft. of air per minute, depending on the full-load rating and maximum ambient-air temperature.

It will be appreciated that the mercury-pool cathode excitron possesses several advantages over the hot-cathode type of tube. There is, for example, no limitation of life due to loss of emission from the cathode; it can withstand backfire and heavy overload currents without damage. and furthermore, no initial filament-heating time is required.

Typical High-voltage Excitron Equipment

Single-anode rectifiers permit the use of the three-phase, full-wave bridge circuit and the transformers are thus of the 3/3 phase type, the primary being connected either in star or delta and the secondary in star. The use of this connection results in a capital saving over the delta/diametric or delta/fork connected transformers used to supply the high-voltage steel-tank units.

One of the major advantages of the high-voltage single-anode rectifier unit. is its compactness. For a typical 260-kW 12-kV unit the floor space occupied by the cubicle (2 ft. 6 in. by 6 ft.) compares most favourably with the 16 ft. by 14 ft. enclosure required by a steel-tank equipment.

The six excitrons are each mounted on a separate insulated panel, which also carries all the ignition and excitation circuit auxiliary apparatus associated with the tube. Cooling air is drawn generally through a filter, from outside the cubicle, by a motor-driven blower unit.

On closure of the A.C. breaker the tubes are automatically excited, the fan starts, and the equipment is immediately ready to supply full load. A failure of the fan motor results in automatic arc suppression. As the D.C. output of the rectifier has usually to be heavily smoothed, it is generally advisable to limit the charging current taken by the smoothing circuit in order to be able to maintain reasonable settings of the overload relays. This is achieved by gradually applying the voltage in steps, by exciting first two, then four and finally all the tubes. This latter operation is very rapid, and full volts are applied to the D.C. circuit in about 0·75 seconds from the closure of the A.C. breaker.

Variation of the D.C. output voltage is achieved by suitably delaying the firing of the excitron with a negative bias voltage on the control grid. At the correct instant of each cycle this bias is removed and the grid made positive with respect to the cathode, allowing the anode to carry current.

Any of the three normal methods of grid control can be used, namely, impulse D.C. biased sine wave or variable-phase sine wave.

The excitrons are arc suppressed by applying a large negative bias to the control grid, so preventing any of the tubes from firing. This feature is used to give extremely rapid protection against severe overloads, whether rectifier or transmitter, and to remove the H.T./D.C. supply in the event of transmitter faults, thereby minimizing possible damage to apparatus or valves

VOLTAGE-STABILIZER AND REFERENCE TUBES

Voltage-stabilizer tubes (often termed voltage-regulator tubes) and voltage-reference tubes are gas discharge tubes, usually gas-filled diodes with unheated cathodes, specially designed to provide steady voltages. Voltage-stabilizer tubes are normally employed in simple circuits to supply reasonably stable voltages to circuits with loads of up to about 40 mA. Typical applications are for the voltages supplied to local oscillators in communications receivers, for master oscillators in transmitters or for the stabilization of screen-grid voltages in audio-frequency Class B amplifiers. Voltage-reference tubes are intended for applications requiring extremely accurate stabilization; these tubes deal directly with only negligible current, but are often used as a source of reference in units delivering considerable power by comparing the output with the voltage across a reference tube and then employing the error signals to vary the effective impedance of the supply.

The following are terms used in the specification of the characteristics of reference and stabilizer tubes:

Ignition Voltage.—Voltage at which the discharge in the tube will be initiated. This characteristic is affected, in some tubes, by the ambient light, and delay may occur in the firing of a tube in complete darkness.

Burning Voltage.—When the tube has fired, the voltage across it drops to a value known as the burning voltage. Over a long period this value tends to change slightly, and the long-term stability is determined by this characteristic. With most tubes there is a tendency for voltage stability to increase during their life, so that an overall increase in stability may be obtained by ageing the tube.

Burning Current.—This is the current flowing through the tube and the manufacturer provides maximum and minimum limits for satisfactory operation. Temporary overloads of several times the normal maximum current are generally permissible on starting, provided that these do not re-occur at intervals of less than a few minutes.

Regulation Voltage.—This voltage is the difference in burning voltage between maximum and minimum burning currents.

When designing for voltage-stabilizer tubes it is essential to ensure that the minimum ignition voltage is applied to the tube even under the most unfavourable conditions of high loads and low mains voltages, etc., and that the burning current should not fall outside the specified limits under any likely conditions. Manufacturers do not recommend the connection of gas-filled tubes in parallel in order to increase current ratings, since slight variations between individual tubes produce unequal current distribution. Stabilizers may, however, be connected in series, in order to stabilize higher voltages than that for which they are designed, in which case it may be advisable to connect 0·2–1-MΩ shunt resistors across one or more of the tubes in order to reduce the minimum ignition voltage.

ZENER DIODES

If a silicon junction diode with a low breakdown voltage in the reverse direction is biased to just beyond the turnover point, current will flow limited only by the low slope resistance, any further increases in current will result in only a negligible increase in voltage across the diode, the variation in voltage being a function of the slope resistance. Hence, in the turnover region the voltage across a semi-conductor diode will be substantially constant for reasonably small changes in current.

This characteristic of semi-conductor diodes may therefore be used as a voltage reference source or as a voltage regulator or as a voltage-surge limiter in much the same way—only at lower voltages—as voltage-regulator tubes. Diodes intended for this application are termed Zener diodes, and are available in a number of ranges with turnover voltages between about 3 and 15 volts. In practice, both the inverse and forward characteristics may be used as reference sources.

Since the slope resistance of a Zener diode decreases as the inverse current increases, it is desirable to operate the diodes at as high a current as possible in order to obtain the optimum regulation; this current is, however, limited by thermal effects. A typical operating current for optimum regulation is of the order of 20 mA.

For a given current, the inverse-reference voltage will vary with ambient temperature. The effect may be either an increase of voltage with increasing temperature (positive coefficient) or a decrease of voltage (negative coefficient), depending on the nominal voltage of the diode. The temperature coefficient of voltage remains fairly constant over the ambient temperature range for which it is designed, and is often expressed as a percentage value for unit temperature variation. Zener diodes may be operated in series, and it then becomes possible to select diodes whose combined temperature coefficient may be almost zero over a very wide temperature range. For additional stability, Zener diodes may be operated in an oil bath. Parallel connection of Zener diodes is not possible, as there would always be unequal sharing of the current.

The usual precautions should be taken when connecting these semi-conductor devices into circuit. Zener diodes may also be used as low-voltage power rectifiers.

26. TRANSISTORS

26. TRANSISTORS

Human control over nature has been accomplished in the main through the instrumentality of machines, in which a relatively small human effort is able to control a much greater output of work by the machine. In the science of electronics the use of a small input to control a larger output is termed "amplification", and the devices used to accomplish this end may be divided into four broad categories :

(a) servo mechanisms, in which a small torque controls a proportionate output;

(b) transductors, in which the non-linear magnetic characteristic of ferro-magnetic materials is utilized to control the flow of an electric current;

(c) vacuum tubes and gas-filled valves, in which the potential on an electrostatic screen is used to modulate an electron stream;

(d) transistors, in which a small input current is used to vary the conductivity of a semi-conducting element, and hence to determine the flow of current through the element.

Amplification is not an end in itself, and it is used in complicated equipment to accomplish some utilitarian purpose; it follows therefore that the above four methods do not necessarily compete with each other, but each has its own particular realm in which it is the most convenient agent to employ. The present article will discuss the nature of the transistor, rather than the purposes for which it is best fitted, and will give data from which the reader may judge in any specific case whether its use will be more advantageous than that of an alternative method. Practical circuitry, for available types of transistors, is given later in this section.

For many years prior to the discovery of the transistor, certain materials which were semi-conductors had been used to construct rectifiers. The first valve amplifier was constructed by inserting a third electrode in a rectifier diode, and by analogy it was thought that the semi-conductor rectifier (often called in one form a crystal diode) might also be converted into a crystal triode. The analogy proved worthless, but in pursuit of it new discoveries were made which eventually resulted in a device which had many similarities to the thermionic triode. Since then several forms of crystal triode have been developed into a production proposition, and more than one form of tetrode has been constructed in the laboratory.

Semi-conductors

Normal constructional materials may be broadly divided into two categories, those which are electrical conductors and those which are electrical insulators. Those in the first category possess a large number of charges (electrons) which are not firmly attached to any particular nucleus, and hence may be thought of as an electron-gas which by diffusion through the body of the conductor serves to convey an electric current. This concept is only a crude attempt to visualize a mechanical model of the conduction process, and for more exact work a mathematical model in which functions replace material particles must be used. In a similar way insulators may be considered as materials in which so few charges are unattached to any specific nucleus that electrical conduction is infinitesimal.

A small number of materials fall into an intermediate category, and are termed semi-conductors. They are so constituted that the number of free charges available for conduction is much smaller than that in a conductor, but much larger than that in an insulator. Only a few of the semi-conducting materials are usable for transistors, namely, the group-four class-three materials such as silicon and germanium, or stoichiometric alloys of their neighbours in the periodic table of elements. In a pure semi-conductor the number of free charges increases rapidly with increase of temperature over a critical range, giving it a large negative temperature-coefficient of resistivity. The material used for the manufacture of transistors has a critical amount of impurity which provides a number of free current-carrying charges still small in number compared with those available in a conductor, but large compared with those available due to thermal energy in the pure semi-conductor. Such materials are called impurity semi-conductors.

Two types of impurity may be used to lower the resistivity of a semi-conductor; one type provides an excess of electrons over the critical number required for valency bonding inside the crystal, and because the current-carrying charges are electrons, this class of material is called an *n*-type semi-conductor; the second type of impurity which may be used causes a deficiency of electrons in the valency bonds of the crystal, and each such deficiency is termed a "positive hole" because it also can provide a conduction current, behaving as though it were a particle with a positive charge due to the interchange of electrons between such deficiencies, and such material is called a *p*-type semi-conductor. A semi-conductor without impurity is called an intrinsic semi-conductor, and, providing the total impurity content is very small, it is possible to so balance the *n*- and *p*-type impurities that intrinsic-conductance is obtained in the region where such balance is maintained.

An added complexity to the simplified picture above, is responsible for the operation of the transistor. If a current be caused to pass through a junction in an impurity semi-conductor, it can cause a modification to the conductance of the material in the neighbourhood of the flow by modifying the density of the charges available for current-carrying. This effect is responsible also for the ability of the material to "rectify", increasing current in the forward direction causing a decrease in resistivity, and increasing current in the reverse direction causing an increase. In the case of the transistor a small current flowing in the forward direction is used to control the conductivity of the semi-conductor, and hence to modulate a second current flowing in the reverse direction between independent electrodes. This effect may be achieved by a number of differing mechanical arrangements, each of which has its own peculiar electrical advantages and disadvantages; in consequence, it seems probable that the variety of transistors available will be as great as that of electronic valves, and a knowledge of their characteristics is essential to choose the optimum type for a particular purpose. Those types at present under development will be described later in this section.

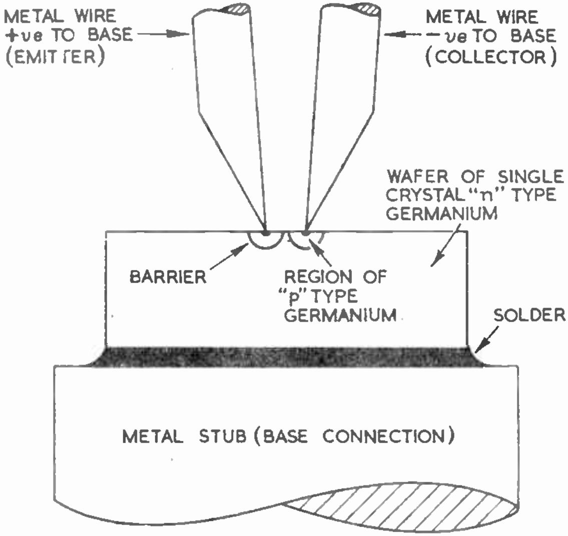
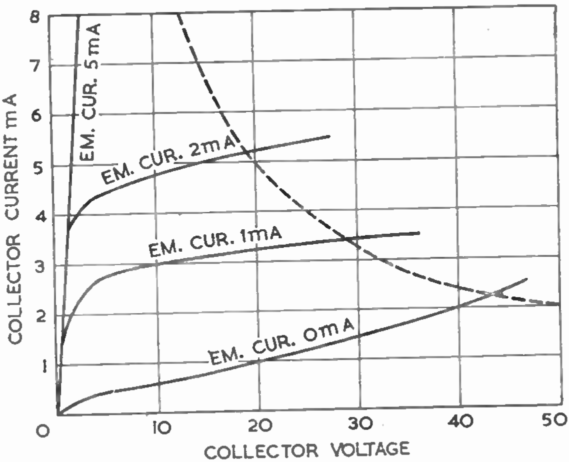


Fig. 1.—Construction of Point-contact Transistor.

Point-contact Transistors

The point-contact transistor consists essentially of two metallic points contacting a plate of semi-conductor within a few thousandths of an inch of each other. Several physical arrangements were tried in the early days, but today only one form has survived, that shown in Fig. 1. Two metallic wires, cut at an angle to their axis to form a point, are lowered on to a flat surface of single-crystal germanium, and the whole assembly is sealed to prevent the entry of moisture and other contaminating elements. The germanium is prepared as an n-type impurity semi-conductor of approximately 10 ohms/cm. cube at room temperature (this value depends on application, and may be higher), and after the contacts are lowered pulses of current are passed through to cause a redistribution of impurities in the neighbourhood of the points. This redistribution of impurities generates a minute volume of p-type material in the neighbourhood of each point, and an approximately hemispherical boundary where n and p types meet, which will be only a few atoms thick of germanium having intrinsic conductance (called a "barrier"). A still smaller region, which is probably low-resistivity n type, is formed around the contact between metal and germanium, and a somewhat indeterminate modification of the surface also occurs. A third, low-impedance connection is made to the bulk of the germanium by soldering, and the two currents flow between each of the points and this base contact.

When a negative potential is applied between collector and base, the reverse current which flows depends upon the conductivity of the



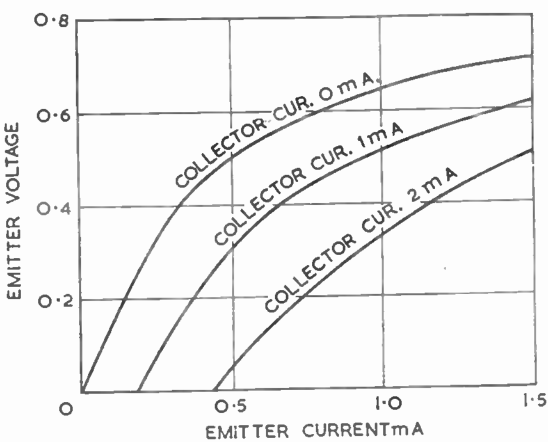


FIG. 2.—STATIC CHARACTERISTICS OF TYPICAL POINT-CONTACT TRANSISTOR.

germanium. When a positive potential is applied between emitter and base, the forward current which flows lowers the resistivity of the germanium in the neighbourhood of the emitter point, and, if this point is sufficiently near to the collector point, the conductivity in that region is increased, with consequent increase of collector current. Since the impedance from emitter to base is considerably lower than that from collector to base, if the two points are sufficiently close a power gain may be realized from emitter to collector circuits. If all the charges flowing into the base from the emitter were to flow out again into the collector, the increase of collector current would be equal in magnitude to that of the emitter; in a good point-contact transistor the increase of collector current is very much greater than that of the emitter (current gain greater than unity, and usually greater than two), due to the trapping of charges in the neighbourhood of the point with the consequent generation of still more charges.

Characteristics

The static characteristics of a typical point-contact transistor are shown in Fig. 2. It will be seen that the collector voltage/current characteristics are similar to the anode current characteristics of a pentode valve, except that emitter *current* has replaced grid *voltage* transistor circuits may be designed in a manner analogous to that in which valve circuits are designed, but in the case of the point-contact transistor a current gain greater than unity enables a single transistor with only resistive and capacitative elements to exhibit negative resistance, and in consequence greater care must be exercised to ensure stability. If the base be grounded, input be applied to the emitter and output taken from the collector, then the input impedance will be less than that of the output circuit—a condition which is a disadvantage in many amplifier circuits. If the emitter be grounded and input applied to the base, the input impedance will be high, but such connection is not very desirable in the case of the point-contact transistor, because the greater current gain tends to cause instability in the stage.

A convenient equivalent circuit is shown in Fig. 3. When the temperature increases there is a large decrease in r_e and an increase in α, while r_b and r_e will vary either up or down in a manner which tends to differ with the individual transistor; the overall result is always a decrease in amplification as temperature rises. Most point-contact germanium transistors will be permanently damaged if the temperature is allowed to rise as high as 85° C., and will not return to their original characteristics. Both the diffusion of current-carrying charges into the semi-conductor, and the decay of such charges once there, occurs at a finite rate, and when frequency rises to such a value that the periodic time is comparable in value with the time-constant of this process, a fundamental change of behaviour occurs. The major effect is a rapid decrease of current-gain as the frequency increases above a critical value, and the frequency at which this factor is 3 db down is called the "cut-off frequency". The value of the cut-off frequency may be increased by decreasing the spacing between the two points, but this causes an increase in r_b which may result in instability. If a germanium with a lower resistivity be used, this results in a lower value of r_b for a given spacing, but, unfortunately, reducing the resistivity of the germanium also reduces r_e, which it is desired to maintain as high as possible. In

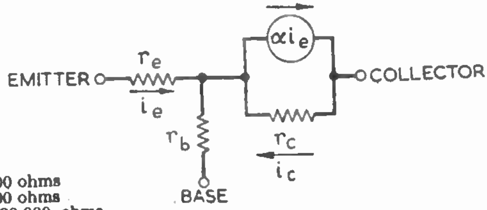


FIG. 3.—CONVENIENT EQUIVALENT CIRCUIT FOR POINT-CONTACT TRANSISTOR.

Typical values are as follows:

α (current gain) = 2–4
r_e = 100–500 ohms
r_b = 100–500 ohms
r_c = 2,000–30,000 ohms

consequence of this it is usual to manufacture two basic types of point-contact transistors, one with wide spacing, high-resistivity germanium and high α at medium frequencies, and a second type with low-resistivity germanium and small contact spacing with a high cut-off frequency.

A point-contact transistor has a much higher noise figure (ratio of equivalent input noise to thermal noise in the input circuit) than has a normal valve amplifier; a typical value is 50 db, and this is independent of the operating point on the characteristic. The distribution of this noise with frequency is fundamentally different from that of thermal noise, and the noise content of a band 1 c/s wide varies inversely as the frequency to a first approximation. For this reason, it is obvious that the noise figure will be a function of frequency and band-width, and the 50 db figure quoted above is measured at 1,000 c/s with a band-width of 1 c/s, this method having been adopted as a provisional standard for transistors. The source of this noise is still unexplained theoretically, and active research proceeding at present might well call for a different standard once the origin of the noise is understood.

Junction Transistors

Alloyed Junction Transistors

One of the earliest forms of point-contact transistor to be made was formed by etching two pits on opposite sides of a thin wafer of germanium, and bringing two whiskers into contact. In a modified version this form gave the earliest, and still most common, type of

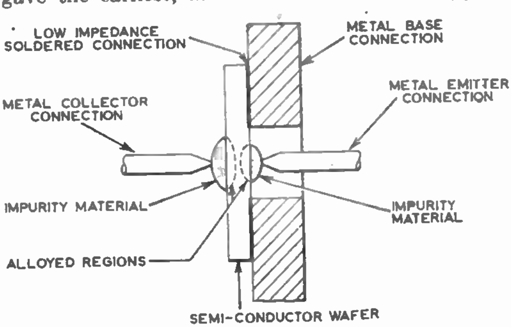


FIG. 4. TYPICAL ALLOYED JUNCTION TRANSISTOR.

junction transistor. The general construction is illustrated in Fig 4 A thin wafer of single-crystal semi-conductor is soldered to a metal tab to give a low-impedance connection. Two small circular "dots" of an appropriate impurity material are placed exactly opposite to one another and made to wet the surface over a circular area by application of heat. The heat is continued until they have alloyed into the wafer and the two resulting junctions are parallel and only about 10–20 μ apart (1 μ is a thousandth part of a millimetre). The collector is made about twice the diameter of the emitter to give maximum current gain. For high-frequency working, the diameters are made as small as possible. High-power versions are very difficult to manufacture because the impurity has a different coefficient of thermal expansion from the semi-conductor, and, if the area of junction is made too large, fracture occurs, giving a discontinuity which ruins performance. Alloying occurs at differing rates along the different directions of the crystal growth, and the wafers are cut with the direction of quickest penetration perpendicular to the major plain to ensure that the two junctions are parallel and approach within a minimum distance of each other to give maximum frequency response. The cheapest available form of junction transistor is of this type, constructed with a germanium wafer. Its output is only a few milliwatts, and it will operate up to a few hundred kilocycles per second only. Versions with a silicon wafer are also available. These give higher output powers, and will work at higher temperatures. A specially constructed version in germanium will operate at up to 10 Mc/s as an amplifier and higher as an oscillator.

Surface Barrier Transistors

The major causes of the frequency limitation of the alloyed junction transistor are the irregularity of the junction face and the irregularity of the rate of alloying, which limits the minimum distance which can be obtained between the two junctions. This is overcome in the surface-barrier type of unit. The surface-barrier transistor is very similar in construction to the alloyed-junction type, the difference being in the method of forming the junctions. In the surface-barrier type a thin wafer of high-resistivity doped semi-conductor has two circular pits etched by a special electrolytic jet process in the positions where the junctions will be, and this etching is continued until the thickness of the material between the two is only 2–6 μ. The etching current is then reversed to plate a suitable impurity material on to the two faces, and finally connection is made to them. Constructed from n-type germanium, with emitter diameter about 50 μ and collector about 100 μ, this device will operate up to about 100 Mc/s, but gives only very low power, about 10 mW. Constructed out of silicon, and with the impurities lightly alloyed into the surface to give a spacing of about 7 μ, it is possible to operate up to about 200 Mc/s. The theoretical analysis of this type is slightly different from that of the normal alloyed junction due to the extreme thinness of the material between the junctions and to the special nature of the impurity semi-conductor interface, but the general characteristics are similar except for the much higher cut-off frequency.

Diffused Junction Transistors

The desire for large area junctions led to the development of diffusion techniques for forming the junctions. It was soon realized that the extreme thinness of the layers which could be obtained in this way gave an even greater advantage, and many of the latest forms of device utilize diffusion. The principle is simple; the semi-conductor is exposed at a high temperature to an atmosphere containing a concentration of a suitable impurity material which is allowed to diffuse into the surface to the desired depth. The depth is controlled by time and temperature, while the penetration is kept even by cutting the wafer with the direction of quickest diffusion normal to the plain of the desired junction. The type of junction is normally specified by naming the successive types of impurity, proceeding from emitter through base to collector. Those which have been described up to now have been either *npn* or *pnp* types, the most common germanium ones being of the *pnp* variety and the most common silicon ones being *npn* types, although the others are available if required for special purposes.

The diffusion technique has made possible a further type in which a fourth layer of near-intrinsic semi-conductor is interposed between the base and collector, giving a *pnip* type with considerable advantages. The construction of this type is shown in Fig. 5. A recess is etched into a thin wafer of semi-conductor to leave a wall thickness of about 25 μ. The semi-conductor is as free from impurity as commercially producible, the remaining traces being balanced to give intrinsic characteristics tending very slightly towards the *n* type. A layer of *n*-type impurity is diffused into one face (for practical reasons the diffusing is usually done before etching the recess) to a depth of about 2·5 μ. A *p*-type emitter junction is then alloyed into this layer to leave a gap of less than 1 μ between the junction and the interface between the intrinsic and *n*-type materials. A low-resistance connection is made to the layer for the base connection, and a further (and larger) *p*-type collector junction is made opposite the emitter. The diffused layer is made to have low resistivity, so that there is a much greater field in the intrinsic material and the velocity of the carriers is great in the comparatively large gap. The major time of passage is thus in the *n* material, where the gap is less than 1 μ. This construction gives an upper operating frequency above 100 Mc/s and a power output up to about 100 mW. The Bell laboratories have constructed a *pnip* type, in silicon, with both

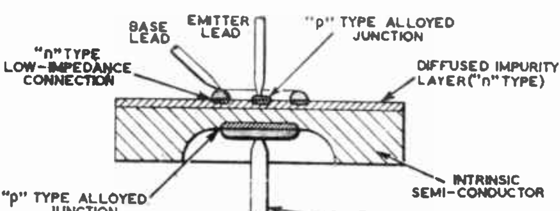


FIG. 5.—*pnip*-TYPE TRANSISTOR.

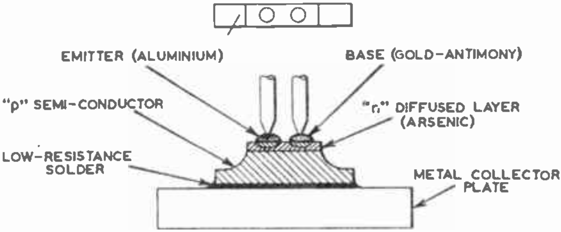


FIG. 6.—DRIFT TYPE TRANSISTOR.

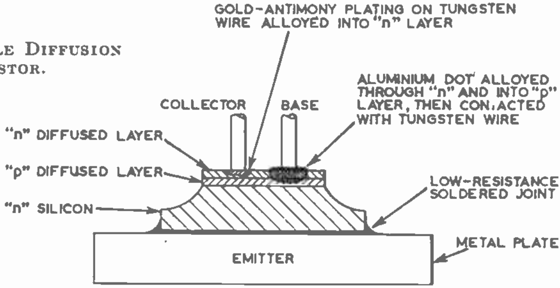
emitter and base diffused. This will deliver 5 watts of power at 10 Mc/s, and will give 1 watt at 100 Mc/s when operating as an oscillator.

"Drift" and "Mesa" Types

A second type in which the diffusion and alloying techniques are combined is illustrated in Fig. 6. This type can have an upper frequency above 1,000 Mc/s—the highest frequency unit yet made. The wafer is of low-resistivity p-type material about 0·8 ohm-cm. value, and the n-type layer is made by diffusing arsenic into the wafer to a depth of a little under 1 μ. The emitter is formed by alloying a dot of aluminium, and the base by alloying a similar dot of gold–antimony alloy. This type has been called a "drift"-type transistor, and a somewhat similar design, called a "mesa" type, has also been made. This has a similar frequency range, and in this the emitter is a narrow strip, the base consisting of two strips of equal width on either side, spaced from the emitter by half the width of the strips. The major obstacle to extending this latter design to frequencies above 1,000 Mc/s using germanium material is the difficulty of processing wafers sufficiently small to keep down transit time to a fraction of the periodic time.

A further type in which a double-diffusion technique is used is illustrated in Fig. 7. An n-type silicon wafer is exposed to an atmosphere containing both aluminium and antimony at a temperature which causes both to diffuse inwards. The p impurity diffuses at a faster rate than the n, giving rise to an outer layer of n which is separated from the

FIG. 7.—DOUBLE DIFFUSION TRANSISTOR.



original silicon by a continuous layer of p material. The p layer is made about 2–3 μ thick, and contact is established by alloying a dot of aluminium straight through the n and partly into the p layer. In operation, the n layer is so biased that the junction with the electrode is a rectifier in the non-conducting condition. A tungsten wire plated with gold–antimony alloy is pulse-welded on to the n layer, and a second tungsten wire is used to contact the aluminium base electrode. This unit is capable of giving 400 mW output up to a frequency of 200 Mc/s. At present it is in development production only.

Note that in Figs. 6 and 7 the rectangular wafer is ground away at the edges adjacent to the electrodes. This is done because the diffusion techniques necessitate the diffusion of the impurities over the whole surface area of the wafer. By subsequently grinding the edges as shown the useful area of diffusion is separated from the remainder, which is not shown in the diagrams.

Junction Power Transistors

To deliver a power of several watts to a load, the area of the junction must be increased so that the current density can be reduced to a reasonable value. Despite this increase of area—and consequent decrease of maximum working frequency—the absolute value of the waste energy dissipated in the semi-conductor wafer presents a difficult problem. As the wafer is so small, the dissipation of a watt or more would raise the temperature well above the safe maximum if normal constructional techniques were used. The solution which has been adopted is to bond the semi-conductor on to a block of metal, such as copper, having high thermal conductivity, and then to bolt the metal firmly to the metal chassis of the equipment. In this way the thermal energy is conducted away from the device. Since the wafer is bonded to the metal, the metal itself is one of the transistor electrodes: where this electrode must be insulated from the chassis, a very thin sheet of mica, which has a reasonably high thermal conductivity, is interposed between the block and the chassis. The principle of one form of these devices is shown in Fig. 8. In this, a wafer of n material has a p layer diffused in all over. This forms the collector, which is ground off on one face to allow a p material to be alloyed in to form the emitter, and a ring of n material to be alloyed in to form the base connection. The diffused layer is soldered

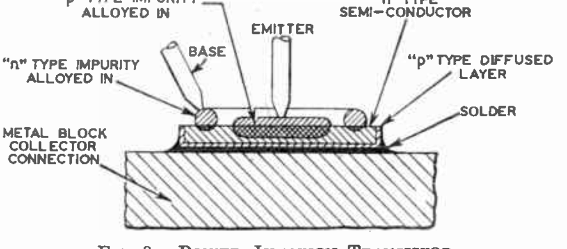


FIG. 8.—POWER JUNCTION TRANSISTOR.

to the metal base-plate with a solder containing a p-type impurity to make a connection of low impedance, both electrically and thermally.

Triode Characteristics

The characteristics of a junction-type triode transistor such as those already described differ somewhat from those for the point-contact type. The junction is simpler than the welded point of the whisker contact, and the current gain for the grounded-base connection is usually

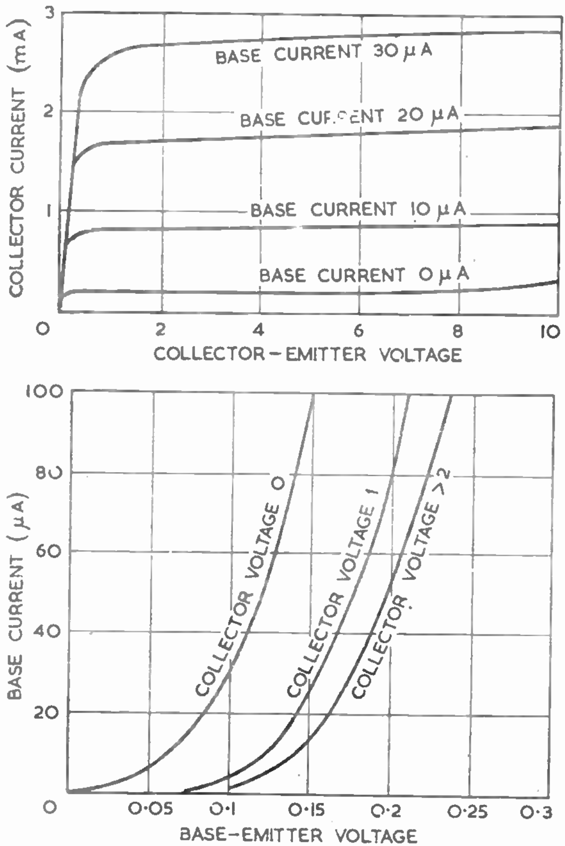


FIG. 9.—TYPICAL STATIC CHARACTERISTICS OF JUNCTION TRANSISTORS.

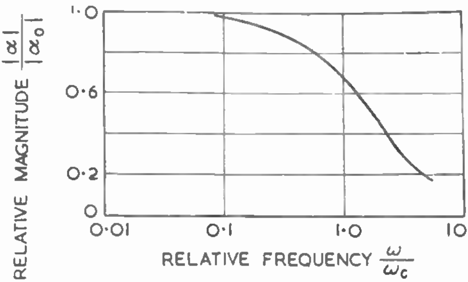
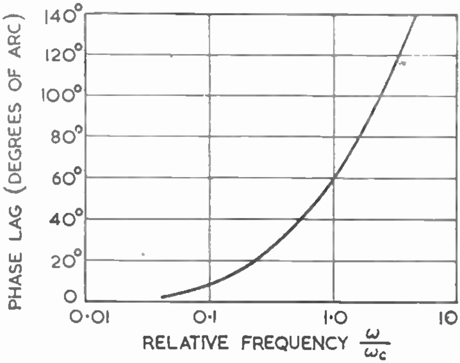


FIG. 10.—VARIATION OF CURRENT AMPLIFICATION FACTOR WITH FREQUENCY FOR JUNCTION TYPE TRANSISTORS.



less than unity, allowing the grounded emitter circuit to be used with stability. The general form of characteristics is similar for all types of junction triode, but differs considerably in the magnitude of the parameters. Those of a typical audio-frequency, alloyed-junction germanium transistor are shown in Fig. 9, for the grounded-emitter connection in which they are most frequently used. The noise figure of a junction transistor is considerably lower than that of a point-contact type, of the order of 10 dB, or even lower.

The junction-type transistor is more susceptible to exact mathematical prediction of characteristics than is the point-contact type, because the geometry of the active elements is more exactly known and controlled, as are also the electrical characteristics of the materials. When the mathematical processes of quantum physics cannot be followed, however, the following crude and imperfect picture may aid in understanding the mode of operation. When a junction is biased in the reverse direction, as between the base and collector of a transistor, the reverse current will not increase indefinitely as the reverse voltage is increased, but will rapidly increase to a saturation value and thereafter remain constant as the voltage increases. The transistor is operated at a voltage well above that at which saturation occurs, and the current is a function of the free-charge density in the base material, being limited because all available charges are engaged in carrying it. As emitter current is increased, however, this represents a current flow in the forward direction across the other junction, which increases the free-charge density in the base in the neighbourhood of the emitter junction. If the two junctions are so near to one another that the increase of

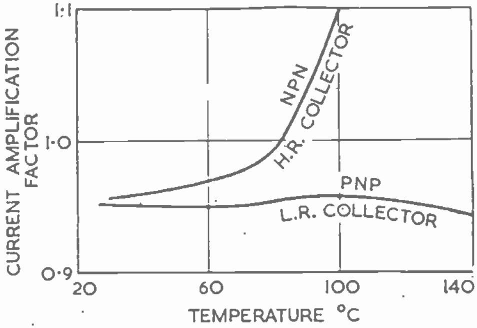


Fig. 11.—Typical Variation of Current Amplification Factor for Junction Type Transistors.

charge density near the emitter represents also an increase in the neighbourhood of the collector, it is obvious that the collector current will increase in consequence. In practical transistors the width of base between the two junctions is made so small that the two densities are almost identical, and current gain is almost unity.

When the transistor is used in the grounded-emitter circuit, the phase of the current-amplification factor is of importance as well as the magnitude, a lagging phase angle causing a serious fall in gain from the stage at a frequency where the magnitude is still fairly high. A carefully constructed transistor will exhibit a variation closely following the theoretical performance which is illustrated in Fig. 10, where it will be seen that the phase angle varies appreciably at a frequency well below the cut-off value.

The characteristics vary with the temperature of operation. The maximum temperature at which a silicon unit may be used is usually limited by the methods of construction (melting point of solder, etc.); that at which germanium units may be used by deterioration of the junctions, which occurs around the neighbourhood of 100° C. The resistance of the semi-conducting materials falls with rising temperature, and the standing current rises. The variation of the current gain depends upon the junction characteristics, the variation for two germanium units being shown in Fig. 11. That with a low-resistivity collector is typical of an alloyed-junction construction, and that with a high-resistivity collector typical of the grown-junction construction.

Grown-junction Transistors

The earliest form of germanium junction transistors was obtained by growing the junctions in the single crystal. The ease of alloying junctions soon caused this type to fall out of favour. The difficulty, however, in the case of silicon of alloying junctions which will have an area permitting large power dissipation has made the grown-junction the favoured construction for this latter material. A single crystal of semi-conductor is slowly grown from a reservoir, into which the appro-

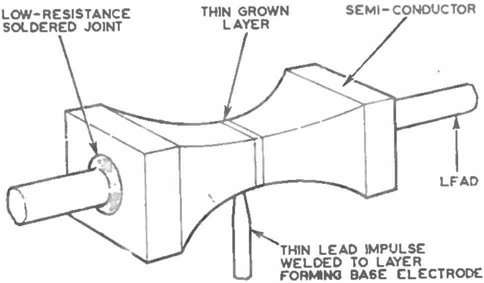


FIG. 12.—GROWN-JUNCTION TRANSISTOR.

priate impurities are introduced during the growing, to yield a thin layer (about 10 μ or less) of either n or p type sandwiched between thick layers of the opposite impurity type. The composite crystal is then cut into a number of blocks which are ground or etched to reduce the area of the junction, as shown in Fig. 12, and three leads are attached to form a transistor. As it is impossible to confine the base connection entirely to the width of the layer, it is allowed to overlap slightly on the collector side, where the electrical bias will be such as to give a non-conducting junction. In some cases the junctions are grown by varying the rate of withdrawal from a reservoir containing both n and p impurities, the type of material depending upon the rate. This method allows alternate layers to be grown along a long crystal. Operation up to about 10 Mc/s may be obtained.

Grinding away the body to reduce the area (and hence the self-capacitance) of the junction gives a worth-while increase in maximum usable frequency. Where still higher-frequency working is required, the "effective" area of junction may be reduced to much smaller dimensions by joining two leads to opposite sides of the base layer and passing a transverse current through the base to so bias it that only a

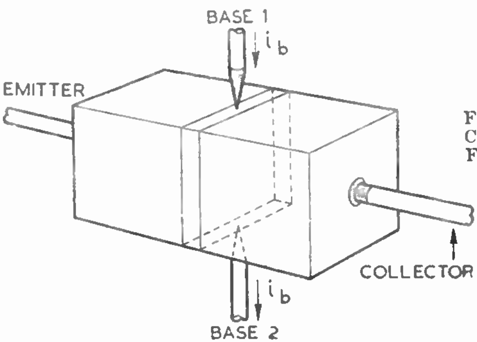


FIG. 13.—DIAGRAMMATIC CONSTRUCTION OF HIGH-FREQUENCY JUNCTION TETRODE.

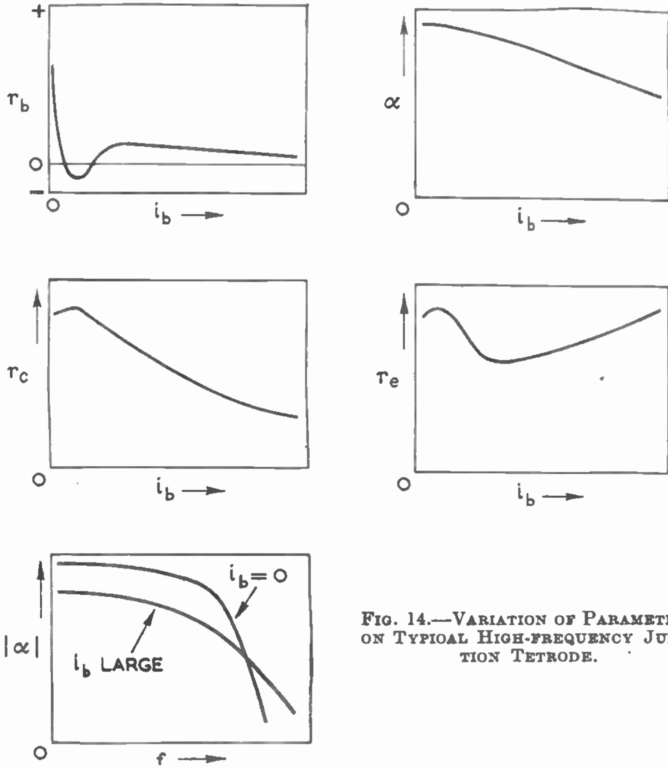


FIG. 14.—VARIATION OF PARAMETERS ON TYPICAL HIGH-FREQUENCY JUNCTION TETRODE.

narrow strip adjacent to one of the leads is conducting. This raises the frequency both by reduction of effective capacitance and by reduction of base resistivity. In Fig. 13 this construction is shown diagrammatically; for clearness, the grinding away of the bar is not shown although it is employed in the actual device. The manner in which the parameters vary with the transverse base-current, and the current gain varies with frequency, is shown in Fig. 14. Because of the fourth lead, this type is often called a " junction tetrode ", and operation up to 100 Mc/s is possible.

For uses where a current gain greatly in excess of unity (typical of a point-contact transistor) is required, this can be obtained in a single unit by growing two thin, closely adjacent layers, instead of one, thus giving an additional block of *n*-type material on the collector end of a *pnp* type, as shown in Fig. 15, resulting in three symmetrical junctions in the centre of the block. The centre junction is the reversed biased collector junction as before, the left-hand junction is the forward-biased

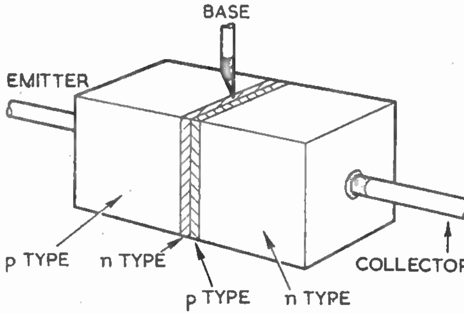
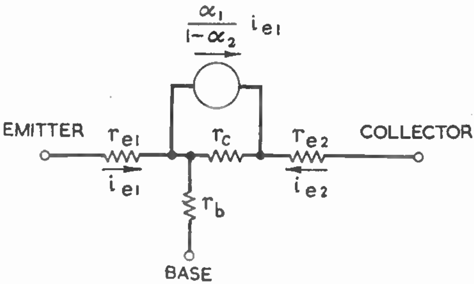


FIG. 15 (*right*).—TYPICAL CONSTRUCTION OF *pnpn* JUNCTION TRANSISTOR.

FIG. 15 (*a*) (*below*).—EQUIVALENT CIRCUIT OF *pnpn* JUNCTION TRANSISTOR.



emitter junction and the new right-hand junction (called a *pn* hook) is also biased in a forward direction to form a second emitter. If the current gain of the original emitter is called α_1 and that of the new hook is called α_2, then the equivalent circuit may be drawn as shown in the figure. Since the value of α_2 may be made near to unity, it is obvious that very high equivalent current-amplification factors of the order of 10–100 may be obtained. As a crude qualitative picture of the mechanism, the positive holes injected from emitter to base pass through into the second *p* region, biasing the hook so that it conducts in the forward direction; thus the complete unit is seen to be equivalent to two triodes in series with direct coupling " built in ", and such transistors have been simulated in the laboratory by suitably connecting a *pnp* and an *npn* transistor.

Comparison of Junctions

The different forms of junction described differ in the ease with which they may be constructed, but more important than this is the variation of electrical characteristic due to the differing distribution of impurity level along the direction of current flow. An excellent example of this is seen in the case of the *pnip* transistor, where the abrupt change of resistivity allows a high working frequency despite comparatively large physical dimensions. In general, a grown junction exhibits a gradual change in type from *n* to *p*, or *vice versa*, as the added impurity slowly diffuses through the reservoir and affects the growing crystal. The alloyed junction, on the other hand, shows a more sudden step, the transition from one type to the other being almost discontinuous. The diffused junction occupies a position intermediate between these two, depending upon the process used in construction. The gradual transition of the grown junction is best for the collector, because of the lower effective self-capacitance, but the sudden transition of the alloyed type is the best for an emitter because of the greater efficiency obtainable with it. An additional advantage of the alloyed type is the lower value of resistivity obtained in emitter and base electrodes.

Unipolar Transistors

A new mode of operation is found in the "unipolar" transistor, in which high gain at very high frequencies is possible once the extreme difficulties of fabrication have been overcome. Although only in the initial stages of investigation at present, this unit, which should be capable of giving considerable gain at frequencies of several hundreds of megacycles per second, will probably become one of the most important types in the future. The constructional principles are illustrated in Fig. 16, from which it will be seen that the form and operation are fundamentally different from any type described as yet. A block of high-resistivity *p*-type semi-conductor is faced on two opposite sides, with a layer of low-resistivity *n* type formed by diffusion, and two strips of low-resistivity *p*-type material are alloyed across the ends. A second

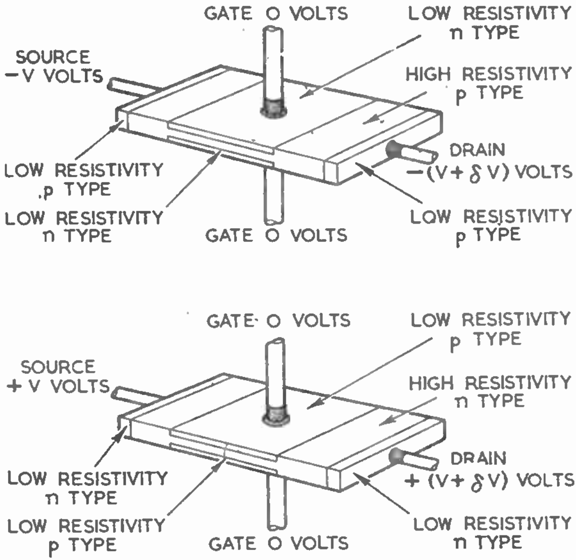


Fig. 16.—Diagrammatic Construction of Unipolar Transistors.

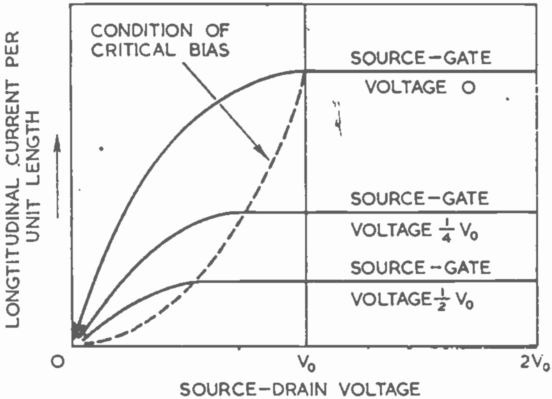


FIG. 17.—TYPICAL CHARACTERISTIC OF UNIPOLAR TRANSISTOR.

form of this type may be made from a block of n-type semi-conductor as shown. The pn junctions formed between the layers and the block are biased in the reverse direction, which results in a region in the block adjacent to the junctions in which the carrier concentration is negligible. As a result of this denudation of charges, the longitudinal current between the two end electrodes is constrained to flow along a limited cross-section in the centre of the block where free charges are still available. As the reverse junction voltage is made larger, the charge-free layer widens and the area capable of conduction becomes smaller. The potential drop δV along the block which causes the longitudinal current flow will also cause the reverse potential across the junction to vary by this amount along the length of the block, thus causing the charge-free volume to vary in width and form a wedge-shaped conducting volume. When the junction voltage is increased, there will be a critical value V_0 at which the wedge will just close at the narrowest end. In Fig. 17 the static characteristics of this type are shown, together with a dotted curve indicating the condition of critical bias at which the maximum junction voltage just succeeds in closing the conduction gap at one end. It will be seen that if the source is made earthy, and the signal applied to the gate, an amplified output may be obtained from the drain. Once the technique of manufacturing these units sufficiently small has been mastered, it should be possible to obtain considerable gain at a frequency of well over 100 Mc/s, and the upper limit depends upon the limiting size of unit possible.

The Junction " Fieldistor "

A further form of transistor which differs fundamentally from the normal junction triode, but which is of academic more than of practical interest, is the " fieldistor ". This is very easy to construct, but usable only over a very limited frequency range. It is illustrated in Fig. 18. A block of germanium half n type and half p, is surrounded near the junction by a metallic electrode spaced away from the block, and hence

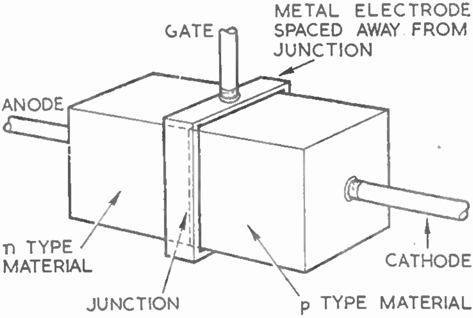


Fig. 18.—Diagrammatic Construction of Junction Fieldistor.

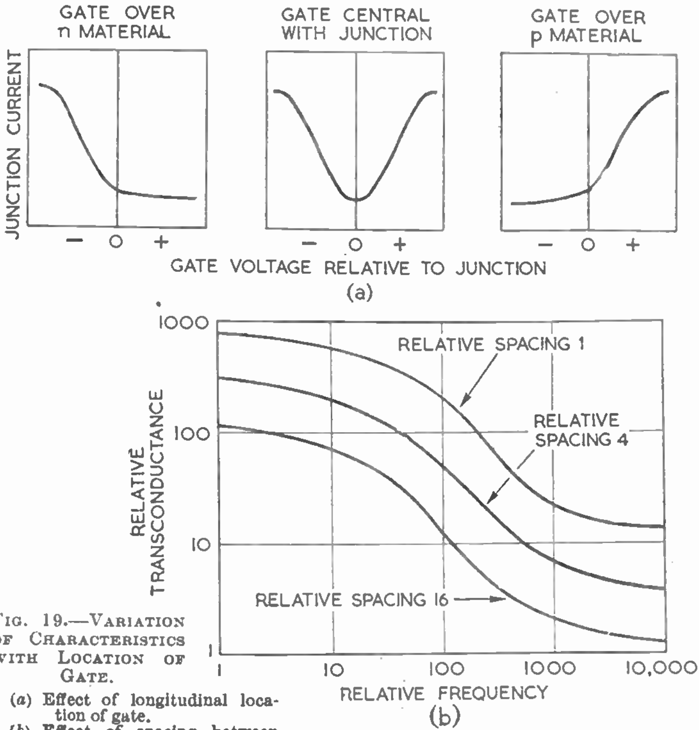


Fig. 19.—Variation of Characteristics with Location of Gate.

(*a*) Effect of longitudinal location of gate.
(*b*) Effect of spacing between gate and junction.

insulated from it. The junction is biased in the reverse direction, and variation of potential difference between the junction and the gate will cause variation of charge density along the edges of the junction, and hence variation of reverse current through it. In Fig 19 the effect of varying the position of the gate is shown. Variation along the axis of the block will cause departure from symmetry as the gate becomes asymmetrical with the junction, while increase of spacing between gate and block will result in a fall of effective transconductance. The variation of transconductance with frequency is also illustrated in the figure. The gap between the two may be artificially reduced by immersion of the unit in a polar dielectric, thus increasing the capacitance between gate and block. Such transistors may be constructed with an input impedance of several megohms in parallel with several picofarads, and an output impedance of several thousands of ohms; the transconductance at a frequency of 10 c/s may be as high as 1 mA/volt, but this will have fallen by 40 dB at 100 kc/s; the noise figure is in the region of 70 dB. The very limited frequency range will seriously restrict the future uses of this type.

The Grain-boundary Transistor

The grain-boundary transistor represents a completely new principle, which has recently been proposed to overcome the troublesome variation of transistor characteristics with varying ambient temperature. A few units have been constructed in the laboratory, and these prove that the principle can be translated into a practical device. The advantages are such that this type would appear to have a very favourable future, although it may be several years before they are commercially available. The performance is independent of temperature to the extent that good operation can be obtained from 100° C. down to 2° K.: for the first time a device is possible which could operate at the extreme temperatures of outer space.

A special type of crystal structure is used, as shown in Fig. 20. Instead of a single crystal of semi-conductor being grown from a carefully orientated seed of homogeneous material, the crystal is grown from a composite seed such that the "100" crystal plane is identical throughout the material, but the "100" planes are tilted relative to one another

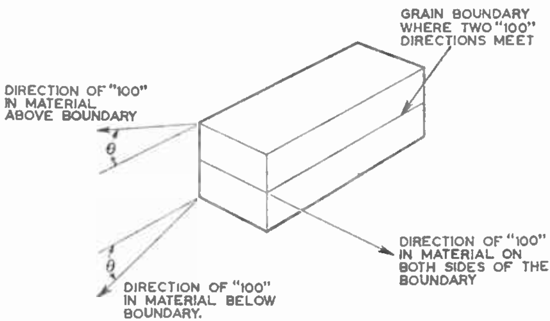


Fig. 20.—Semi-conductor with Grain-boundary.

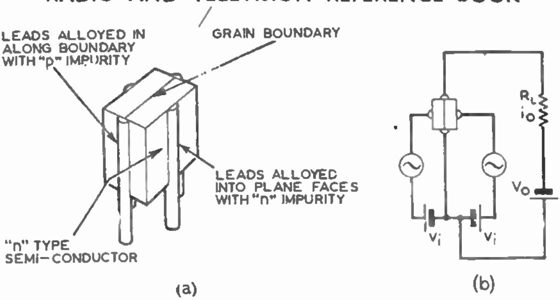


FIG. 21.—(*a*) GRAIN-BOUNDARY TRANSISTOR; (*b*) CIRCUIT OF GRAIN-BOUNDARY TRANSISTOR.

in the two halves of the block, thus forming a "grain-boundary" plane surface of infinitely small thickness throughout the block.

The first form of device constructed from such a block is shown in Fig. 21. Two strips of conductor are soldered to the plane faces of the n-type block with an n-type impurity, and two strips of p-type impurity are alloyed along the opposite faces of the grain boundary. These strips along the grain boundary are so biased, relative to the others, that they form a high-impedance, reverse-biased junction to the n material on either side of the boundary, but they serve to pass a transverse current i_0 through the grain boundary and load R_L as a result of potential V_0. If an A.C. potential is superimposed on the junction between n material and the grain boundary (see circuit, Fig. 21 (*b*)) the effective electrical conductivity of the grain boundary is altered, and the circuit i_0 is modulated. Typical characteristics of the device are shown in Fig. 22. A mutual conductance of more than 3 mA/V has already been achieved. Since the input is a reverse-biased junction, the input

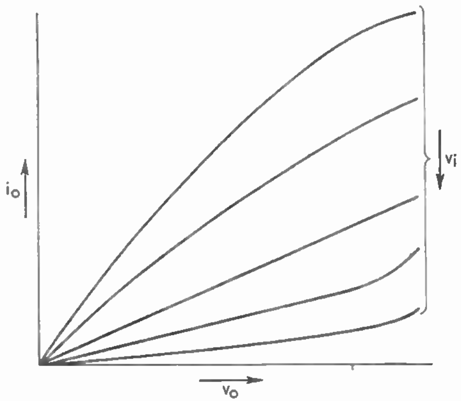


FIG. 22.—CHARACTERISTICS OF GRAIN-BOUNDARY TRANSISTOR.

impedance is high and the amplification considerable. The conductivity of the grain boundary is independent of ambient temperature, and it is this feature that makes the device so attractive. It is too early yet to know what the noise and frequency characteristics of the developed device will be like.

Transistors as Circuit Elements

An equivalent "T" circuit for a transistor considered as a three-terminal active network has already been given, and will be found useful in the design of transistor circuits. Unfortunately, the values of the parameters become frequency dependent at the higher values of frequency, and it is this variation which will be considered very briefly here. The most serious change is that of the current gain factor α, and for the junction-type transistor this may be shown on theoretical grounds to approximate to the form :

$$\alpha = \operatorname{sech} \frac{W}{L_b} \sqrt{1 + j\omega \tau_b}$$

where W = width of base region;
τ_b = lifetime of minority carriers in base region;
L_b = diffusion length of minority carriers in base.

A further series of approximations allows the effect of the complex current-gain factor to be drawn as an additional network attached to the low-frequency equivalent as shown in Fig. 20. This approximation is of great value for the transient analysis of transistor circuits.

The only effective capacitance which need be included is that of the collector C_c as shown shunted across r_c, while the variation in magnitude of α is accounted for by a second resistance and capacitance R and C. The difference of phase between this RC circuit and the actual value of α is accounted for by a time-delay network having zero attenuation. The time constant of R and C is that of the cut-off frequency as shown in the figure. In present types of alloyed junction transistors the time-

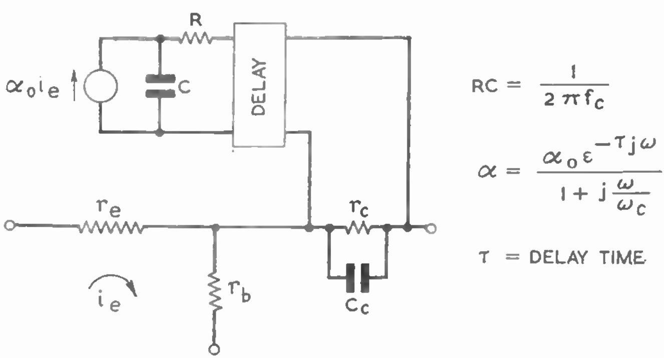


FIG. 23.—APPROXIMATE EQUIVALENT CIRCUIT FOR TRANSIENT ANALYSIS OF JUNCTION TRANSISTORS.

constant r_eC_e tends to be greater than RC, and the delay time τ lies between the two in value.

SEMI-CONDUCTING MATERIALS

The two materials in common use today are germanium and silicon, both of which have advantages and disadvantages, so that choice of material is always a compromise depending upon which characteristics are of most importance in the desired unit. Germanium is easily processed, which means that good life-time is readily obtainable, but the energy-gap is so low that the maximum temperature of operation is unduly restricted. Silicon has a much higher energy gap, allowing operation up to 200° C., but mobility and processing difficulties make it much more difficult to obtain a suitably long life-time of minority carriers. Alloys of these two have been tried, but no great advantages are found. The alloy of silicon with the other group 4 material, carbon, however, offers the possibility of operation up to very high temperatures (approaching 1,000° C.), and a serious attempt is being made to process silicon carbide to give transistors suitable for control functions in space vehicles.

A very promising field of enquiry lies in the use of stoichiometric alloys of group 3 and 5 materials, and considerable success has already been obtained in the preparation of a number of these. The principle characteristics of these materials are listed in Table 1, the values not shown being (at the moment) known only very approximately. A further field which is being explored, but without so much success, is that of alloys containing not only groups 3 and 5, but also groups 2 and 6, and as the importance of semi-conductors warrants ever-greater sums spent in research a considerable extension of characteristics should become available through the use of such materials.

R. T. Lo.

TABLE I.—CHARACTERISTICS OF SOME STOICHIOMETRIC ALLOYS

Material	*Phosphorus*	*Arsenic*	*Antimony*	*Characteristic*
Aluminium	—	—	1,080	Melting point
	3·0	2·16	1·60	Energy gap
	—	—	100	*e* mobility
	—	—	200	*p* mobility
Gallium	1,340	1,240	720	Melting point
	2·25	1·35	0·67	Energy gap
	—	4,600	4,000	*e* mobility
	—	300	800	*p* mobility
Indium	1,070	936	523	Melting point
	1·25	0·33	0·17	Energy gap
	3,400	2,800	70,000	*e* mobility
	650	—	800	*p* mobility

TRANSISTOR CIRCUITRY

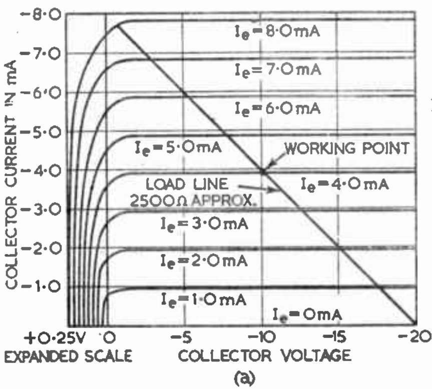
Most of the applications where transistors are already established or envisaged in the near future are ones where thermionic valves were previously employed, and there is a certain basic similarity in the circuit arrangements. There are, however, a number of important differences in detail which are best understood by comparing and contrasting the characteristics of valves and transistors.

In a thermionic valve the anode current is controlled by variation of the potential of the control grid. Although in certain circumstances this electrode may draw current, this does not contribute to the fundamental action of the device and, in most cases, merely has a nuisance value. In the case of the transistor, control of the collector current is brought about by variations of the current in the emitter-base path, and its flow is fundamental to the operation of the device. This is the most significant difference between valves and transistors as far as circuitry is concerned; and the need for current in the control circuit of transistors must always be recognized when devising amplifiers or oscillators around them.

Common-base Operation

One method of operating a transistor is the common-base arrangement, that is, with the signal fed in between emitter and base, and taken out between collector and base. The characteristic curves relating to this form of operation are given in Fig. 24 (*a*). They are for a *pnp* junction type, and the type of circuit in which the measurements may be taken is shown in Fig. 24 (*b*). The variation of collector current with collector voltage is shown for a number of different values of emitter current. The basic similaritv to the anode characteristics of a thermionic pentode is apparent.

The curves show that the collector impedance is very high (typically above a megohm) and that the collector current changes by 0·975 mA for each change of 1·0 mA in the emitter current. A load line of approximately 2,500 ohms is shown drawn through an operating point



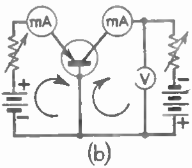


FIG. 24.—(*a*) COMMON BASE CHARACTERISTICS OF *pnp* JUNCTION TRANSISTOR TYPE EW 53/2. (*b*) CIRCUIT ILLUSTRATING CURRENT FLOW IN TRANSISTOR IN COMMON BASE ARRANGEMENT.

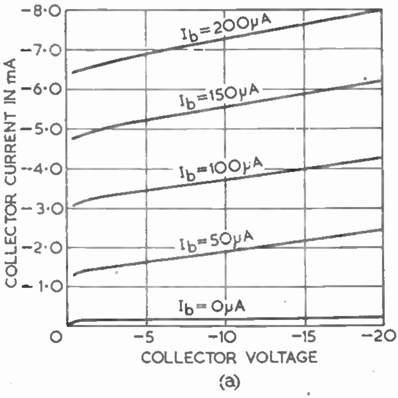
of collector current 3·9 mA, collector voltage 10·0 volts. This point is reached by supplying a current bias to the emitter of 4·0 mA. A sine-wave current of peak value 4·0 mA supplied to the emitter will cause a similar current in the collector load of peak value 3·9 mA and peak voltage 11·0. There is nothing in the curves to show that amplification has taken place, because there is no indication of the voltage required to produce the emitter current. The input impedance of a common-base junction transistor is typically about 25 ohms, so the voltage required is about 100 mV peak, and in this case the power gain is approximately 150 times (22 db).

In this calculation there has been a simplification in giving the input impedance as 25 ohms. In fact, it is highly dependent on the emitter current, and it also varies according to the value of collector load. This variation of input impedance with emitter current can cause serious distortion when handling large signals, and will be dealt with later when considering practical circuits.

Common Emitter Operation

Common-base operation of junction transistors is relatively infrequent, and has been mentioned mainly as an introduction to the more usual common emitter operation. In this arrangement the signal is fed in between the base and emitter and taken out between the collector and emitter.

Reference to Fig. 24 (*b*) shows that both emitter and collector currents flow in the base lead of the transistor, but their directions are opposite, so the net current is their numerical difference. For instance, when the emitter current is 2 mA the collector current is 1·95 mA, so that the base current is 0·05 mA (50 μA), and when the emitter current changes to 3 mA the collector current becomes 2·925 mA and the base current is 0·075 mA (75 μA). A variation of base current of 25 μA therefore corresponds to a change in collector current of 0·975 mA. This is the relationship which is important when the base is used as the input electrode, and it is therefore usual to plot this in separate curves instead of deriving it from those in Fig. 24 (*b*).



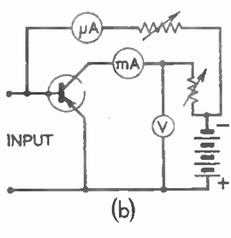


FIG. 25.—COMMON EMITTER OPERATION.

(*a*) Characteristics of *pnp* junction transistor EW 53/2. (*b*) Circuit.

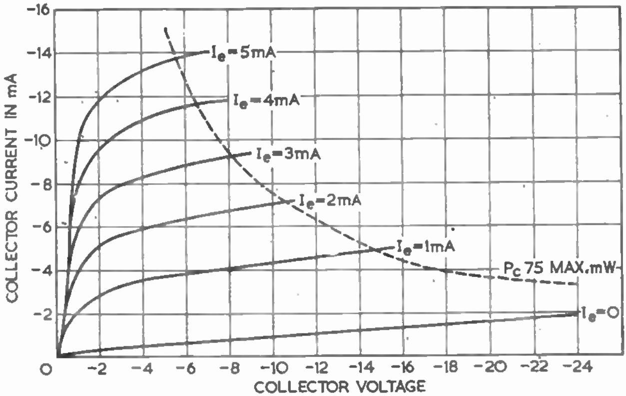


FIG. 26.—COMMON BASE CHARACTERISTICS OF A TYPICAL POINT-CONTACT TRANSISTOR, TYPE GET 2.

Fig. 25 (*a*) gives common emitter curves for a *pnp* transistor, and Fig. 25 (*b*) shows the type of circuit which may be used to obtain them.

As compared with common-base operation, the collector impedance is lower (50,000 ohms is typical), the input impedance is higher (500–1,000 ohms is typical) and there is a considerable current gain between input and output (40 is typical). The overall effect with correct matching is higher stage gain (35 db typical).

Common Collector Operation

A third circuit configuration is possible where the signal is fed in between base and collector, and taken out between emitter and collector. This is referred to as common-collector operation, and corresponds loosely to the ther mionic-valve cathode-follower circuit. It gives less gain than either o the two other arrangements (12 db typical), but can be made to have igh input resistance and low output resistance, so that it is useful fc matching purposes.

Circuits for Point-contact Transistors

The point-contact transistor is usually confined to common-base circuits, and curves illustrating this operation are given in Fig. 26. Apart from the curves being less perfect than those of the junction type in the shape of the knee and the magnitude of the cut-off current (that is, the collector current for zero emitter current), there is a very important difference, in that a change in emitter current of 1 mA results in a change in collector current of several milliamperes. At first sight this might suggest that the point-contact transistor would give higher gain than the junction type, but due to its higher input resistance (250 ohms typical) and its lower output resistance (20,000 ohms typical), the gain is limited to about 20 db, in spite of this current gain. The true importance of the current gain being greater than unity is that any resistance in the base circuit gives positive feedback from collector to emitter circuit. This results in base-input circuits being inherently unstable, and accounts for the infrequent use of the common emitter and common collector configurations. It also renders possible some interesting oscillator circuits, both of the sine-wave and relaxation types. Finally, it means that even with emitter input (common-base operation) care must be taken to avoid instability when resistance is present in the base lead, and there is always some due to the resistance of the germanium itself, even when none has been deliberately introduced for such reasons as provision of bias.

Bias and Power Supplies

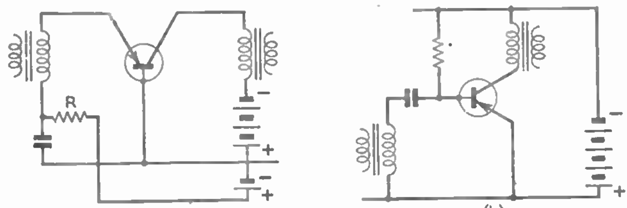
The provision of power supplies for transistors presents some points of difference in comparison with valves. In the first place, of course, no heater or filament power is required. Secondly, the H.T. supply rarely exceeds 20 volts, and may be either positive or negative, according to the type of transistor used; point-contact and *pnp* types requiring a negative collector polarity, whilst *npn* types require positive. Finally, the bias for the control electrode is current rather than voltage, and its direction depends on the type of transistor and the circuit configuration used. Thus in common-base circuits point-contact and *pnp* transistors require positive current bias to establish an operating point suitable for amplification, whilst *npn* types require negative current bias. When the base is the input electrode the polarities listed above are reversed.

A typical bias circuit for a point-contact or *pnp* grounded-base amplifying stage is shown in Fig. 27 (*a*). A separate bias battery is used and the bias current is determined by the battery voltage divided by the sum of the input resistance of the transistor and the feed resistance R. The secondary winding of the transformer feeding the signal to the emitter has its lower end decoupled to earth by a capacitor, which should, at the lowest operating frequency, have a low impedance compared to the input impedance of the transistor. A 32-μF electrolytic of low working voltage is typical for audio-frequency work, 0·005 μF for intermediate-frequency and radio-frequency.

In the case of the point-contact transistor, where the collector current is greater than the emitter current, it is possible to obtain bias without a separate supply by a resistance in the base lead, in a manner somewhat similar to cathode bias in a thermionic valve. This arrangement is not to be recommended, because of the positive feedback previously mentioned; it may be reduced for A.C. by shunt capacitance, but D.C. feedback remains, which may cause the transistor to trigger to a high value of current and destroy itself.

Bias Circuits for *pnp* Transistors

The usual bias arrangement for a *pnp* transistor with common emitter connection is shown in Fig. 27 (*b*). Here the negative current is derived from the negative H.T. line through the resistor, and the capacitor prevents its passage direct to earth through the transformer winding. Here again the value of capacitance should be chosen to give an impedance which is low compared to that of the input of the transistor. Changes in the transistor characteristics due to temperature changes will result in tho operating point shifting with the simple circuit just described, and elaborations are desirable to give greater stability. One method is to add resistance in the emitter lead, shunting it by a



(a) (b)

FIG. 27 (*above*).—(*a*) BIAS CIRCUIT FOR COMMON BASE CONNECTED TRANSISTOR. (*b*) BIAS CIRCUIT FOR COMMON EMITTER CONNECTED TRANSISTOR.

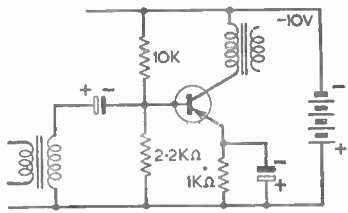


FIG. 28 (*right*).—JUNCTION TRANSISTOR WITH STABILIZED BIAS CURRENT.

capacitor to prevent reduction of gain. This is quite effective, but there is some sacrifice of output power due to the voltage drop in the resistor. Further stabilization at the expense of additional current drain is given by an additional resistance between base and earth. The combined arrangement is shown in Fig. 28, which enables a sufficient degree of stabilization to be obtained for any practical purpose by a suitable choice of values. To give good stability without undue sacrifice of power, circuits may be devised using temperature-sensitive resistors in the bias network.

For the sake of simplicity the circuits which follow will not in general incorporate bias stabilization, but its use is advisable in any apparatus which must work over a large range of ambient temperatures: even domestic equipment such as a personal radio receiver may meet extremes, corresponding to an unheated bedroom in winter and a beach in blazing sun. Some industrial and Services equipments, of course, have to operate in much wider extremes than this, so that different values would be necessary in each of these cases.

Radio-frequency and Intermediate-frequency Amplification

When transistors are used in radio-frequency or intermediate-frequency amplifying stages the inter-stage couplings differ somewhat from those in common use with valves, mainly because of the low input impedance of transistors. Because the transistor is a current-operated device there are good arguments for connecting its input circuit in series with the tuned circuit; however, there are drawbacks to this procedure, and the most common practice at present is to match to the low emitter impedance by means of a step-down transformer with an untuned winding. Fig. 29 (*a*) shows two point-contact transistors in radio-frequency amplifier and detector stages. The aerial is coupled to the tuned winding in the usual way, and the emitter is fed from an untuned winding closely coupled to the tuned winding and giving a

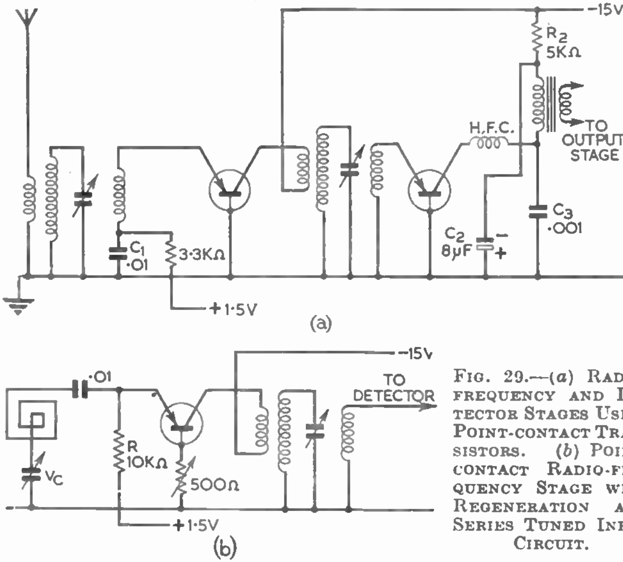


FIG. 29.—(*a*) RADIO-FREQUENCY AND DETECTOR STAGES USING POINT-CONTACT TRANSISTORS. (*b*) POINT-CONTACT RADIO-FREQUENCY STAGE WITH REGENERATION AND SERIES TUNED INPUT CIRCUIT.

step-down ratio of about 5–1. Bias of about 0·5 mA is provided through 3·3kΩ and a path to earth for the radio-frequency currents is completed through C_1.

The collector presents too low an impedance for direct connection across the tuned circuit, and a coupling winding is shown. As an alternative it could be connected to a tap on the tuned circuit. The detector gives rectification because in its unbiased condition it can only respond to the positive excursions of the signal current. It is coupled to the tuned circuit by a step-down untuned winding similar to that used in the radio-frequency stage. In its collector, radio-frequency filtering is given by RFC and C_3, and coupling to the next stage is provided by the audio-frequency transformer. R_2 and C_2 are for audio-frequency decoupling and, apart from this, R_2 is necessary to avoid any danger of triggering which might otherwise occur in the absence of a limiter in the emitter circuit. In the radio-frequency stage this point is looked after by the inclusion of 3·3kΩ in the emitter path. Point-contact types are shown in this circuit because, for the time being, they provide the only ready method of achieving radio-frequency gain. When radio-frequency junction types become readily available their use will be preferable, since they give higher gain and lower noise.

The arrangement shown in Fig. 29 (*b*) is interesting for several reasons. The variable resistor in the base lead gives positive feedback to the emitter circuit, and, by a suitable adjustment, the input impedance of the transistor may be reduced to zero so that this may be connected in series with the elements of the tuned circuit without introducing damping. An increase in the base resistance will make the input

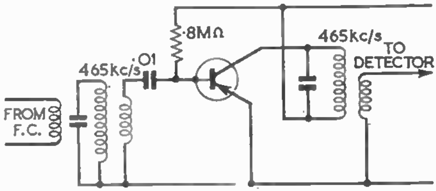


FIG. 30.—INTERMEDIATE-FREQUENCY AMPLIFYING STAGE USING A *pnp* JUNCTION TRANSISTOR.

Not all types of junction transistor are suitable for this use.

impedance negative, so reducing the damping of the tuned circuit to the point of self-oscillation. Bias is provided through R, and the blocking capacitor is to prevent damage to the transistor in the event of a short-circuit in V_c. This type of circuit is valuable to the experimenter who wishes to obtain utmost sensitivity at the expense of critical adjustment. Its use should be restricted to receivers with very small frame or ferrite-rod aerials.

Intermediate-frequency Amplification with Junction Transistors

Some presently available junction transistors give reasonable performance at intermediate frequency, and the circuit in Fig. 30 illustrates the usual arrangement to be adopted. It will be seen that it follows closely the technique already shown for the point-contact radio-frequency stage, with the bias supplies modified for common emitter operation. The value of the bias resistor should be chosen to make the collector current 1 mA. Its actual value depends on the current gain factor of the transistor and, assuming this to be in the neighbourhood of 40, a good starting point would be 0·8 MΩ. This figure is arrived at by taking the base current as being equal to the collector current divided by the current gain (this is approximately true over a large range of values), and to draw this current of 0·025 mA from the 20-volt H.T. line obviously requires R to be 800,000 ohms, the resistance of the transistor being low enough to be ignored.

The collector, being of reasonably high impedance in the common emitter arrangement of a junction transistor, may be connected direct to the tuned circuit, provided that a low *L–C* ratio is used. With conventional *L–C* ratios a tap is desirable for the best combination of gain with selectivity.

Audio-frequency Amplification

The higher noise and lower gain of the point-contact type made their use in audio-frequency amplifying stages appropriate only whilst junction types were not readily available, and the latter type will therefore be the only ones to be considered here. In Class A arrangements the only configuration normally used is the common-emitter one, but in Class B output stages all three of the possible connections may be suitable under various circumstances.

Fig. 31 shows the detector of a small radio set, transformer coupled into an audio-frequency stage which, in its turn, is resistance-capacitance coupled into a small Class A output stage. The first transistor is working at low signal level, so that it need not be biased to a high current: a collector current of 0·5–1 mA gives a reasonable compromise between gain and economy. The second transistor, intended to operate a

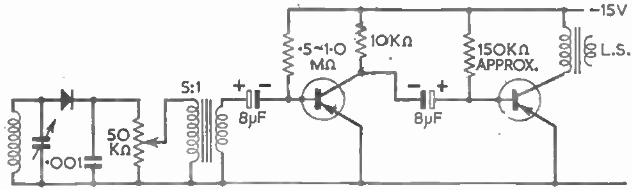


Fig. 31.—Crystal Detector Followed by a Two-stage Transistor Audio-frequency Amplifier.

loudspeaker at low level, is biased to take 4·0 mA, giving—with an H.T. line of 15 volts—a collector dissipation of 60 mW and a power output approaching 30 mW. The correct load impedance for this stage is 3,750 ohms, and is easily calculated by dividing the collector voltage by the collector current.

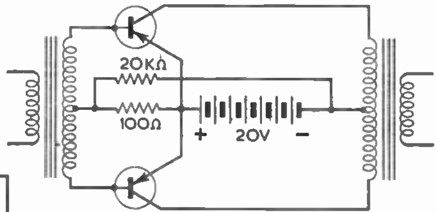
Maximum gain is given by transformer inter-stage coupling, a current step-up being achieved by a turns step-down ratio. Another consideration, however, makes the use of resistance–capacitance coupling preferable. It has already been mentioned that the input impedance of a transistor is highly dependent on the emitter current : therefore, when a large signal is being handled, the impedance varies over a large range with the swing of the signal. If the signal derives from a low-impedance source (a constant-voltage source), the current waveform will be distorted in the emitter circuit, and so will the resulting current in the collector. If, on the other hand, the emitter is driven from a high-impedance source (constant current) the current in the emitter circuit will scarcely be affected by variations in emitter impedance, and distortion due to this will be avoided.

Class B Power-output Circuits

For maximum power output with a given collector dissipation, Class B circuits should be used : these have the additional advantage of extreme economy. The transistor really comes into its own here, for whilst the valve Class B arrangement gives great economy in H.T. consumption, there is a constant L.T. drain whatever the strength of the signal. In the transistor arrangement, consumption is only appreciable when a signal is being handled. The common-base circuit is capable of good linearity, but, by the time it has been fed from a high-impedance source to avoid the type of distortion referred to in the previous section, the power gain is very low, so that its use is not attractive, except where low distortion is the prime requirement.

The common-emitter circuit also suffers from distortion due to emitter non-linearity, but, having a higher fundamental gain, there is still a useful measure remaining, even after taking the precautions needed to reduce the distortion. Fig. 32 shows a Class B output stage capable of an output of 300 mW without exceeding the 70-mW dissipation of the individual transistors used. Two steps have been taken to reduce the type of distortion mentioned above, which, in push–pull circuits, is sometimes referred to as cross-over distortion because it occurs when the signal is transferring from one transistor to the other. The first is the provision of a little bias so that the transistors are not completely

FIG. 32 (*right*).—COMMON EMITTER CLASS B OUTPUT CIRCUIT.



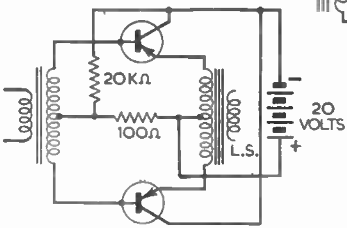


FIG. 33 (*left*).—COMMON COLLECTOR CLASS B OUTPUT CIRCUIT.

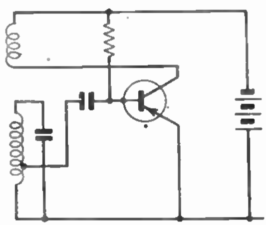
cut-off in the absence of a signal; the other is the deliberate mismatching into the bases of the transistors to mask the variations in input impedance.

There is also a further form of distortion which occurs in common-emitter circuits due to the falling-off in the value of current gain at high values of emitter current. The only simple way of avoiding this is not to drive the transistor to high values of emitter current, which, of course, limits the power output that can be obtained. Negative feedback can be used to reduce the magnitude of this distortion, and the easiest way to achieve this is to use the common-collector configuration, which has inherent feedback. Such a circuit is illustrated in Fig. 33. The gain given is less than that of the grounded-emitter circuit, but the output for a given degree of distortion is higher. The same arrangements for a little bias and an increased source resistance are made as in the previous circuit. One point worth noting in both these circuits is that the bias must be supplied from a low-resistance source, otherwise the rectifying action of the emitter disturbs the bias conditions.

Oscillator Circuits

With junction transistors, oscillator circuits bear a close resemblance to those used with thermionic valves; the emitter taking the place of the cathode, the base that of the grid and the collector that of the anode. Fig. 34 shows a typical arrangement. A transistor capable of

FIG. 34.—JUNCTION TRANSISTOR OSCILLATOR CIRCUIT. VALUES OF TUNING CAPACITANCE AND INDUCTANCE ARE CHOSEN TO SUIT THE FREQUENCY REQUIRED.



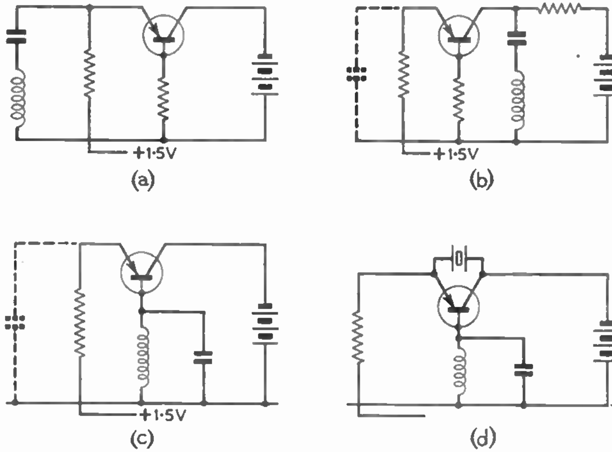


FIG. 35.—(*a*) POINT-CONTACT TRANSISTOR OSCILLATOR WITH SERIES TUNED EMITTER CIRCUIT. (*b*) POINT-CONTACT TRANSISTOR OSCILLATOR WITH SERIES TUNED COLLECTOR CIRCUIT. THE COLLECTOR RESISTANCE MAY BE REPLACED BY A RADIO-FREQUENCY CHOKE. IN SOME CASES THE EMITTER RESISTANCE SHOULD BE BY-PASSED. (*c*) POINT-CONTACT TRANSISTOR OSCILLATOR WITH PARALLEL TUNED CIRCUIT IN THE BASE LEAD. IN SOME CASES THE EMITTER RESISTANCE SHOULD BE BY-PASSED. (*d*) CRYSTAL CONTROLLED POINT-CONTACT TRANSISTOR OSCILLATOR.

intermediate-frequency amplification will oscillate readily at higher frequencies, and may be used as the oscillator in a super-heterodyne circuit for medium-wave operation.

Point-contact transistors enable circuits of a different nature to be used for oscillators. If resistance is inserted in the base lead, the feedback so produced will cause oscillations to be sustained in a series-tuned circuit in either emitter or collector. Alternatively, a parallel-tuned circuit in the base causes feedback at its resonant frequency, and oscillation occurs. These three oscillators are illustrated in Fig. 35 (*a*), (*b*) and (*c*). The addition of crystal control is possible, a satisfactory method being shown in Fig. 35 (*d*). This would operate with the tuned circuit replaced by a resistor, but the tuned circuit prevents oscillation at spurious frequencies.

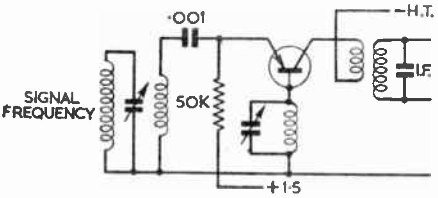


FIG. 36.—POINT-CONTACT TRANSISTOR USED AS A FREQUENCY CHANGER.

Frequency Changers

By using a parallel-tuned-base-circuit oscillator, an effective frequency-changer stage may be made with a point-contact transistor. This is shown in Fig. 37, where the signal is fed into the emitter through the usual step-down transformer, and the intermediate frequency is taken from the collector. For good frequency stability, it is desirable to use a low *L–C* ratio for the oscillator circuit. Alternatively, if more normal values are required to enable use of ganged tuning, the base should be connected to a tap on the coil. This arrangement is quite effective in a receiver of moderate sensitivity, but would give too much noise in a really high-gain set. For this latter use, a junction transistor mixer with separate oscillator is better. Such a mixer stage is shown in Fig. 30.

Experimental Receivers

By combining the various stages in different ways it is possible to devise a number of super-heterodyne receiver circuits, and the designing of them is an interesting way of gaining practical transistor-circuit experience. One such arrangement, which works well within 25 miles of a main broadcasting station, is illustrated in Fig. 38. It uses a ferrite-

FIG. 37 (*right*).—BASIC FREQUENCY CHANGER CIRCUIT USING SEPARATE JUNCTION TRANSISTORS AS MIXER AND OSCILLATOR. THE VALUE OF C_1 SHOULD BE CHOSEN TO GIVE A MIXER COLLECTOR CURRENT OF 0·5–1·0 mA.

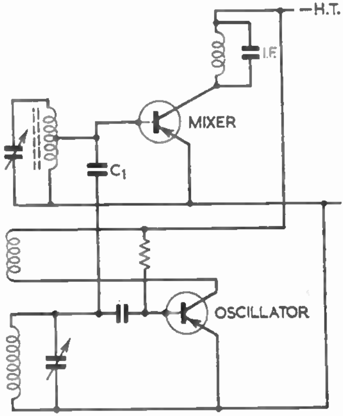
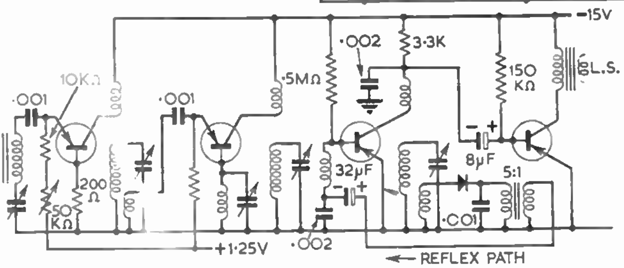


FIG. 38 (*below*).—EXPERIMENTAL SUPERHETERODYNE RECEIVER USING TWO POINT-CONTACT AND TWO JUNCTION TRANSISTORS. FIXED TUNING IS EMPLOYED.



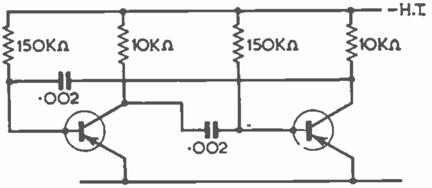


FIG. 39 (*left*).—MULTI-VIBRATOR CIRCUIT USING TWO JUNCTION TRANSISTORS.

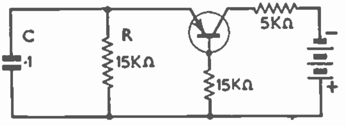


FIG. 40 (*right*).—RELAXATION OSCILLATOR USING A SINGLE POINT-CONTACT TRANSISTOR.

rod aerial and achieves extra gain by reflex use of one transistor for both intermediate-frequency and audio-frequency amplification. The total current consumption is 7 mA at 15 volts for the H.T., plus 1 mA at 1·25 volts for the bias. Fixed tuning is employed and an intermediate frequency of 165 kc/s is chosen, so that long-wave coils may be used for the intermediate-frequency transformers.

For commercial use, higher sensitivity would be required, and the use of point-contact types would be precluded because of noise.

Transistor portable and car receivers are discussed further in Section 14.

Multi-vibrator Circuits

Besides the sine-wave oscillators mentioned above, both junction and point-contact types may be used in relaxation oscillators. A pair of junction types are back-coupled to produce a multi-vibrator circuit in much the same way as thermionic valves. This is shown in Fig. 39. With point-contact types, single transistor circuits are possible, and a typical one is given in Fig. 40. Feedback is produced by the base resistance, and the time constant controlling the relaxation rate is provided by the combination *R* and *C*. This is the basic circuit from which many computor circuits are derived.

npn Transistors

In the foregoing circuits, where junction transistors are used, the *pnp* type is shown. This is because it is at present the more common. The circuits are, however, equally applicable to the *npn* type if the polarity of supplies is reversed. The combining of *pnp* and *npn* types in complementary symmetrical circuits has great possibilities, but has been deliberately excluded in this treatment as being still some way removed from general practical application.

B. R. A. B.

27. CRYSTAL DIODES

27. CRYSTAL DIODES

Semi-conductors

A small group of materials of recent development has assumed a particular interest to the radio and television engineer. They are known as semi-conductors, and the group with which we are here concerned, notably silicon and germanium, are intermediate in electrical conductivity between conductors and insulators, there being a wide divergence between different types.

Perhaps the most notable feature of these materials is that they have non-linear electric-circuit characteristics, i.e., their resistance is much greater when current passes in one direction than in the other. Some of these substances contain electrons in the free state, and under certain conditions are also able to perform the functions so long associated with thermionic valves, viz., signal amplification, oscillation, etc. This latter characteristic is covered more fully in Section 26 (Transistors).

One of the first practical applications of semi-conductors was that of high-frequency rectification, where a metal electrode (" cat's whisker ") was fixed by light pressure in contact with the surface of one of several crystalline substances, the pair forming the detector of the early radio receiver. These early crystals were, by modern standards, inefficient and unstable, and were soon displaced by thermionic valves, which were able to function not only as rectifiers but were capable of voltage and power amplification as well.

The development of radar and the increasing use of ultra-high-frequency transmission brought with them problems that taxed the resources of valve designers. Even special designs of the thermionic valve had their limitations, since the upper frequency was set by the capacitance between the electrodes and by the transit time of the electrons—that is, the time taken for an electron to pass from the valve cathode to one of the other electrodes. At very high frequencies the time of one oscillation becomes comparable with the transit time, and the valve can no longer completely perform its function.

In the crystal type of diode the two electrodes are virtually in contact, so that the electrode spacing is practically zero, and the transit time thus becomes very brief. Consequently, with the need in radar practice for a mixer valve operating at about 200 Mc/s, the crystal valve was resurrected and was rapidly developed into a robust and reliable component of controlled characteristics; later, in conjunction with cavity magnetrons, it was used on frequencies of the order of 3,000 Mc/s.

Extensive search was made for a satisfactory semi-conducting contact rectifier, and many of the combinations that had been used in the early days of radio were investigated. along with a number of new materials: carborundum, silicon, galena, copper pyrites and germanium. Of these the silicon–tungsten combination and germanium have been developed farthest.

Silicon is eminently satisfactory as a frequency-converter, and units

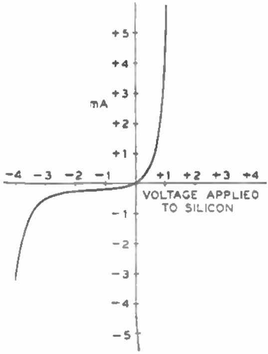


FIG. 1.—VOLTAGE/CURRENT CURVE OF TYPICAL SILICON RECTIFIER.

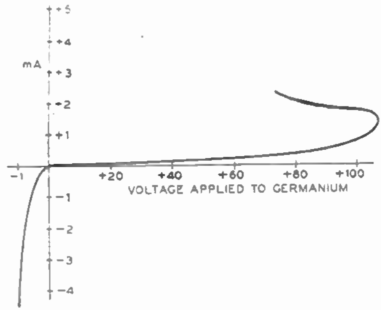


FIG. 2.—VOLTAGE/CURRENT CURVE OF TYPICAL GERMANIUM RECTIFIER.

have been developed for use up to about 60,000 Mc/s, but it is rather easily damaged by overload.

Germanium is chemically a near relative of silicon. Both are in the same family as selenium, tellurium and boron. Germanium is a silver-white metal of atomic number 32, specific gravity 5·4 and atomic weight 72·5. Though still a comparatively rare substance, means have now been developed for the extraction and purification of germanium from flue dusts in gasworks and industrial concerns.

Characteristic Properties

A typical voltage/current characteristic of a silicon–tungsten diode rectifier is shown in Fig. 1. It will be noted that as the semi-conductor is made positive with respect to the metal point electrode, the current increases rapidly, but when the polarity is reversed, the current is small for voltages up to about 3 volts, beyond which the reverse current increases rapidly.

Fig. 2 shows a typical voltage/current curve of a germanium crystal rectifier, although the difference of the positive and negative abscissa scales should be noted. In this case the resistance in the reverse direction is high, differing markedly from silicon in this respect, and is of the order of several megohms, a feature which is evident only when the metal is present in a highly purified state. It will be noted that for a reverse voltage of about 4 volts the resistance, in the case of silicon, falls rapidly, whereas with germanium it does not break down until about 100 volts is reached.

In the forward direction, the slope measured at +2 volts may be of the order of 100 ohms (10 mA/volt), and the conductance is such that the current is much higher than would be expected in a small component. The slope in the reverse direction reaches a maximum at about −2 volts, beyond which it falls slowly, finally culminating in an elbow in the curve known as the turnover voltage, beyond which the

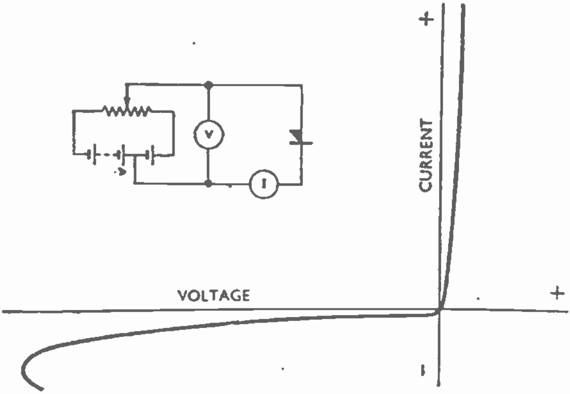


FIG. 3.—USUAL FORM OF PRESENTATION OF TYPICAL GERMANIUM RECTIFIER CHARACTERISTIC.

(*G.E.C.*)

resistance falls and then becomes negative. In the reverse direction, a small current represents a considerable dissipation, and in view of the high impedance a current limit of 1 or 2 mA is usually imposed.

The turnover voltage is an indication of the purity of the metal. Beyond this point the rectifier loses control and the current is limited by external resistance. If such resistance does not exist, the rectifier may be destroyed by excessive current. In general, the turnover voltage is reduced as the temperature rises.

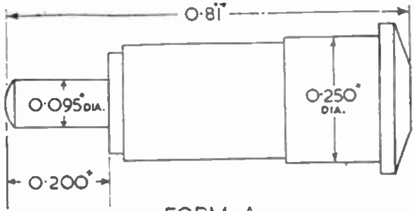
The voltage/current curve of the germanium rectifier is usually depicted in the form shown in Fig. 3, where the abscissæ represent potential applied to the whisker. This is, of course, the same information as that shown in Fig. 2 with the potential co-ordinates reversed and the condition of easy current passage indicated as positive for both potential and current.

The direction of easy current passage is opposite for the two semiconductors cited. Chemically they are near relatives and are operated under similar conditions, so that they might be reasonably expected to behave in the same way. As they do not, it is clear that the theoretical basis of the mechanism of current transfer will be complex. In fact, no completely satisfactory theory has yet been postulated.

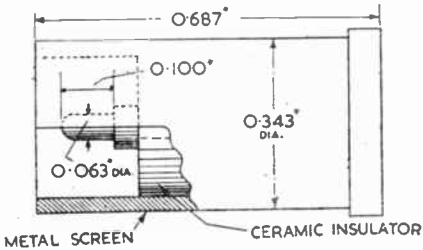
An outline of the conduction principles is given in Section 26 (Transistors).

Silicon Crystal Rectifiers

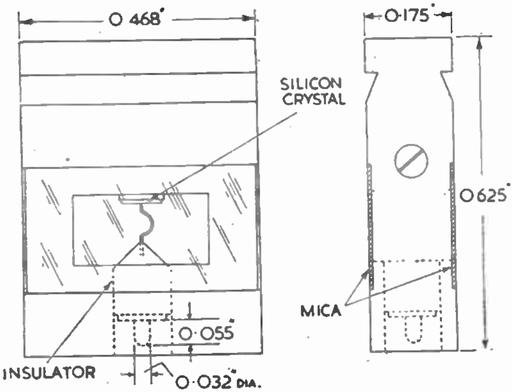
Prepared as a *p*-type semi-conductor, the silicon crystal rectifier has been in general use for some years for detection, frequency-changing or mixing circuits handling frequencies above 500 Mc/s. It has uniform H.F. impedance, ability to withstand moderate overload, and low conversion loss and capacitance. These features are attained through



FORM A



FORM B



FORM D

Fig. 4.—Standard Dimensions of Silicon Cartridges.
(*B.T.H.*)

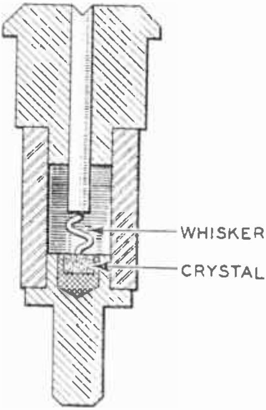


FIG. 5.—CONSTRUCTION OF TYPICAL SILICON DETECTOR.

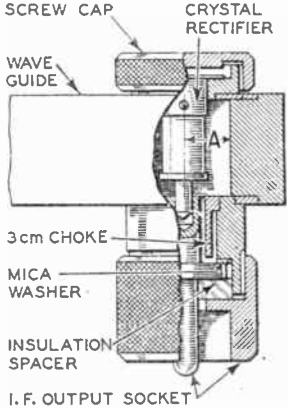


FIG. 6.—EXPLODED VIEW OF WAVEGUIDE MIXER SHOWING CRYSTAL RECTIFIER.

(*B.T.H.*)

small contact area, low forward and relatively high reverse resistance. Its overload capacity is increased by the addition of minute quantities of aluminium and beryllium in addition to the boron impurity which gives the crystal its *p* characteristics.

The boron–silicon crystal is normally prepared by grinding and polishing followed by a heat treatment to produce surface oxidation, which is subsequently removed with acid to leave a pure surface. Crystals of this type, shown in Fig. 5, are used by connection across a resonant coaxial line, which is tuned and matched to the input signal, or are connected across a waveguide with the shorting plunger adjusted behind the crystal.

To avoid these somewhat critical adjustments it has been possible to produce these units with fixed dimension and sufficiently close tolerances to match waveguides of given dimensions, and thus permit the use of interchangeable components.

Another type (Form B) is the coaxial unit in which the dimensions of the components are fairly easily controlled, resulting in greater uniformity of electrical characteristics. Being double-ended, this construction enables the unit to function as a dipole and absorb electromagnetic energy of wavelength twice its own length. The shielding of the outer conductor obviates damage from external high-power fields.

Here the whisker forms a continuation of the inner conductor of a coaxial line and is soldered to a hole in a brass bush fixed at the centre of a ceramic disc located by a shoulder on the outer tube. The crystal is fixed to a nicked disc, which is forced into the outer tube and pushed forward until the correct pressure of the whisker is secured. Fig. 6 shows a waveguide mixer, pre-tuned to a local oscillator wavelength of 3·20 cm. using a B.T.H. type CS3-A. Optimum coupling corresponds to values

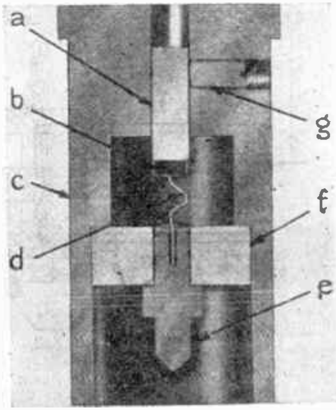


FIG. 7.—CROSS-SECTION OF CS3-B SILICON RECTIFIER UNIT.

a—Plunger.
b—Crystal.
c—Head-end.
d—Tungsten contact.
e—Plug terminal.
f—Ceramic insulator.
g—Grub-screw.

(*B.T.H.*)

of rectified current between 0·3 and 1·0 mA, depending upon design features of the mixer and the efficiency of the intermediate-frequency amplifier.

Fig. 7 is an enlarged cross-section of a Form B rectifier. Although Forms A and D differ in size, shape and arrangement of components, the principle of assembly of all three is basically similar. On the end of plunger *a* is fixed a flake of silicon crystal *b*, the plunger being a running fit in head-end *c*, which is extended to form a screen for the assembly. Tungsten-alloy wire contact *d* is soldered to plug terminal *e*, which is supported by ceramic insulator *f*. Predetermined characteristics are obtained by adjusting the contact pressure between the crystal *b* and the wire contact *d*, after which plunger *a* is locked by a grub-screw *g*. The contact pressure is such that the assembly will withstand normal handling without change of characteristics.

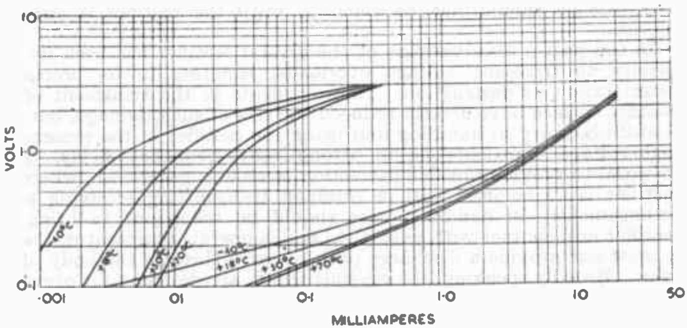


FIG. 8.—EFFECT OF TEMPERATURE VARIATIONS ON AVERAGE CRYSTAL RECTIFIER.

(*B.T.H.*)

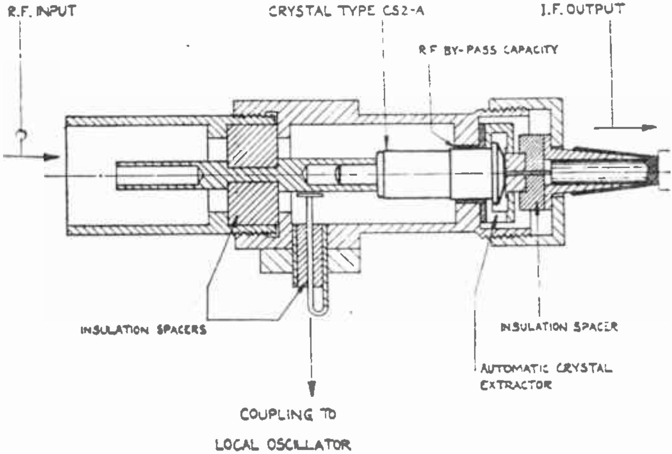


FIG. 9.—B.T.H. CS2-A RECTIFIER USED AS A MIXER.

Fig. 9 shows a B.T.H. CS2-A rectifier used in one of the many types of crystal-holder as a mixer; connections are shown from the signal source and to the beat-frequency local oscillator and to the intermediate-frequency amplifier. Low impedance, small input and output capacitances and negligible electron transit time are features which make the crystal extremely suitable for this type of application.

PRECAUTIONS. Unless already protected by all-metal construction, these rectifiers are usually supplied enclosed in lead capsules as a protection against damage by strong radio-frequency fields, etc. The capsule should not be removed until the rectifier is put into service.

In the past a disadvantage of the crystal rectifier has been its sensitivity to transient voltage overloads, resulting in its premature deterioration or destruction. Improvements in the treatment of the contact surface have greatly reduced the risk of such damage, but care is still necessary in handling and using the rectifier in the presence of high-voltage transients or in strong radio-frequency fields. It is essential to provide suitable protection, such as a gas switch, where the rectifier is to be used with a common transmitting–receiving aerial arrangement. In handling, care should be taken not to bring the rectifier into contact with unearthed mains-operated apparatus, that is, it must not provide a discharge path to earth through the body of the user. Broadly speaking, the crystals which are designed for operation at the highest frequencies require the greatest protection against electrical overloads and mechanical shock. In general, crystal rectifiers should be handled with no more and no less care than would be used with a thermionic valve. Wires must not be soldered direct on to the electrodes, but connection must be made by means of the mounting.

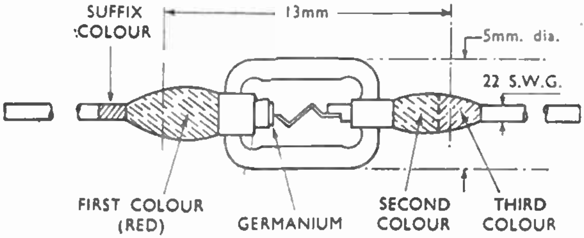


FIG. 10.—ARRANGEMENT OF TYPICAL GERMANIUM CRYSTAL UNIT. (*G.E.C.*)

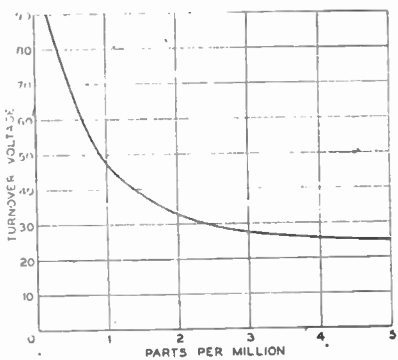
Germanium Crystal Rectifiers

Judged by normal chemical standards, the germanium required for a high back-voltage rectifier is pure. Experience shows that germanium containing more than about 0·3 part per million of impurities is unsuitable. The control of impurities, mainly arsenic, has not yet reached the stage where the characteristics in any given case can be exactly predetermined, and are thus not completely known until manufacture is completed. The grading of crystal rectifiers for turnover voltage and back resistance are therefore matters of selection and segregation from batch manufacture.

In general, the higher the turnover voltage, the greater the back resistance, which may be as high as 1 megohm at a reverse voltage of 50 volts. The magnitude of the reverse voltage also reflects itself in the production cost, so that it is advisable to use a unit with the lowest back resistance which the application will tolerate. Most high-reverse-voltage rectifiers have forward currents up to 50 mA at 1 volt, though this can be increased with a corresponding reduction of back resistance and turnover voltage. Exceptionally low forward resistance is always offset by lower turnover voltage.

Since the resistance becomes negative after the turnover voltage is exceeded, it is sometimes suggested that this region be utilized to secure

FIG. 11.—RELATIONSHIP BETWEEN TURNOVER VOLTAGE AND ARSENIC IMPURITY IN A GERMANIUM CRYSTAL RECTIFIER.



oscillation. Operation under such conditions is not desirable, since the permissible reverse current would probably be exceeded and operation would be unstable due to changes of the characteristics caused by the resulting temperature rise. The turnover voltage should be limited for rectifiers designed to give their best performance at high frequencies. The capacitance is usually less than 1 pF, but varies slightly at high frequencies according to the voltage applied.

The efficiency of a germanium rectifier in the region of 100 Mc/s falls as the frequency is raised, and appears to suffer more when the turnover voltage is large than when it is small. In this respect it is rather less satisfactory than silicon, in which the rectifier efficiency remains constant to frequencies of several thousand megacycles.

Where the load impedance is constant, this tends to nullify any temperature variation in the rectifier impedance, which can be kept low, as the input voltage is of the order of 10 volts. Even where the load is a low resistance, for example, with a measuring instrument with an input of less than 1 volt, the performance is at least equal to that of a metal rectifier.

Germanium diodes will function satisfactorily over a range of $-100°$ to $+200°$ Centigrade.

Junction Diodes

In addition to the point-contact types described in this section, various forms of junction diodes have become available recently: see Sections 25, 26.

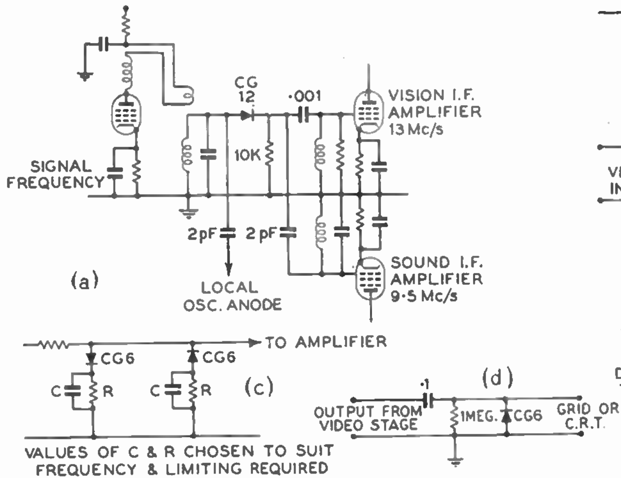
A recent development in germanium junction diodes has been the reduction of the junction area to produce diodes having a capacitance of a few picofarads (compared with some hundreds of picofarads in a conventional junction diode), a high reverse resistance and a hole storage time of a fraction of a microfarad. The forward conductance of such diodes at for example 1 volt may be over 1 amp., compared with the 3–5 mA of a point-contact type. Apart from applications such as waveform clipping, diode modulators, etc., these diodes may also be used as miniature mains rectifiers for transistor supplies, valve-bias supplies, etc.

Another development is the use of a gold wire in place of tungsten, bonded into position by discharging a large pulse of current. These diodes have a forward conductance intermediate between point-contact and small-area junction diodes, but have an even smaller reverse capacitance; a typical figure for forward conductance at 1 volt would be 300 mA.

SEMI-CONDUCTOR VARIABLE CAPACITORS

The capacitance of a germanium or silicon junction diode is a function of the reverse voltage applied across it; this property can be utilized to provide a circuit element in which changes of voltage are used to produce a variable capacitance.

The principle of this device can be explained in simple terms as follows: If a reverse voltage is applied across a p–n junction, the positive voltage tends to attract the electrons in the n region and the negative voltage attracts the positive holes, producing an increase in the space-charge region between the two conductors, thus decreasing the effective



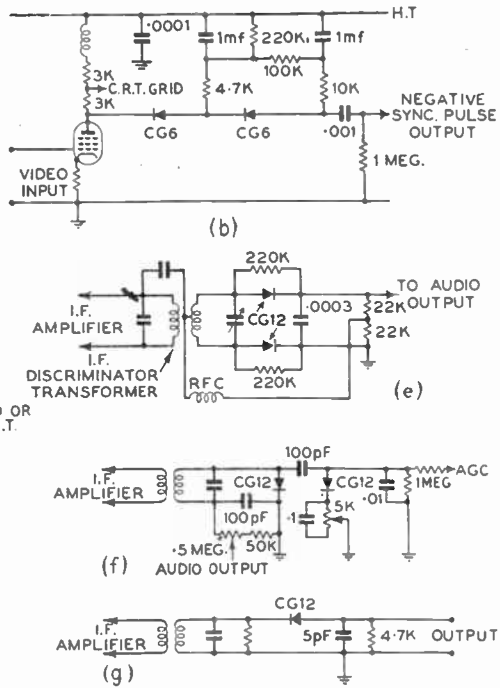


Fig. 12.—Some Radio and Television Applications of the Germanium Crystal Diode.

(a) Frequency conversion circuit.
(b) Synchronizing impulse separator circuit.
(c) Limiter circuit.
(d) D.C. restorer circuit.
(e) U.H.F. discriminator circuit.
(f) Radio 2nd. detector and A.G.C.
(g) 2nd. detector circuit in television band.

capacitance of the diode. As the reverse voltage increases the space-charge region continues to widen, and hence the capacitance of the diode continues to decrease until the point is reached where the reverse voltage is sufficient to cause a breakdown of the diode.

Over its working range the capacitance of a junction diode will approximate $C \propto 1/\sqrt{V}$ and the capacitance of a typical junction diode, such as the OA11, varies from about 13 pF at 3 volts to 7 pF at 13 volts. Diodes already developed specifically for this purpose have an effective range of the order of 3–40 pF, and much higher ranges are theoretically possible.

The Q of the device depends upon the equivalent series and shunt resistances, the Q being increased by a decrease in series or base resistance or by an increase in shunt resistance. A Q of 1,000 at 1 Mc/s can be readily achieved, and diodes may be used for this purpose up to the U.H.F. region for applications in which a high Q is unimportant.

Applications so far described include automatic frequency control for A.M., F.M. radio and television receivers, remote tuning of receivers and transmitters, simple frequency and phase modulators, and for variable reactance and parametric amplifiers.

TUNNEL DIODES

The discovery of a new form of semi-conductor device—termed a tunnel diode—was reported in 1958 by the Japanese scientist Leo Esaki. Although a two-terminal device, these diodes can be used for amplification, oscillation or high speed switching. Experimental models produced during 1959 by the General Electric Co. in the United States are capable of oscillation up to 2000 Mc/s and it is believed that this figure should be increased shortly to about 10,000 Mc/s. As an amplifier, its noise figure compares with the klystron and travelling-wave tube. For switching applications, the tunnel diode is said to be from 10 to 100 times faster than existing transistors.

The diode comprises a heavily doped junction semi-conductor having a negative resistance characteristic over part of its operating range due to so-called " quantum mechanical tunnel " theory which postulates a tunnelling effect in a semi-conductor junction allowing electrons to cross the barrier without change of energy. In a tunnel diode, equal and opposite tunnel currents flow at zero voltage. When a forward voltage is applied, the current in the diode increases owing to the tunnel currents becoming unbalanced. However, when this forward bias is increased the tunnel currents become very small, giving a negative resistance characteristic which can be used for amplification, oscillation and switching.

28. RESISTORS

28. RESISTORS

The resistors used in radio and television can usefully be considered under three heads :

(a) composition types;
(b) wire-wound types;
(c) potentiometers.

The great majority of resistors for radio and television use are for receivers. The power to be handled is usually small, and in general the voltages applied to the resistors are not high. Consequently, the resistors are of small physical size, and these sizes, preferred resistance values, marking and other parameters have been standardized for some time. These standards have met with general acceptance by component manufacturers and set makers alike, with the result that standards of performance have risen, availability has increased and the cost of components has been reduced. It must be understood that resistors for radio and television purposes are mass produced, and they have both the virtues and weaknesses which result from this method of production. Standardization of values and tolerances has greatly aided mass production. Standard tolerances on nominal values range from ± 20 to ± 1 per cent, but tolerances of ± 5, ± 10 or ± 20 per cent are the commonest.

Resistor Values

The normal range of resistance values used in radio and television is from 10 ohms to 10 MΩ, and the "preferred values" associated with

TABLE 1.—FOUR-BAND COLOUR FOR FIXED RESISTORS

Colour	*B.S. 381C No.*	*1st Figure "A"*	*2nd Figure "B"*	*Multiplying Value "C"*	*Tolerance "D" (%)*
Silver . .	—	—	—	10^{-2}	± 10
Gold . .	—	—	—	10^{-1}	± 5
Black . .	—	—	0	1	—
Brown . .	412	1	1	10	± 1
Red . .	538	2	2	10^{2}	± 2
Orange . .	557	3	3	10^{3}	—
Yellow . .	355	4	4	10^{4}	—
Green . .	221	5	5	10^{5}	—
Blue . .	166	6	6	10^{6}	—
Violet . .	796	7	7	10^{7}	—
Grey . .	632	8	8	10^{8}	—
White . .	—	9	9	10^{9}	—
None . .	—	—	—	—	± 20

particular tolerances will be found in Publication No. 63 of the International Electrotechnical Commission. It should be noted that the preferred values for tolerances of ± 5 per cent are generally used for finer tolerances.

Values may be marked in plain figures or by means of a colour code. The colour code employed is set out in *the I.E.C. Publication No.* 62. It may also be found in B.S. 1852 : 1952, in The Government Interservice Specifications RCL. 111 and RCL. 112, and in the appropriate Radio Industry Council Specifications RIC. 112 and RIC. 113.

For convenience of reference, both the preferred values and the colour code are reproduced in Tables 1 and 2.

Colour Coding

The components shall be clearly and indelibly colour coded in accordance with Fig. 1 and Table 1.

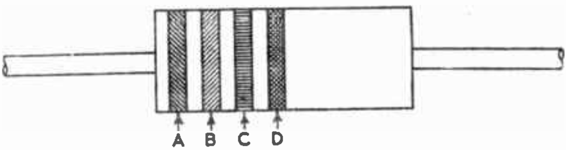


Fig. 1.—Resistor Colour Coding.

General Scheme :

1st Colour " A " denotes 1st digit of resistance value.
2nd Colour " B " denotes 2nd digit of resistance value.
3rd Colour " C " denotes Multiplier.
4th Colour " D " denotes Tolerance.

Table 2.—Series of Preferred Values and their Associated Tolerances

Tolerance			*Tolerance*		
±5%	±10%	±20%	±5%	±10%	±20%
1·0	1·0	1·0	3·3	3·3	3·3
1·1			3·6		
1·2	1·2		3·9	3·9	
1·3			4·3		
1·5	1·5	1·5	4·7	4·7	4·7
1·6			5·1		
1·8	1·8		5·6	5·6	
2·0			6·2		
2·2	2·2	2·2	6·8	6·8	6·8
2·4			7·5		
2·7	2·7		8·2		
3·0			9·1		

With the large expansion of electronics and the use of Grade 1 composition resistors there is a tendency to use the finer tolerances, and there is a considerable body of opinion which considers that the preferred-value series used for ± 5 per cent is not adequate for close-tolerance resistors. In the U.S.A. a new table has been set out for high-stability resistors, using a Renard series which gives values with smaller intervals between them than those given by the Renard series used for the ± 5 per cent tolerance in the I.E.C. publication. Those interested will find the table set out in Specification MIL-R-10509A.

The world-wide use of radio and radar means that radio components have to meet a variety of climatic conditions. These, of course, are no different for resistors than for any other component. The conditions may be high temperature, high relative humidity, a combination of both of these, extreme cold and low air pressure. All these conditions are catered for by the various specifications covering resistors for radio and radar and television use.

COMPOSITION RESISTORS

There are two classes of composition resistors referred to as Grade 1 and Grade 2. This grouping makes it appear that the Grade 2 resistor is much inferior to Grade 1, and in point of fact this is not true. A better classification would be: High-stability Resistors and Normal-stability Resistors. Definition of these grades is to be found in the Government Specification RCG. 110, which reads as follows:

> "*Resistors Composition—Grade* 1.—Resistors whose stability under the test conditions specified in RCS. 11 is within the limits of ± 5 per cent of the initial value.
>
> "*Resistors Composition—Grade* 2.—Resistors whose resistance stability under the test conditions specified in RCS. 11 is within the limits of ± 25 per cent of the initial value."

These definitions are reasonable, but the limits of change allotted to Grade 2 resistors require adjustment, since no resistors are made which have characteristics quite so bad.

Grade 1 Resistors

These may be insulated or non-insulated and consist essentially of a ceramic rod or tube upon which a crystalline film of carbon has been deposited by the cracking of a hydrocarbon gas at high temperature. The resistance range of the film is not great, and it is therefore usual to increase the resistance value of any particular specimen by cutting a spiral groove on the rod with a diamond wheel. The result is a resistor of considerable electrical length, but of small width and thickness. As a result of this construction, Grade 1 resistors are not able to stand much overload, and the film is easily damaged in handling. The resistors are also liable to failure from breaking away of the spiral track owing to thermal working of the ceramic. The pitch of the spiral groove imposes a limit on the voltage which may be applied to the resistor, and all these factors together are affected by the convention that a resistor of given wattage has the same overall physical dimensions, no matter what its resistance value may happen to be.

Manufacturing details are discussed briefly later on, but reference should be made to one important point. The resistor trade urgently needs more and better ceramics for high-stability resistors. The ceramic mus be of high-quality porcelain, and unfortunately the high quality required demands special manufacturing facilities. One of the chief reasons for the lack of good ceramics for high-stability use is economic, since if the whole of the trade demand was supplied from one source, it would not provide a very good living for the maker. All resistor makers investigate hopefully any new source of supply which comes along, but at the time of writing there does not seem to be much hope of larger supplies of good resistor ceramic in the near future.

Grade 2 Resistors

This type forms the great majority of resistors used for radio and television circuits. There are two general types of these resistors, one consisting of a solid rod of composition containing carbon and the other a rod or tube sprayed with a paint containing carbon. As in the case of Grade 1 resistors, all Grade 2 resistors of a given wattage are required to have the same overall physical size, and therefore the specific resistance must vary throughout the range.

Manufacture of Grade 1 Resistors

It is not intended that this shall be a detailed description of manufacture, only a very general outline being given. Grade 1 resistors are of insulated or non-insulated types, and the insulated types vary in their insulating coverings. All manufacturers have differences in their techniques, therefore the process described here does not claim to be exclusive, since it is merely the process of one particular maker. In this case the resistors consist of high-quality ceramic rods upon which the carbon film is deposited. The resistors are terminated with brass caps carrying high-conductivity copper connections. These resistors are insulated, the resistor assembly being put into a high-quality ceramic tube, the ends of which are sealed with a special plastic seal. The completed resistors are vacuum impregnated with a special compound, and so are capable of a very high degree of resistance to humidity.

The carbon film is deposited in a special type of furnace, which may be of batch or continuous-feed type. There is some degree of control of the film in cracking, but as before mentioned not enough to cope with all the values of resistance that are required. To get the necessary degree of control the electrical length of the resistor is altered whilst the physical dimensions are not changed. This is accomplished by grinding a helical groove through the carbon film. A diamond wheel is used for this purpose, and either hand or automatic control of resistance value may be used. After the grinding process the resistors are cased, coded and impregnated.

The standard range of Grade 1 resistor is from 10 ohms to 10 MΩ, and the usual tolerances are ± 1, ± 2 and ± 5 per cent.

Manufacture of Grade 2 Resistors

There are two main kinds of Grade 2 resistors: (*a*) those consisting of a solid composition rod; (*b*) the sprayed-rod variety. We will deal first with the solid carbon composition rod type.

In Grade 2 resistors the standard range is from 10 ohms to 10 MΩ, and the standard tolerances are ± 5, ± 10 and ± 20 per cent. There are 255 preferred values and tolerances in the standard range, and the first problem to be faced by the resistor maker is the number of primary powders he will make. The usual practice is to make as small a number of these as will economically fit the manufacturing process. Most makers of resistors have several wattage sizes in their ranges, which means several sets of physical dimensions.

From first principles we have that

$$R = \rho \times l/a$$

where R = resistance in ohms;
ρ = specific resistance of material used; and
l and a = length and cross-sectional area, expressed in the same units as used for ρ.

It is evident that a powder having any particular value of ρ will give different results when made up into resistors of different physical sizes. Manufacturing experience shows that if a powder gives a resistance of A ohms on a particular set of dimensions, then the resistance given on another set of dimensions will be XA ohms, where X is a factor depending mainly upon the relations between the dimensions. In practice, the factor X is also affected by the material and the manufacturing process used. The result is that only one range of primary powders is necessary. As many as possible of the qualities desired in the completed resistor are built into the primary powders, and they can be blended together to produce values other than those of the straight-run primaries.

The primary powders also contain a resin binder. After moulding to a target value, the resistor pins have the usual heat treatment for polymerization of the resin. The process is usually a continuous one, mechanized as far as may be. The relation of moulding value to finished value differs with different makers. In some cases the finished shape of resistor pins is as moulded, but in others, particularly those of circular cross-section, the pins are ground to finished dimensions after moulding. Terminations are then applied to the pins, which are now ready for grading into finished value. Most manufacturers of solid-rod types vacuum impregnate their resistors to improve humidity characteristics. After the provision of the insulating cover, if any, the resistors are coded and then ready for despatch.

The Sprayed Carbon Types.—These are made up with sprays of various values. Here again the manufacturer makes up a range of sprays having values appropriate to his process. Finishing is upon the same lines as that used for solid-rod resistors. There is one class of sprayed resistor where the spray contains a low-melting-point glass: this resistor is heat treated at a temperature high enough to vitrefy the glass, and variation of value in this case is achieved by the cutting of a spiral groove on the resistor in addition to the range of sprays.

Specifications Used for Resistors

In this country at the present time there are two (almost) parallel sets of specifications for resistors used for radio and television. These are the Government Radio Component Standardization Committee Specifications and the Radio Industry Specifications. The R.C.S.C. Specification is RCS. 112, which covers both grades of resistors. The

R.I.C. Specifications are No. 112 for Grade 1 resistors and 113 for Grade 2 resistors. There are no British Standards Institution specifications for resistors at the present time, but it is intended that there shall be such specifications. The work done by the Standardizing Committees of the Radio Industry which resulted in the R.I.C. specifications had, and still has, ultimate B.S.I. specification in view. It is hoped that when a specification has been produced by the B.S.I. there will be no further need for either R.C.S.C. or R.I.C. specifications.

These specifications can be obtained from:

R.C.S.C. Ministry of Supply,
Castlewood House,
77/91 New Oxford Street
London, W.C.1

The Radio Industry Council,
59 Russell Square,
London, W.C.1.

Besides the individual resistor specifications, there are general specfications for climatic, robustness and durability testing, which are common to all radio and television components. One important point must be made clear. None of the specifications mentioned here are manufacturing specifications; they are performance specifications.

Conditions To Be Met In Service

Quite apart from the question of resistance value and tolerance for a particular job, there is a series of general conditions to be met by resistors during their service, and these will be more arduous in some cases than in others. The operating conditions for a resistor in a domestic radio in this country are very different from those met by a similar resistor in the radio of an air-liner, or even from the conditions met in a domestic radio receiver in the tropics. All of these conditions can be grouped under the following heads:

(*a*) changes of temperature;
(*b*) changes of humidity;
(*c*) changes of air pressure;
(*d*) vibration and bumping;
(*e*) stresses during assembly.

Other necessary conditions are:

(*f*) voltage coefficient;
(*g*) low noise;
(*h*) reasonable ageing characteristics;
(*i*) a reasonable working life;
(*j*) stability under operating conditions.

Under all these conditions the resistors must have a definite standard of performance. Breakage and noise excepted, the response of a resistor to all the changes of conditions quoted can be assessed in terms of its permanent resistance change under those conditions.

Soldering of Resistors.—Resistors are almost invariably connected into circuit by soldering, and it is essential that the leads can be soldered easily up to a point near the body of the resistor. It is also essential that the permanent change of resistance, because of soldering, shall be small. The solder test in the specifications is designed to simulate very severe soldering conditions. In both Government and R.I.C. specifications the test method is the same, but in the Government specification it is laid down that the soldering-iron shall be in contact with the termination for 15 seconds, whilst only 10 seconds is called for in the R.I.C. specifications. The change of resistance is to be measured not less than ½ hour after soldering, i.e., when the resistors have cooled down to room temperature.

Noise

Signal-to-noise ratio is a matter of ever-increasing importance in radio and television, and it is of great importance that the resistors used in the circuit add as little to the noise as possible. All composition resistors have an inherent noise which is proportional to the absolute temperature and the resistance in ohms.

On load this noise is increased owing to rise of temperature and contact variations between adjacent particles of the composition. In the case of Grade 1 resistors the absence of non-carbon particles means more regular contact between the particles, and the noise is low. Between the limits of the normal range of resistors, noise in Grade 1 resistors is considered to vary directly with the applied voltage. In Grade 2 resistors the non-carbon constituents cause greater noise on load, and it is proportional to the log of the resistance value. There is a great deal of controversy about the measurement of resistor noise. The R.C.S.C. specification quotes a "preferred" noise meter, but the other specifications simply state that: "Pending the development of a suitable standard test apparatus for this purpose, the noise shall be measured by a method to be agreed between the purchaser and the supplier." At the time of writing the main approaches to the subject seem to be given by the following references:

1. *R.A.E. Technical Note—R.A.D.* 465: "Measurement of Noise in Resistors", by J. Burgess.
2. American Standards Association—C.75.17—1944: "Method of Noise Testing Fixed Composition Resistors" (being revised).
3. *Alta Frequenza*, Volume 18, pages 254–267: "Noise Level of Resistors"; by E. Paoline and G. Canegalle.
4. *Proceedings of the Physical Society*, 1st August 1953: "The Electrical Conductivity and Current Noise of Carbon Resistors", by I. M. Templeton and D. K. C. MacDonald.

There are significant differences between the methods used, and the results are not directly comparable. The situation to date may be reasonably summed up as follows:

The present specification limits for noise in resistors are met without much difficulty, and there is no pressure from the radio and television industry for resistors with less noise than the specifications demand. There is no widely agreed method or apparatus for the measurement of resistor noise, but there is need for such apparatus. To this end more research on resistor noise and its measurement is necessary.

The specification limits for resistor noise are 0·5 μV/D.C. volt applied for Grade 1 resistors, and $\log_{10} R$ ohms μV/D.C. volt applied for Grade 2 resistors.

Ageing

All composition resistors change in value during storage, and a limit to the amount of change in a given time is included in both R.C.S.C.

and R.I.C. specifications. The test is called " Shelf Life " in RCS. 112. The amount of change depends on the storage conditions in some degree, and in practice resistors, in common with other components, are stored under conditions as good as possible. In the test laid down in RCS. 112 the conditions are deliberately made as bad as can be without actual exposure of the resistors to the weather. In the opinion of the trade these conditions are not representative, and in practice the Government requires resistors to be specially packed if the storage conditions are as bad as those laid down in the test.

TABLE 3.—LIMITS OF CHANGE UNDER SPECIFICATION TEST CONDITIONS

	Grade 1		*Grade 2*	
	R.C.S.C.	*R.I.C.*	*R.C.S.C.*	*R.I.C.*
Temperature coefficient	0·04–0·1%	0·04–0·1%	0·12%	0·12%
Soldering	0·5%	0·3%	3·0%	2·0%
Voltage coefficient	0·002%	0·002%	0·025%	0·025%
Noise	0·5 μV/V	0·5 μV/V	$2 + \log \frac{R}{1{,}000}$ μV/V	$\log R$ μV/V
Climatic	1·0–3·0%	1·0–3·0%	10–15%	10–15%
Working life	1·0–3·0%	1·0–3·0%	25%	10–15%
Tropical exposure	1·0–3·0%	1·0–3·0%	15%	10–15%
Ageing	1·0% 6 months	0·5% 94 days	5% 6 months	2% 3 months

Future of Composition Resistors

In the writer's opinion the composition resistor will not go out of use in the foreseeable future. There will undoubtedly be changes. All composition resistors have greatly improved during the last decade, and this improvement will continue, probably to a point where there will not be two grades of composition resistors. Miniaturization problems will be as troublesome as ever, but perhaps we shall get miniature watts to operate with. It does seem that the transistor will help the miniaturization position, because of the small amounts of power required. Printed circuits are not exactly new problems to the resistor manufacturer; they chiefly mean a change of physical form. These circuits are still to some extent in the experimental stage, although they have already become established as normal technique in the radio and television industry. What their final place will be cannot yet be foreseen.

WIRE-WOUND RESISTORS

Very large numbers of wire-wound resistors are used in radio and television. In the smaller sizes they are usually used where the greater load is beyond the capacity of a carbon resistor. In the larger sizes mains droppers form the majority of wire-wound resistors.

The type of wire-wound resistor used in radio and television consists of a high-quality ceramic former upon which is a winding of iron-free resistance wire, usually a nickel alloy. The winding is nearly always single-layer straight solenoid, but Ayreton Perry winding is used where very low inductance is essential. The terminals may be ferrule, tags or wire tails.

The standard resistance range is from 10 ohms to 100 kΩ, but values higher, and lower, are offered by a number of makers. Tapped units may have as many as ten taps according to make and type. The standard tolerances range from ± 1 to ± 10 per cent, the most common being ± 5 per cent. The great majority of wire-wound resistors used in radio and television have the winding covered with vitreous enamel or some other compound, and most are capable of operating at surface temperatures up to 300° C. Since the surface temperature is high, the resistors must be suitably mounted and care taken that the heat dissipated does not cause damage to other components. The most suitable mounting method for large wire-wound resistors is vertical, and so arranged that air can circulate through the bore. It is standard practice in the industry to make use of the same preferred-value series as used for carbon resistors, though other values can be supplied. Wire-wound resistors are usually marked in plain lettering, and on account of the high surface temperature special care is necessary in the selection of the materials used for marking.

Manufacture of Wire-wound Resistors

The formers upon which the wire is wound are of high-quality ceramic. All but the smallest sizes are tubes. Special machines have been developed for winding, and the process is semi-automatic. The winding may be connected to the terminations by welding, brazing or by pressure joints. Tag and wire terminations are more common than ferrules, and the present tendency seems to be for ferrule terminations to be confined to the largest sizes. Vitreous enamelling processes vary a great deal among manufacturers, some using continuous and others batch coating. Silicone coatings have been developed which can work at a surface temperature of almost 300° C., and these have the advantage from the manufacturers' angle that they do not require as high a temperature for curing as is required for vitreous enamel. The lower temperature means much less scrap due to thermal shock. Wire-wound resistors are also lacquer covered, but these have a much lower working temperature.

Most wire-wound resistors are wound on to formers of circular cross-section, though there are some other forms. Winding wires range in size from about 30 S.W.G. to finer than 0·001 in. diameter. Copper–nickel or nickel–chrome alloys are used and supplied under various trade names, such as Cupron, Nichrome, etc. On the smaller sizes, usually below 15 watts, the terminations may be axial or radial.

POTENTIOMETERS

There are two types of potentiometers used in radio and television. These are wire-wound and carbon-track types. Wire-wound potentiometers are not usually used in radio receivers, but both types are used in television receivers. Carbon-track potentiometers form the biggest proportion of these components.

Carbon-track Potentiometers

These are used in large numbers for volume and tone controls in radio receivers, and for various other controls in television receivers. They consist essentially of a resistance element in the form of a plastic or ceramic ring sprayed with a carbon mixture. These resistance elements are contained in a circular case furnished with a rotating arm for variation of resistance. Contact can be made through tags to each end of the resistance element, and to the moving arm.

These potentiometers are made in values up to about 5 MΩ with linear, logarithmic and reverse logarithmic resistance laws. The usual wattage ratings are up to 2 watts, and this wattage is taken over the whole of the track. Dual potentiometers mounted concentrically can be obtained, arranged for operation either separately or together.

Where carbon-track potentiometers are used as volume controls, an on-and-off switch is usually fitted to the back of the control. The switch is operated by the control spindle, and this operation takes up some of the angular rotation on the potentiometer and cuts down the effective resistance rotation. The resistance sprays are made up in values in the series 1, 2, 5, 10, etc., and the standard tolerances are ±10 per cent and ±20 per cent.

The mechanical details fixed by specification are :

Overall	Diameter, depth including tags and switch, projection of tags and angular rotation.
Fixed Bush	Length, thread length, diameter and thread details.
Fixing Nut	Thickness and diameter across flats.
Spindle	Length, diameter, length of flat, dimension perpendicular to flat, angular position of flat, details of slot or cross hole, and spindle-end chamfer.
Locating Peg	Position, height and diameter.
Switch	Angle of operation.

The switch rating is 1 ampere at 250 volts A.C. or D.C. and 3 amperes 125 volts A.C. or D.C. The contact resistance of the switch is not to exceed 0·01 ohms, and the rating is to be marked on the switch.

Performance Tests

Spindles must be insulated from the moving contact, and the insulation resistance between the spindle and the live parts of the potentiometer—switch included—must not be less than 500 MΩ measured at 500 volts D.C. It is proposed to increase this insulation resistance to 1,000 MΩ. Carbon-track potentiometers must be able to withstand, without breakdown, 2,000 volts r.m.s. 50 c/s applied for 15 seconds between the live parts of the switch and the spindle, and 1,000 volts r.m.s. 50 c/s between the whole of the live parts of the control and the spindle. For these tests the spindle and case are to be connected together electrically. Besides their resistance, potentiometer tracks are required to have low "hop-on" and "hop-off" resistance. "Hop-on" is to be less than 50 ohms or 0·5 per cent of the total track resistance, whichever is the less. The "hop-off" is to be less than 1 per cent of the track resistance. Limits are laid down for law variation, and these are shown in Fig. 2.

A very important point concerning carbon-track potentiometers in service is rotational noise, and RIC/122, Appendix "A", describes the apparatus to be used for the measurement of this noise. During test the potentiometer is to be worked at approximately thirty operations per minute, and the noise measured between the slider and the anti-clockwise tag. The noise must not exceed 50 mV.

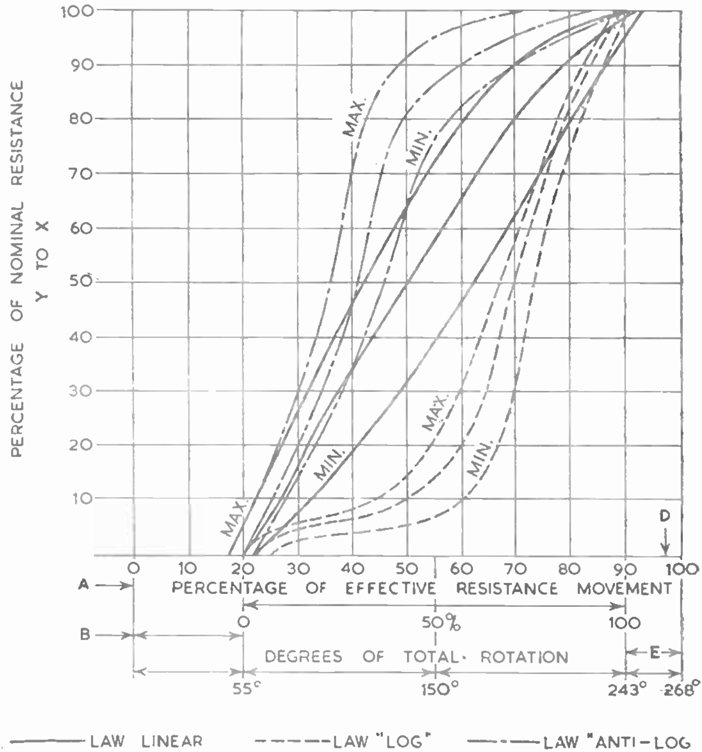


Fig. 2.—Resistance Law Limits for Carbon-track Potentiometers.

There are various mechanical requirements to be met, such as operating torque, end play, side play, robustness of stops and tags.

Endurance tests which the potentiometers must be capable of meeting, include 1,000 hours application of the rated wattage across the track, 50,000 cycles of operation at thirty operations per minute and 10,000 operations of the switch under conditions where the full rated current is broken. At the end of these tests there are specification requirements covering wear, change of resistance, etc., to be met. Lastly, there is a soldering test to ensure that the control can be satisfactorily connected into the circuit.

The 100 per cent production tests are: general examination for workmanship; measurement of overall resistance, hop-on and hop-off; law; rotational noise; flash insulation; functional test of switch.

Carbon-track potentiometers are relatively complicated pieces of apparatus, and in spite of their limitations give good service. This is best when there is no switch attached.

The trend of development at the present time is towards reduction of size and better humidity characteristics.

Wire-wound Potentiometers

These are used in considerable numbers in television receivers, and two forms are in use at the present time. One consists of a resistance winding on a ceramic former with a sliding contact arranged to move along the length of the coil. In the other, and more numerous, form the resistance element takes the form of a strip of plastic material carrying a winding of resistance wire. This is mounted on edge and, bent into a circular form, is placed in a plastic moulding. A moving contact attached to a spindle varies the resistance by rotation of the spindle.

Many of the mechanical details fixed for carbon-track potentiometers apply equally to the rotary type of wire-wound potentiometer. The usual range of these potentiometers is up to 100,000 ohms with tolerances of ±5 per cent and ±10 per cent. For television purposes wattage

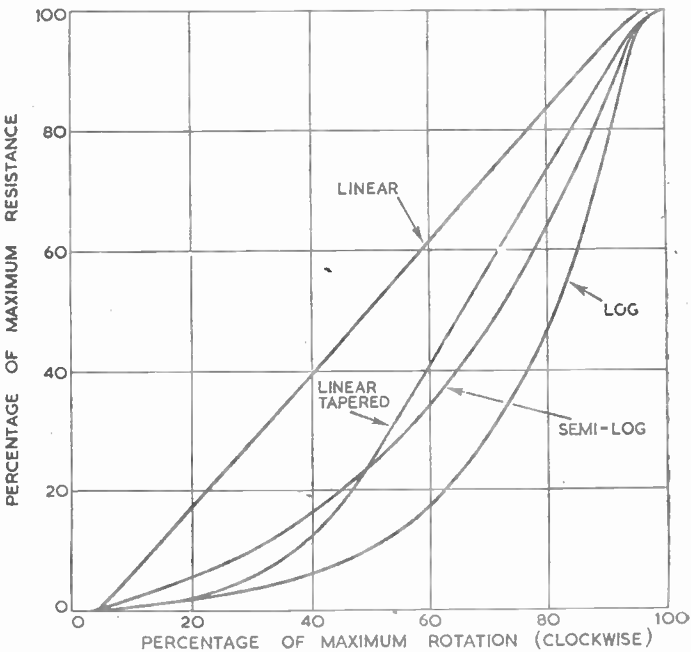


FIG. 3.—WIRE-WOUND RESISTANCE LAWS

ratings of up to 3 watts are used. The nominal wattage is considered as applied over the whole element. The limiting working temperature of potentiometers having plastic formers is 130° C., the wattage ratings being stated at 70° C.

Performance Tests

The commercial specification covering performance of wire-wound potentiometers is RIC. 1·21. The robustness and durability tests laid down in RIC. 11 also apply. In these potentiometers the spindle must be insulated from the contact arm and the resistance element. Winding wires are of nickel alloy, varying in size according to the resistance.

The resistance laws usually required are shown in Fig. 3. Of these the most common in use is linear. The mechanical and robustness standards for wire-wound potentiometers are approximately the same as those of carbon-track potentiometers, but there are extra end-thrust and side-thrust requirements to be met. The temperature coefficient must not be greater than 0·025 per cent per ° C.

Endurance tests for wire-wound potentiometers are divided into electrical and mechanical sections. The electrical requirement is that the rated power shall be applied to the element of the potentiometer and the maximum rated voltage for the insulation shall be applied between the spindle and mounting plate connected together and the element. The test period shall be eighty-four days, with the voltage applied to the potentiometers for 12 hours continuously within each 24-hour period.

Mechanical endurance tests for wire-wound potentiometers differ for panel-mounting and preset controls. Panel-mounting types must have at least 30,000 operations at the rate of thirty operations per minute, each operation traversing not less than 90 per cent of the total angular movement. The maximum rated voltage for the insulation is to be applied during the whole of the test. During half of the test the rated wattage is to be applied to the element with no current through the slider, and during the other half, current is to be applied to the slider through a resistor of the same value as that of the potentiometer under test. At the end of the tests there are electrical and mechanical standards to be met. For preset potentiometers the test described above is reduced to 2,000 operations. For full details of the electrical and mechanical standards to be met at the end of the tests the specifications should be consulted. Climatic tests differ in severity with the grade of the potentiometer. There are three grades: red, green and yellow, red having the highest climatic qualities.

All the tests described above are type-approval tests, which are also applied as batch tests. In production every control has to pass the following tests: (1) general examination; (2) dimensions; (3) continuity; (4) resistance value; (5) voltage proof.

Wire-wound potentiometers, though complicated in construction, are very reliable components. The trend of development at the present time is towards simplification of mechanical detail and smaller sizes.

D. F. U.

29. CAPACITORS

29. CAPACITORS

Any two conductors which are electrically insulated from one another have the property of storing an electric charge when a potential difference is maintained between them, and the charge is proportional to the potential difference. The ratio of the charge to the potential difference is the capacitance between the conductors. Capacitance may be defined in various ways, and British Standard 205 : 1943 gives the following definitions :

(1) The property of a conducting body by virtue of which a quantity of electricity has to be imparted to it to produce a difference of potential between it and the surrounding bodies.

(2) The ratio of the charge on a conductor to its potential when all neighbouring conductors are at zero potential.

(3) The ratio of the charge of a capacitor, i.e., the total electric flux between its electrodes, to the potential difference between them.

(4) Of a circuit-element at a given frequency—the quotient of the susceptance divided by the angular frequency.

The practical unit of capacitance is the farad (F), which may be defined as the capacitance of a capacitor which stores a charge of one coulomb when the potential between its electrodes is maintained at one volt. Commonly used subdivisions of the farad are the micro-farad (μF) equal to 10^{-6} farad and the micro-micro-farad ($\mu\mu$F) or pica-farad (pF) equal to 10^{-12} farad. In electrostatic calculations the centimetre is frequently used as the unit of capacitance (1 pF = 0·899 cm.).

Dielectric Constant

The dielectric constant or permittivity of a material is defined as the ratio of the electric flux density produced in the material to that produced in free space (i.e., a vacuum) by the same electric force (see B.S. 205). The dielectric constant of air is 1·006, which for practical purposes may be taken as unity. Thus the dielectric constant of a substance is the ratio in which the capacitance between two electrodes is increased when the space between them is filled with the substance in question instead of air.

The Capacitor

A capacitor is a circuit element in which two conductors or electrodes are arranged to have a specified value of capacitance. The capacitance of a single electrode or conductor is sometimes quoted, and the second electrode is then assumed to be earth. The energy stored in a capacitor is actually stored in the insulating material between the electrodes in the form of an electric stress, and in a capacitor this insulating material is usually called the "dielectric", the term "insulator" being used when the function of the material is only to prevent the passage of current.

Capacitance and Reactance

The capacitance of two simple geometrical forms is given by the following formulæ :

PARALLEL PLATES

$$C = \frac{\kappa A}{4\pi t}\text{ cm.} = 0{\cdot}0885\frac{\kappa A}{t}\text{ pF}$$

where κ = dielectric constant;
A = area of one plate, in sq. cm.;
t = distance between plates, in cm.

CONCENTRIC CYLINDERS

$$C = \frac{\kappa l}{2\log_{\epsilon}\frac{r_2}{r_1}}\text{ cm.} = \frac{0{\cdot}242\ \kappa l}{\log_{10}\frac{r_2}{r_1}}\text{ pF}$$

where κ = dielectric constant;
l = length of cylinder, in cm.;
r_1 = outside radius of inner cylinder, in cm.;
r_2 = inside radius of outer cylinder, in cm.

When an alternating potential is applied to a capacitor a current passes proportional to its capacitance (C) in farads, the applied potential (E) and the frequency of alternation (f) according to the relation $I = 2\pi fCE$. This property, together with the fact that the dielectric resists the passage of D.C., provides the basis of a means of separating alternating and direct currents. The quantity $X = \frac{1}{\omega C} = \frac{1}{2\pi fC}$ is the reactance of the capacitor.

Power Factor and Insulation Resistance

In addition to capacitance it is often desirable to know two other characteristics of a capacitor, viz., power factor and insulation resistance or insulance.

The power factor of a capacitor indicates the loss of energy which occurs when an A.C. is passed through it. Thus if a voltage E across a capacitor produces a current I through it and the resultant loss in the capacitor is W watts, then the power factor is W/EI.

The loss in a capacitor may be due to several factors which may be represented by equivalent shunt and series resistances, but at a given frequency it is convenient to lump all the components of loss together as a single equivalent series resistance r. Assuming the applied voltage to be sinusoidal, we may use the vector triangle shown in Fig. 1 to derive the power factor.

The loss angle δ is usually small in a capacitor, so that for all practical purposes we may take $\cos\phi = \tan\delta$.

In electronic circuits the power factor of a capacitor should be considered from two aspects. First, in a tuned circuit the losses will increase the damping, and the value of the equivalent series resistance should be considered from this point of view. Secondly, when a heavy A.C. is passed, as in the case of a reservoir capacitor following a rectifier

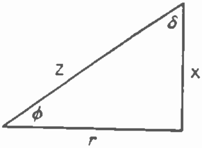


FIG. 1.—POWER FACTOR OF CAPACITORS.

X = reactance of capacitor = $1/\omega C$;
r = equivalent series resistance representing total losses;
Z = impedance of capacitor = E/I;
ϕ = phase angle;
δ = loss angle;
$\cos\phi$ = power factor = r/z;
$\tan\delta$ = loss tangent = r/x.

or a by-pass capacitor in a large transmitter, the heat generated in the capacitor may be considerable, and since it is proportional to the power factor, the latter must be sufficiently low to prevent overheating.

Insulation resistance of a capacitor is usually measured under D.C. conditions by applying a steady voltage and observing the current which flows. Since this leakage current is proportional to capacitance (other things being equal) the product of resistance and capacitance is generally quoted for a capacitor and expressed in megohm micro-farads or ohm farads. The insulation resistance varies with time of application of the voltage, increasing rapidly at first and then tending asymptotically to the final long-time value (Fig. 2). It is usual to quote the value as measured after one minute's application of voltage.

If a capacitor has an insulation resistance of R ohms and a capacitance C farads, the quantity CR is called the "time constant" of the capacitor, and is, in fact, the time in seconds taken for a charge in the capacitor to decay to 0·3679 of its initial value, assuming R to remain constant. As explained above, R will not usually remain constant unless the charging voltage has been applied for a considerable time. However, the method of timing the discharge may often prove useful as a means of estimating CR, and hence R; especially when R is very large, as is the case with polystyrene dielectric capacitors. To make the measurement the capacitor is charged to a potential V_0 and then isolated from the charging circuit. After a time t seconds the voltage V across the capacitor terminals is measured (using a meter of suitably high resistance) and the value of CR calculated from the equation $CR = t/2{\cdot}3 \log \dfrac{V_0}{V}$.

If V_0 is nearly equal to V, then a reasonable approximation is given by $CR = \dfrac{V_0\,t}{V_0 - V}$.

Series and Parallel Connections

If a number of capacitors having capacitance values C_1, C_2, C_3 . . . etc., are connected in series, the resulting capacitance value C is given by the relation

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} + \ldots \text{etc.}$$

Alternatively, this may be expressed by saying that the reactance of the series combination is the sum of the separate reactances, or

$$X = X_1 + X_2 + X_3 + \ldots \text{etc.}$$

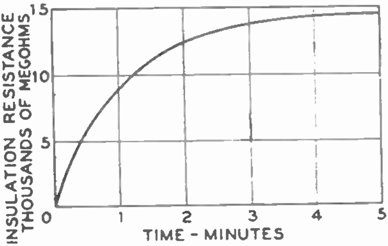


FIG. 2.—CURVE SHOWING VARIATION IN INSULATION RESISTANCE WITH TIME AFTER THE APPLICATION OF VOLTAGE. TYPICAL CURVE FOR A 0·5-μF 500-VOLT WORKING CAPACITOR.

When capacitors are connected in parallel the resulting capacitance is the sum of the separate capacitances, i.e.,

$$C = C_1 + C_2 + C_3 + \ldots \text{etc.}$$

Two or more capacitors are sometimes connected in series for working at voltages higher than their individual ratings, and when this is done it is important to remember that for D.C. the voltage will be shared between the capacitors in proportion to their insulation resistance, while for A.C. the voltage will be divided in proportion to the reactance of the capacitors. In the D.C. case the voltages across the capacitors may be equalized by connecting resistors of equal value across each capacitor. The value of the resistors should be small compared with the insulation resistance of the capacitors, but otherwise as large as can conveniently be used. A value of 1–10 MΩ is usually satisfactory.

J. H. C.

PAPER DIELECTRIC CAPACITORS

Paper as a dielectric has been in use since the earliest days of capacitor manufacture, though it is always used in conjunction with an impregnant such as oil, petroleum jelly, wax, or a synthetic compound having suitable properties. It is uniquely suited to the manufacture of capacitors, since it can be made in long lengths of any suitable width, and may readily be rolled with foil interleaving to form a capacitor having any given value.

Construction

Paper dielectric capacitors are normally made by winding from reels of the tissue and reels of metal foil on to a small-diameter former of suitable length, sufficient paper and foil being wound to give the required capacitance; at appropriate intervals terminating lugs are inserted, these lugs making intimate contact with the foils. With winding complete, the wound capacitor is dried, impregnated, assembled into a protective case, terminating wires attached to the lugs and the assembly finished off.

Capacitors employing metal-foil conductors are never wound with less than two dielectric tissues between foils, the reason for this being that no dielectric tissue is perfect, there being many small discontinuities and conducting particles along its length. To wind a capacitor with a single tissue would inevitably lead to either a complete short-circuit or to an extremely low breakdown voltage, and to obviate this two dielectric tissues are used, since it is considered unlikely that the discontinuities or conducting particles in the respective tissues would coincide. By this means a satisfactory capacitor can be made. For higher-voltage operation the number of dielectric tissues may be increased, and up to five or even six are used commercially.

It may thus be stated that so far as the working voltage of a paper capacitor is concerned, one of the papers must always be considered to be a short-circuit. As proof of this, the breakdown voltage of a capacitor wound with three tissues is approximately double that of a capacitor with two tissues. This effect does not, however, hold for the dielectric properties of the capacitor, and the capacitance will be strictly dependent on the overall thickness of the dielectric.

Single dielectric tissues may be used if metallizing is employed for the electrodes instead of metal foils, since the thickness of the metallizing film is such that in those areas where conduction occurs sufficient energy is fed in to cause the metallizing to be vaporized around the conducting particle or gap, and the breakdown cleared. Due, however, to the inherent weakness of a single tissue, breakdowns can occur at any time during the operating life of the capacitor, with the result that pulses of current are caused to flow, and in many circuits would be undesirable, there is also a small fall in capacitance due to a reduction in electrode area whenever a breakdown is cleared.

For low-voltage D.C. operation, two or even three paper tissues may be used, increasing to up to five or six papers for voltages of a few thousand volts, but for higher voltages it is necessary to series-connect numbers of capacitors, the voltage dividing according to the number of series sections. It is essential to ensure that the potential across any one section does not exceed its rating, and series connected capacitors are therefore individually measured for insulation resistance, all the capacitors in a series stack being matched to within ± 10 per cent of each other. In the case of high-voltage A.C. operation, series stacks will also be necessary, but the voltage across the stack will divide inversely in proportion to the capacitance and not to the insulation resistance. Because of this, the sections are matched for capacitance to within limits of ± 10 per cent of each other, so that again the voltage across each section is kept within the specified limits. These methods of series connection apply essentially to the capacitor manufacturer and not to the user, who, if he has a large number of individual finished and sealed capacitors and wishes to connect them in series so as to operate at a higher D.C. voltage, must use bleeder resistors across each capacitor, the resistance of the bleeder chain being low in comparison with the insulation of the capacitors; there is then no fear of the voltage distribution being such as to become excessive across any one section. Similarly, for A.C. operation the user must measure the capacitance of each capacitor and grade them so as still to maintain a voltage across each section, which is within the specified rating.

The smaller types of paper tubular capacitors as used in radio and television receivers have tin–lead lugs inserted when the capacitor is being wound, the lugs making contact with the respective foils. Wires are then soldered to each lug and the capacitor vacuum dried and impregnated and fitted into its protective tube, which may be made of either metal, bakelized paper or cardboard, and sealed off, the capacitor being then complete.

TABLE 1.—DIELECTRIC PROPERTIES OF IMPREGNANTS

Impregnant	*Melting Point (°C.)*	*Power Factor*	κ	*Insulation Resistance ($M\Omega/\mu F$)*
		For Impregnated Paper		
Petroleum jelly	40	0·003	3·8	10,000
Hydrocarbon wax	50–80	0·003	3·8	5,000
Chlorinated wax	90–130	0·005	5·4	1,000
Transformer oil	—	0·003	3·8	10,000
Chlorinated oil	—	0·003	3·4	10,000
High-viscosity mineral oil	—	0·003	3·8	10,000

With larger capacitors it is necessary to wind in several lugs, which are connected together so as to reduce the contact resistance with the foil, and thus obviate high power factor and at the same time improve the current-carrying capacity of the lugs, this latter point being of great importance in the case of A.C. operation, where appreciable currents may flow. This method of termination is also necessary for pulse operation, where the capacitor may be charged and discharged rapidly; though in many cases, to reduce the resistance of the lugs even further, the foils themselves are extended beyond the edges of the dielectric and contact made direct to them.

Petroleum jelly or a wax impregnant is invariably used on the smaller types of capacitors, particularly those in a waxed-paper or cardboard-tube construction, since at the temperatures obtaining in domestic receivers the impregnant is still a solid and does not flow. Petroleum jelly is also used extensively for larger capacitors in metal-can construction, though oil is coming into more general use, particularly for higher voltages, since voids which might cause ionization are obviated. Special viscous oils, both mineral and synthetic, have been produced in recent years, and have done much to increase the voltage ratings and reliability of capacitors.

The electrical properties of paper dielectric capacitors are dependent mainly on the impregnant which is used, and for convenience their properties are classified in Table 1. The important features for the small types as used in radio and television are the capacitance and insulation resistance, whilst for A.C. operation (particularly for power-factor correction) the power factor must also be included, since a high power factor will mean excessive dissipation, causing a temperature rise which may ultimately lead to deterioration of the impregnant and breakdown of the dielectric.

High-temperature operation will also affect the insulation resistance of the dielectric, the resistance approximately halving for each 10° C. rise in temperature. This is clearly indicated in Fig. 3.

The importance of this fall in *IR* becomes more apparent when the curve is examined closely, as it may be seen that a capacitor having an insulation of around 50,000 MΩ at 20° C. will have an insulation of only 200 MΩ at a temperature of 100° C., and if used as a coupling

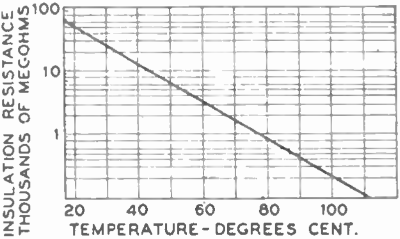


FIG. 3.—CURVE INDICATING VARIATION OF INSULATION RESISTANCE WITH TEMPERATURE.

capacitor between the anode and grid of two valves in an amplifier stage, the operating characteristics of the valves are likely to be changed, possibly with some detrimental effect.

As a direct result of the fall in insulation resistance, the breakdown voltage of the dielectric is also reduced, and for that reason capacitors intended to operate at high temperatures are always derated, an example of a derating for a particular dielectric being 500 volts working at 71° C. and 350 volts working at 100° C. Temperatures of the order of 100° C. are unlikely to be met in domestic equipment, but are frequent in Service equipment, where physical size is frequently limited and ventilation may not be possible because of sealing of the complete equipment.

Interference Suppression Capacitors

Capacitors required for suppression of interference from electrical equipment have particularly rigid requirements, since not only must there be a large safety factor against breakdown of the dielectric, but the insulation resistance should also be very high and the series inductance as low as possible. For this reason many suppressor capacitors are constructed with the foils extended at each end instead of having lug terminations. In this way the inductance may be kept to a minimum. The importance of the series inductance is indicated in Fig. 4, which gives the impedance of a typical capacitor over a range of frequencies.

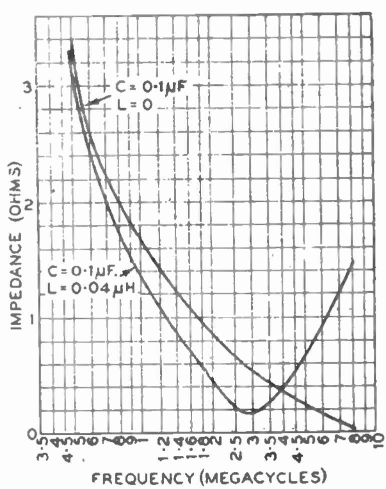


FIG. 4.—EFFECT OF RESIDUAL INDUCTANCE ON THE IMPEDANCE OF A 0·1-μF PAPER TUBULAR CAPACITOR.

As many suppressor capacitors are required to be of small size and have to be fitted into electrical equipment, such as motors, which operate at a high temperature, the impregnant used is usually a chlorinated hydrocarbon which has a higher κ than petroleum jelly or mineral wax and gives an increase of approximately 30 per cent in capacitance. The melting point is also high, being no less than 90° C., so that there is little likelihood of the impregnant seeping.

Tropical Requirements

Mention has already been made of the smaller types of paper capacitors in tubular construction, large numbers of which are used in domestic receivers and other small electrical apparatus. The capacitance values required for these applications, as well as the voltage ratings, enable capacitors of small size to be employed. The protection may be relatively cheap, and will usually consist of a thick paper or cardboard tube, the whole assembly being impregnated in wax, with an additional covering of wax on the outside.

For tropical use, where temperatures and humidity are high, or where greater reliability is required, as for example in equipment for the Services, marine radar, aircraft, etc., an improved protection becomes essential. Experience has shown that wax-protected capacitors will fail due to the ingress of moisture, the wax absorbing moisture from the atmosphere, which gradually penetrates, and ultimately causes a fall in insulation resistance, leading to breakdown. An improved method of sealing, capable of withstanding extreme humid conditions, may consist of a metal tube with synthetic-rubber end seals, these being either bungs or corks tightly spun into the ends of the tube: an alternative construction may consist of a rubber-and-bakelite laminate in the form of a disc securely fitted at each end of a metal tube.

Where the highest insulation resistance is required, the seals may be made of glazed ceramic discs, metallized and soldered into position. Even better results have been achieved recently by the use of P.T.F.E., which may be operated at temperatures in excess of 100° C. and which has outstanding water-repellent properties. Ceramic terminals of various shapes and sizes are also used on larger metal-box capacitors, and are gradually replacing the moulded bakelite terminals which were standard for many years, the bakelite failing under the extreme humidity conditions called for in current specifications.

Smoothing Capacitors

For smoothing D.C. supplies of perhaps 250–50,000 volts, capacitors are assembled in metal boxes with terminals, the construction of the box varying according to the size of the capacitor; with small capacitance, low-voltage types, the case may be made of tinned iron folded and soldered to shape with integral mounting feet. With higher voltages and capacitances having a large bulk, heavy-gauge iron is used for the containers: for radio transmitters, where capacitors having a volume of several cubic feet may be used, metal tanks are employed, these being for either floor mounting on castors or for fitting into suitable racks. The smaller-capacitance, high-voltage types as used for E.H.T. smoothing in television receivers are fitted in bakelite or other plastic cases having a tubular construction so as to increase the leakage path and prevent flash-over.

Very large numbers of capacitors are also employed for power-factor correction of a reactive load, so as to reduce the load on the incoming mains and at the same time to reduce the operating costs of the equipment. Power-factor-correction capacitors may vary in size from small tin boxes necessary for a few VA, as for instance in fluorescent-lighting equipment, up to large oil-filled tanks having a volume in excess of 50 cu. ft. and operating on loads of 250 kVA.

A development of recent years has been the use of capacitor discharges in gas tubes to produce high-intensity, short-period flashes for photographic work: capacitors for this application have a somewhat high voltage rating for the dielectric, and an impregnant having a higher κ is used. These factors combine to give a large capacitance suitable for operation at 2 or 3 kV D.C. yet having relatively small bulk.

Development work on the production of capacitors having increased voltage ratings, higher operating temperatures and smaller volume are continuing, though in some cases equipment manufacturers' requirements are mutually contradictory.

W. I. F.

MICA CAPACITORS

Mica is a mineral which occurs in various parts of the world and, of the several varieties known, that called Muscovite has the best electrical properties from the point of view of capacitor manufacture. The best Muscovite mica is found in Bengal.

Construction

A characteristic of mica is that it may be readily split into thin, uniform, flexible sheets, and these, in thicknesses ranging from 0·001 to 0·003 in., are used as dielectric in capacitors. The mica sheets are cut to shape, usually rectangular, and may be used singly with a metal foil as electrode on each side or, to achieve larger capacitance values, several sheets may be stacked with metal foils inserted between each layer, alternate foils projecting from opposite sides of the stack for connection to the terminal wires.

In the silvered-mica capacitor the electrodes comprise thin layers of silver fired on and firmly adhering to the mica sheet. Here also single mica sheets or multiple layers may be used.

Contact with the silver layer is made by inserting metal foils to which the terminal wires are welded or soldered, or the terminal wires are sometimes attached to the silver layer by soldering or sintering.

The capacitor element is enclosed in a protective casing, which usually consists of either a phenol–formaldehyde resin moulded directly on to the element or a layer of micro-crystalline hydrocarbon wax applied by dipping. These finishes provide insulation and protection against atmospheric moisture. Sometimes the mica capacitor element is finished with a coat of enamel only, which results in economy of space but gives only a poor degree of insulation and protection against humidity. The maximum resistance to humidity is attained in a relatively recent range, in which the elements are embedded in a thermosetting resin of the epoxy type.

Properties

The mica capacitor has inherently high insulation resistance and low power factor, although these properties may be modified by atmospheric humidity and by the container or coating compound used to insulate and protect the element.

Silvered-mica capacitors have good capacitance stability and a small positive temperature coefficient (about $30 \times 10^{-6}/°$ C.).

The power factor of good-quality mica is about 0·0003, and this is maintained at radio frequencies: thus mica is a good dielectric for use in radio-frequency tuned circuits where high Q values are required and where large amounts of power are handled. For radio-frequency current ratings greater than 100 mA, the thin electrodes of the silvered mica are usually unsuitable, and the foil electrode is preferred. The foil-type capacitor can be made as stable as the silvered mica by suitable clamping, but this increases the size considerably.

A single mica element should never be operated on A.C. at a voltage greater than 250 volts r.m.s. If the applied voltage is appreciably in excess of this value a very marked increase in power factor will occur, due to the ionization of air which cannot be completely removed from the mica sheets. This applies even to units having a high D.C. voltage rating of, say, 1,000 or 2,000 volts.

To cater for high A.C. voltages a number of capacitor elements are connected in series so that no element is subjected to more than 250 volts r.m.s.; transmitting-type capacitors designed to operate at large radio-frequency voltages are built on this principle.

The inductance of a mica capacitor is usually small: of the order of 0·01–0·02 μH for the small types used in radio receivers.

Applications

Mica capacitors in capacitance values ranging from 50 to 5,000 pF are commonly used in radio receivers for coupling and decoupling in radio-frequency stages and for padding and tracking in tuned circuits, silvered micas being mostly used because of their smaller size and their stable properties.

In radio transmitters the mica capacitor plays a very large part. It handles very large powers, and the high values of current and voltage involved are allowed for by suitable series–parallel arrangements of elements. Frequently air blast or oil cooling is employed.

Special Types

A type of mica capacitor of special importance in ultra-high-frequency circuits is the disc-type lead-through capacitor. This consists of a stack of mica rings interleaved with foil ring electrodes, the latter projecting alternately on the inside and on the outside of the mica rings. The inner projecting electrodes are connected to a conductor passing axially through the centre, while the outer projecting electrodes are connected to the case, which is arranged for connection to a metal plate, such as a chassis or screen. Such a capacitor is most useful for de-coupling at high frequencies, because it has no measurable inductance and appears to be quite free from the resonances which mar the performance of paper dielectric capacitors of the lead-through type. Capacitors of this type have proved very successful for decoupling radio-frequency stages of television receivers, and for interference suppression on D.C. rotary generators and D.C. or universal motors, where they can form the actual terminals inserted in the frame of the machine.

Future Trends

Future developments are likely to be confined to improving the methods of protecting these capacitors by means of new resins and plastics and improvements in the technique of using these materials. Mica has been synthesized on a small scale in the laboratory, but the process is very expensive. It seems most unlikely that this process will be developed to the state where synthetic mica could compete with the naturally occurring mica, or with other synthetic dielectrics.

J. H. C.

CERAMIC DIELECTRICS

Ceramics are perhaps the most interesting of the dielectrics in use today, and a wide variety of different types are available, each having properties suitable for particular applications. For convenience they may be classified under two broad headings, these being the low dielectric constant (low κ) types and those having high dielectric constant (high κ).

Low κ Types

Among the useful properties of the low κ types there is a wide range of dielectric constants varying from 6 to a figure of around 260 and power factors at radio frequency of approximately 0·0001, though this figure is greater at low frequencies. There are also wide variations in the temperature coefficient of capacitance, varying from positive 120 parts/million/° C., to negative 3,000 parts/million/° C. These properties may be used for stabilizing tuned circuits, and particularly oscillator circuits, where there would otherwise be a shift in the resonant frequency due to temperature variations (see Section 14).

High κ Types

Whilst the low κ types are used essentially for coupling and tuning of radio-frequency circuits, the high κ types are intended for decoupling purposes, the very high dielectric constant of between 1,400 and 3,000 permitting a large capacitance to be obtained in a small volume. Power factors in the range of high-κ dielectrics are appreciably greater than those of the low-κ dielectrics, but since for decoupling purposes relatively small powers are handled, this is no disadvantage.

TABLE 2.—COLOUR CODE FOR CERAMIC CAPACITORS

Colour	*End Colour (Temperature Coefficient)*	*1st Dot (1st Significant Figure)*	*2nd Dot (2nd Significant Figure)*	*3rd Dot (Multiplier)*	*4th Dot (Tolerance)*	
					10 pF or Less	*More than 10 pF*
Black	NP0	0	0	1	±2·0 pF	±20%
Brown	N030	1	1	10	±0·1 pF	± 1%
Red	N080	2	2	100	—	± 2%
Orange	N150	3	3	1,000	—	± 2·5%
Yellow	N220	4	4	10,000	—	—
Green	N330	5	5	—	±0·5 pF	± 5%
Blue	N470	6	6	—	—	—
Violet	N750	7	7	—	—	—
Gray	P030	8	8	0·01	±0·25 pF	—
White	P100	9	9	0·1	±1·0 pF	±10%

TABLE 3.—PROPERTIES OF CERAMICS

Basic Material	*Trade Names*	κ	*Power Factor*	*Temp. Coeff.*
Clinoenstatite	Frequentite Frequelex Calit	6	0·0005	$+ 120 \times 10^{-6}/°$ C.
Magnesium titanate	Tempradex Tempalex Tempa S	14	0·0001	$+ 100 \times 10^{-6}/°$ C.
Rutile (Titanium dioxide)	Faradex Permalex Conda C	90	0·0003	$- 750 \times 10^{-6}/°$ C.
Strontium titanate	—	260	0·0005	$-3{,}000 \times 10^{-6}/°$ C.
Barium titanate	—	1,400	0·01	—
Barium–strontium titanate	—	3,000	0·02	—

Ceramic dielectrics, together with their properties and trade names, are listed in Table 3.

Ceramics for Radio-frequency Operation

Table 3 shows that titanium dioxide and magnesium titanate have exceptionally low power factors at radio frequencies, and for this reason are used very extensively in transmitters and dielectric heating oscillators. A favoured design is to make the capacitors in the form of hollow tubes so that they may be air cooled by either convection or air blast. In many cases the capacitors are cooled in the same air blast as is used for air-cooled transmitting valves, enabling a compact and economical design to be achieved.

Capacitance values of 100–4,000 pF are included in the range of transmitting-type capacitors, enabling radio-frequency working loads of 25–300 kVA to be handled. When necessary, numbers of these capacitors may be connected in series or parallel or series–parallel to give an increased loading. In many transmitters and dielectric heaters a total of thirty or forty capacitors may be connected in series–parallel.

Titanium dioxide, apart from having a low power factor, is also frequently used for temperature-compensating purposes. It possesses a temperature coefficient of capacitance of negative 750. Where, due to circuit limitations, only a very small capacitance can be used—as, for example, in the oscillatory circuit in a television receiver—then strontium titanate with a temperature coefficient of negative 3,000 may be used. Where a temperature coefficient intermediate between these figures or even lower than negative 750 is required, then a combination of the 3,000 material with either other ceramic dielectrics or silvered micas may be employed, two capacitors being connected in either series or parallel, the combined temperature coefficient being modified to suit each particular case.

Ceramic capacitors for use in receivers are made in a wide variety of shapes and sizes, the shape of the dielectric body being chosen for its suitability for the particular capacitance, voltage rating and temperature coefficient, and amongst shapes in current use are tubes, discs, cups, rods, and pearls, each shape being made in a range of varying sizes.

Silvering

The electrodes of ceramic capacitors are usually of silver. The silver is deposited by painting, spraying or printing a solution of silver oxide in an organic base on to the dielectric body, which is then fired at a temperature of around 700° C., the oxide being reduced to metallic silver. Where heavy currents are likely to flow, as in transmitting types, then the silvering process may be repeated so as to give a heavier coating. The adhesion and quality of the electrodes are largely dependent on the quality of the silver solutions, on the use of the correct solution for a particular dielectric or body shape, and on the cleanliness and care in processing. For special purposes additional electro-plating may be carried out, so as to increase the thickness of the layer, and plating with metals other than silver has also been done to a limited extent.

After silvering, terminating wires are attached, wherever possible a mechanical joint being made; this is then finished off by soldering. In the case of disc-type capacitors, where a mechanical joint cannot be made, then greater care is required in silvering and soldering of the terminating wires to ensure a robust termination which will not fail under normal handling or during the tensile tests which may be specified. Where a close tolerance of capacitance is required, adjustment of the silvered area may be carried out by suitably grinding away the silver, care being taken not to leave discrete islands, since this may give an unstable capacitance with a high power factor.

Tubular ceramics having wire terminations have been in use for a large number of years. Their popularity is largely due to the fact that the ceramic bodies may be produced by means of an extrusion process, whilst the silvering and attaching of terminating leads may be done with the aid of automatic machinery, enabling an economic design to be achieved. Capacitances from a few to several thousand pica-farads are included in this construction, the capacitance being dependent on the dielectric constant of the material, together with the dimensions. Tube diameters vary from 3 to 6 mm., and lengths from 10 to 30 mm.

Disc-type Construction

Discs in low-κ material do not normally exceed 50 pF, the cup shapes extending the range to 100 pF. For very small values of the order of 0·5 pF to 10 pF, solid rods of a ceramic material of varying diameter and length are employed. These are referred to as "pearls", and are in effect very short rods. Ceramic discs of high-κ material are now in general use, and permit capacitances of 1,000 pF to be obtained on discs no greater than $\frac{1}{4}$ in. in diameter and $\frac{1}{32}$ in. thick, yet capable of operating up to 500 volts D.C. The range may be further extended up to capacitances of 0·01 μF on discs of approximately 1 in. in diameter.

The relatively high power factor (about 2 per cent) is not generally of importance for decoupling purposes, with the result that high-κ ceramic capacitors are widely used for de-coupling at radio frequencies of 40–70 Mc/s, as well as at intermediate frequencies of between 18 and 35 Mc/s. They take up little space and, by virtue of their construction, have a low series inductance. The disc type is preferred to the tubular type because of the lower inductance. For use in television tuners and converters operating at 200–400 Mc/s a co-axial tubular construction becomes necessary, consisting of a ceramic tube having an inner and outer conductor, the inner conductor usually being taken right through so as to act as a feed-through, whilst the outer electrode has a flange intended to fit into a hole in a chassis. By this means the series inductance is reduced to the lowest possible value and decoupling efficiency maintained at ultra-high-frequencies.

Protection against humid conditions is normally done by painting the capacitor with a suitable lacquer or coating with wax, or even combining both processes. Where a tougher, more robust protection is required, capable not only of withstanding humid conditions but also of preventing breakdown between the capacitor and a metal chassis, then a phenolic resin coating is used. After curing, this is impregnated in wax.

The necessity for protecting ceramic capacitors against humid conditions is because a film of moisture across the capacitor will have the effect of lowering the insulation resistance and increasing the power factor. Where the best possible performance is required it is usual to dry the capacitor thoroughly, vacuum impregnate it, then seal it in a protective tube having water-repellent end seals.

CERAMIC TRIMMERS

Ceramic materials have been used with success in the manufacture of trimmer capacitors of the pre-set type. These consist of flat silvered discs, the silvering being semicircular in area; by suitably adjusting the overlap a variation in capacitance is achieved. In the smaller types the minimum and maximum capacitance varies from 2–7 pF, increasing to 20–120 pF for the largest sizes. Materials having temperature coefficients of +100 and −750 are used, being chosen for particular circuit applications.

For ultra-high-frequency operation, even smaller capacitance trimmers are required; ceramic materials have again proved to be most suitable for this application. Small tubular bodies threaded internally are favoured, the centre screw acting as the inner electrode as well as for mounting the capacitor on the chassis, the outer electrode being a conventional silvered area with a wire or tag termination. In these types capacitances of 0·5–3, 1–5 and 3–9 pF will generally cover the requirements for 200–400 Mc/s operation.

ELECTROLYTIC CAPACITORS

The principle of the electrolytic capacitor has been known for close on a hundred years, though it is only during the last twenty-five or thirty years that practical use has been made of this knowledge and reliable capacitors produced. The outstanding feature of the electrolytic capacitor is the large capacitance which can be obtained in a given volume, particularly at the lower operating voltages. This large capacitance is brought about because the dielectric, unlike paper or mica, is formed on the surface of the positive electrode, which is usually aluminium, and consists of an extremely thin oxide film, the thickness of the oxide film being proportional to the voltage used in forming the film.

An oxide film may be formed on aluminium if the aluminium is made the anode of an electrolytic cell containing a solution of ammonium borate and the cell connected to a D.C. supply. When first connected to the supply, a current will flow which is limited only by the resistance of the cell and the voltage source. The current flow will then diminish to a low value until a stable condition has been reached, and it is the formation or build up of the oxide film behaving as a dielectric that causes the current to fall. The oxide film is produced by oxygen combining with the aluminium, the oxygen having been liberated by electrolysis.

Since the thickness of the dielectric film is proportional to the forming voltage used during manufacture, the capacitance per square inch of foil will be inversely proportional to this forming voltage, and hence capacitor size will vary with rated working voltage right down to a rating of 6 volts. This is unlike the paper dielectric type, where no reduction in capacitor size is obtainable for ratings below about 150 volts, because the paper used for this voltage is the thinnest available. It should be noted, however, that once the manufacturing process is complete the capacitance of an electrolytic capacitor shows only a negligible change with applied voltage, provided that this does not exceed the rated voltage.

The nature of the oxide film merits some thought, since it is so thin as to be virtually transparent, and in many cases can only be detected visually by the presence of interference colours, and from the appearance of these interference colours the thickness of the film may be determined. Its dielectric constant is around 8·6, and bearing in mind that the capacitance of a parallel-plate capacitor is inversely proportional to the thickness of the dielectric and also that the dielectric constant of the oxide film is appreciably greater than that of paper, it will be readily understood how the electrolytic capacitor can have such a large capacitance for its volume.

It was early realized that if the area of the aluminium foil could be increased artificially, then an even greater capacitance per unit volume could be obtained, and various methods were developed to achieve this object, amongst them being mechanical roughening of the aluminium foil, chemical etching and electrochemical etching, and it is these latter two methods that are mainly in use to-day. The increase of capacitance by increasing the surface area of the foil is applicable to the electrolytic capacitor because the oxide film is so thin that it readily follows the contours of the anode, and the electrolyte, which is the true cathode and is semi-liquid, is also able to follow the contours and therefore make contact over the whole area. An interesting point that arises with etched electrodes, again confirming that the oxide film varies in thickness according to the formation voltage, is the fact that etched foil anodes formed to a low voltage may have a ratio of capacitance as compared to plain foil of up to 10 to 1, whereas if the same foil is formed to a high voltage the ratio falls to between 3 and 4 to 1. This is explained by the fact that the thin oxide film follows the contours accurately, whereas the thicker oxide film bridges the smaller gaps or valleys, and thereby reduces the surface area. This effect is also shown in the accompanying diagram, Fig. 5.

A further method of obtaining a large capacitance per unit volume is

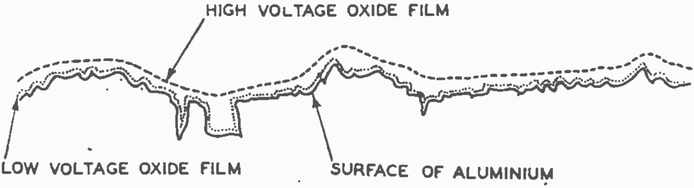


FIG. 5.—ETCHED-FOIL CROSS-SECTION SHOWING LOW- AND HIGH-VOLTAGE OXIDE FILMS.

to spray aluminium on to paper or gauze so that an extremely rough surface having a large area is obtained, and by this means an increase in capacitance in comparison with plain foil of 12 to 1 may be obtained. The problems involved in spraying the aluminium are, however, considerable, and for some years reliability was doubtful, though it has now been improved and is comparable with etched-foil electrolytics.

Early types of electrolytic capacitors were essentially wet or aqueous types, the electrolyte being a liquid necessitating upright mounting, and though these wet types were popular for many years, they were ousted by the modern dry types, since not only were they more convenient to handle, but the electrical performance was improved, the power factor, particularly, being better and enabling higher ripple currents to be passed. Furthermore, the capacitance per unit volume was far greater than could be achieved with the wet types, and as a result the wet types are not now manufactured.

Manufacture

The first process in the production of electrolytic capacitors is to produce an oxide film on the surface of the aluminium foil which is to be the positive electrode. This is done by running a continuous length of foil through a number of tanks containing electrolyte and applying a positive voltage to the foil. A heavy current flows when the foil is first immersed in the electrolyte, the current gradually falling until a suitably low figure is reached. The foil is then removed from the last forming tank and re-wound. The purpose of using a number of tanks is to grade the applied voltage so that the current at no time will be so great as to fuse the foil, and for convenience the various tanks are arranged to have an increasing negative potential, the foil being the common positive.

The general form of a dry electrolytic capacitor is very similar to that of a paper dielectric capacitor, two aluminium foils (the positive foil having an oxide film) are interleaved with paper or other suitable materials and wound into a cylindrical form, lugs are brought out, and the capacitor impregnated in a suitable electrolyte containing ammonia in combination with boric acid and either ethylene glycol or glycerol.

The paper or other separators used are more absorbent than is normal for paper dielectric capacitors, since in the electrolytic capacitor the function of the paper is mainly to hold in suspension the electrolyte, which is the true cathode of the capacitor, the negative foil being convenient for making contact with the electrolyte.

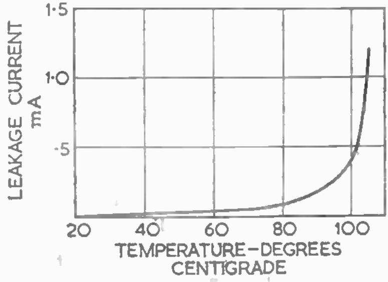
Connections to the foils are usually made by folding back a strip of the foil, which may be riveted direct to the tag terminations, though with the thicker foils used for etched-foil capacitors an alternative method is adopted, consisting of sewing lugs to the foils, the sewing in this case being done, not with the aid of cotton or aluminium wire, but by bursting a series of holes through the lug and foil and hammering over. Hammering of the joint is important, as it is essential to completely break up the oxide film on the surface of the foil so as to ensure an intimate contact between the aluminium surfaces. If this oxide film was not broken up, then a high-resistance connection or possibly a complete electrical disconnection could occur, even though mechanically the joint would appear to be sound.

Following winding and attaching of the lugs, the capacitors are impregnated in an electrolyte which is suitable for the particular working voltage. Various methods of impregnation are in use, amongst them being soak, vacuum and centrifugal impregnation. In all cases the main object of impregnation is to force the electrolyte into the fibres of the paper separators so as to be in suspension between the foils, and in that way give a low resistance path. Poor impregnation will give high P.F., poor low-temperature performance and possibly low capacitance. The capacitors are then ready to be assembled in their cases, which are usually of aluminium and may be tubular or can types with rubber or bakelite ends.

Due to handling of the formed foil, some damage is caused to the oxide film, which must be repaired before the capacitor can be considered suitable for incorporating into electrical equipment. There are also cut edges on the foils where lugs have been attached, as well as at the ends of the positive foil, where it has been cut from the reel, and these cut edges require to have an oxide film formed on them. This formation, as well as the necessary repair to the oxide film mentioned previously, is done by applying the working voltage to the capacitor through a resistance of such a value as to limit the current and prevent over-heating; this process is known as ageing or re-forming. The time required for ageing or re-forming will vary according to the capacitance and working voltage of the capacitors, and for low-voltage types may take only 1 to 2 hours, whereas for high-voltage types, 12 to 24 hours will be required. The main limitation in ageing is to ensure that excessive heat should not be developed internally due to the wattage generated by the flow of current. Excessive heating has the effect of increasing the leakage current, the increased leakage current generating a further temperature rise, which again will cause the current to increase, the result of this vicious circle being to cause an excessive generation of gas, with the possibility that the internal pressure may rise to such an extent as to cause the end discs to be blown out, Fig. 6. In many cases where high-voltage, high-capacitance electrolytic capacitors have to be aged, this process is done before final sealing of the container has been carried out and a high internal pressure obviated.

To prevent excessive pressure causing a dangerous condition when the capacitor is in use, it is usual for the larger types of electrolytic capacitors to have a vent or a weak spot which can give way should the internal pressure become excessive. The increase in leakage current, which is clearly indicated in Fig. 6, limits the maximum operating temperature of electrolytic capacitors, but whereas a few years ago maximum temperatures of 50° and 60° C. were specified, operating temperatures of up to 85° C., and even 100° C., are now becoming usual for Services equipment.

Fig. 6.—Typical Leakage Current–Temperature Curve for Capacitor Suitable for Operation at 85° C.



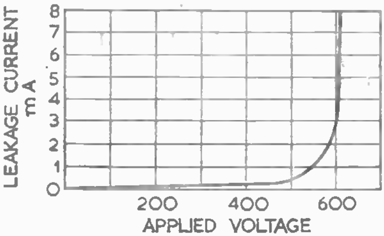


Fig. 7.—Typical Leakage Current–Voltage Curve for Capacitor Rated at 500 Volts D.C. Working 600 Volts Surge.

When the leakage current falls to a sufficiently low figure the capacitor is tested, the tests consisting of a measurement of leakage current at the working voltage, and a measurement of capacitance and power factor.

Typical values of leakage current would be :

8 μF	350 V	0·4 mA
8 μF	450 V	0·5 mA
32 μF	450 V	2 mA

with other capacitances and voltages being in proportion, though for the very large capacitances, such as 100–200 μF, a lower rate of leakage current is adopted. A convenient guide for calculating the permissible leakage current for plain-foil electrolytics is :

$$0{\cdot}15CV \text{ in } \mu\text{A}$$
$$\text{and } 0{\cdot}05CV \text{ in } \mu\text{A for etched foil.}$$

where C = capacitance, in μF; and V = working voltage. See also Fig. 7.

Due to the nature of the oxide film the capacitance tolerance of electrolytic capacitors is wider than that which is usual for paper dielectric capacitors, and for low-voltage types a tolerance of -20 per cent $+$ 50 per cent of nominal, or even -20 per cent $+$ 100 per cent, may be used, whilst for high-voltage types the tolerance is invariably -20 per cent $+$ 50 per cent of nominal capacitance.

Surge Ratings

An examination of Fig. 7 shows that the leakage current rises rapidly above the working voltage, and it is this property which is made use of in the electrolytic capacitor to absorb some of the D.C. current when first switching on a radio or television receiver, and thereby prevent a dangerous rise in voltage in the receiver. The surge rating of an electrolytic capacitor is usually 100 volts above the working voltage for high-voltage types and 75 volts for medium-voltage types.

Re-Forming

A feature of electrolytic capacitors that is not always recognized is that if a capacitor has been in store for a long period the leakage current will take longer to drop to its normal value when it is first put into circuit than would be the case with a new capacitor or one in constant use. Normally this is of no importance unless the storage time has been so long that the leakage current does not decrease sufficiently rapidly to prevent overheating of the capacitor. Broadly speaking, capacitors should be reformed every twelve months if stored under good conditions. Wherever the storage is doubtful, as in an unduly high ambient temperature or the tropics, then re-forming must be done more frequently, intervals of three to six months being necessary. Re-forming may best be done by applying the working voltage to the capacitor in series with a resistor of such a value as to limit the initial current to a safe value and for an 8-μF, 450-volt capacitor a resistance of 10,000 ohms will be found satisfactory.

The leakage-current figures mentioned earlier should not be regarded as an absolute limit for the rejection of the capacitor, and so long as the leakage current is still falling at the expiration of, say, 1 hour, then the capacitor may be considered as suitable for being put into use.

Special Types

Though so far we have dealt with polarized electrolytics, for many applications a non-polarized type is required, and this is achieved by winding the capacitor with two formed foils instead of only one, so that either electrode may be connected to H.T. positive. As two formed foils are used, the capacitance is halved, as in effect there are two capacitors in series, and to compensate for this it is necessary to double the length of each foil.

An application where non-polarized electrolytic capacitors may be used is for H.T. smoothing in a D.C. mains radio receiver, so that no harm will be caused by connecting the receiver incorrectly to the mains. A similar application may arise where filtering of a 6- or 12-volt supply for a car radio is necessary, and where otherwise an incorrect connection to the car battery could lead to failure of the capacitor. There are also frequent applications for low-voltage A.C. operation; the limitation in this case being that the temperature rise due to the power factor of the capacitor should not be so high as to cause overheating.

A.C. working electrolytics are also produced for high-voltage working where the operating period is very short, and large numbers are employed in conjunction with "capacitor start motors" where operating voltages of 100–300 r.m.s. are usual, but in these applications the voltage is applied to the capacitor for only a fraction of a second.

Tantalum

Aluminium is only one of many metals which have the property of forming a protective oxide film when made the anode of an electrolytic cell, and more recently the use of tantalum has been under investigation, and increasing numbers of tantalum electrolytic capacitors are now being produced. The oxide film produced on tantalum is tougher and more resistant to damage and corrosion than aluminium oxide, with the result that the working life is very much greater and little, if any, re-forming is required during storage. In addition, highly conducting electrolytes particularly suitable for operation at the extremes of temperature may be employed satisfactorily, and tantalum is used where operating temperatures of $-50°$ and $+125°$ C. are specified.

The capacitance obtained per unit volume with tantalum is slightly greater than that of aluminium, but the maximum operating voltage is limited to 150 volts D.C., and for higher-voltage working it becomes necessary to connect two or more capacitors in series.

The most recent development in tantalum capacitors is the solid-electrolyte type, sometimes called the semi-conductor type. In this variety the anode is a porous block made by pressing and sintering tantalum powder. This anode is given a dielectric coating of tantalum oxide in an electrolytic bath, and it is then impregnated with a solution of manganese nitrate, which is subsequently heated until it decomposes, leaving a coating of semi-conducting manganese dioxide on the oxide dielectric layer. This coating is equivalent to the electrolyte in the conventional electrolytic capacitor; finally, a coating of graphite is usually applied which serves to make contact with the manganese dioxide and is analogous to the negative foil in the conventional type of capacitor.

The great virtue of these capacitors is that they are completely dry and contain no liquid or paste electrolyte, to seep at high temperatures or freeze at low temperatures. Thus operation at very low temperatures is possible, the capacitance at $-100°$ C. being only about 15 per cent less than the 25° C. value; while the power factor is rather better at $-100°$ C. than at 25° C. The upper temperature limit is 85° C. at present, but types suitable for 125° C. are expected in the near future. The maximum working voltage which has been achieved so far is 60 volts D.C. Owing to the very large effective area of the sintered block, these capacitors can be made in very small sizes.

A limit to the general use of tantalum is its price, due to the relative scarcity of the metal, as well as to the difficulty in producing the foil, which is extremely hard and difficult to roll thin. For this reason it is unlikely that tantalum electrolytic capacitors will come into general use, and it is probable that their applications will be restricted to special equipment required for the Fighting Services and for industry, where cost is not of prime importance.

W. I. F.

PLASTIC-FILM CAPACITORS

Plastic-film capacitors are generally made by winding metal-foil electrodes interleaved with films of the dielectric material, in exactly the same way as paper dielectric capacitors are made. In the case of the plastic film, however, no impregnation is required, except for very high-voltage operation.

With certain of the dielectric materials the film may be made to shrink slightly by suitable heat treatment after the capacitor element is wound, and this results in a very high degree of capacitance stability.

In a recent modification of the plastic-film capacitor the electrodes consist not of metal foils but of a very thin layer of metal deposited on the dielectric film by chemical deposition or by vacuum evaporation. When an electrical breakdown occurs in such a film the metal deposit, because it is very thin, evaporates at the point of breakdown and over a small area around it, so clearing the fault. Thus the metallized film has a self-healing property, and a single layer may be used in place of the multi-layer dielectric normally employed to guard against weak spots. This, in turn, leads to a considerable reduction in the size of the finished capacitor compared with one of similar rating employing metal-foil electrodes because of the fact that the volume of the capacitor element is roughly proportional to the square of the dielectric thickness.

Few plastics have been used as dielectrics, chiefly because of the difficulty of producing them in the form of continuous lengths of suitably thin film (0·001 in. or less). Cellulose, cellulose acetate and more recently cellulose triacetate have been used to a small extent in the U.S.A., but these materials have no advantage over impregnated paper, except possibly the triacetate, which has good stability at temperatures up to 120° C. Polystyrene is the only plastic film which has been used to any great extent in the United Kingdom.

To maintain the excellent electrical properties of these capacitors it is essential that they should be hermetically sealed, and to ensure this the larger values are housed in rectangular metal boxes with glass or ceramic terminals, while the smaller units are contained in metal tubes with P.T.F.E. insulators.

Polystyrene with Foil Electrodes

The most outstanding characteristic of the polystyrene film capacitor is its extraordinarily high insulation resistance, the value for a 1-μF unit being usually greater than a million megohms after 1 minute's electrification, and after longer periods the value may exceed 10^7 MΩ. As an

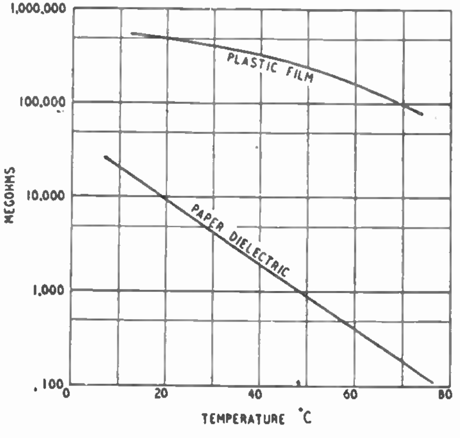


Fig. 8.—Variations of Insulating Resistance with Temperature for Typical 1-μF Capacitors of the Plastic Film and Paper Dielectric Types.

(Courtesy of The Telegraph Cond. ser Co. Ltd.)

illustration of the very high insulation figures which can be obtained with this dielectric, a 1-μF unit was charged initially to 500 volts, and when tested after a period of two years it was found to have lost only 8 per cent of its charge. As with all dielectrics, the insulation resistance decreases with increasing temperature, but for a given temperature increase the percentage drop in resistance is appreciably less than with other capacitor dielectrics. The curves in Fig. 8 show the effect of temperature change on the insulation resistance.

When the capacitor elements have been given suitable heat treatment the capacitance value is very stable and, even with repeated temperature changes, the drift in the capacitance value measured at a standard temperature is less than 0·1 per cent. The maximum temperature rating for reliable operation is 60° C., and, although operation is sometimes permissible up to 70° C., the high stability mentioned above may be lost if the temperature exceeds 60° C. The temperature coefficient of capacitance lies in the range -100 to -200×10^{-6} per ° C.

The power factor of this type of capacitor is in the range 0·0002–0·0005 at all frequencies, and the dielectric absorption is less than that of any other known type of capacitor. Dielectric absorption may be defined as the percentage difference between the quantity of electricity required to charge a capacitor to a given voltage and the quantity of electricity given up by the capacitor when it is then rapidly discharged.

The physical size of these capacitors is rather large, being three to four times that of a paper dielectric capacitor of equivalent voltage rating.

The use of these capacitors is indicated wherever very long-time constants are needed or a quantity of charge has to be stored with negligible leakage. Thus they may be used to measure ionization current by allowing the current to flow into the capacitor and observing the rate of voltage rise; and the ionization current in a gas may be used to indicate the intensity of the radiation from a radioactive material.

The good capacitance stability and low power factor over the whole frequency range render these capacitors very suitable for use as bridge standards and for tuned filters where high Q-values are required. In the latter application the negative temperature coefficient is useful to offset the temperature coefficient of the inductance, which is usually positive.

Metallized Polystyrene

These capacitors, after suitable heat treatment, have the same high stability of capacitance as the foil type, and the capacitance temperature coefficient is -80×10^{-6} per ° C. The dielectric absorption is also very low, as in the case of the foil types.

Power factor is rather higher than for the foil type, owing to the resistance of the thin deposit of metal, and it tends to rise with frequency. Typical values are 0·0007 at 50 c/s and 0·001 at 1,000 c/s.

Insulation resistance is also somewhat lower in the metallized capacitor, owing to the damage to the dielectric which results from the self-healing action at weak spots. The value is usually of the order of 25,000 megohm micro-farads.

The maximum recommended operating temperature for these capacitors is 70° C., and their physical size is comparable with that of a paper dielectric capacitor of equivalent voltage rating.

These capacitors are much smaller in size than the foil-electrode type, while having a similar capacitance stability, although their power factor and insulation resistance are not so good.

They are used chiefly in tuned filters for carrier-frequency telephony, but may also be used to replace paper dielectric capacitors, where rather better properties are required and the working temperature does not exceed 70° C.

Metallized Melinex

Melinex is the trade name for polyethylene terephthalate, which, when used in textiles, is known as Terylene. This material has great mechanical strength and can be processed into films down to 0·00025 in. thick. The dielectric constant is 3·2, so that Melinex capacitors can be made comparable in size to paper dielectric types. The capacitors at present available mostly employ a metallized dielectric, the self-healing characteristic of which enables a single 0·00025-in. layer to be used, and thus allows a very small unit to be produced. For example, a 1-μF metallized Melinex capacitor rated for 150 volts at 125° C. and sealed for tropical conditions can be made $1\frac{5}{8}$ in. long × $\frac{1}{2}$ in. diameter.

The insulation resistance of these capacitors is appreciably better than that of paper dielectric capacitors, and the power factor is about 0·005. The maximum permissible temperature for continuous operation is 125° C., but 150° C. for short spells can be tolerated.

Polytetrafluorethylene (P.T.F.E.)

Capacitors having a dielectric of P.T.F.E. film are now available and are useful for extra-high-temperature applications, since they may be used in ambient temperatures up to 200° C. Unfortunately, capacitor sizes are rather large because the dielectric constant is only 2·0, and rather greater film thickness is required than with polystyrene or Melinex.

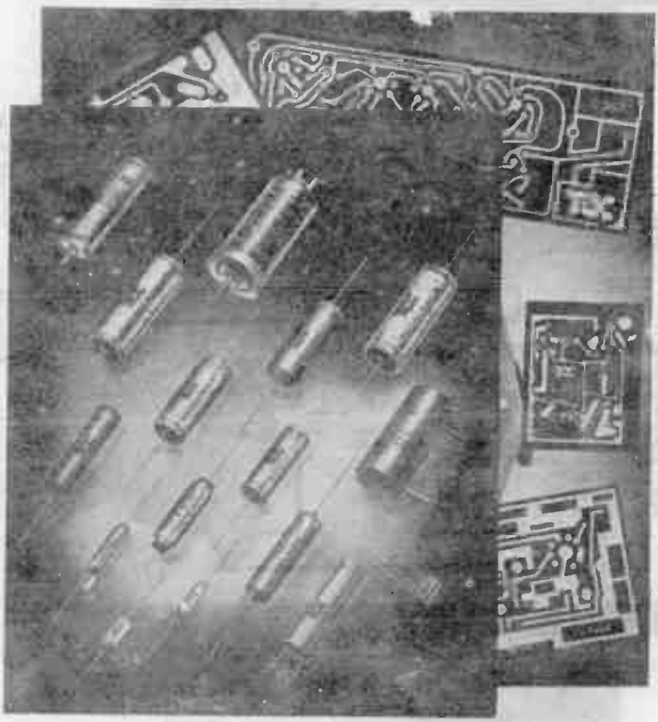
The insulation resistance and power factor are extremely good, the values being comparable with those for polystyrene. The temperature coefficient of capacitance is about -200×10^{-6} per °C.

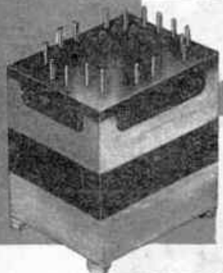
Future Trends

The electrical properties of the polystyrene film in current use are of a very high standard, and it is unlikely that they will be improved upon. The maximum permissible working temperature, however, will be increased by the production of polymers of higher molecular weight and of modified compounds, such as the co-polymer of styrene and alpha-methylstyrene, which is at present in the experimental stage.

Some other new dielectric materials in the experimental stage are high-density polythene (which has a higher melting point than the normal polythene), polypropylene (which has similar properties to high-density polythene), and polycarbonate, which shows great promise and has electrical properties very similar to those of Melinex, although its power factor is better and its maximum safe operating temperature is very slightly lower.

J. H. C.



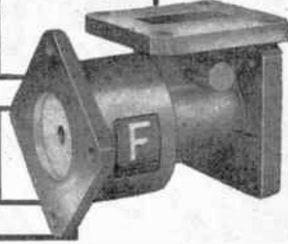


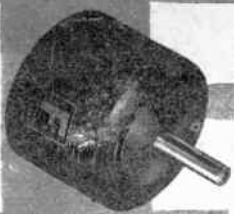
Forth Series Transformers

Since their introduction, Forth Series Resin Cast Transformers and Chokes have been extensively adopted for airborne and missile applications in view of the significant saving of weight and volume which can be achieved over conventional types. The range is designed to meet inter-service specifications.

Ferrite Circulators

The Ferranti Ferrite Circulator has the following characteristics when used as a duplexer in low power radar systems.
Bandwidth Frequency: 400 Mc/s at X Band. During Transmission: Isolation transmitter to Receiver > 25 db. Insertion Loss transmitter to aerial < 0·2 db. During Reception: Isolation Aerial to Transmitter > 15 db. Insertion Loss Aerial to receiver < 0·5 db.





Precision Potentiometers

The Ferranti range of Precision Potentiometers is designed to provide analogue conversion from mechanical rotation to an electrical signal. They have a wide range of uses in flight simulators, flight trainers, airborne flight instruments, computers and similar applications.

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30. INDUCTORS AND TRANSFORMERS

30. INDUCTORS AND TRANSFORMERS

The transmission of electrical power from one point to another takes place, contrary to general opinion, through space, and is accompanied by an electromagnetic field. The metallic conductor which is so often associated with this transmission serves mainly as a "guide" to the field, and the flow of electrons within it is a coincident phenomenon resulting from the cutting of the conductor by the field. In the elementary stage of instruction it is usual to consider the electrostatic component of the field as causing the flow of current, and the current as causing the magnetic component, and for simplicity the language of this article will be conformed to that viewpoint, whereby much difficult mathematics may be avoided. The reader should remember, however, that this rather naïve view is not strictly accurate, and breaks down when applied to systems in which the dimensions are comparable with the wavelength: in such cases, either the simplification of "distributed parameters" must be adopted or recourse must be had to the fundamental propagation equations formulated by Maxwell.

Fundamental Considerations

We may consider then that the establishment of a potential difference in a conductor will cause the flow of current within it, and that the flow of current will generate a magnetic field. Variation of current, which will always be accompanied by variation of potential difference, will cause variation of the magnetic field enclosing the conductor. If, however, a non-current-carrying conductor also be enclosed by the varying field, an e.m.f. will be generated in it by variation of the field, and if the circuit be complete, current will flow consequent upon this e.m.f. A similar e.m.f. will also be generated by variation of the field in the conductor carrying the generating current, and this e.m.f. must be considered as connected in series with that generating the current, and it will thus modify the current within the circuit. Thus while a circuit carrying an invariant current will be subject to Ohm's Law, and will pass a current entirely dependent upon externally applied e.m.f. and upon its own resistance, as soon as that current is caused to vary, the value of current flowing during the period of variation will be a function of the magnetic-field configuration in its neighbourhood as well as of the other two parameters; also, during the period of variation, any other neighbouring conductor will have generated within it an e.m.f., and may, as a result, carry current.

The relationship between the variation of current within a circuit and the resultant e.m.f. generated within it by the varying magnetic field is expressed in terms of a coefficient termed the "self-inductance" of the circuit. Similarly, the relationship between the variation of a current in a circuit and the induced voltage in a neighbouring circuit is expressed in terms of a coefficient termed the "mutual inductance" between the two circuits. Both these quantities are measured in terms of a unit named the "Henry", and standard sub-multiples of it in frequent use are the milli-henry and the micro-henry. The quantity has the dimensions of length, and in fundamental work is often expressed in centimetres. The symbol most frequently used for self-inductance is " L ", and that used for mutual inductance is " M ", while the standard abbreviations for the unit and its sub-multiples are : H (Henry), mH (milli-henry), μH (micro-henry).

Inductance

When current flows through a circuit the strength of the magnetic field generated will depend upon the geometry of the circuit, and the magnetic permeability of the medium in which the circuit is immersed. Thus we may write :

$$\phi = K\mu i$$

$$\frac{d\phi}{dt} = K\mu\frac{di}{dt}$$

where the symbols have the following significance :

ϕ = magnetic flux in maxwells;
K = magnetic coefficient of circuit;
μ = magnetic permeability of medium;
i = electric current in amperes;
t = time in seconds.

When the flux varies, the induced e.m.f. may be calculated from the following law :

$$e = -N\frac{d\phi}{dt} \times 10^{-8}$$

in which the negative sign indicates that the induced e.m.f. is of such polarity that it will oppose the current change which induces the change of flux, and the additional symbols have the following significance :

e = e.m.f. in volts;
N = number of turns in circuit linked by flux.

When the variation of flux is caused by variation of the current, it is obvious that the following relationship may be written :

$$e = -NK\mu\frac{di}{dt} \times 10^{-8}$$

In terms of self-inductance and of mutual inductance, the induced e.m.f. may be expressed thus :

$$e_1 = -L\frac{di_1}{dt}$$

$$e_2 = -M\frac{di_1}{dt}$$

in which :

e_1 = induced e.m.f. in first circuit;
e_2 = induced e.m.f. in second circuit;
i_1 = current in first circuit.

and from these it becomes obvious that L is the flux per ampere linking the first circuit divided by 10^8 and multiplied by the number of turns, while M is the flux per ampere of current in the first circuit linking the second circuit, divided by 10^8 and multiplied by the number of turns in the second circuit. A third parameter, known as the coefficient of coupling and usually designated by the symbol k, is defined thus :

$$k = \frac{M}{\sqrt{L_1 L_2}}$$

and it has a maximum value of unity when all the flux linking one circuit due to a current flowing in that circuit links also all turns in the second circuit.

Since the value of M will be the same whichever circuit is designated number one, it is obvious that the relationship is reversible thus :

$$e_2 = -M\frac{di_1}{dt}$$

$$e_1 = -M\frac{di_2}{dt}$$

and in particular, when two circuits are connected in series so that the same current flows through both of them, the total induced e.m.f. in both circuits will be :

$$e_1 + e_2 = -L_1\frac{di}{dt} - M\frac{di}{dt} - L_2\frac{di}{dt} - M\frac{di}{dt}$$

$$= -(L_1 + L_2 + 2M)\frac{di}{dt}$$

from which it becomes obvious that the total inductance of both circuits in series will be given by :

$$L = L_1 + L_2 + 2M$$

A little thought will show that this is so only when the circuits are so connected that e.m.f. generated by current variation in the other circuit is of similar sense to that produced by current variation in itself; it is possible for circuits to be so connected that they have a negative mutual inductance, and in these circumstances the total inductance will be :

$$L = L_1 + L_2 - 2M$$

This fact gives rise to a method of measuring the mutual inductance between two circuits. If they are connected in series and the total inductance measured both aiding and opposing, the difference between the two values of total inductance will obviously be four times the mutual inductance between them.

Fundamental Formulæ

Before proceeding with a consideration of the practical circuits built with the object of obtaining inductance, a few fundamental formulæ will be reviewed, upon which the composite formulæ to be considered later are based : where small inductances are required in

V.H.F. circuits, the elementary geometrical forms here considered are sometimes used, however. Due to "skin effect" the inductance of most circuits varies with frequency, and the two limiting values at very low and very high frequencies are given below; the actual inductance will usually lie between these two values, and if it is essential that its exact value be calculated, suitable methods may be found in more detailed texts. In each case it is assumed that the circuit is surrounded by a medium of unity permeability, but that the conductor itself has a permeability μ. A straight cylindrical conductor of length l cm. and diameter d cm. has an inductance:

LOW FREQUENCY. $$L = 2l\left(\log_e \frac{4l}{d} - 1 + \frac{\mu}{4}\right) \times 10^{-9}$$

HIGH FREQUENCY. $$L = 2l\left(\log_e \frac{4l}{d} - 1\right) \times 10^{-9}$$

A straight conductor of rectangular section with sides b and c cm. of material having unity permeability has a low-frequency inductance given by:

$$L = 2l\left[\log_e \frac{2l}{b + c} + \frac{1}{2} + \frac{0{\cdot}2235(b + c)}{l}\right] \times 10^{-9}$$

A circle having diameter D cm. formed from conductor with a circular cross-section has inductance given by:

LOW FREQUENCY. $$L = 2\pi D\left[\log_e \frac{8D}{d} - 2 + \frac{\mu}{4}\right] \times 10^{-9}$$

HIGH FREQUENCY. $$L = 2\pi D\left[\log_e \frac{8D}{d} - 2\right] \times 10^{-9}$$

A rectangle of sides a_1 and a_2 formed from conductor of circular cross-section, has inductance given by:

$$g = \sqrt{a_1^2 + a_2^2}$$

LOW FREQUENCY.

$$L = 4\left[(a_1 + a_2)\log_e \frac{4a_1a_2}{d} - a_1 \log_e (a_1 + g) - a_2 \log_e (a_2 + g)\right.$$
$$\left. + 2\left(g + \frac{d}{2}\right) + \left(\frac{\mu}{4} - 2\right)(a_1 + a_2)\right] \times 10^{-9}$$

HIGH FREQUENCY.

$$L = 4\left[(a_1 + a_2)\log_e \frac{4a_1a_2}{d} - a_1 \log_e (a_1 + g) - a_2 \log_e (a_2 + g)\right.$$
$$\left. + 2\left(g + \frac{d}{2}\right) - 2(a_1 + a_2)\right] \times 10^{-9}$$

A rectangle formed from conductor of rectangular cross-section having unity permeability has a low-frequency inductance:

$$L = 4\left[(a_1 + a_2)\log_e \frac{2a_1a_2}{b + c} - a_1 \log_e (a_1 + g) - a_2 \log_e (a_2 + g)\right.$$
$$\left. + 2g - \frac{a_1 + a_2}{2} + 0{\cdot}447(b + c)\right. \times 10^{-9}$$

The formulæ for mutual inductance between two circuits are very complicated, and only two special cases will be given here for straight circular conductors of unity permeability at low frequencies. Two parallel conductors of equal length spaced D cm. apart, with D much larger than d have a mutual inductance :

$$M = 2\left[l \log_e \frac{l + \sqrt{l^2 + D^2}}{D} - \sqrt{l^2 + D^2} + D\right] \times 10^{-9}$$

Two conductors of lengths l and m, with axes in the same straight line with ends spaced n cm. apart, will have mutual inductance :

$$M = [(l + m + n) \log_e (l + m + n) + n \log_e n - (l + n) \log_e (l + n) - (m + n) \log_e (m + n)] \times 10^{-9}$$

INDUCTORS

Air-cored Inductors

Air-cored inductors are used for radio-frequency circuits, and their physical form depends mainly upon the value of inductance desired, the voltage which they must withstand and the power which they will be called upon to dissipate. The resistance and capacitance will also be a factor determining the construction.

Single-layer Solenoid

The simplest form of air-cored inductor is the single-layer solenoid, a helix of conductor either supported on a former or else self-supporting without former. The former material most frequently used is either S.R.B.P. or a cellulose-filled P.F. moulding; these materials, however, have a fairly high power-factor, and only a moderate dimensional stability in an atmosphere of varying humidity. Where greater stability is required a ceramic former is used, and the conductor is tension wound in a helical groove on the outer face; for maximum stability a helix of silver is deposited on the ceramic and fired in position, being then reinforced by an electro-plated covering of copper, which is finally silver-plated overall. The inductance of such a solenoid may be calculated thus :

$$K_n = \frac{1}{1 + 0{\cdot}45\left(\frac{D}{l}\right) - 0{\cdot}005\left(\frac{D}{l}\right)^2}$$

$$A = \log_e 1{\cdot}73dn$$

$$B = 0{\cdot}336\left(1 - \frac{2{\cdot}5}{nl} + \frac{3{\cdot}8}{n^2l^2}\right)$$

$$L = \pi Dnl[\pi DnK_n - 2A - 2B] \times 10^{-9}$$

where : D = effective diameter of winding in centimetres;
l = length of winding in centimetres;
d = effective diameter of conductor in centimetres;
n = winding pitch in turns per centimetre.

For approximate results the terms A and B may be ignored, while if much work is to be done, all three terms K_n, A and B may be found tabulated to several places of decimals in many standard texts.

The effective self-capacitance of a single-layer solenoid will vary slightly with frequency (as will also its inductance and resistance), but an approximate guide to its value is to consider the value in pica-farads equal to the radius in centimetres. This value will be increased if any form of lacquer, varnish or other insulant such as wax is applied to the winding to cement it in position; such materials may well double the capacitance, and the least-objectionable is polystyrene cement, which has the lowest permittivity of them all, but which does not form a very effective seal against ingress of moisture to the winding. It should be noted that the capacitance is almost independent of the length of winding, and where the minimum capacitance is desired the diameter should be made as small as practicable, and the winding composed of several sections in series spaced from one another; there is little gain in using more than three such spaced windings. Unfortunately such constructions, which minimize the capacitance, increase the resistance for a given inductance value, and most inductors are a compromise between the desire for minimum resistance and that for minimum capacitance.

If the minimum resistance is required for a given inductance, the diameter of winding should lie between 2 and 2½ times the length, and the inductor be close-wound to use the largest possible diameter of conductor. As frequency increases, skin-effect causes the current to be concentrated over only a fraction of the conductor cross-section, with consequent increase of resistance, which is greater in proportion for the larger-diameter conductors; it is usual therefore to use a stranded enamel-covered copper conductor in the high-frequency range to obtain a more even distribution of current and a smaller effective resistance. Use of such conductor will materially decrease resistance in the range 100 kc/s to 5 Mc/s, but at higher frequencies than this the capacitance between strands defeats the object of stranding. The greatest decrease in resistance is obtained with "Litz" wire stranded in threes, and then in sets of three threes, but this conductor is expensive so that it is used only where essential, and the greater use is of a cheaper "bunched and stranded" conductor.

Multi-layer Solenoid

Where an inductance larger than may readily be obtained with a single-layer solenoid is required, recourse must be had to a multi-layer winding, and several forms of this are common, depending upon expense and the relative importance of stability, resistance and capacitance. For maximum stability of inductance, the best form is a series of deep, narrow slots, tightly wound with conductor, in a ceramic former; if care is taken to fill each layer in the slot before proceeding to the next, so that outer turns cannot lie adjacent to inner turns, and at least six slots be used, the capacitance of this type will be found fairly low, but resistance will be fairly high. A form having moderate stability, but high capacitance and resistance, is the multi-layer solenoid with length greater than diameter. The form having lowest resistance and capacitance is a short multi-layer solenoid with winding height equal to its length and mean diameter approximately three times the length. Capacitance can be decreased, with consequent increase of resistance, by increasing winding depth and decreasing winding length of this form. Where such proportions are used, it is usual in most radio and television applications to decrease the capacitance and increase the resistance by using "wave-winding", a technique which feeds the conductor on to a rotating former over an oscillating shuttle; the shuttle oscillates for the full width of the winding at a rate which is not quite a multiple of the rotational period, thus giving a winding which is only partly occupied by conductor, and which gives appreciable spacing between a proportion of each turn and the next. By causing the former to progress axially at a slow rate past the shuttle, a progressive wave-winding is obtained, with length greatly in excess of diameter, but this type is not so common as the short, high type described above. Due to the smooth surface of enamelled conductor, sufficient adhesion cannot be obtained with it for wave-winding, and it is usual to use single silk-covered enamelled copper wire; in many cases bunched stranded conductors are used to increase the ease of winding and also to decrease the effective high-frequency resistance.

Inductance of Multi-layer Solenoids

Two approximate formulæ may be used for calculating the inductance of multi-layer solenoids, one for length greater than diameter, and one for diameter greater than length. The same symbols will be used as before, with the following additions:

N = total number of turns;
D_1 = inner diameter of winding;
D_2 = outer diameter of winding.

Then, if the following factors be calculated for convenience:

$$R = \frac{D_1 + D_2}{4}$$

$$h = \frac{D_2 - D_1}{2}$$

The inductance of the solenoid long relative to diameter is given by:

$$L = \frac{4\pi^2 R^2 N^2 K_e}{l \times 10^9}$$

$$K_e = \frac{1}{1 + 0{\cdot}9\dfrac{R}{l} + 0{\cdot}32\dfrac{h}{R} + 0{\cdot}84\dfrac{h}{l}}$$

and that for the form with short length by:

$$L = \frac{4\pi^2 R^2 N^2 F_1 F_2}{\left(l + h + \dfrac{D_2}{2}\right) 10^9}$$

$$F_1 = \frac{10l + 12h + D_2}{10l + 10h + 0{\cdot}7D_2}$$

$$F_2 = 0{\cdot}5 \log_{10}\left(100 + \frac{7D_2}{2l + 3h}\right)$$

Where a complete inductor comprises a number of sections connected in series, the inductance of each section must be computed and added to the components due to the mutual inductance between them.

Where a minimum external magnetic field is required from an air-cored inductor, a former of torus shape is wound with a single layer of conductor. If the following symbols be used :

r = radius of torus cross-section ;
R = radius of torus ;
N = number of turns.

Inductance is given by the following formula :

$$L = \frac{\pi N^2 r^2}{R\left[1 + \sqrt{1 - \left(\frac{r}{R}\right)^2}\right] \times 10^9}$$

When these toroidal inductors are wound with several layers, the inductance is much more difficult to calculate precisely, and it is wise to make an approximation with the above formula, wind a test inductor and measure its inductance, and then compensate accordingly.

The mutual inductance between air-cored solenoids may be calculated readily, but the formulæ are involved, and the process long and laborious. In practice, the solenoids are so easy to wind that it is quicker and more accurate to wind test inductors and measure mutual inductance for a series of spacings, than to construct a graph to illustrate the law relating mutual inductance to spacing. From such a curve the values required for final design may be read off and used.

Iron-cored Radio-frequency Inductors

If the medium surrounding an inductor had a magnetic permeability greater than unity, the inductance would be increased without increase of winding resistance and capacitance; unfortunately all materials having high permeabilities also exhibit hysteresis, which introduces a loss equivalent, so far as the circuit is concerned, to an increase in ohmic resistance of the windings. In addition to this, most of the ferric and ferrous materials have appreciable conductance, which adds a further loss due to eddy-currents and also increases capacitance of the inductor due to capacitance between core and winding. For many years it has been customary to employ these materials in the form of finely divided powder, with the particles bonded together and insulated from one another by a resin binder. Lately, a new material of ferritic type has been made available which has such low conductivity that it may be used in the solid form for magnetic cores operating at radio frequency. In using dust cores, a compromise must be made between effective permeability and loss; the binding agent introduces a gap between the particles of magnetic material, and if very small particles are used the ratio of gap to iron increases and effective permeability decreases. The higher the frequency, however, the smaller must be the individual particles to maintain losses within reasonable bounds, and in consequence it is an unfortunate fact that the higher the working frequency of a dust-core, the lower the effective permeability. It is this fact which limits the frequency range within which a reduction of losses may be achieved through the use of dust cores.

Restriction of Magnetic Field

Two further advantages are to be obtained from use of magnetic cores. It is usual for several inductors to be connected in succeeding stages of amplification, and magnetic coupling due to mutual inductance between them can introduce instability and distortion of the pass-characteristic. The obvious method of preventing this coupling is to confine the magnetic field of each inductor within such limits that it will not cut the conductors of the other inductors. In the case of air-cored inductors only two methods are available for restricting the external field; either the inductor must be of the toroidal form, which has high losses and is difficult to wind, or a "screening can" must be used to enclose the conductor in such a way that the field will cut the screen and induce in it an eddy-current system which will neutralize the field external to it; unfortunately the screen will introduce additional losses into the inductor due to its own ohmic loss; in general, the nearer the screen to the inductor, the greater the reflected loss, while the use of large screens spaced well away from the inductor so increases the size that connections must be longer (and lead inductance higher) than can be tolerated at high frequencies. The use of an iron core, however, will concentrate the majority of the field within its own small compass, and either no screen will be necessary or it may be placed so near to the inductor without materially increasing losses that its use is no disadvantage. The second great advantage of magnetic cores lies in the possibility of varying the location of the core relative to the inductor, and hence, by varying the effective permeability of the inductor field, varying also the inductance of the unit. So useful is this ability to vary inductance by movement of a core, that cores may be used in cases where there is no decrease of loss, purely for the facility of adjustment. Where a very wide range of inductance is desired, several narrow solenoids may be arranged on a common axis, and a long cylindrical core be caused to slide axially through them; not only does this increase the permeability of the space occupied by the field of each individual inductor, but it also increases the mutual inductance between the individual solenoids, thus giving a very great ratio of maximum to minimum inductance.

Types of Dust Cores

Dust cores are made in four principal types. cylinders with brass threaded stem projecting axially from one end, by means of which they may be advanced along the axis of a solenoid; cylinders with coarse external thread moulded on, by means of which they may be axially advanced along a solenoid wound on a former with internal thread to engage that on the core; cores moulded in two halves to form a cylindrical assembly completely surrounding the solenoid; and toroids, usually of rectangular cross-section. The method of manufacture is such that close control of effective permeability is difficult, and in addition the overall temperature coefficient of winding plus core is usually more erratic than that of a well-constructed air-cored solenoid. The effective permeability of dust cores for very high frequency may be less than two, but for frequencies below 1 Mc/s it is possible to increase to as high a value as ten. Even higher values are available in the frequency range below 100 kc/s. The effective Q (ratio of reactance to resistance) of the inductor may be as high as 500 at 100 kc/s, falling to

300 at 1 Mc/s, and so to a value in the neighbourhood of 100 at 30 Mc/s. To obtain maximum Q the losses in the conductor should equal those in the core, a rule which enables the optimum proportions of inductor to be obtained for any frequency and material.

Ferrites

Cores made of ferrites are of more recent origin, and to a certain extent are still under development. Nevertheless, many excellent cores are commercially available now, and but for their high cost would have replaced the dust core over the frequency range in which they are advantageous. Physically the ferrites are analogous to some ceramic materials, and being very poor conductors have practical values of permittivity as well as permeability. A wide range of electrical characteristics are possible by suitable compounding and processing; the lower-frequency types have a resistivity of the order 50 ohm-cm. and a permittivity falling from 150,000 in the audio-frequency range to 50,000 in the radio-frequency range, while the higher-frequency types have resistivity falling from 5 MΩ-cm. to 0·2 at the higher frequencies and permittivity falling from 400 to 15. The initial permeability of the lower-frequency types varies from 850 to 1,500, while that of the higher-frequency types can be obtained from 20 to 650. As a low-loss core in an inductor resonant in the band 1 kc/s to 50 Mc/s this material is unequalled in performance, and it can be used in non-resonant inductors and transformers up to 200 Mc/s. Many physical sizes and shapes are available—rods, tubes, pots, laminations and the like—and due to the high value of permeability it is possible to obtain greater values of coupling factor in the radio-frequency range than by other methods. Due to difficulty in controlling the effective permeability during manufacture, it is usual to build the core with a small air-gap, and to adjust the overall permeability by grinding this gap to obtain the desired inductance value from the inductor. The price which must be paid for the high permeability, however, is the non-linearity inherent in all high-permeability materials. Not only does the effective incremental permeability vary with induction, but it also varies with frequency, and for this reason also, in cases where the inductance of the inductor must remain within close limits, it is essential to employ an air-gap in series with the core, and thus minimize the variation at the expense of reducing the overall permeability.

Protection Against Moisture

Where materials with finite conductance and permittivity are used in juxtaposition to the conductor, it becomes additionally important to prevent the ingress of moisture to the assembly, with its attendant disadvantages of leakage and other losses. For this reason it is essential that some form of protection be employed, but the constructional materials employed may seriously limit the choice of protective media. It is common practice to use a moulded polystyrene bobbin to contain the winding of dust-cored inductors, and while this material is admirable for that purpose, its low softening point prevents the use of wax or varnish impregnants, while its solubility in commercial solvents prevents the use of cold-setting lacquers. The so-called " casting resins " which have recently become popular make an ideal impregnant and enclosure for cored inductors, but where polystyrene formers are used the polyester resins may not be used because of the solubility of the former in the monomeric styrene used for cross-linking; this is to be regretted, because the polyester resins may be compounded with excellent electrical properties, and if they cannot be used, the most practical alternative in present supply is an ethoxyline resin compounded with a hardener capable of cold-setting. One alternative used where maximum protection is not essential is to cool-dip the complete assembly in molten wax, giving a skin of wax over the exterior without raising the internal temperature of the bobbin to a dangerous level.

Iron-cored Audio-frequency Inductors

In the audio-frequency range the eddy current losses may be suitably minimized by building up the core from a series of thin plates (laminations), each insulated from its neighbours. These plates are obtainable in various shapes and sizes, the most common being U, E, T and I shapes. These are formed either from a silicon steel with low hysteresis loss and high saturation or from nickel–iron alloys possessing very high permeabilities. Recently great progress has been made in rolling silicon-iron sheets to have greater permeability in one preferred direction than in the others, and such sheet is then cut into strip, with the direction of maximum permeability along the major axis; such strip is then wound into oval cores with a suitable binding agent to make it homogeneous, and each core is then cut into two equal U shapes to allow the winding to be inserted on the limbs. This mode of construction is known as C core, and is becoming increasingly popular for the larger physical sizes. Where it is desired to include an air-gap in

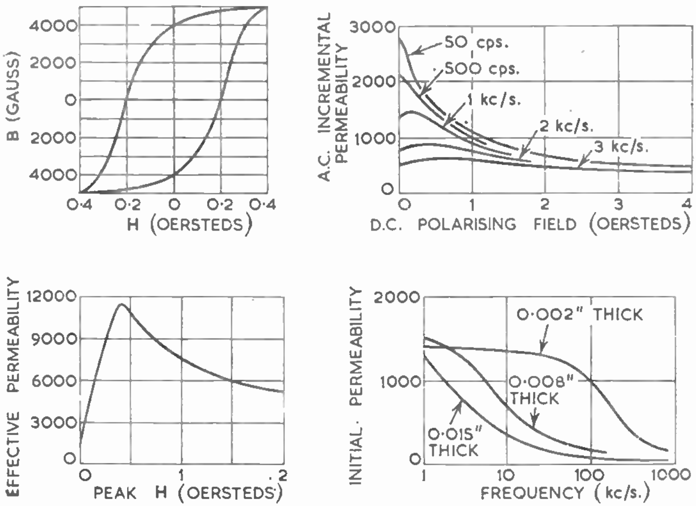


Fig. 1.—Typical Characteristics of Magnetic Laminations.

the core, the gap can be inserted between the two halves of the C core, or at an appropriate position in an assemblage of laminations. Where no air-gap is desired, however, the C core can only be butted, but in the case of laminations it is usual to assemble the core from two different shapes such as E and I or U and T, or to use an irregular E with short top limb and long bottom limb, and to so assemble them that the butt-joints on adjacent pairs of laminations are not adjacent to one another. This interleaving of pairs gives an effective air-gap smaller than a plain butt-joint.

Magnetic Characteristics of Alloys

Typical magnetic characteristics for a representative alloy are shown in Fig. 1, and it will be seen that the characteristics are non-linear. The hysteresis loss is proportional to the area included in the loop, and to minimize losses (and consequent heating of the inductor) it is desirable to choose alloys with a minimum area; unfortunately those with the minimum area may saturate at too low a flux density, or may have too low a permeability, and actual choice is always a compromise between these conflicting criteria. Many inductors are called upon to exhibit a high reactance with large values of D.C. flowing through them; in such cases the effective permeability is the tangent of the curve at the working point; it will be seen from the curves that the flux induced by the D.C. may well be in the saturation range of the material, and in consequence the effective permeability so low as to be useless. In such cases an air-gap is included in the iron circuit to limit the induced flux, and although inclusion of such a gap will reduce the effective permeability for a given tangent slope, due to the greatly increased slope the overall permeability will be greater with the gap. It is possible with any given core to calculate inductance versus gap for several values of gap, and then by constructing a curve of the relationship to choose the optimum value. In other cases it is desired that the effective inductance to small values of A.C. shall vary with the value of D.C. flowing through the winding (swinging choke), and this compromise may also be obtained by constructing a family of inductance/gap curves for several discrete values of D.C., and choosing the condition giving the desired spread of values.

Calculation of Inductance

To calculate the inductance of an iron-cored inductor requires knowledge of the effective permeability of the core. This is difficult to obtain theoretically; even for small A.C. components superimposed on a larger D.C. component, where the permeability might be expected to equal the tangent of the B.H. curve, eddy current effects will modify the penetration of the flux throughout the cross-section of the lamination, and effective permeability will vary with frequency and with thickness of lamination, being highest for thinnest laminations. When the A.C. component is so large that the portion of the curve traversed each cycle is not effectively linear, distortion of the waveform will occur, and the permeability will be an indeterminate " average " slope of the curve. This is further complicated by the inclusion of insulation between laminations, and also by the existence of tooling burrs on the laminations themselves, which will give a packing factor which may be as low as 0·9. It is wise therefore to pack a standard winding with typical laminations and by measurement of effective inductance to calculate the effective permeability of the core assembly complete; if this be done for each of the assemblies in which one is interested, the value so obtained may be used as the design permeability in calculations. The inductance of such an inductor may be calculated from the following formula:

$$L = \frac{4\pi N^2 A \mu}{10^9(l_c + \mu l_a)}$$

where N = number of turns;
A = cross-sectional area of core in square centimetres;
μ = effective permeability of core assembly;
l_c = effective length of core in centimetres;
l_a = effective length of air-gap in centimetres.

This formula assumes that the permeability of the core is so high that the cross-section of the core may be assumed the total area of flux, and while this will not be strictly true, a first-order allowance for the error will have been made in measuring the effective permeability of the core assembly.

Where losses must be minimized, and a high Q obtained, it is essential for any given core assembly that the winding losses be approximately equal to the core losses. In most interleaved assemblies it will be found that the core losses are much higher than the winding loss, and the usual method of adjustment is to introduce an air-gap into the core circuit, thus reducing peak flux density and increasing winding resistance; a variation which both reduces core loss and increases winding loss. By measuring Q on a few inductors with differing gaps and constructing a curve to pass through the points relating gap and Q, the optimum gap for a particular core and frequency may be found.

TRANSFORMERS

Transformer Design Methods

If two windings are so located that a portion of the flux from one encloses also the other, variation of current in one will generate an e.m.f. in the other. Since the first differential of a sinusoidal curve is also sinusoidal, the steady-state solution for a sinusoidal current flowing through one winding is a sinusoidal e.m.f. in the other, and the complete assembly is called a transformer. In a perfect transformer all the flux

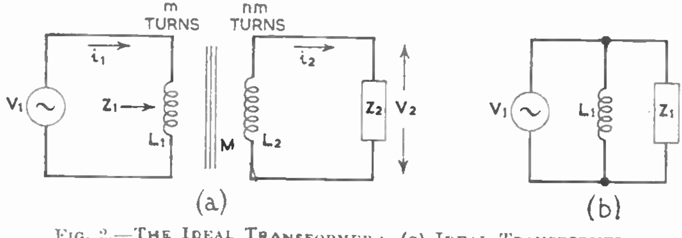


Fig. 2.—The Ideal Transformer: (a) Ideal Transformer; (b) Equivalent Circuit.

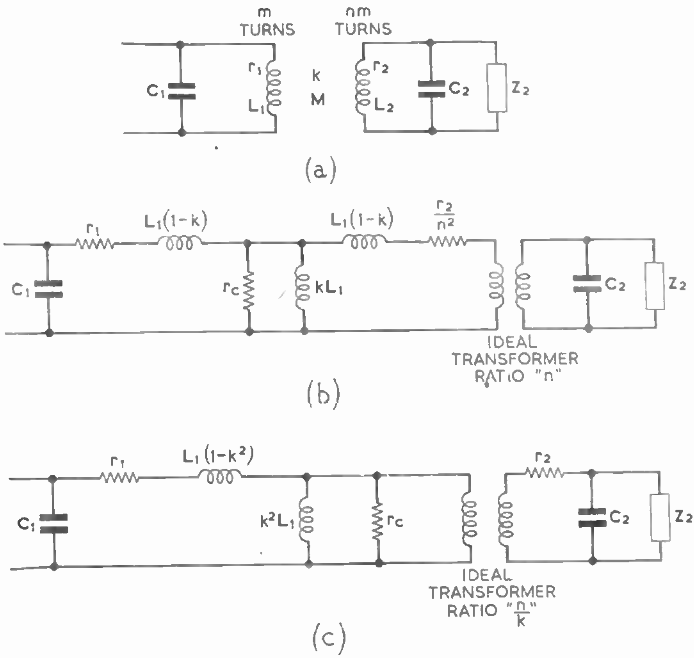


Fig. 3.—Network Approximations to the Practical Transformer.

from one winding will enclose all of the other winding, and the two windings will have a coupling factor of unity. Such a transformer is shown diagrammatically in Fig. 2, and for such a system the following relationships will hold :

$$L_2 = n^2 L_1$$

$$Z_1 = \frac{Z_2}{n^2}$$

$$V_2 = nV_1$$

$$i_1 = ni_2$$

$$M^2 = L_1 L_2$$

Analysis of Practical Transformers

A practical transformer with losses may be analysed into a passive network plus an ideal transformer as shown in Fig. 3 and calculation eased by treatment of the network alone. If the inter-winding capacitances must be considered, the more elaborate network shown in

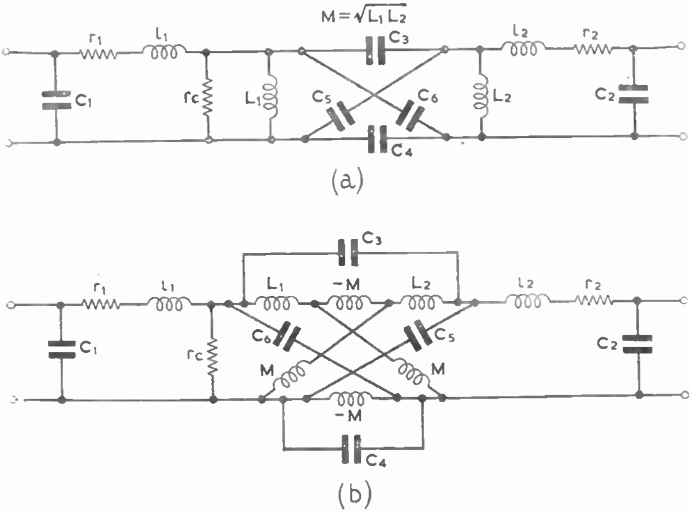


FIG. 4.—REPRESENTATION OF ALL STRAY CAPACITANCES IN A TRANSFORMER: (*a*) GENERAL REPRESENTATION OF TRANSFORMER; (*b*) EQUIVALENT CIRCUIT OF TRANSFORMER.

Fig. 4 must be used, but this is so ungainly for practical calculation that the approximations of Fig. 3 are to be preferred where the inter-winding capacitances can be ignored. Since these capacitances are usually very undesirable, it is good practice in any case to so design the transformer that they are so small as to be negligible, and thus achieve a desirable performance. When a transformer has to pass a very wide frequency range, the effect of strays is to cause an attenuation of the high and low frequencies relative to the mid-band components, and to

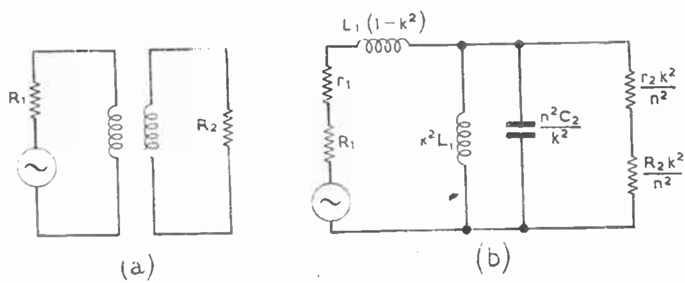


FIG. 5.—EQUALIZATION OF TRANSFORMER PASS-BAND: (*a*) LOADED TRANSFORMER; (*b*) EQUIVALENT BAND-PASS CIRCUIT.

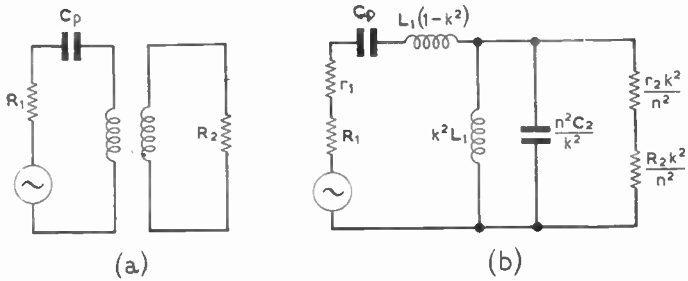


Fig. 6.—Series Resonance of Primary: (a) Loaded Transformer; (b) Equivalent Band-pass Circuit.

minimize this effect and obtain a flat characteristic it is necessary to so proportion the strays that those of importance form a band-pass filter in the equivalent network. This involves designing the transformer parameters to have the correct relationship to one another, and the equivalent circuit for the transformer alone is shown in Fig. 5. The band-width which may be attained with this connection is:

$$\frac{\omega_2}{\omega_1} = \sqrt{\frac{2 - k}{2(1 - k)}}$$

and it is obvious that unless k can be made very near to unity, high ratios cannot be obtained. Where a high ratio is required it is sometimes possible by inserting a series capacitor in the primary circuit to obtain a constant k type of filter as shown in Fig. 6. Where a fairly wide band is

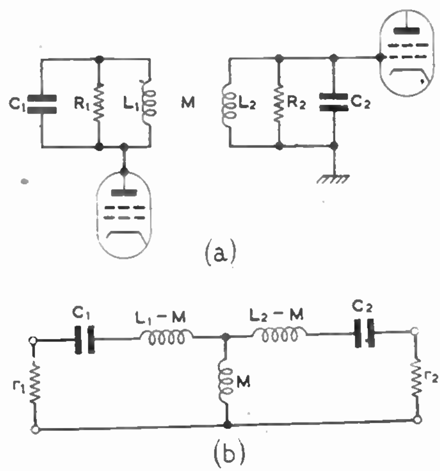


Fig. 7.—Transformer as Band-pass Filter: (a) Practical Application of Transformer as Band-pass Filter; (b) Approximate Equivalent of Tuned Band-pass Transformer.

required, up to about 50 per cent of the mean frequency, as is the case in many radio-frequency circuits, the arrangement shown in Fig. 7 is frequently used, the two capacitances being the self-capacitance of the windings plus the electrode capacitances of the valves; this is, of course, an application of the arrangement in Fig. 6. The resistances R_1 and R_2 are partly losses in the windings and partly resistors shunted across the windings.

Power-transformers

When designing power-transformers, in which a considerable power is to be transferred at a single frequency between two low impedance circuits, the loss of the core usually assumes greatest relative importance. The loss, expressed as watts per pound of core material, is a function of maximum flux density when sinusoidally excited, and if very large flux densities are used, the temperature rise of the core due to this power loss may become excessive. The alternative to using high flux density is to increase the number of turns, which for a given core size entails decreasing the conductor diameter and increasing conductor resistance and loss. If therefore it is found that the conductor losses are too high when a reasonable flux density is used and that they in their turn cause excessive temperature rise, it is necessary to use a larger core with greater area and winding space. The use of a larger core, however, will increase the weight by a greater ratio than the surface area, and in consequence the temperature rise due to core losses will be higher for a given flux density than with the smaller unit. It is obvious therefore that the design of a power transformer will be a compromise between opposing factors, and it is usually desirable to obtain empirical data on each of the standard core assemblies which it is intended to use.

Linearity of Core

In cases where the frequency range is so limited that no great difficulty due to frequency characteristic is experienced, and where the power is so low that core losses will not cause excessive temperature rise, it is possible that a third factor may become of prime importance. The magnetic characteristic of the core is non-linear, and in consequence the primary current contains harmonics of the primary voltage waveform; the secondary voltage is a function of primary current, and will also contain these harmonics, and where such "distortion" is not permissible, it is frequently necessary to "linearize" the core characteristic by the inclusion of an air-gap. Where a D.C. is to be superimposed on the A.C. component, an air-gap is included in any case to prevent saturation of the core, but it becomes necessary in such circumstances to work at a point on the characteristic where the signal will not carry the core into a saturated portion of its characteristic.

Constructional Methods

In designing a transformer there are a number of independent requirements which must be met. In each case there are a number of constructional techniques which have become more or less standard, and the practical transformer consists of a combination of these to meet requirements. The possible permutations are extremely large, but by considering the individual points separately a large range may be covered concisely.

Control of Capacitance

Where high frequencies are concerned, or high-impedance circuits are connected to a transformer, the capacitances of the windings are of vital importance. Self-capacitance of a winding may be reduced by winding in narrow sections, or, in extreme cases, by wave-winding a number of narrow sections, later joined in series. Where this is done, however, it is essential to space the sections from any neighbouring conductor to prevent such conductor from forming a shunt across the total winding by virtue of its capacitance to either extremity. In general, it may be ruled that reduction of self-capacitance is always wasteful of winding space, and raises the winding loss in every case. The capacitance between windings may be minimized by enclosing the windings in an electrostatic screen which is connected to a low-potential point in the circuit; this may reduce it to almost zero, but inevitably increases the self-capacitance of the windings; it is essential that such screens be sectioned to prevent the establishment of circulating currents due to variation of the magnetic field through the core, and such sectionalization often leads to very complicated physical construction. In other transformers, used in circuits balanced to "earth", it is necessary to so control capacitance between windings and screens that the value is almost exactly equal for two different windings on the same transformer, and this involves controlling the dimensions of the insulation to very close limits, and operating a quality-control system on the product.

Control of Inductance

Close control of inductance can also present great difficulty. In the case of air-cored inductors and transformers, if a great number of turns are present as in wave-windings, they may be over-wound and a few turns removed to bring within limits, and if only a few turns are present as in single-layer solenoids, the spacing between adjacent turns may be varied with the same result. Radio-frequency windings with magnetic cores are adjusted most readily by variation of the effective air-gap in the magnetic circuit, thus varying its permeability. In the case of laminated cores, the inductance may be varied by adjustment of an air-gap, or by adjustment of the number of laminations packed into the former, any spare space being occupied by a non-magnetic lamination such as a shaped strip of S.R.B.P., but neither of these methods is easy, and close tolerances are avoided in this type of construction wherever possible. Variation of mutual inductance between two windings is possible by variation of the relative spacing if air-cored, or of the interconnecting magnetic circuit if cored, but in either case variation of mutual inductance almost inevitably involves a variation also of the self-inductance.

Magnetic Screening

When it is desired to prevent magnetic interaction between two adjacent inductors or transformers, either the windings may be so constructed (astatic winding or toroid) that the magnetic circuit is symmetrical with respect to the surrounding space, or the complete units may be screened. Screening may consist either of a non-magnetic, highly-conductive material (as for high frequencies) or of a magnetic material of high permeability (nickel–iron alloy) when only audio frequencies and steady fields are concerned.

Atmospheric Deterioration

One of the most difficult tasks is to prevent deterioration due to the surrounding atmosphere, and this is more difficult when a component has to function in a tropical or arctic climate than when its use will be restricted to a temperate zone. The most effective method is to totally enclose the unit in a metal case and bring out the connections through hermetic seals, but this method is by far the most expensive, and is used only where essential. When this is done, it is possible to provide additional insulation for high-voltage working, and additional heat-conductance for high power dissipation by filling the unit with transformer oil, leaving, of course, sufficient air-space to form a pneumatic buffer against variation of volume with temperature. A more usual method is to vacuum-impregnate the windings in either varnish, wax, bitumen or polyester resin, and then to cover the whole assembly with a coat of either wax, varnish, bitumen, polyester resin or ethoxyline resin. Any one of the impregnating mediums may be used in conjunction with any one of the coating mediums, but those in most frequent use today are wax and wax, wax and bitumen, varnish and bitumen, varnish and varnish, polyester resin and ethoxyline resin. Where self-capacitance must be maintained at an absolute minimum and an impregnating medium cannot be used, the only really effective protection is a sealed cover filled with dry air.

As has been mentioned already, when high powers must be dissipated, oil may be used as a filling medium to conduct heat from the interior of the windings to the outer case. Whatever the construction, however, it is vital in such cases that the overall colour should be a dark, matt to assist radiation of heat, and that the unit should be clamped to a metal chassis by heavy metal clamps in good thermal contact with core and windings to assist in conduction. For the same reason the unit should be surrounded with matt, dark-coloured objects to absorb radiated heat rather than reflect it back again into the unit.

R. T. Lo.

31. MICROPHONES AND GRAMOPHONE PICK-UPS

31. MICROPHONES

A transducer is a device for receiving energy in one form and passing it on in another form. Typical examples are a fluorescent powder which receives invisible radiations and passes on visible light; and a microphone which receives audible radiation and passes on inaudible electric voltages. It is therefore necessary to make a three-fold study. First, of the energy received by the microphone; second, of the mechanism of transformation, and third, of the energy passed on.

ACOUSTICS

A microphone receives its energy from sound radiation in the air, and it may help in the understanding of the operation of a microphone if the mechanism of sound radiation is first considered. A gas, air is a typical one, is made up of an enormous number of molecules. They are separate discrete particles which move about at quite high velocities, 500 m./sec. depending on their temperature, expressed in degrees Kelvin above absolute zero. Since there are so many of them and they move so fast, there will be many collisions, in fact the average path length before a collision at normal room pressure and temperature will be about $\frac{1}{10000}$ mm. However, they "bounce" perfectly and under normal conditions they do not mind in what direction they travel. A sound wave occurs when a number of molecules are all pushed in the same direction, they jostle into their neighbours in that particular direction, which move away from them and interfere with others beyond them, thus producing a travelling wave. It can be shown that the speed of travel c in cm./sec. in this wave is given by

$$\frac{\gamma \rho_0}{\rho} = c^2 \qquad \qquad (1)$$

where $\rho_0 =$ the static pressure in dynes per sq. cm.;

$\rho =$ the static density;

$\gamma =$ the ratio of specific heat at constant pressure to that at constant volume. It has a value of 1·4 for air.

For a plane wave, assume that a progressive wave proceeds along the axis of x. Then the velocity potential ϕ is a function of x and t only, and the wave equation may be written

$$\frac{\partial^2 \phi}{\partial t^2} = c^2 \frac{\partial^2 \phi}{\partial x^2} \qquad \qquad (2)$$

A solution to this equation for a simple harmonic wave travelling in the direction x is

$$\phi = A \cos k(ct - x) \qquad \qquad (3)$$

where

$$\begin{aligned} A &= \text{the amplitude of } \phi; \\ K &= 2\pi/\lambda; \\ \lambda &= \text{the wavelength in cm.;} \\ c = f\lambda &= \text{velocity of sound in cm./sec.;} \\ f &= \text{frequency in c/s.} \end{aligned}$$

The particle velocity u in a plane sound wave is given by

$$u = \frac{\partial\phi}{\partial x} = kA \sin k(ct - x) \quad . \quad . \quad . \quad (4)$$

The pressure p in a plane sound wave is given by

$$p = kc\rho A \sin k(ct - x) \quad . \quad . \quad . \quad (5)$$

Note that pressure and particle velocity are in phase in a plane wave.

Particle amplitude in a plane sound wave is given by

$$E = -\frac{A}{c} \cos k(ct - x) \quad . \quad . \quad . \quad (6)$$

The characteristic impedance is equal to the ratio of the sound pressure to the particle velocity in a free plane wave. It is equal to the density times the velocity of propogation ρc. It is expressed in rayls (1 dyne/cm.^{2} produces a velocity of 1 cm./sec.) and is approximately 42 rayls for air. The characteristic impedance of the free air is modified by the presence of the microphone, and this modification is revealed in the electrical output of the microphone.

General Observations on Microphones

It has been shown that a sound wave consists of two components, particle velocity and pressure. A microphone should make use of both components, but it has not been possible to construct a mechanical system with the low impedance necessary to match to the air and so respond accurately to the velocity component. The nearest approach is the ribbon velocity microphone with its gossamer-thin ribbon, but the acoustic impedance of even this is more than 200 rayls, and is a very considerable mis-match to the 42 rayls of the air. This fact, coupled with the almost universal use of voltage-amplifying valves, gives rise to the common treatment of a microphone as a transducer of sound pressure into electrical pressure.

Pressure-operated Microphones

Pressure-operated microphones include all those types where the sound pressure operates on one side of the diaphragm only; it includes the condenser, crystal, carbon and the moving coil. Since the pressure in a fluid or a gaseous medium is transmitted equally in all directions, it follows that the sensitivity of microphones operating on sound pressure is independent of the angle at which the sound strikes the microphone.

Diffraction and Reflection

At low frequencies, where the size of the microphone is small compared with the wavelength of the sound, the obstruction caused by the presence of a microphone whose acoustic impedance does not match that of the air will cause slight diffraction of the sound wave; but where the wavelength is comparable to or smaller than the size of the microphone, reflection takes place. This modifies the response of the microphone to high-frequency sounds arriving from the front of the microphone. In use, a microphone having polar response which is independent of frequency is to be preferred.

Velocity-operated Microphones

There are no microphones available operating strictly on the velocity component of the sound wave due to the difficulty of making a mechanical system with a sufficiently low impedance, but a fairly close approximation can be obtained by making a microphone to operate on the "pressure gradient" of the sound wave, which is proportional to velocity in those cases where the characteristic impedance is unchanged. The term "velocity microphone" will be used here to indicate both the true velocity microphone and a practical first-order pressure-gradient microphone. Since the air-particle velocity in a sound wave is in the direction of propagation of the wave, it follows that microphones operating on this principle will have a very different polar response to pressure-operated microphones.

In all velocity microphones both sides of the moving member are exposed to the sound field, and movement is caused by the sound-pressure difference between the two sides of the moving member. There will be a maximum pressure difference perpendicular to the diaphragm or ribbon, decreasing towards the sides with the cosine of the angle of incidence of the sound wave. The relationship is expressed by:

$$\theta = \sqrt{\theta_0^2 \cos^2 x} \qquad \qquad (7)$$

where θ_0 = the sensitivity at maximum excitation; and
x = the angle of incidence of the sound wave.

The magnitude of the effective pressure gradient depends upon the wavelength of the sound, and on the length of path from the centre of the front of the diaphragm or ribbon, round the edge of the capsule, to the centre of the back of the diaphragm or ribbon. Such a microphone has a "figure-of-eight" polar response in which the phase of the electrical output from the microphone reverses for sounds coming from the rear.

Self Noise

With the notable exception of the carbon microphone, the self noise generated by the microphone is not measurably greater than the noise produced by thermal agitation in the output impedance of the microphone, and it is this thermal noise which finally limits the minimum sound field which may be detected by the microphone. For microphones whose output impedance is substantially resistive, the thermal agitation noise in the output impedance of the microphone is given by

$$e = 1{\cdot}27 \times 10^{-10}\sqrt{R(f_2 - f_1)} \qquad \qquad (8)$$

where e = the r.m.s. open-circuit voltage in the band between f_1 and f_2 c/s;

$R(f)$ = the resistive component of the electrical impedance of the microphone, in ohms, sometimes a function of frequency;

f_1 and f_2 = the limiting frequencies of the pass band.

In the case of condenser and crystal microphones

$$R(f) = \frac{R}{1 + (\omega CR)^2} \qquad (9)$$

where $\omega = 2\pi$ times the frequency f.

It can be shown that a high-pass filter should be used wherever possible with these types of microphones to eliminate the large low-frequency noise whenever possible. If a high-pass filter is used, or if the amplifier naturally cuts off the low frequencies, less noise is obtained with a higher value of grid resistor.

Sound Levels

Acoustic power is expressed in watts, and covers a vast range, about 1,000 billion to one. It is thus almost essential to use a logarithmic unit, and the decibel is commonly used.

However, since most microphones are pressure operated, and all are mis-matched to the air, it is customary to measure acoustic pressure rather than acoustic power. This is conveniently expressed in decibels above a fixed reference level. The reference level chosen is 0·002 dynes/cm.² and is almost the pressure produced by a sound power of 10^{-16} watts. It so happens that it is also about the limit of audibility of a 1,000-c/s note for young ears.

Table 1 shows the relationship in pressure between some everyday sounds.

TABLE 1.—PRESSURE OF TYPICAL EVERYDAY SOUNDS

Pressure	*Typical Source*
140	Pneumatic riveter
130	Boiler shop
120	Thunder
110	Motor horn (at 3 ft.)
100	Loud Orchestra; Inside airliner
90	Inside motor bus
80	Voice shouting
70	Noisy office
60	*Conversational speech* (at 3 ft.)
50	Average office
40	Quiet speech (at 3 ft.)
30	Very quiet dwelling
20	Whisper (at 5 ft.)
10	Watch ticking (at 2 ft.)
0	Threshold of hearing for young ears

The above table is only intended to serve as a guide to those not acquainted with acoustics; there is, of course, a very considerable "spread" in the sound pressures made by various items.

CHOICE OF MICROPHONE

Microphones are used for a very wide variety of purposes, and as a result, a large number of types have been developed to meet particular requirements. The purpose of this section is to show what are the factors to be considered in choosing a microphone to fit a particular set of conditions.

Most types of microphones have some particular characteristic which makes the type unsuitable for certain applications. Examples are the polarizing voltage necessary for a condenser microphone, which may not be readily available; and the high noise level of the carbon microphone, which makes it unsatisfactory for the reproduction of music.

The Electrical Output of the Microphone

Sensitivity.—The sensitivity of the microphone must be sufficient to give the required result with the external amplification available. The carbon microphone, which has self-amplifying properties, is used in Post Office circuits because no external amplifier is available. For measurement purposes, the sensitivity of a microphone must remain constant over long periods of time.

Frequency Response.—Both the extent of the frequency range of the electrical output from a microphone and the variation in sensitivity for different frequencies within the frequency range must be taken into account.

Polar Response.—Some acoustical considerations of the polar response of microphones are given later. The polar response is sometimes altered in order to correct the frequency response of the microphone, which is usually considered to be the more important. There is no doubt that, in the case of a good microphone, both the frequency response and the polar response should be independent and uniform.

Non-linearity.—The output from the microphone may not be a linear function of the input to it. In this case it will give rise to the generation of harmonics of the original frequencies present in the sound wave, and, in addition, if more than one frequency is present at the same time, sum and difference tones will be produced.

Phase Distortion.—This distortion may frequently be neglected unless two or more microphones are employed together to obtain special directional effects.

Self Noise.—The noise generated by the microphone may be acoustic in origin, as when a microphone is used in a wind, or may be discontinuous electrical currents as in the carbon microphone, or thermal noise in the output circuit of the microphone, as is the case with a condenser microphone.

The Environment of the Microphone

Sound-pressure Level.—The microphone must be capable of withstanding the maximum sound pressures that may be experienced in its particular application. In addition, it must withstand barometric changes in pressure, and direct breath from the users lips. This last requirement is severe in the case of a hand microphone.

Temperature and Humidity.—The microphone must survive the climates through which it is to be shipped, and the final destination. This has particular importance in connection with some forms of crystal microphone.

Mechanical Shock.—The "ruggedness" of the construction of the microphone must be adequate, but it generally means that the mismatch between the air and the diaphragm or ribbon which invariably takes place is made even worse if the design is "rugged".

Susceptibility to Stray Electro-magnetic Fields.—For some particular purposes, such as measuring sound pressures near electrical machinery, it is essential for the microphone to have no appreciable external field, which may easily give an output comparable with that from the sound field.

Size.—With the widespread use of motion-picture television cameras, it is necessary to make the microphone as small and inconspicuous as possible.

Weight, reliability and *cost* must also be taken into consideration.

THE CONDENSER MICROPHONE

The condenser microphone was first introduced by E. C. Wente in 1917 (*Phys. Rev.*, **10**, 39–63, 1917), and was improved by D. A. Oliver (*J. Sci. Inst.*, Vol. 7, No. 4, April 1930) and also by v. Braunmühl and Weber (*Hochfrequenztechnik u. Elektroakustik*, Vol. 46, 187–192, 1935). This type of microphone can be designed to cover a very wide range of frequencies, and its performance accurately predicted. Its stability of calibration can be made so great that a number of these microphones made by D. A. Oliver in 1930–31 are still in use by standardizing laboratories.

Mode of Operation

This microphone is essentially a condenser, one plate of which is made in the form of a very thin diaphragm, which is moved towards or away from the fixed electrode by the incident sound pressure. When a positive sound pressure displaces this diaphragm inwards, the capacitance of the condenser increases; and since the resistance in series with it and the polarizing potential are very large, 10–1,000 MΩ are used. The

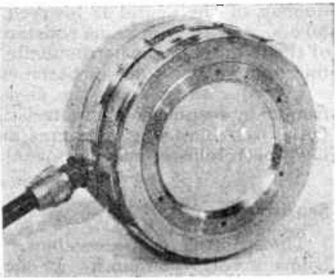


FIG. 1.—LABORATORY STANDARD CONDENSER MICROPHONE.
(*The General Electric Co. Ltd.*)

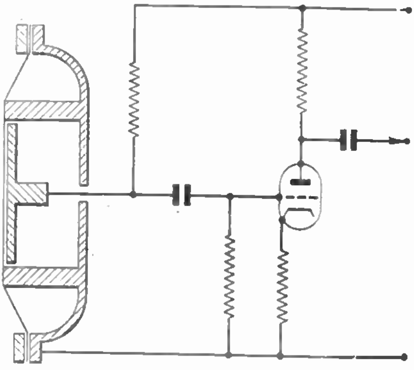


FIG. 2.—SCHEMATIC DIAGRAM OF A CONDENSER MICROPHONE AND CIRCUIT COUPLING TO A VALVE.

charge on the condenser remains constant except at very low frequencies. Hence the potential across the condenser drops, being equal to the charge divided by the capacitance. When the cycle reverses, the diaphragm moves away from the back plate, and the potential across the condenser increases. The output voltage is proportional to the displacement of the diaphragm. In the case of microphones which are intended to be used for measurement, the diaphragm is made of metal, usually stainless steel or duralumin, and subjected to very considerable tension. The tension stiffens and flattens the diaphragm, and in conjunction with the effective mass of the diaphragm it produces a mechanical resonance which has to be controlled by mechanical or acoustical damping. In the case of D. A. Oliver's microphone this damping was provided by a very thin layer of air 0·025 mm. thick, which was trapped between the diaphragm and the perfectly flat back plate which forms the second plate of the condenser. The normal resonant frequency of the diaphragm by itself without the back plate was about 8,000 c/s., but in the assembled microphone, it was quite impossible to find any trace of this resonance.

On the other hand, the viscosity of this thin lamina of air is quite low at low frequencies, and it provides no damping. The result is that the electrical output of the microphone rises gently as the excitation frequency of the sound is lowered. It reaches its final value at about 50 c/s., and then remains constant at all lower frequencies. If the shell of the microphone is hermetically sealed, it is possible to use a condenser microphone as a form of barometer or pressure gauge.

In order to increase the electrical output, and still retain a good frequency-response characteristic, some designers have used thin metallized plastic diaphragms and special forms of back plate which have been relieved with grooves, small holes and other devices

Sensitivity

The sensitivity of a condenser microphone is of the order of -60 db below 1 volt/dyne/cm.^{2}. It should be noted that since the electrical output impedance of a condenser microphone is a capacitance, the effect of any capacity associated with a connecting cable will be to form a capacity potential divider which will reduce the sensitivity of the microphone in the direct ratios of the capacitances, but it will not alter the frequency response. The very high electrical output impedance makes it necessary to couple the microphone immediately to a valve or other device for reducing the impedance.

Frequency Response

The frequency response of a condenser microphone can be made to be very good, and for certain forms of microphone construction it can be computed with a very high degree of accuracy. Condenser microphones have been used at frequencies up to 100 kc/s.

Polar Response

Pressure Microphone.—The polar response of a normal condenser microphone, being pressure operated, has a spherical distribution for all frequencies of sound up to that frequency where the wavelength of the sound becomes comparable with the dimensions of the microphone; above this point the microphone will have greater sensitivity to sound falling on the front of the instrument.

Velocity Microphone.—Until 1935 nearly all of the condenser microphones available had been sensitive to the pressure component of the sound wave, but von Braunmühl and Weber showed how to construct a condenser microphone to operate on the "velocity" component of the sound wave. This was done by providing a fixed electrode which was perforated so that both sides of the diaphragm were now exposed to the sound field, and the driving force was changed from the single sound pressure on one side of the diaphragm to the sound-pressure difference between the front and the back of the diaphragm. Such a microphone will have a "figure of eight" polar response.

Cardioid Polar-response Microphone.—If the output from a microphone having a spherical response is combined with the output from a microphone having a "figure-of-eight" response and the sensitivities are adjusted to be exactly equal for sounds arriving on the common axis, it is possible to obtain a cardioid polar response diagram. However, difficulties arise at the higher frequencies where it is impossible to make two microphones occupy the same space at the same time in order that they may receive the same sound wave. The problem may be solved by constructing a condenser microphone with two diaphragms separated by a perforated and drilled electrode. It is not necessary for the second diaphragm to be electrically connected, or even conducting; its presence is necessary to make the electrically connected front diaphragm respond to both pressure and "velocity" at the same time. If the microphone is very accurately constructed, so that the two types of operation result in equal outputs for sounds arriving on the axis, a cardioid type of polar response will be obtained.

A microphone of this type uses very thin plastic membranes which are rendered conducting by gold spluttering. The resonant frequency of the two diaphragms by themselves is quite low, but in use the two diaphragms are tightly coupled together by the stiffness of the air enclosed in the small space between them.

Variable Polar-response Microphone

It has been shown that if two cardioid-type microphones are mounted back to back, and the sensitivity and polarity of one is varied by altering its polarizing potential, it is possible to obtain a spherical, "figure-of-eight", or cardioid polar response at will (H. Grosskopf, *FZT*, p. 398, September 1951, and F. W. O. Bauch, *Wireless World*, February 1953).

Since, in the design of the cardioid microphone the rear diaphragm was required for acoustic reasons only and no electrical connection was made to it, it may conveniently be combined with the main electrically connected diaphragm of the rear-facing cardioid microphone, whose rear unconnected diaphragm is now the front diaphragm of the forward-facing cardioid microphone.

The change over from one type of polar characteristic to another is very easily obtained by altering the potential of the rear diaphragm with respect to the front diaphragm.

Cardioid.—Rear diaphragm earthed, effectively one microphone only is in operation.

" *Figure-of-eight.*"—Rear diaphragm opposite polarity to the front, effectively two cardioids in opposition.

Spherical.—Rear diaphragm same polarity as the front, effectively two cardioids in addition.

Non-linear Distortion

The condenser microphone is usually made in an unbalanced form with one diaphragm and one fixed electrode. Owing to the high acoustic impedance of the diaphragm, its displacement is very small and the total non-linear distortion is less than 1 per cent even for very loud sounds above 120 db above 0·002 dynes/cm.^{2}.

Special Applications

The condenser microphone is mainly used for very high-quality sound pick-up. Its low output, coupled with its high electrical impedance and the necessity for providing a completely hum-and-noise-free polarizing voltage, prevent its more general use.

It is largely used for measurement purposes because of its stability, and may be fitted with a probe tube to reduce its effective size.

THE CARBON MICROPHONE

The carbon microphone is used in greater numbers than any other type of microphone because of its sensitivity, cheapness and rugged construction.

The carbon microphone is fundamentally a very simple device consisting of a diaphragm actuated by sound pressure carrying a plate which compresses a loose pack of carbon granules. Pressing the granules together decreases their resistance; when the pressure is released, the resistance is restored. When a constant current passes through the microphone, the potential drop across the resistance of the pack is inversely related to the instantaneous sound pressure. Since the carbon resistance acts as a control valve rather than as a generator,

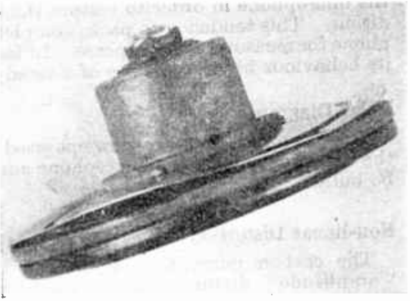


FIG. 3.—SIDE VIEW OF A CARBON MICROPHONE SHOWING CONTAINER FOR GRANULES.

(The General Electric Co. Ltd.)

it is quite possible for the microphone to give out more power in an electrical form than it receives in an acoustic form.

Sensitivity

The carbon microphone is generally used at a distance of only a few inches from the lips of the speaker, and at this distance it will give an output of 0·1 volts. The output impedance of the microphone is resistive and only about 100 ohms, so that it is quite possible to use a step-up transformer with a high ratio, bearing in mind that the primary of the transformer will carry the polarizing current of the microphone. The carbon microphone is not suitable for use in very weak sound fields because of the large amount of self-noise.

Frequency Response

The frequency response of a carbon microphone is limited to the range from 200 to 3,500 c/s for use in telephone systems. Special carbon microphones have been made with a very much greater and smoother response, but this is achieved only at the expense of sensitivity, and they still have a large amount of self-noise. Further, this frequency response is not a constant, but depends upon the condition of the microphone; the carbon granules pack together if the microphone is used in the same position with the bias current passing through it for a considerable time. When this happens the frequency response is usually improved to a small extent, but the sensitivity of the microphone may be reduced to a very large extent. It then becomes necessary to loosen the carbon granules by shaking or vibrating

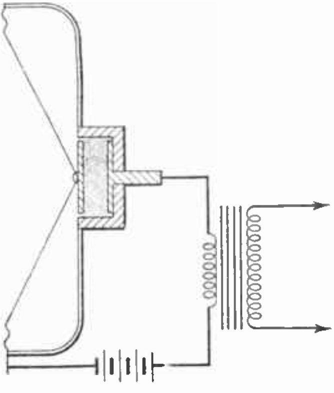


FIG. 4.—CARBON MICROPHONE AND CIRCUIT.

the microphone in order to restore the microphone to its original condition. This tendency to pack, completely rules out the carbon microphone for measurement purposes. In fact, it is very difficult to measure its behaviour because the use of a steady tone also causes it to pack.

Polar Diagram

This is normal for a pressure-operated microphone, but is modified by the size and shape of the microphone and the handset into which it may be built.

Non-linear Distortion

The carbon microphone is so bad in this respect that the term "amplitude" distortion is often used to describe the very non-linear relationship which exists between the acoustic input and the electrical output. It gives rise to gross amounts of cross modulation even at the level of normal speech. Push–pull carbon microphones have been made with the intention of reducing this form of distortion, but if, as may easily happen, one side of the microphone packs and refuses to "push" the resultant distortion may be worse than in the case of a single microphone.

Special Applications

The self-noise generated by a carbon microphone is so great that this type of microphone can only be used on low-quality circuits. The noise is largely due to the passage of the polarizing current through the carbon granules; more accurately, it is the variation in resistance of the various points of contact between one grain and the next, which are always altering due to the heat generated in the microscopic areas of contact. There are other causes of noise of a different kind, due to the heating up of the case of the microphone as a whole, which may make the container of the granules expand and so reduce the pressure on them.

Hearing Aids

Owing to the amplifying action of the carbon microphone, it has been used in a number of cases where valve amplifiers were impracticable. A notable case was in the older types of hearing aid, where a second carbon microphone was sometimes coupled to a miniature telephone earpiece to pick up and amplify the weak sounds made by it. It is rapidly being superseded by valve and transistor amplifiers, and is not liked even as a microphone because its self-noise becomes very trying when the aid is worn for long periods, particularly if the wearer suffers from tinitus.

The carbon microphone "pack" without its acoustic diaphragm is still used as a vibration pick-up in connection with the amplification of some instruments, but only to produce strange effects.

Power Microphone

If the ability to amplify is carried to the extreme limit, and a very short life is acceptable, it is possible to build a carbon microphone assembly which is capable of supplying enough power into an efficient horn loudspeaker to give understandable speech at a distance of ¼ mile.

FIG. 5.—GENERAL VIEW OF A MOVING-COIL MICROPHONE.
(*Standard Telephones & Cables Ltd.*)

THE MOVING-COIL MICROPHONE

A moving-coil microphone is a more complicated mechanical structure than either the condenser or the carbon type. However, it is comparatively rugged, and can be used without any polarizing voltage or current, their place being taken by a permanent magnet. The electrical output impedance is low, permitting the amplifier to be situated at some distance from the microphone.

The moving-coil microphone is basically a moving-conductor microphone, having the conductor in the form of a coil which may be made to cut the lines of force of the permanent magnet by movement imparted to the coil by the diaphragm. The voltage produced by the microphone is proportional to the rate of cutting lines of force: that is, for a given flux density, it is proportional to the velocity of the coil.

The mechanical system consists of a diaphragm to which is attached

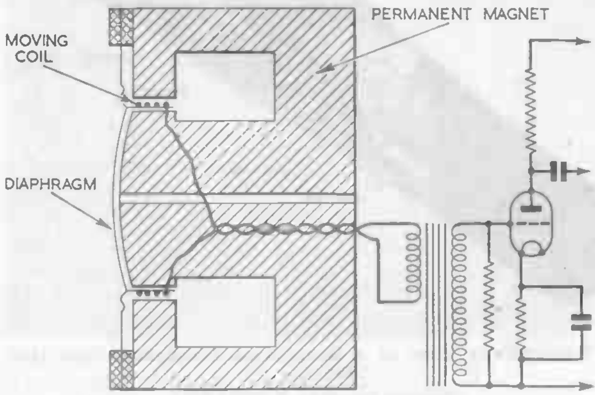


FIG. 6.—A MOVING-COIL MICROPHONE IN CIRCUIT.

a circular coil which dips into an annular air gap having a high magnetic field. Various acoustical and mechanical elements have been devised to control the response of the microphone, which is usually designed for pressure operation.

Sensitivity

The sensitivity of the moving-coil microphone is usually considerably better than that of the condenser microphone. The output impedance is between 15 and 50 ohms, and the open-circuit sensitivity is about 80 db below 1 volt/dyne/cm.^{2}, for a high-quality microphone with a good frequency response. Owing to the low output impedance of the microphone, it is customary to feed it into the grid circuit of a voltage amplifier via a step-up transformer, which may increase the output voltage by as much as 40 db.

Frequency Response

Moving-coil microphones are manufactured in such a wide range of prices and sensitivities that the frequency response varies considerably for different designs.

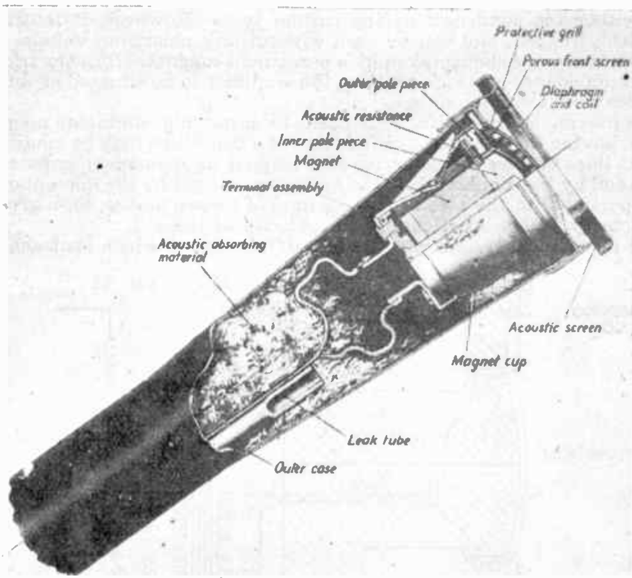


Fig. 7.—Cut-away View of a Moving-coil Microphone (see also Fig. 5).

(*Standard Telephones & Cables Ltd.*)

Polar Response

The majority of moving-coil microphones are pressure operated, giving a nearly spherical polar diagram, except in the higher frequencies. However, one or two designs have appeared of microphones having a diaphragm exposed to the sound field on both sides. Such a microphone has a polar diagram of the figure-of-eight form at the low frequencies. Because of the particular way in which they are used, very close to the source of sound, they are sometimes known as noise cancelling microphones, the pressure difference on the two sides of the diaphragm being much greater for near sounds than for distant ones, see "Frequency Response of Ribbon Microphones". The moving-coil pressure microphone may be combined with a velocity ribbon microphone to produce a cardioid polar diagram.

Applications

In general, moving-coil microphones produce little distortion.

The moving-coil microphone has many applications because it needs no polarizing voltage or current and its output is of low impedance, permitting the use of quite long lines between the microphone and the transformer—it is very widely used in the public-address field, and the best types are also used for broadcasting and television.

THE RIBBON MICROPHONE

In contradistinction to most other types, the ribbon microphone is nearly always made to operate from the pressure gradient or "velocity" component of the sound wave.

The ribbon is another form of moving-conductor microphone. In this case the moving conductor takes the form of a narrow strip or "ribbon" of aluminium foil which has been rolled down as thin as possible and then beaten out in the same manner as gold until it is so thin that it is translucent. This ribbon is mounted vertically between the poles of a powerful magnet. If the poles are narrow, the sound is able to reach both sides of the diaphragm, even at the highest frequencies, thus causing it to be moved by the "velocity" component of the sound wave.

As in the case of the moving-coil microphone, the electrical output is proportional to the rate of cutting lines of force.

Sensitivity

Since the ribbon is virtually a coil with only half a turn, the electrical output impedance is much lower than that of the moving-coil microphone; so low in fact that a transformer is built into the microphone case to increase the impedance to 600 ohms (in some cases 50 ohms, so that the ribbon may be interchanged with a moving-coil microphone). The output from the built-in transformer is about 0·2 mV/dyne/cm.^{2}. A second transformer is used to match the output of the ribbon transformer to the grid circuit of the first valve. This second transformer will increase the output voltage by about 25 db. It is interesting to note that the acoustic input impedance may be as low as 200 rayls, the lowest impedance of any microphone, but still not a good match for the 42 rayls of the air.

Frequency Response

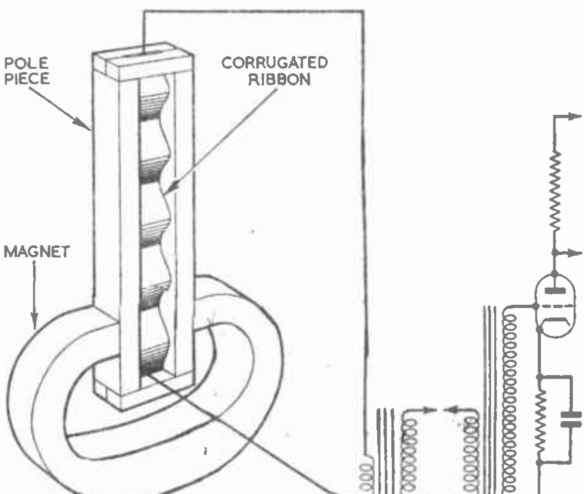
Theoretical considerations of the sound pressure produced on the two sides of the ribbon by a free plane wave of constant sound pressure show that the effective force available to move the ribbon is proportional to frequency over a very wide frequency range; that is, it is proportional to pressure gradient of a free plane wave. In order to make the velocity of the ribbon directly proportional to particle velocity it is necessary to make the ribbon behave as a mass.

The frequency response of a ribbon microphone may be made exceptionally good up to about 12,000 c/s. Beyond this frequency, secondary effects make it difficult to maintain the output. The

FIG. 8 (*left*).—GENERAL VIEW OF A RIBBON MICROPHONE.

(*The General Electric Co. Ltd.*)

FIG. 9 (*below*).—SCHEMATIC ARRANGEMENT OF A RIBBON MICROPHONE IN CIRCUIT.



low-frequency end is usually maintained by "corrugating" the ribbon to lower its frequency of resonance.

Since the ribbon is very light and has a very low restoring force, it is easily displaced from its normal position in the middle of the air gap by wind of quite low velocity. Never blow on to a microphone to ascertain if it is "alive"—it may easily displace the ribbon. Several layers of fine gauze are used to protect the ribbon from the wind, but even so it is more subject to wind noise than other forms of microphone, which are a worse match to the impedance of the air.

If a ribbon (velocity) microphone has a flat frequency response in a free plane sound wave, i.e., one whose wave front is a flat plane, it cannot have a flat frequency response in a free spherical sound wave when it is brought close to the source. The expression for the particle velocity in a free spherical sound wave is

$$V = \frac{P_0 r_0}{r \rho c} \sqrt{1 + \left(\frac{c}{2\pi f r}\right)^2}.$$

where V = particle velocity in cm./sec. from the origin;
P_0 = sound pressure in the wave in dynes/cm. at a reference distance r_0 cm.;
r_0 = reference distance in cm. from the geometric centre of the source;
r = distance from the origin of the spherical wave in cm.;
ρ = density of the air, in gm./c.c.;
c = speed of sound, in cm./sec.;
f = frequency.

It will be seen that for a "velocity" microphone the low-frequency response will be increased if the microphone is brought near to the source of sound. This increase in low-frequency response is particularly noticeable on speech. In other respects the frequency response quite justifies the extensive use of this type of microphone for broadcasting.

Polar Response

If the design is good, an excellent figure-of-eight polar response can be obtained up to the highest frequencies. It is possible to obtain a cardioid response by closing the back of half the ribbon with an acoustic resistance so that it becomes pressure operated, giving a circular response from one half which, when combined with the figure-of-eight of the other half, produces a cardioid. The greatest difficulty is to make a small compact acoustic resistance at all frequencies.

Non-linear Distortion

Under normal conditions the ribbon microphone has exceptionally low distortion, but if the ribbon gets "blown" out of the linear magnetic field, or touches the pole pieces very lightly, the distortion may increase considerably. There is also the danger of distortion being introduced into the electrical output due to the iron core of the microphone transformer becoming magnetized by the permanent magnet and distorting the wave form.

Special Applications

The ribbon microphone is extensively used in broadcasting, and for all cases where a "velocity" microphone is required. It is sometimes combined with some type of pressure microphone to obtain a cardioid polar response. It may be used with a series of lines or pipes to produce a highly directional microphone.

PIEZO-ELECTRIC OR CRYSTAL MICROPHONE

This class of microphone makes use of the piezo-electric properties of Rochelle salt, ammonium dihydrogen phosphate and lithium sulphate crystals. Two general types of microphone are manufactured: in one case the sound pressure directly actuates the crystal, giving a good frequency response and relatively low output, in the other case the crystal is driven by a diaphragm which is actuated by the sound pressure. The advantages of the crystal microphone are: fairly simple construction, light in weight and not subject to interference from stray magnetic fields.

Mode of Operation

It is first necessary to study the transducing element, which consists of a slab of material cut from a single crystal which has three axes, X, Y and Z, mutually at right angles in space. If the X-axis is perpendicular to the flat face of the thin slab, the slab is said to be "X" cut. If a thin rectangular X-cut slab of Rochelle salt has foil electrodes cemented to the two flat faces and a voltage is applied between the foils, the rectangle will be distorted into a parallelogram; that is, one of the diagonals will shorten and thicken and the other will lengthen and become thinner. It would be possible to make a microphone by attaching a diaphragm to the end of a bar cut from the diagonal of an X-cut plate, but unfortunately the mechanical impedance of all available crystals is extremely high. In order to reduce this mechanical impedance, a construction similar to a bimetallic strip, such as is used in thermometers and relays, is employed. It consists of two diagonal cuts from X-cut slabs, the cuts being at right angles to each other, so that when subjected to electrical stress, one gets longer and thinner, and the other gets shorter and thicker. These two diagonal cuts are cemented together with suitable electrodes to form a bi-crystal strip. If a voltage is applied to this composite strip, one side will get longer and the other side will get shorter, thus causing the device to bend. The mechanical impedance of the "bender" is very much less than that of the direct compression rod, and a diaphragm attached to the tip of the "bender" can be made into a satisfactory microphone. Strips can also be arranged so that the resultant movement is one of twisting rather than bending.

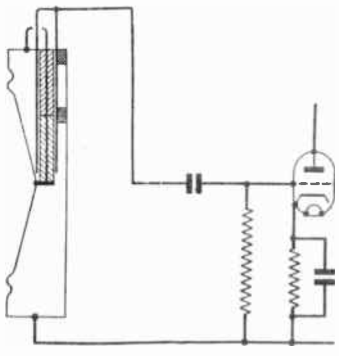


FIG. 10.—"DIAPHRAGM"-TYPE CRYSTAL MICROPHONE IN CIRCUIT.

Two or more pairs of these bi-crystal plates can be mounted in such a way that they are caused to bend by the action of the sound pressure. A microphone made in such a way is called a "sound cell" to differentiate it from the type having a diaphragm. The sound-cell microphone has a lower output but a better frequency response than the diaphragm type.

Sensitivity

The sensitivity of the "sound cell" type of construction is -65 db relative to 1 volt/dyne/cm.^{2}.

The sensitivity of the diaphragm type is -55 db relative to 1 volt/dyne/cm.^{2}. The electrical output impedance is roughly equivalent to a capacitance of 0·002 μF (0·8 MΩ at 100 c/s).

Frequency Response

The frequency response is good from about 30 up to 10,000 c/s for the sound-cell microphone; it is not quite so good, and does not extend beyond about 7,000 c/s in the case of the diaphragm type, and not beyond about 3,000 c/s for the simplest forms.

Polar Response

Both types of microphone are pressure operated with the result that a circular polar response is obtained for the low frequencies. Since the microphone can be made very thin and small, the pressure rise due to reflection will occur at a higher frequency than on most other types of microphone.

Non-linear Distortion

Rochelle-salt-crystal microphones exhibit small hysteresis effects in relation between the applied force and the resulting voltage. It arises in the electrical capacitance. and is very small if the microphone operates into a high impedance.

Special Applications

Due to its light weight, small size and ease of construction, the crystal microphone is very largely used in deaf-aid sets. Its principal disadvantage, at least for Rochelle salt, is its instability at high temperatures; it is permanently damaged at temperatures above 110° F.

The sound-cell type of construction has been used with success for the broadcasting of concerts where a spherical polar response is required.

SPECIAL TYPES OF MICROPHONE

From time to time special microphones have been developed for some particular purpose. Amongst such types are:

Highly directional microphones can be made by means of distributed arrays of sound-sensitive elements. Reflectors and horns can also be used in conjunction with a pressure-operated microphone, as can a multi-tube array. Multi-tube arrays with phase delays can be used with a velocity microphone to obtain high directivity.

A "moving-iron" microphone has been used successfully for short-distance communication in which the voice power was all that was employed. For this reason, they are sometimes described as "voice-power microphones".

F. H. B.

GRAMOPHONE PICK-UPS

Gramophone pick-ups may be classified roughly into two groups: (1) general-purpose, and (2) high-fidelity types. Usually, the latter generate an output voltage proportional to velocity, whilst the former give an "equalized output", i.e., the output from the pick-up can be fed directly into an amplifier with a linear frequency response characteristic without electrical equalizing being necessary. These general-purpose pick-ups are usually of the crystal type.

Crystal Pick-ups

Until recently, the crystal material was usually Rochelle Salt. This material has the advantages of high dielectric constant, extremely high coupling coefficient and being relatively easy to manufacture. It has the drawbacks of a large variation of permittivity with temperature; is deliquescent and must therefore be protected from extremes of humidity; and it has an upper working temperature of 45°–50° C.

Also now used in pick-up manufacture is an electro-strictive material in the form of poly-crystalline barium titanate. It has the advantages of high dielectric constant practically unvariable up to 70° C.; it is not humidity conscious, but because it can be produced only in "bender"-type bimorphs the frequency range of the pick-up is restricted; also, the sensitivity of the pick-up and the compliance are usually considerably lower than equivalent Rochelle Salt units.

Typical crystal pick-ups are shown in Figs. 11 and 12.

Output of a Crystal Pick-up

A piezo-electric crystal behaves as a generator of constant voltage in series with a capacitance (the electrical self-capacitance of the crystal). The voltage generated is proportional only to the force applied to the driving point of the crystal.

It is usually assumed that the groove walls of a record are infinitely stiff and that they therefore behave as the equivalent of a constant current (velocity) generator (they depart only very slightly from this concept, provided that the groove walls do not collapse due to excessive pressure). In practice, this must be slightly modified by the addition

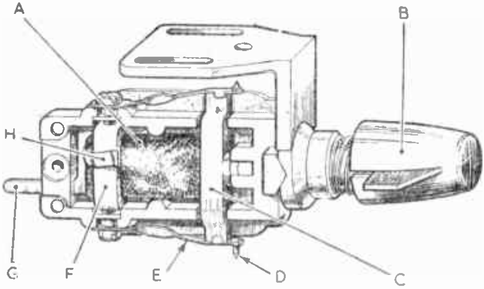
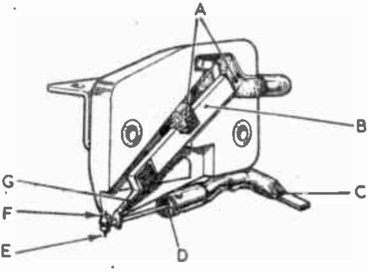


FIG. 11.—SCHEMATIC VIEW OF A CRYSTAL PICK-UP.

A, crystal bimorph; B, turnover knob; C, plastic driving member; D, stylus; E, stylus arm; F, plastic "clamping block"; G, contact pin; H, foil electrode from crystal to contact pin.

FIG. 12.—SCHEMATIC VIEW OF A CERAMIC CRYSTAL CARTRIDGE.

A, plastic clamping and supporting blocks; B, ceramic crystal bimorph; C, styli turn-under lever; D, rubber styli support and damper; E, standard stylus; F, L.P. stylus; G, reed.



of a parallel capacitance representing the compliance between the stylus point and the record groove.

If the mechanical termination of the stylus is purely resistive, the force generated across this termination will be independent of frequency and proportional only to the velocity of the stylus. On the other hand, should the termination be capacitative (that is, a mechanical compliance) the force will decrease by 6 dB per octave. This is analogous to a constant current through a capacitor. Finally, if the termination is that of a mass (inductance in electrical terms) the force will increase at 6 dB per octave. The effect of these terminations on a normalized fine-groove recording characteristic is shown in Fig. 13.

The output from a piezo-electric crystal is proportional to the force applied across it. The equivalent circuit in Fig. 14 shows that the back clamp of the crystal consists of a compliance in series with a resistance.

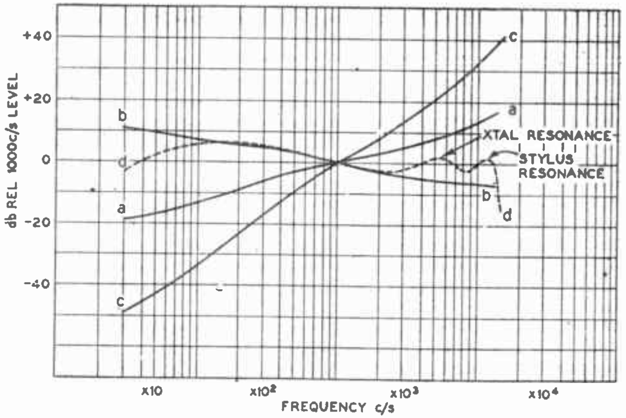


FIG. 13.—EFFECT OF CRYSTAL TERMINATION ON OUTPUT FROM FINE GROOVE RECORDINGS.

a, Resistive termination; b, compliance termination; c, mass termination; d, practical pick-up.

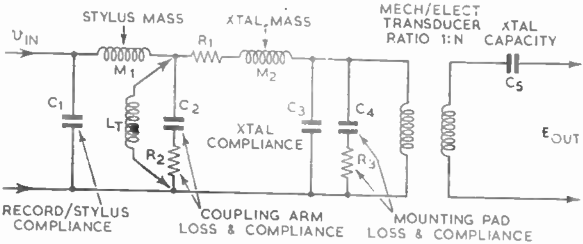


FIG. 14.—SIMPLE ANALOGUE FOR CRYSTAL PICK-UP CARTRIDGE.

R_1, R_2 and R_3 added to reduce mechanical resonance.

Resistive termination: R_1, R_2 and R_3 predominate.

Compliant termination: R_1 M_1 and M_2 short-circuited. C_2, C_3, C_4, R_2 and R open-circuited.

Mass termination; R_1 short-circuited. C_1, C_2 C_3, R_2 and R_4 open-circuited.

Note: L_T is tone arm mass and resonances. Total compliance to the L.F. resonance ≃ 30 c/s.

These are physically realizable from a plasticized co-polymer of P.V.C. in which, generally, the loss component is provided by the plasticizer and the reactive component by the resin. Thus, by suitably compounding the clamp, the time constant can be such as to provide low-frequency equalization. The driving member (or coupling arm) is fashioned in a similar manner, and again, by correct dimensions and ratio of plasticizer and resin, high-frequency equalization can be achieved. Clearly the equalization is never completely accurate, but with a well-designed and manufactured pick-up the variations should not be greater than the tolerance found in recording characteristics on current records.

In general, crystal pick-ups give an output voltage of between approximately 0·1 and 1 volt for low-frequency compliance of approximately 4·5–0·5 × 10^{-6} cm./dyne. They are reasonably flat and smooth to about 7 or 8 kc/s, and the output then attenuates at up to 18 dB per octave. Part of their high frequency response is due to deliberate introduction of resonance in the cantilever stylus system, generally between 9 and 12 kc/s. In some turn-over types of cartridge a reduction of output between 3 and 5 kc/s is apparent. This is due to the resonance of the free stylus absorbing energy from the driving circuit. Whilst this was a noticeable feature of early turn-over cartridges, its effect has been reduced considerably, so as to be almost negligible, by correct proportioning of the driving member. This type of cartridge is used practically to the exclusion of all others in popularly priced record reproducers and radio-gramophones, and the frequency response and distortion characteristics are more than adequate for the type of loudspeaker normally fitted in such equipment.

Stereophonic Types

The arrangements shown in Fig. 15 (*a*) and (*b*) form the basis of most current stereophonic crystal cartridges. In both designs separate crystals are used to provide the two outputs.

The principles of stereophonic disc recording are outlined in Section

A—clamps.
B—stylus arm pivot.
C—stylus arm.
D—stylus.
E—crystal A torsion arm.
F—crystal B torsion arm.

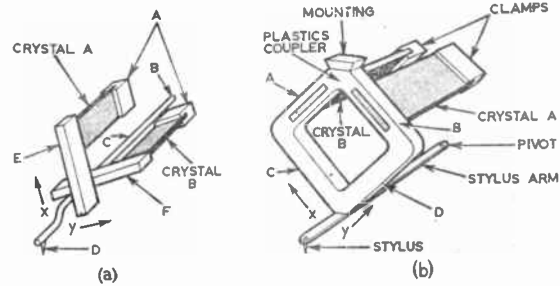


FIG. 15.—CRYSTAL STEREO PICK-UPS.

37. Both the designs shown in Fig. 15 are for use with recordings made on the standard " 45/45 " system.

To consider, first, the arrangement shown in Fig. 15 (*a*): Movement of the stylus (*D*) and stylus arm (*C*) in direction *x* will move tension arm *F*, thus bending crystal *B*, which is clamped at its farther end, thus causing an electrical output from this crystal. This movement will not affect tension arm *E* and crystal *A*. Movement of the stylus in direction *y* will, similarly, result in an electrical output from crystal *A* but not from *B*. Thus the complex movements of the stylus in directions *x* and *y*, as it follows the complex-cut track of the stereophonic disc, will provide a dual-channel output from the two crystals.

In the design shown in Fig. 15 (*b*) the two crystals are arranged in a " diamond structure resolver ". Movement of the stylus and stylus arm in direction *x* (parallel to sections *C* and *B* of the coupler) will cause movement of sections *A*, *C* and *D* of the coupler; thus crystal *B* (which is clamped at its other end) will produce an electrical output. During this movement, section *B* of the coupler remains stationary, so that crystal *A* is not affected. Movement of the stylus in direction *y* will, in the same way, provide an output from crystal *A*, section *A* of the coupler then remaining stationary.

The treble response of most stereophonic crystal cartridges extends to only about 8 kc/s, and the dynamic mass referred to the tip is of the order of 13–14 mg.

Because of the small dimensions of a stylus for stereophonic discs, the playing weight of a stereophonic cartridge must be kept low, though it must be sufficient to keep the stylus in the track, which will at times be much shallower than the standard monophonic microgroove track. To keep record wear to reasonable limits, and delay the development of flats on the stylus, under these conditions, a diamond stylus is advisable.

Constant-velocity Pick-ups

The constant-velocity cartridge is usually based on one of the following magnetic generating systems: (*a*) moving conductor; (*b*) vibrating reed supported at the remote end; (*c*) vibrating reed supported at the front end (this is now obsolete); (*d*) vibrating reed, usually known as the variable-reluctance type; (*e*) balanced armature; (*f*) moving magnet.

The basic equation for any magnetic generating system is

$$E = N \times \frac{d\theta}{dt}$$

where θ is the magnetic flux and N the number of turns. Therefore, in essence, all that is required to obtain the output voltage is to vary the flux linkage with pick-up coil in a predetermined manner. In practice, there are many difficulties in attempting to achieve this.

Moving-coil Pick-ups

The moving-coil generator is magnetically and mechanically the simplest. There are two generally available forms. One, the "long coil", was developed by Voigt in the early 1930s, and is shown schematically in Fig. 16. On well-designed units, the effective mass can

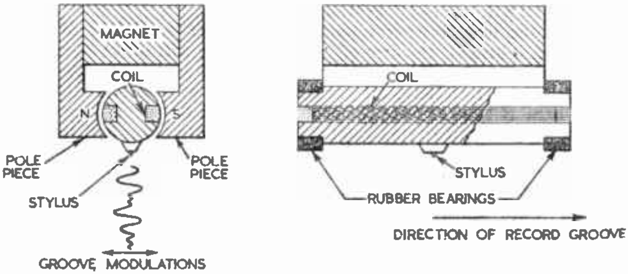


FIG. 16.—THE "LONG COIL" MOVING-COIL PICK-UP.

be reduced to 2 mg.; the low-frequency compliance will be of the order of 5×10^{-6} cm./dyne; and, if correctly assembled and balanced, the resonant frequency will be in excess of 20 kc/s. It suffers from the disadvantage that the stylus is rigidly coupled to the coil, with the result that a considerable mass must be vertically accelerated due to "pinch effect" at high frequencies. This can result in the generation of "even" harmonic distortion and considerable amounts of "needle chatter".

The other form is schematically shown in Fig. 17. In this the coil is mounted vertically and is driven from a short cantilever stylus. Because there are no clearances to worry about between the bottom of the coil and the record, the magnetic system can be increased in size, with the result that the number of turns, and hence the dynamic mass of the coil system, can be considerably reduced. Thus the added mass due to the cantilever is offset. In the best examples of this type, needle talk is practically absent, the upper resonant frequency is about 22 kc/s and the compliance is of the order of 5×10^{-6} cm./dyne. A low-impedance coil must be used with this type to keep the number of turns, and therefore the mass of the winding, to a minimum. It is almost always used with a matching transformer, with a turns ratio between 30 and 100 : 1. The output at the secondary terminals is usually about 3–5 mV/cm./sec. recorded velocity.

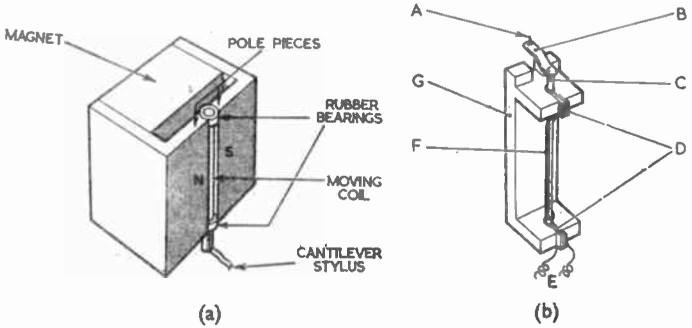


FIG. 17.—CANTILEVER MOVING-COIL PICK-UP.

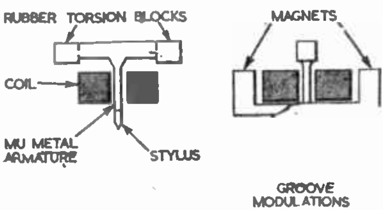
(a) Schematic diagram; (b) practical arrangement with magnet removed: A, stylus; B, barium copper cantilever 0·100 in. long; C, ferrous rod; D, rubber bearings; E, leads, F, coil: G, polystyrene bracket.

Moving-armature Pick-ups

A schematic diagram of a moving-armature pick-up is shown in Fig. 18. In one well-known design using this principle, introduced shortly after the Second World War the frequency response was true velocity to about 8 kc/s, rising by about +6 to +10 dB at the resonant frequency, which varied between 9 and 14 kc/s. Because the pick-up coil was stationary, it was quite massive and outputs of the order of 0·1 volt were common. The disadvantage was that the effective vertical mass of the stylus referred to its point was four times that of the lateral mass; the vertical stiffness was very many times that of the lateral stiffness, which varied between 1 and $2{\cdot}5 \times 10^{-6}$ cm./dyne, and resulted in considerable amounts of needle chatter being generated.

Developments of this type have produced tiny micro-armatures, in which the resonant frequency has been brought to an upper limit of 22 kc/s. Whilst the vertical compliance has been considerably increased by using a back bearing of nylon or some other plastic material, the needle talk is, however, always greater (because of the unavoidably high vertical to lateral dynamic mass ratio) than with a cantilever type of construction with the same frequency response.

FIG. 18.—SCHEMATIC DIAGRAM OF A MOVING-ARMATURE PICK-UP.



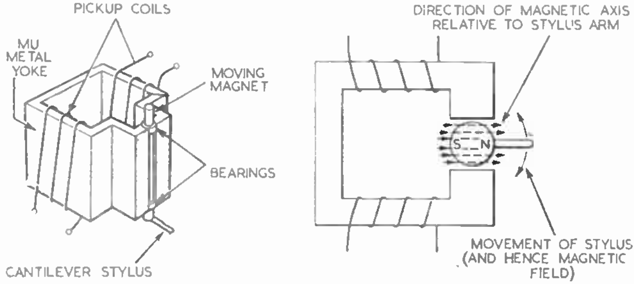


FIG. 19.—SCHEMATIC VIEW OF MOVING-MAGNET PICK-UP.

Moving-magnet Pick-ups

The moving-magnet pick-up is the inverse of the moving coil, having a stationary coil with a moving magnet. A schematic diagram of this system is shown in Fig. 19. In one version, the magnet consists of a tiny rod of ferrite material, magnetized across its diameter, and driven with a cantilever stylus arm. The resonant frequency of this unit before damping is of the order of 15 kc/s, the response being within 2 dB up to that limiting frequency, after which it "dies" at about 18 dB per octave. Sensitivity is approximately 4 mV/cm./sec., giving an output of 24 mV r.m.s. for a peak velocity of 6 cm./sec. The upper bearing is rubber, and the lower one P.V.C., this arrangement giving a lateral compliance of 5×10^{-6} cm./dyne.

Variable Reluctance Pick-ups

This is one of the most popular magnetic cartridges. Whilst all magnetic cartridges other than the moving-coil type are by definition variable reluctance, the term has come to be associated particularly with the system shown schematically in Fig. 20. The magnetic circuit consists of a magnet, from which the flux is carried by a cantilever stylus arm symmetrically disposed between the pole pieces of two pick-up coils. The remote end of the pole pieces are magnetically connected to the other end of the magnet. Displacing the stylus from its position

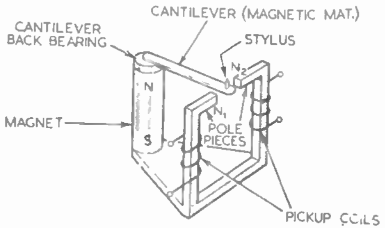


FIG. 20.—SCHEMATIC DIAGRAM OF A VARIABLE RELUCTANCE CARTRIDGE.

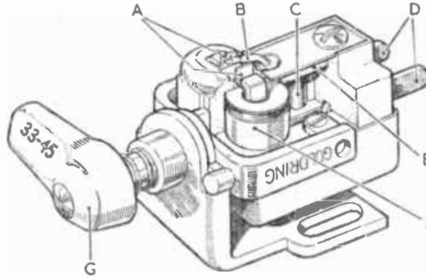


FIG. 21.—"TURNOVER" TYPE VARIABLE RELUCTANCE CARTRIDGE.

A, Mumetal pole pieces; B, ferrous stylus arm; C, magnet; D, contact pins; E, rubber stylus arm clamping block; F, coil (other one is in moulding); G, turnover knob.

of rest increases the flux in one core and reduces the flux in the other. The coil windings are so phased that the induced e.m.f.s are series aiding. In current types the stylus arm is usually less than $\frac{3}{16}$ in. long, and can be made almost mechanically aperiodic by the judicious use of damping material, either on the stylus pedestal or along the arm. The effective stylus mass is usually of the order of 2–4 mg.; the lateral compliance 5×10^{-6} cm./dyne; and the resonant frequency would be, if the damping material were removed, about 17–23 kc/s.

The form of design makes it very suitable for the production of "turnover" or "turn-around" cartridges.

It is unfortunate that with the variable-reluctance magnetic cartridge the magnetic efficiency is not very great, with the result that several thousand turns of wire must be accommodated on each pick-up coil. It is impossible to obtain exact magnetic symmetry, so that the unit is susceptible to hum-pick-up. It is therefore usual to fit a Mumetal magnetic shield completely enclosing the cartridge.

Stereophonic Cartridges

Both the variable-reluctance and moving-coil systems have been successfully adapted to the requirements of a stereophonic pick-up.

Fig. 22 shows a variable-reluctance arrangement for a stereophonic pick-up. The permanent magnet is not shown, but occupies a similar position to that in the monophonic arrangement shown in Fig. 20. Movement of the stylus and armature in the direction $x'x''$ induces a voltage across coils 1 and 3 (in series), the output being from terminals A and C. As, during this movement, the armature remains parallel to pole pieces L and O, no voltage will be induced across coils 2 and 4. Similarly, movement of the armature in direction $y'y''$ will induce a voltage across coils 2 and 4, with output from terminals A, B, there being no effect on coils 1 and 3.

In moving-coil arrangements the two moving coils are of different physical dimensions, and are placed one within the other, at 90° to each other, in the gap of a suitably designed magnet. Arranged thus, the composite movement of the assembly consisting of armature, stylus and two coils provides at the terminals of the two coils the resolved outputs for the two channels.

The mass referred to the stylus tip for a well-designed magnetic

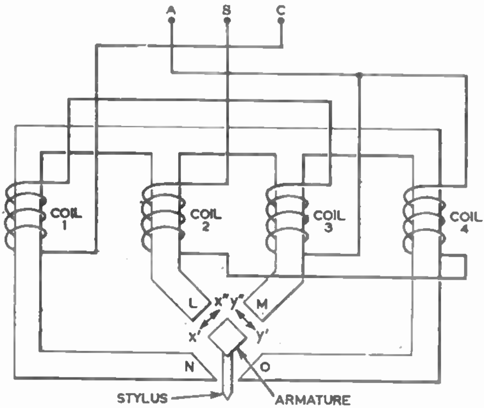


FIG. 22.—ONE ARRANGEMENT OF VARIABLE-RELUCTANCE STEREO CARTRIDGE.

stereophonic cartridge can be 0·5–2 mg. The vertical and lateral masses should be equal, and the compliance of the order of 5×10^{-6} cm./dyne.

Some degree of cross-modulation between the two channels is almost certain to occur in the pick-up, as it is difficult to prevent one transducer from having some response to the other channel. Cross-modulation should not be serious, provided that rejection of the unwanted channel is of the order of 25 dB at 1 kc/s.

32. LOUDSPEAKERS

32. LOUDSPEAKERS

TYPES OF LOUDSPEAKERS

Electromagnetic Types

These include the reed-armature and balanced-armature units, which both depend for their action on the mechanical forces resulting from the magnetic variation due to a fluctuating field and a constant polarizing field (Fig. 1).

The reed- and balanced-armature types usually employ either a simple bar or horsehose type of magnet, the driving force being dependent upon the strength of the magnetic fields.

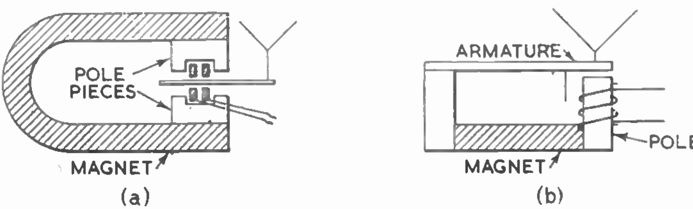


FIG. 1.—ELECTROMAGNETIC LOUDSPEAKERS.

Condenser Type

In the condenser system mechanical forces are set up due to electrostatic action.

The system can either be unilateral (Fig. 2) or bilateral. In the bilateral type the moving plate is situated between two fixed plates.

A polarizing field is required for this type of loudspeaker.

Crystal Type

The crystal type (Fig. 3) is principally used for extremely high frequencies. A crystal driving system is one in which the mechanical forces are set up by the deformation of the crystals which have converse piezoelectric characteristics.

Electro-dynamic

The dynamic or moving-coil type of loudspeaker is one in which mechanical forces are developed as a result of the interaction of two magnetic fields, one the field due to a permanent magnet and the other due to a current-carrying coil (Fig. 4). When a current is passed through the coil, it moves axially either into or out of the magnet, depending on the direction of the current. The force, in dynes, due to the interaction of the field resulting from the current in the moving coil and the magnet's field is

$$F_m = Bli$$

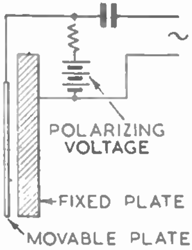


FIG. 2.—CONDENSER TYPE LOUDSPEAKER.

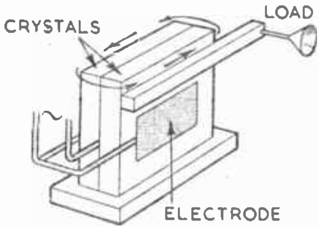
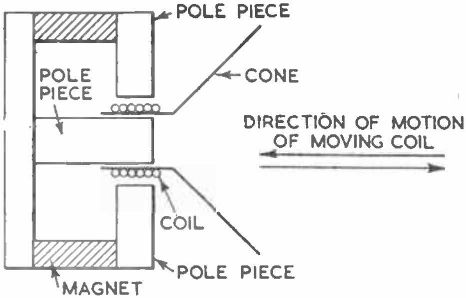


FIG. 3.—CRYSTAL LOUDSPEAKER.

where F_m = forces in dynes;
B = flux density in gauss;
l = length of conductor in centimetres; and
i = current in absolute amperes.

The electromotive force in absolute volts developed by the motion is $e = Blv$, where v = velocity, in cm./second, of the coil.

FIG. 4.—ELECTRO-DYNAMIC OR MOVING-COIL LOUDSPEAKER.



The electro-dynamic type is the loudspeaker generally used today, and this will be dealt with in detail.

From the foregoing, it can be seen that the driving force, and hence the acoustic output, is proportional to the flux density. It will be appreciated, therefore, that with the increase in efficiency of magnet steels it is possible to manufacture loudspeakers which are much more efficient with regard to performance, and also to obtain a given performance with smaller and lighter magnets.

MAGNETS

Economical Working Limit

There is a limit to the flux density utilized in the new magnet steels above which the magnet becomes uneconomical. From the graphs of Fig. 5 it will be seen that the economical working point of Alnico is 10,000 gauss, and for Alcomax is 12,000–14,000 gauss, when considering a magnet having a 1-in. pole. It is necessary from an

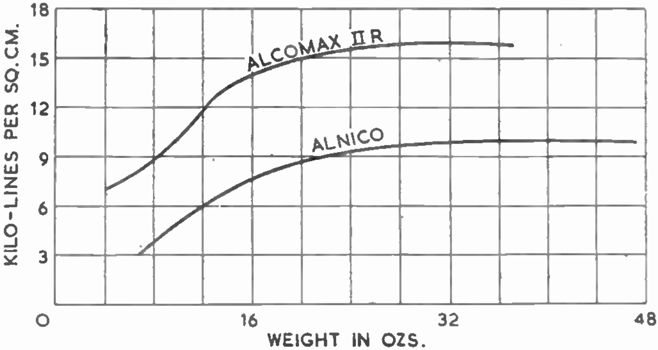


Fig. 5.—Flux Densities of Typical Magnetic Materials.

economical point of view to take into consideration the cross-section, and hence the weight of soft iron required. Whilst the curve for Alnico commences to flatten out before 10,000 gauss, it is possible to obtain up to 10,000 gauss without unduly increasing the iron section, thus making it economical to work at this flux density.

Types of Magnets

Various forms of magnets used on moving-coil loudspeakers are illustrated in Fig. 6 : the shaded portion is the actual magnet steel, and the unshaded parts are soft iron. The claw type of magnet, Fig. 6 (*a*),

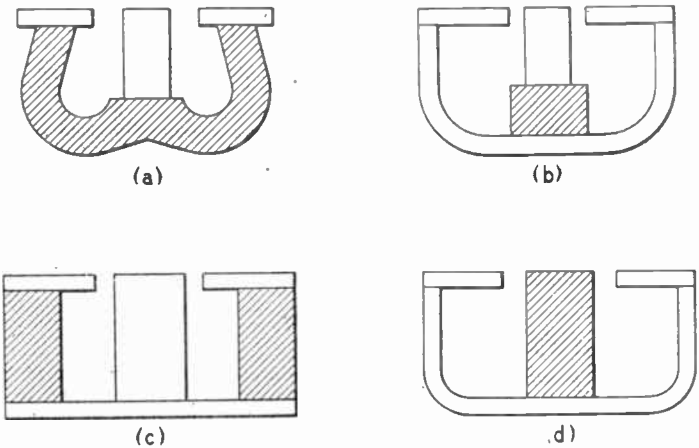


Fig. 6.—Types of Magnets.

was used extensively between the years 1930 and 1933, and represented one of the earliest commercial forms of permanent magnets for use on moving-coil loudspeakers. The slug type of magnet (Fig. 6 (*b*)) was developed during the year 1933, and was used at that time on loudspeakers of the smaller type. The ring type of magnet shown in Fig. 6 (*c*) is one which is very extensively used today, being used on almost every moving-coil loudspeaker, irrespective of size. This shape is very convenient for manufacture.

The type of magnet shown in Fig. 6 (*d*) is a recent development, and is known as the centre-pole type. At present the flux density which is obtained on this type of magnet is limited, but the external field is almost negligible. This factor, coupled with the use of a non-magnetic chassis, makes it ideal for television receivers, where the external field can cause distortion of the picture.

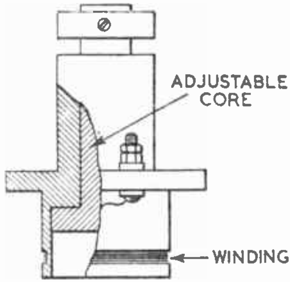


FIG. 7.—SEARCH COIL USED TO TEST THE FLUX DENSITY.

Heat Treatment

The full properties of magnet materials are realized only after correct heat treatment. The treatment necessary may range from a single heating and quenching to a complicated process requiring three different heatings, each followed by a controlled cooling.

The heat-treatment magnetizing of the Alcomax type of alloy is carried out either by means of large electro-magnets or permanent magnets, the field strength necessary being of the order of 1,500 oersteds.

Magnetizing

Magnetization is invariably carried out electrically, and in its simplest form is done by placing the magnet in, or along, a solenoid, which is then energized. A modification of this method is by the use of a solenoid having a soft-iron core. A very heavy current is passed through the coil, and so produces a large magnetizing force.

Another method is to use a magnetizing transformer which has the secondary in the form of a single turn.

Testing Methods

Measurements of total flux or flux density are usually made by a search coil (Fig. 7) in conjunction with a flux meter. The search coil has a known number of turns, and is placed in the magnetic gap and

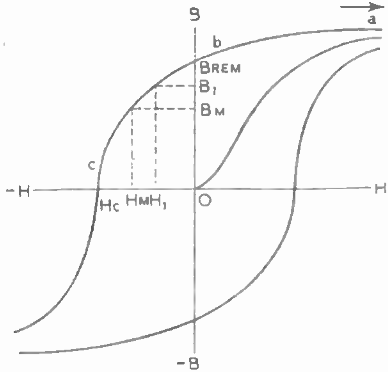


FIG. 8.—HYSTERESIS LOOP.

removed from a region of high flux to one of negligible flux and the change of flux linkage noted.

$$\text{Flux Density} = \frac{\text{Maxwell turns} \times \text{Deflection on meter}}{\text{Area cut by search coil in sq. cm.}}$$

Interpretation of Test Figures (Fig. 8)

The portion between B_{rem} and H_c is the only part of the full loop in which the magnet manufacturer and user is interested. A flux density in a magnet equal to B_{rem} could be obtained only if the magnetic circuit external to the magnet had zero reluctance. If there is a gap, flux lines will appear at the face and across the gap. This flux obviously requires a capacitance of some picafarads to maintain it, and this is taken from the magnet itself. The working point is no longer B_{rem}, but has dropped along the curve to some point B_1 (Fig. 8), where the corresponding value H_1 is that required to maintain the gap flux. (The result is as if there were a force tending to demagnetize the system.)

The energy required to maintain the flux in the gap is proportional to the area of the rectangle $B_1 \times H_1$. Innumerable such rectangles may be drawn inside the demagnetizing curve, and it will be seen that at some point between points B_{rem} and H_c the area is a maximum.

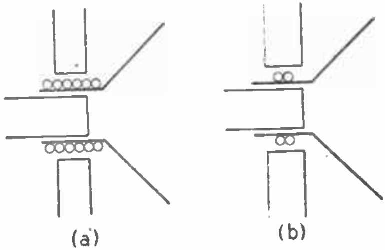
If the rectangle of maximum area has co-ordinated B_m and H_m, then it will be obvious that maximum energy will be obtained from the magnet if the gap length is such that B is equal to B_m, and H to H_m. This area B_mH_m is called the $BH_{max.}$ of the material, and is the true criterion of the performance of a magnet alloy.

Design Considerations Relating to the Magnet

Effect of Non-uniformity of the Magnetic Field

For a moving-coil loudspeaker to radiate even a small amount of power, the amplitude of movement of the coil will be appreciable. The excursion may be large enough, particularly at low frequencies, for the coil to move a considerable distance out of the gap, and so into the

FIG. 9.—METHODS OF REDUCING DISTORTION.



leakage field of the magnet, the flux density of which rapidly decreases as the distance from the gap increases. It has been stated that the force $F = Bli$, and therefore if the value of B varies in the space traversed by the moving coil, the driving force will not always be proportional to the current in the coil, and hence distortion will result.

In order to minimize this effect, one of two methods is generally used. The first is to make the speech-coil longer than the flux face of the magnet, thus maintaining a constant inter-linkage of the flux and the moving-coil turns (Fig. 9 (*a*)), irrespective of the position of the coil. This method has the disadvantage of making the speech-coil heavier, but it is inexpensive to apply.

An alternative is to make the magnet so that the length of the flux face is much greater than the length of the moving-coil, thus providing a constant flux throughout the amplitude of the coil excursion (Fig. 9 (*b*)). This method, which is very efficient, is of course costly, as to maintain the required flux density in the increased gap requires a larger amount of magnet steel. It is, however, becoming an economical proposition due to the more efficient magnet steels now available.

Relation of Acoustic Output to the Magnet

Consider two magnets, both having a flux density of 10,000 gauss, but one with a gap length of 0·050 in. and the other a gap length of 0·025 in. It will be obvious that, if the same number of turns are to be accommodated on the moving-coil, wire of smaller diameter will have to be used in the case of the smaller gap, and this in turn results in a higher coil resistance. The losses in the latter case would be greater, with a corresponding loss in acoustic output.

Again, considering two magnets, both having a flux density of 10,000 gauss (and the same air-gap) but one with a centre pole of 1·5 in. diameter and the other with a centre pole of 1 in. diameter, it will be appreciated that in the former case a larger coil, with a greater handling capacity and more total driving force, can be accommodated.

It follows, therefore, that flux density (lines of force per square centimetre) is not a real indication of the loudspeaker's potential performance, and the factor to be considered as a measure of this must also include some reference to the area and volume of the air-gap.

The term which more nearly gives an indication of the expected performance is the total flux of the magnet, which is proportional to the flux density multiplied by the mean area of the air-gap.

Table 1 gives details of various types of loudspeakers, and shows the size of the magnet usually associated with them.

TABLE 1.—TYPICAL RANGE OF LOUDSPEAKERS

Type	*Flux Density (gauss)*	*Pole Diameter (in.)*	*Gap Length (in.)*	*Flux Face (in.)*	*Total Flux*	*Speech-coil Impedance (ohms)*	*Handling Capacity (watts)*
18·14	14,000	2·5	0·063	0·312	227,000	15	30
12·135	13,500	1·5	0·050	0·25	106,000	15	15
10·12	12,000	1	0·043	0·187	47,400	3 or 15	10
9·10	10,000	1	0·043	0·187	39,500	3 or 15	6
8·10	10,000	1	0·043	0·187	39,500	3 or 15	5
8·08	8,000	1	0·043	0·187	31,600	3 or 15	5
6·10	10,000	0·75	0·040	0·125	20,000	3	3·0
6·07	7,000	0·75	0·040	0·125	14,000	3	3·0
5·135	13,500	0·75	0·040	0·125	27,000	3	3·0
5·07	7,000	0·75	0·040	0·125	14,000	3	2·5
3·57	7,000	0·625	0·035	0·125	11,500	3	2·0
2·57	7,000	0·375	0·033	0·093	5,285	3	0·3

Types 12·135 and 5·135 are good examples as to why total flux, and not flux density, should be considered as the indication of a loudspeaker's performance. Both loudspeakers have magnets with a flux density of 13,500 gauss, but the total flux of the 12·135, which has a 1·5-in. diameter centre pole, 0·050-in. gap and 0·25-in. flux face, is 106,000 lines, compared with the 5·135, which has a centre pole of 0·75 in. diameter, 0·040 in. gap and 0·125 in. flux face, with a total flux of 27,000 lines.

The Effect of Flux Density upon Performance

Fig. 10 shows the steady-state response curves of a loudspeaker taken with two different magnets, having flux densities of (*a*) 8,000, and (*b*) 12,000 gauss respectively, and the lower part of the illustration shows the impedance characteristics. It will be noted from the impedance curve that the frequency of the bass resonance increases with increase of flux density, moving from 88 c/s with a magnet of 8,000 gauss up to 96 c/s with a magnet of 12,000 gauss and, from the response curves, it will be noted that there is an average increase of 8 db in output. This increase, however, is not constant at all frequencies.

It will be appreciated, therefore, that when designing a loudspeaker to give a certain characteristic, increasing the flux density for the obvious reason of increasing sensitivity, the frequency response will not necessarily remain the same, but will probably be slightly altered, necessitating a modification in the design of the cone and coil. It follows that to obtain the added advantages of higher flux-density magnets, it is necessary to treat the magnet and the cone-and-coil assembly as a composite unit.

Cone and Coil Construction

It is necessary that both the cone and coil should be able to handle the stated input without distortion and break-up, also that they shall reproduce the frequencies evenly over as wide a frequency range as possible. To do this the cone must be of a material which will allow full amplitude

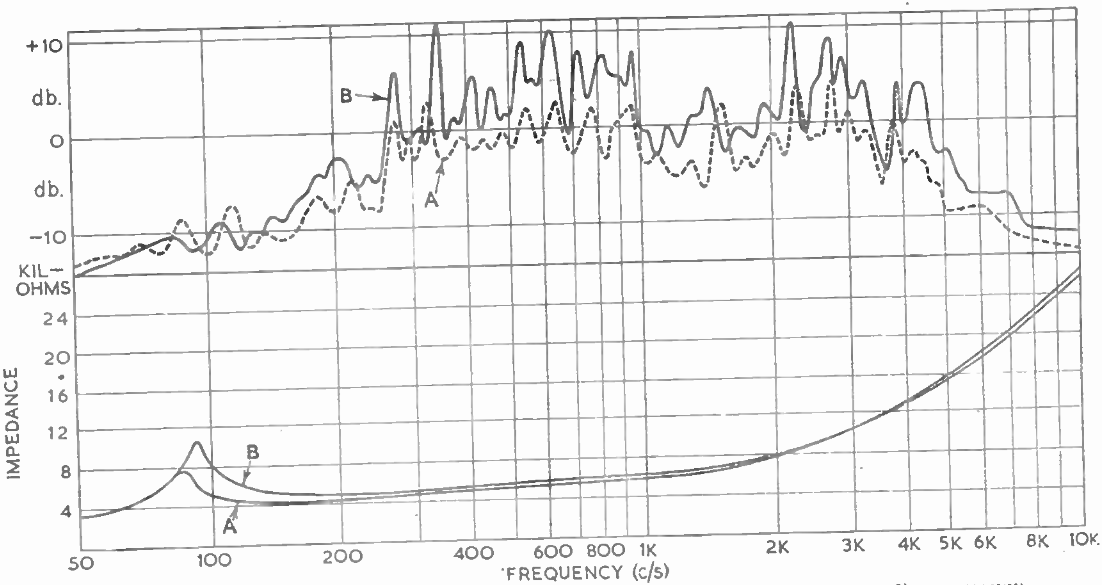


Fig. 10.—Loudspeaker Response Curves showing Effect of Flux Density upon Performance.

at low frequencies; it is necessary, therefore, that the periphery of the cone shall be of a soft and more flexible material than the cone itself. The speech coil should be wound on a very light former which will not introduce losses, and also the former must be very strong to prevent any collapse during movement. This is extremely important for the reproduction of transients.

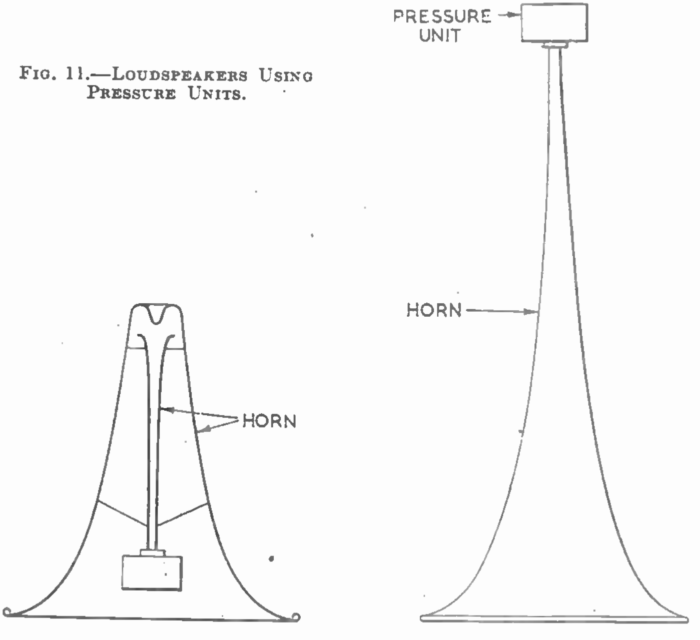
To obtain this performance, the cone is manufactured from small pulp fibres which are allowed to settle on to a basket and thus take up a natural position and not be under any stress or strain. The surround of the cone is usually made thinner by allowing only a small quantity of pulp fibres to settle in this position during manufacture. After the cones have been settled into a normal shape, they are pressed into final shape, then treated to give a desired performance and also to make them suitable for withstanding certain climatic conditions.

The centre-piece of a loudspeaker has to be made of a material which will allow freedom of movement but will not tire over long periods. On the majority of commercial loudspeakers either a cambric is used, which is blanked to a certain shape, or a flexible material is used which is formed not only to be a centre-piece, but also to act as a dust excluder.

Fabric-based Cones

One type of cone, which is now being used to give an improved response, consists of a loosely woven linen or cotton cambric upon which

Fig. 11.—Loudspeakers Using Pressure Units.



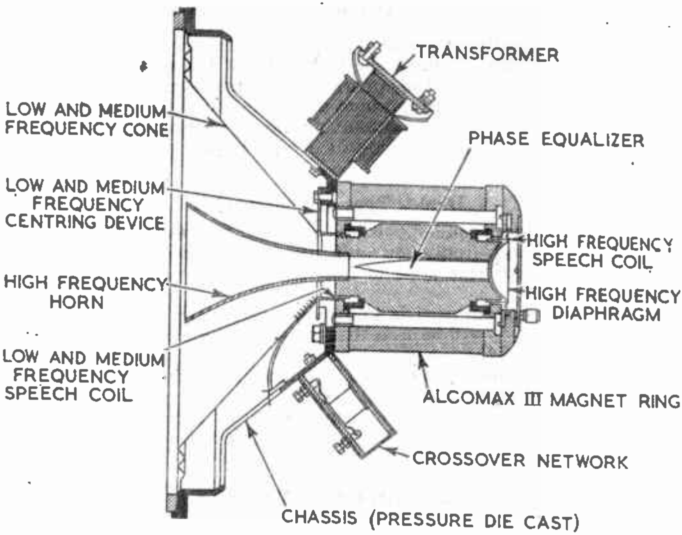


FIG. 12.—DUPLEX TYPE OF LOUDSPEAKER.
This type is usually used in conjunction with a half-section series crossover.

is deposited a relatively compact fibrous material, with the exception of that part of the diaphragm which forms the peripheral corrugations. The corrugations are thus left very flexible, enabling a very low resonant frequency to be obtained, and also a greater uniformity from diaphragm to diaphragm to be secured in manufacture.

PRESSURE UNITS

These units have two principal uses :

(1) for public-address work;
(2) for use as a high-frequency speaker.

The driving mechanism is essentially the same in both cases, i.e., electro-dynamic. An aluminium-alloy diaphragm and an aluminium speech-coil are generally used to reduce the mass.

For public-address work, where low frequencies are required to be reproduced as well as the high, the horn loading must be increased: the lower the frequency required, the longer the horn. This necessitates horns being several feet in length; where they can be permanently fixed no trouble is encountered, but when required for mobile use or where length must be kept short, the re-entrant horn is used, Fig. 11. This consists of folding the horn back within itself; with just one fold the overall length can be halved.

When used simply as a high-frequency speaker (to increase the frequency range of a system), the construction is slightly different. The diaphragm is usually mounted at the rear of the magnet, and the centre pole is hollowed out to form the beginning of the horn, Fig. 12. The phase equalizer is inserted to ensure that all the frequencies coming from the remote parts of the diaphragm arrive at the throat in phase. The horn is completed by non-resonant moulding screwed into the front of the magnet pole, which, as the unit is only for high frequency use, is a few inches long.

Duplex Type of Loudspeaker

The duplex loudspeaker consists of a low-frequency cone unit with a horn-loaded high-frequency unit mounted concentrically within it. It is possible with this type of unit to obtain a very wide frequency response.

The magnet system used is known as a series-gap type, and from the diagram (Fig. 12) it will be seen that the centre pole is hollow to form the commencement of the pressure horn.

Phase equalization of the higher frequencies is carried out by means of the phase equalizer.

It is usual on this type of loudspeaker to fit a crossover network to separate the frequencies which are to be applied to each driving system.

CROSSOVER NETWORKS

Crossover networks are used when two (or more) loudspeakers are to be connected to obtain a wider frequency response. A large loudspeaker is generally used for the lower frequencies, and a pressure unit or small speaker for the higher frequencies.

The crossover is used to separate the units so that inter-modulation distortion is reduced where the loudspeaker responses overlap, also to keep the extreme low frequencies from entering the pressure unit and damaging the diaphragm assembly.

Fig. 13 shows the various types employed. The quarter-section types are the cheapest to manufacture, but have an attenuation of only 6 db/octave away from the crossover frequency.

10 db/octave is obtainable from the networks shown in Fig. 13 (*c*) and (*d*).

From the filter type, half-section (Fig. 13 (*e*) and (*f*)) an attenuation of 12 db/octave is obtainable; and although this type does not provide as constant a resistance as Fig. 13 (*c*) and (*d*), this is offset by the greater attenuation, this characteristic being preferable as the inter-modulation distortion is further decreased.

TABLE 2.—CHOKE AND CAPACITOR VALUES FOR CROSSOVER NETWORKS

Choke		*Capacitor*	
L1	2·4 mH	C1	10·6 μF
L2	1·5 mH	C2	17 μF
L3	1·7 mH	C3	15 μF
L4	3·4 mH	C4	7·5 μF
L5	3·83 mH	C5	6·6 μF

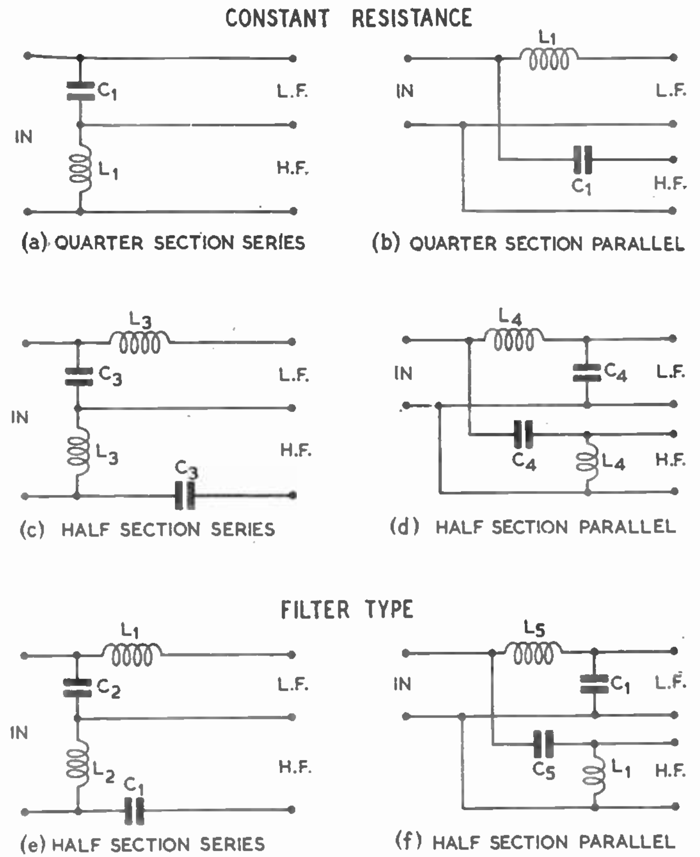


Fig. 13.—Crossover Network Circuits.

The figures in this table are for a crossover at 1,000 c/s with 15-ohm loudspeakers.

The crossover frequency is inversely proportional to the component values; thus we may say :

(1) to double the crossover frequency, halve the values of L and C;
(2) to halve the crossover frequency, double the values of L and C.

When halving the loudspeaker impedance, multiply all C values by two, divide all L values by two and so on.

To keep the insertion loss low, normal electrolytic capacitors should not be used, as they require a D.C. polarizing voltage and will cause a high loss due to the high resistive component. Air-cored coils having low resistance should be used in preference to iron-cored ones.

MATCHING

Line Matching

It is often necessary for a number of loudspeakers to be run from one source and for each loudspeaker to consume only a limited amount of power.

To do this the amplifier used has an output transformer having a relatively low output impedance, say 200 ohms. Assuming the output of the amplifier to be 50 watts, it would be possible to run 100 loudspeakers from it, each loudspeaker to consume 0·5 watts. As the loudspeakers would be connected in parallel, the individual impedance of each one would have to be 200 × 100 = 20,000 ohms.

It is, of course, possible to arrange the loudspeakers to consume varying amounts of power thus,

5 loudspeakers consuming 5 watts
10 loudspeakers consuming 2 watts
5 loudspeakers consuming 1 watt

The relative impedance can then be calculated as follows:

Amplifier output	50 watts
Impedance	200 ohms
Therefore line voltage . . .	100 volts

Therefore:

(1) for 5-watt loudspeaker, $W = \frac{E^2}{Z}$

$$\text{Impedance} = \frac{100^2}{5} = 2{,}000 \text{ ohms}$$

(2) for 2-watt loudspeaker

$$\text{Impedance} = \frac{100^2}{2} = 5{,}000 \text{ ohms}$$

(3) for 1-watt loudspeaker

$$\text{Impedance} = \frac{100^2}{1} = 10{,}000 \text{ ohms}$$

Matching to an Output Stage

To determine the correct transformer ratio to match the loudspeaker to the output stage.

$$\text{Transformer ratio} = \sqrt{\frac{\text{Optimum load impedance}}{\text{Loudspeaker impedance}}}$$

PERFORMANCE TESTS

Performance tests can be divided into "Subjective" and "Objective" tests.

A subjective test is the actual listening to or comparing of loudspeakers by an individual or individuals. Objective tests are ones carried out by means of instruments, the results being such that they can be recorded and used as a means of reference for the specifying and comparison of performance.

Typical objective tests are :

(1) frequency response;
(2) impedance;
(3) polar response;
(4) distortion;
(5) efficiency;
(6) power-handling capacity;
(7) transient response.

The *frequency response* of a loudspeaker is usually taken either outdoors, in such a position as to be away from all reflecting surfaces, or in a soundproof or "dead" room.

The loudspeaker is fed with a constant voltage from an oscillator covering the frequency band over which it is desired to make the test. The output from the loudspeaker is picked up by a microphone, which is connected to an amplifier. The output of the amplifier actuates some form of recording meter.

As the oscillator frequency is varied, the output of the amplifier is noted and a frequency/output characteristic plotted.

Automatic methods are now used, making it possible to view a frequency-response curve on a cathode-ray-tube screen. Such devices usually employ a beat-frequency oscillator with automatic frequency sweep and a logarithmic amplifier which will give a linear decibel scale.

The *impedance* varies with frequency, and is usually quoted at one particular frequency, or a curve of the loudspeaker impedance characteristic is given.

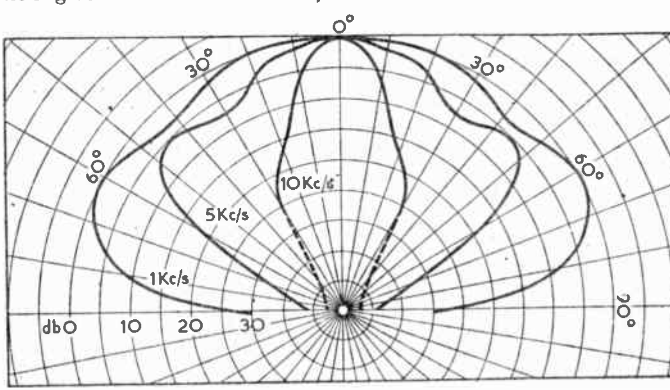


FIG. 14.—TYPICAL LOUDSPEAKER POLAR RESPONSE.

The *polar response* of a loudspeaker is the response measured at varying degrees from the axis of the loudspeaker, and is used to determine the diffusion of the sound. A typical response is shown in Fig. 14.

R. T. La.

LOUDSPEAKER ENCLOSURES

A loudspeaker cone vibrating in free air produces at both its front and rear faces alternate compression and rarefaction in the adjacent air layers; these are propagated as sound waves. At low frequencies, however, where the wavelengths are comparable with the dimensions of the cone, the air compressed at any instant upon one face will flow into the rarefaction occurring simultaneously at the other. The total radiation of sound will thus be negligible. To overcome this, it is necessary to load the loudspeaker. There are a number of systems, each with certain advantages and disadvantages, used in current high-fidelity practice. Cabinet loading as used in domestic radio 14·6 ← practice is also discussed in Section 14. Symbols used in the figures of this section: C_b stiffness of enclosed air; C_c stiffness of cone suspension; f frequency; M_c mass of cone; M_v mass of air in aperture or tunnel; R_a radiation resistance; R_c resistance due to friction in cone suspension; R_d resistance due to electro-magnetic damping; R_v resistance of air in aperture or tunnel.

14-10

Baffles

The most elementary way of separating the front and rear radiation from the loudspeaker cone is to mount the loudspeaker on a flat baffle, which, ideally, should have infinite area. The nearest practical approach to this is to mount the loudspeaker in a wall, e.g., the partition between two rooms. In doing this, precautions have to be taken to ensure that the effective "tube" formed by the wall thickness and loudspeaker aperture does not introduce irregularities in the frequency response. For example, the edges of the aperture may be bevelled, or the aperture may be made considerably larger than the loudspeaker and the latter mounted on a wooden sub-baffle of about 1-in. thickness.

Preferably, the loudspeaker should not be mounted concentrically on the baffle: a circular baffle with the loudspeaker mounted concentrically will have a much poorer response than a rectangular baffle with the loudspeaker mounted off-centre.

In the case of finite flat baffles, back-to-front cancellation occurs at some low frequency depending on the size of the baffle. The minimum length (l) of the shortest side of a rectangular baffle for a given low-frequency extension (f) is:

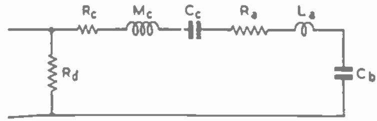
$$l = \frac{C}{2f} \text{ in.}$$

C being the velocity of sound in air (1,120 ft./sec. at 15° C.). In practice there is no point in making f much lower than the cone resonant frequency.

A cone loudspeaker acting as a treble unit should have a baffle large enough to work down to half the cross-over frequency.

As flat baffles generally apply a relatively small amount of acoustic control over the cone movement, loudspeakers having a high degree of magnetic damping and a low resonant frequency suit this type of mounting best.

FIG. 15.—ANALOGOUS CIRCUIT OF A LOUDSPEAKER IN A COMPLETELY CLOSED CABINET.



Flat baffles have the advantage of being free from acoustic resonances, and do not restrict the efficiency of the loudspeaker. The disadvantage lies in the size required for a reasonable low-frequency extension.

Closed Cabinets

An alternative method of preventing front/rear cancellations is to enclose completely the rear of the loudspeaker cone in a cabinet.

The main disadvantage of this system is that the enclosed air behaves as an " elastic cushion ", which will act on the cone in a similar manner, and in addition to, the stiffness of the cone suspension. The effective resonant frequency of the cone is therefore raised, reducing the useful low-frequency response of the loudspeaker. Two approaches have, with a certain degree of success, been tried to overcome this limitation. The first has been to incorporate devices to damp the resonances of the loudspeaker, cabinet and enclosed air. This extends the frequency response to some extent, although tending to make it uneven. The second approach is to develop loudspeakers specially adapted to this form of loading. This can be done by providing the cone with a very free suspension, producing a cone of greater rigidity and providing a generously designed magnet system. There will still be a certain low-frequency restriction, although this approach does provide a smooth response.

The analogous circuit of a loudspeaker in a closed cabinet is shown in Fig. 15.

Vented and Reflex Cabinets

A common type of enclosure is vented by means of an aperture in its wall. Any enclosed or partially enclosed air volume behaves as an elastic cushion. The air within the immediate vicinity of the aperture of a vented cabinet, being capable of flow into and out of the enclosure, may be regarded as a homogeneous mass. If this air mass is disturbed in any way it will oscillate in and out of the enclosure, " bouncing " on the " elastic cushion " of air within the enclosure. This oscillation will take place at one frequency only, determined by the volume of the enclosed air and the dimensions of the aperture. Such an acoustic system is known as the Helmholtz resonator, after the German physicist. At the resonant frequency of the enclosure a maximum impedance will be applied to the rear of the loudspeaker cone, reducing its velocity. The radiation from the aperture will be added to that of the cone, and, provided that the area of the aperture is sufficient, a considerable increase in radiation efficiency may be had at around this frequency.

It is advantageous to arrange for the resonant frequency of the en-

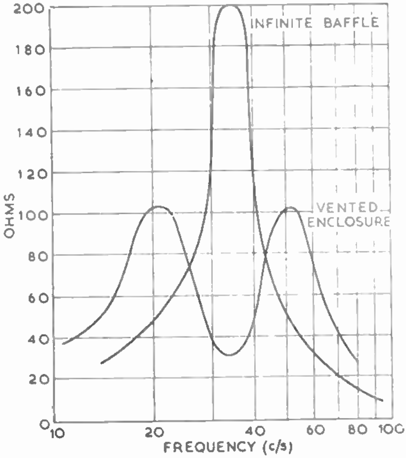


FIG. 16.—IMPEDANCE, FREQUENCY CHARACTERISTICS OF A LOUDSPEAKER MOUNTED ON AN INFINITE BAFFLE COMPARED WITH MOUNTING IN A VENTED ENCLOSURE.

closure to coincide with that of the cone so that the latter receives maximum damping at its resonance.

Below the resonant frequency of the enclosure, the enclosed air has a very high stiffness reactance and will behave as a rigid connecting-rod between the cone and the air mass at the aperture, this air mass thus being effectively added to the cone will reduce its effective resonant frequency. Above the resonant frequency of the enclosure, the air mass in the port becomes too inert to move and the enclosure behaves as if it were completely closed. This has the effect of raising the effective resonant frequency of the cone. Thus two auxiliary resonances occur, which produce points of maximum velocity either side of the enclosure resonance. This is clearly shown in the impedance curve shown in Fig. 16. The upper resonance is a disadvantage, and the lower resonance is of no assistance in extending the range, since the radiation from the aperture is out of phase with that from the cone at this frequency.

In order to match low cone resonances, reflex enclosures are rather large, although their size is usually reduced by extending the aperture inwards by means of a duct or tunnel, thereby increasing the air mass and consequently making possible a smaller enclosure for a given resonant frequency.

In the design of reflex enclosures it is desirable to have the area of the aperture equal to that of the piston area of the cone. Then, if the enclosure resonance is to be equal to that of the cone, the volume of the enclosure is given by

$$V = \pi r^2 \left[\frac{C^2}{4\pi^2 f^2} \times \frac{1}{1{\cdot}7r + l} + l \right]$$

where C is the velocity of sound in air, f is the resonant frequency, r is the piston radius of the cone and l is the length of the tunnel.

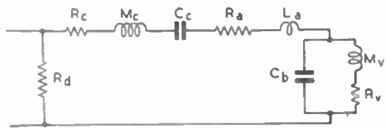
Since the volume of this tunnel must be added to that of the enclosure, a point is reached beyond which a further increase in tunnel length will produce an increase in enclosure volume. The tunnel length for minimum enclosure volume is

$$l = \frac{C}{\omega} - 1{\cdot}7r$$

where $\omega = 2\pi f$.

A further limitation on the length of the tunnel is that it must not exceed one-twelfth of a wavelength at the resonant frequency.

FIG. 17.—ANALOGOUS CIRCUIT OF A LOUDSPEAKER IN A VENTED ENCLOSURE.



The analogous electrical circuit of a vented enclosure is shown in Fig. 17. If the impedance characteristic of this is plotted, it will be found to correspond to that shown in Fig. 16. The impedance of the parallel section, rising to a maximum at resonance, corresponds to that of the enclosure.

Tuned Pipe

The principle of the tuned pipe is based on the principle of the organ pipe, which is that if one end of a pipe is excited at a given frequency and the pipe is exactly a quarter-wavelength at that frequency and open at the far end, the reflected wave will be in phase at the source end of the pipe. If the pipe were half a wavelength long, the reflected wave would tend to cancel that produced by the sound source. If the pipe is closed at the far end, the opposite to the foregoing is true.

One method of applying these properties to a loudspeaker is to use an open pipe with the loudspeaker mounted at one end, the length of the pipe being such that its fundamental anti-resonance coincides with that of the cone resonance, thus procuring some of the advantages of the reflex enclosure.

The length of an open pipe for a given frequency of anti-resonance is

$$= \frac{C}{2f} - 1{\cdot}7\sqrt{\frac{A}{\pi}}$$

where A is the cross-sectional area of the pipe and where all units are of the same denomination.

Such pipes are rather more simple to construct than reflex enclosures. The great disadvantage is the presence of all anti-resonances and resonances occurring at every quarter-wavelength below the fundamental. A way of partially overcoming this is found in the Voigt corner horn, where the loudspeaker is mounted one-third of the pipe length away from the closed end so that the first resonance above the fundamental, that is the third harmonic, is cancelled.

Acoustic Labyrinth

The labyrinth consists of a very long tube or pipe, usually folded and heavily lined with absorbing material, with the loudspeaker mounted at one end. The purpose of a labyrinth is to absorb completely the radiations from the rear of the loudspeaker cone. Its effectiveness is limited at low frequencies by its length, the minimum length being set empirically at a quarter-wavelength, equivalent to the lowest frequency to be produced.

The load applied to the rear of the loudspeaker by the labyrinth is high, and almost entirely resistive. The cone is thus under conditions approaching constant velocity. There will therefore be a fall in low-frequency response, which may be corrected by bass compensation in the amplifier. The advantage of labyrinth loading is that it provides complete freedom from resonances.

The Horn

By far the most efficient method of loading a loudspeaker is by mounting it at the throat of a suitably designed horn. The horn may be considered as an acoustic transformer, matching the relatively high mechanical impedance of the loudspeaker cone to the radiation resistance of the air. We have seen previously how, in the case of direct radiator loudspeakers, the radiation resistance is very much smaller than the mechanical impedance of the loudspeaker system, hence the low efficiency of the direct-radiator type. Horn loading overcomes this to a certain extent, and efficiencies as high as 50 per cent (approximately ten times higher than direct radiator types) have been attained.

The disadvantage of the horn lies in the size required to maintain this efficiency down to low frequencies. A horn capable of presenting a constant-radiation efficiency down to 30 c/s from the cone of a 12-in. loudspeaker would be over 12 ft. long, and have a mouth of about 9½ ft. The greater efficiency, however, means that smaller drive units can be used, thus to some extent reducing the horn length required. although there is still a sharp cut-off. Six and eight-inch units are commonly used to provide the drive in horn loudspeaker systems. It is usual to use the drive unit as a direct radiator for the higher frequencies, or to provide a short high-frequency horn, and to take the lower frequencies from the back of the cone via a rear folded horn.

The greatest efficiency to be obtained in the smallest space is from an exponential horn where the cross-sectional area (A_x) at any distance (x) from the throat is given by

$$A_x = A_0 e^{mx}$$

where A_0 is the throat area, m the flare constant and e base of natural logarithms.

The cut off frequency (f_c) is given by

$$f_c = mC/4\pi$$

All units of the same denomination.

The diameter of the mouth of the horn should not be less than a quarter-wavelength at the cut-off frequency, otherwise standing waves will occur similar to those found in a tuned pipe.

Friction Loaded Enclosures

This type of enclosure consists of a loudspeaker cabinet of predetermined volume, and has a large aperture in one of its walls, a small part of which is left open to form a port, the greater part being covered by an acoustically resistive material. It has been developed to provide a smaller enclosure with good performance.

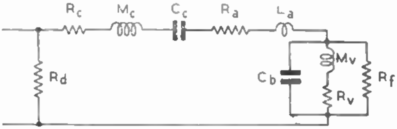


Fig. 18.—Analogous Circuit of a Loudspeaker in a Friction-loaded Enclosure. R_f is the Added Resistive Material.

The port provides additional mass loading of the cone, and therefore reduces the effective resonant frequency; but, since the area of the port is small, it will not have sufficient radiation to produce cancellation of this frequency (see "vented enclosures" earlier). This resonance will be fairly well damped.

The remaining resonances common to vented enclosures are reduced by means of the resistive material, which covers the larger part of the aperture. This is analogous to the resistance R_f shown in Fig. 18. It can readily be seen from this that this resistance will limit the magnitude of the enclosure resonance, and, by careful choice of component values, the upper resonance common to vented enclosures can be reduced to a negligible proportion.

In Fig. 19 the velocity characteristic of this system is compared with that of a loudspeaker in a vented enclosure and on an infinite baffle.

No attempt is made in this system to increase the acoustical impedance applied to the loudspeaker cone at its resonant frequency,

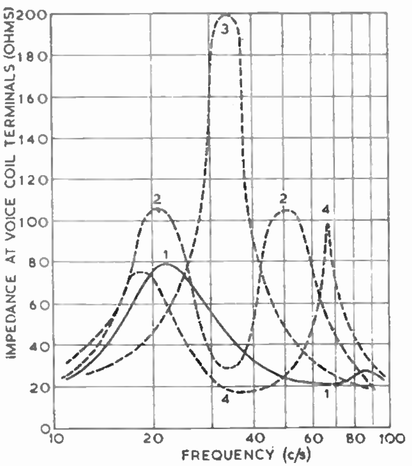


Fig. 19.—Comparative Frequency/Impedance Characteristics: (1) Friction Loaded Enclosure; (2) Reflex Cabinet; (3) Infinite Baffle; (4) as (1) but with R_f Open Circuit.

TABLE 3.—DENSITIES OF VARIOUS MATERIALS

Material	*Density (gm./c.c.)*	*Material*	*Density (gm./c.c.)*
Aluminium	2·7	Rubber	1·0
Brass	8·4	Sand	1·5
Copper	8·9		
Steel	7·8	Woods:	
Lead	11·3	ash	0·70
Nickel	8·7	beech	0·75
Tin	7·3	cork	0·25
Zinc	7·0	elm	0·57
Brick	1·8	fir	0·45
Granite	2·7	maple	0·67
Marble	2·7	mahogany	0·67
Slate	3·0	oak	0·80
Glass	2·6	pine	0·50
Ivory	1·8	poplar	0·50

since the loading is such that in practice the cone resonance does not occur. From Fig. 19 it can be deduced that this system loads the loudspeaker cone in such a manner as to provide a condition of mass control down to the lowest frequency limit of audibility, and to have an absence of resonances above this frequency.

General

Loudspeaker enclosures should be made of as rigid and dense a material as possible, in order to minimize vibration and resonances. To avoid standing-wave effects, it is usual to line all internal faces to a depth of at least 1 in. with soft, sound-absorbent material such as wool or felt.

The preferred position for the loudspeaker system is in the corner of a room, where the low-frequency efficiency may be as much as four times greater than that obtained if the system is adjacent to a plain wall. Although this may excite all the modes of resonance due to the room acoustics, this is more desirable than the excitation of a few sharply defined resonances, which may otherwise occur. Treble diffusion is also better from a corner position.

WIDE-RANGE ELECTROSTATIC LOUDSPEAKERS

The theoretical advantages of the electrostatic loudspeaker are its simplicity and the relatively direct conversion of electrical energy into sound with inherently better linearity. Early simple electrostatic loudspeakers had a number of shortcomings, notably inability to reproduce low frequencies and low sensitivity. Two comparatively recent developments have, however, led to the introduction of the wide-range electrostatic loudspeaker, examples of which are now commercially

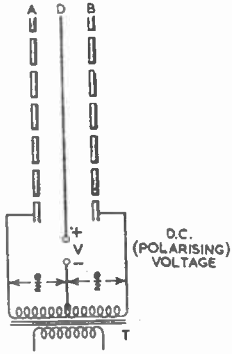


FIG. 20.—BASIC ARRANGEMENT OF A PUSH–PULL ELECTROSTATIC LOUDSPEAKER.

available. These are, first, the idea of the constant-charge, push–pull system, which is the basis of current designs, and, secondly, the appearance of plastics materials of high mechanical strength in extremely thin sheets. Considerable chemical and mechanical problems have attended the commercial development of wide-range electrostatic loudspeakers capable of giving reliable service in all conditions of humidity and temperature.

The basic arrangement of a push–pull electrostatic loudspeaker is shown in Fig. 20. D is the moving diaphragm, and A and B the two fixed plates, which are conveniently driven with equal and opposite signal (A.C. voltages from the transformer secondary). The fixed plates are perforated, or rendered acoustically transparent, for sound radiation. The main difficulty is as follows. If the rest position of the diaphragm is central, application of the polarizing voltages produces two equal and opposite forces which reduce its apparent stiffness, since any deflection is immediately assisted by the unbalancing of these forces. As long as this negative stiffness effect does not exceed the stretch stiffness of the diaphragm, it will not be disturbed. When a signal voltage is also applied, the diaphragm will tend to move towards the plate of "opposite" charge, as the two attractions are now different.

$$\text{Net force} \quad \propto \quad \left(\frac{V + \frac{e}{2}}{d_2}\right)^2 - \left(\frac{V - \frac{e}{2}}{d_1}\right)^2$$

$$\propto \frac{V^2}{d_2^2} - \frac{V^2}{d_1^2} + \frac{e^2}{4d_2^2} - \frac{e^2}{4d_1^2} + \frac{Ve}{d_2^2} + \frac{Ve}{d_1^2}$$

Now while the diaphragm is central, $d_1 = d_2 = d$, say the first four terms cancel, leaving only

$$\text{Force} \propto \frac{2Ve}{d^2} \propto e \text{ (since } V \text{ and } d \text{ are fixed).}$$

i.e., a linear law, which is what is needed. Thus the driving force exhibits non-linearity only when displacement occurs. If the diaphragm movements can be kept small there will be little distortion. Thus the reproduction at higher frequencies will be satisfactory. The low-frequency reproduction will, however, show bad distortion.

The Constant-charge Solution

When the diaphragm moves from its central position (see Fig. 21), the capacitances C_1 and C_2 of the two halves will no longer be equal. C_1, in Fig. 21, will have diminished and C_2 increased. The increase is bigger than the decrease, and the total capacitance $C_1 + C_2$ has therefore increased. If the voltage remains the same, the charge will have increased by a small flow of current on to the diaphragm (flow of electrons from it in Fig. 21), therefore $Q = V \times C$. If, now, the diaphragm is charged to V volts when in its central position, and then isolated electrically, its charge Q must remain the same wherever it moves between the two plates, provided it always remains parallel to them. When displaced from the centre, its capacitance will increase and its voltage or potential decrease. The two plates A and B are more or less parallel, and of dimensions large compared with their separation. The electric field inside them is therefore uniform, and the field strength equals $e/(d_1 + d_2)$. Though d_1 and d_2 vary, $d_1 + d_2$ does not. The force on a fixed charge in a uniform field is a very simple law, force equals charge multiplied by field strength, and this holds irrespective of its position within that field. This really is what is needed: a driving force proportional to the signal voltage.

In addition to this, other limitations of the electrostatic loudspeaker are overcome. It is no longer necessary to restrict the motion to a small fraction of the initial separation; neither is it necessary to restrict the signal voltage to a small fraction of the polarizing voltage.

The constant charge may be obtained by charging the diaphragm through a very high resistance. As long as the charge does not alter appreciably during the half-cycle deflection at the lowest frequency,

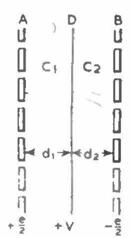


FIG. 21.—THE CONSTANT CHARGE PRINCIPLE.

$$\text{NET FORCE} \propto \left(\frac{V + \frac{e}{2}}{d_1}\right)^2 + \left(\frac{V - \frac{e}{2}}{d_2}\right)^2$$

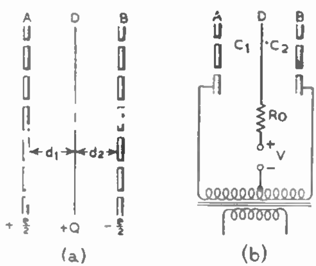


FIG. 22.—APPLICATION OF CONSTANT CHARGE PRINCIPLE.

(a) NET FORCE $\propto \frac{eQ}{d_1 + d_2}$

(b) TOTAL $C = C_1 + C_2$

$R \times C$ to be very large.

the constant charge requirement is satisfied. This means a long time-constant, extending to several seconds. The longer the better, provided that the leakage from the diaphragm does not cause an appreciable voltage drop across the resistance.

Not only must there be a very high resistance between the diaphragm and the polarizing circuit, the diaphragm itself must also be of very high resistance so that the charge cannot " walk about " on its surface. The diaphragm material, in addition to being as uniform as possible, must have a conductive coating of very high resistance.

Distortion

If the diaphragm is centred reasonably accurately, the time constant of the charging circuit is very high, and the elasticity of the diaphragm shows no peculiarities, the acoustical distortion of this type of loudspeaker will be of an extremely low order and the percentage of each harmonic less than that of the next lower order. Apart from this, there are no non-linearities up to the electrical breakdown of air. Also, because of the lightness of the diaphragm, it is not troubled by resonances.

Efficiency

The electrostatic loudspeaker tends to be inefficient mainly because it is predominantly capacitative, so that the current supplied to it is largely wattless. The " apparent efficiency " is the ratio of acoustic power output to a highly inflated volt-ampere input to the loudspeaker terminals. Considerable power is lost in the amplifier output impedance and in any resistances used with it to effect cross-over.

The apparent efficiency also depends on band-width. The low-frequency response is limited by diaphragm stiffness, the high-frequency response by the fact that the electrostatic loudspeaker capacitative reactance falls with rising frequency. Band-width may be traded for apparent efficiency, with the apparent efficiency " lowered " to that of present-day commercial moving-coil loudspeakers. A band-width, for level response, of between four and five octaves is possible. As high-fidelity sound reproduction embraces a span of between eight and nine octaves, a full-range electrostatic loudspeaker must be made up of at least two sections to have an acceptable efficiency. With three sections the apparent efficiency is comparable to that of the best moving-coil loudspeakers.

Radiation Pattern

When the width of a flat sound source is less than one wavelength the sound is emitted in a wide enough beam to cover a room adequately, with the loudspeaker in a corner position. At high frequencies the sound source can easily be several wavelengths wide, and the sound is concentrated into a narrow beam, giving very uneven distribution. If the width of diaphragm radiating sound is always the same proportion of a wavelength, i.e., active width shrinks progressively with rising frequency, then distribution will be independent of frequency. One practical arrangement to give even sound distribution is shown in

Fig. 23. It will be seen that the loudspeaker is constructed with a slight curve, and that it has several sections, the high-frequency sections being in the middle.

With regard to loading, the diaphragm is so light and tenuous that there is little need for the usual loading systems.

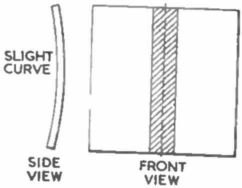


Fig. 23.—Practical Arrangement to Give Even Sound Distribution.

The Doublet

As the electrostatic loudspeaker offers a ready-made match to air, a loudspeaker without any cabinet or horn is possible. Quite obviously, such a loudspeaker, open back and front, ought to have a baffle at least, but why not make the whole assembly a loudspeaker. Thirty inches for the maximum dimension equals one wavelength at 440 c/s, so that for bass notes its dimensions are only a small fraction of a wavelength, and being open both sides it acts as an acoustic doublet. For constant output power the diaphragm amplitude will have to be proportional to the inverse square of the frequency.

The doublet has a number of very useful properties of considerable help in normal domestic surroundings. The sound distribution obeys a cosine law. The polar diagram is a figure-of-eight pattern. The output in the plane of the loudspeaker is nil, none up or down, so that the loudspeaker will excite room resonances only in one of the three directions. Also, anywhere in the "beam" the ratio of direct to indirect sound will be much higher than with most other types of radiator, so that there is far less room coloration. The other main advantage of the doublet type is the complete absence of cabinet, with associated resonances.

General

A polarizing voltage of the order of 5 kV is required; this can be produced by techniques similar to those used in television-receiver practice.

With regard to amplifier loading, a 30 in. × 20 in. push–pull electrostatic loudspeaker has a capacitance of about 0·00025 μF for 0·5 cm. spacing. If this is matched to 15 ohms at 1,000 c/s, it effectively puts 2·5 μF across the amplifier terminals. This need not be alarming, since many amplifiers become stable again at such high capacitances, and, furthermore, most electrostatic loudspeaker systems will use resistors somewhere or other in their circuits.

33. ELECTRICAL INTERFERENCE

33. ELECTRICAL INTERFERENCE

Electrical interference with reception must be carefully considered in almost all branches of radio and television engineering; for example, in domestic entertainment, interference levels play a large part in determining the effective service areas of broadcast and television transmitters; in land, sea and air mobile operation much attention has to be paid to the suppression of engines and auxiliary equipment; in point-to-point communications, receiving stations are usually located in the country to avoid the higher interference levels prevalent in urban areas.

Electrical apparatus which utilizes commutation (e.g., D.C. motors and generators), vibrating contact points (e.g., electric shavers), spark discharge (e.g., automobile ignition) or any mechanism whereby an electric spark, however minute, is produced will radiate damped radio-frequency waves, covering a wide frequency spectrum, unless preventive measures are taken. Certain other electrical appliances may also radiate radio-frequency energy. Radiations of either type may be rendered audible by a radio receiver in the form of continuous crackling, sporadic clicks, buzzing or humming noises, a number of which are characteristic of the source causing the trouble.

Some of the more common sources of electrical interference are: Industrial generators and electric motors, vacuum cleaners, hair dryers, sewing-machine motors, barbers' clippers, farm and dairy machinery, automatic switches and contactors, thermostats, overhead power lines, trolley-bus systems, defective house/factory wiring and switches, lifts, automobile ignition systems, R.F. heating and welding plant, X-ray, ultra-violet ray and diathermy plant, commercial, Service and amateur transmitters, broadcast and television receivers, scraping contacts between metallic objects at differing potentials, door-bells, fluorescent lighting, cash registers, electric typewriters, mercury-arc rectifiers and so on. A number of these, such as diathermy apparatus and automobile ignition systems, are only likely to cause disturbance on short-wave ranges. Corona discharge from high-voltage cables and static-charged wires may, in certain atmospheric conditions, give rise to a high background noise, particularly noticeable on S.W. Static-charged rain may also prove troublesome.

DIRECT RADIATION FROM THE SOURCE. Except for electro-medical apparatus and car interference on H.F. and V.H.F., radiation is likely to occur only over a comparatively short range, especially on the broadcast bands. Arcing between the collector and overhead wires of trolley buses or between the shoes and rails of electric trains may, however, cause interference by direct radiation to programmes transmitted on medium and long wavebands.

CONDUCTED INTERFERENCE. This is probably the most frequent path of transmission on M.W. and L.W. The electrical disturbances are conducted along the power-supply lines and injected into the receiver by way of the mains lead.

MAINS-RADIATED INTERFERENCE. Similar to conducted interference, except that in this case the power-supply leads form a kind of transmitter aerial, the radiations from which are picked up by the aerial or the internal wiring of the broadcast receiver. Both conducted and mains-radiated interference may be borne by the power lines over considerable distances, in extreme cases, affecting an entire district.

SECONDARY RADIATION. Similar to the above, except that two direct radiation links occur and the transmission line may take the form of telephone wiring, neighbouring aerials, gas and water pipes and so on.

Interference can be received in three ways: (1) pick-up by the aerial and lead-in wires; (2) injection into the receiver via the mains supply leads; and (3) by stray electric fields. Apart from cause (2), both mains-operated and battery-operated receivers can be equally affected.

BROADCAST RECEIVER INTERFERENCE

In order to reduce or eliminate interference, it is necessary:

(1) To ascertain the path by which the interference reaches the receiver.

(2) To identify, wherever possible, the source of interference and to ensure that adequate suppression is applied.

(3) Failing this, or where complete suppression at the source is not practicable, to fit interference filters at the receiver, or close to the receiver end of the path.

The manner in which the interference reaches the receiver can usually be ascertained by carrying out a few simple tests, though it should be noted that, in practice, a combination of two or more routes is not uncommon.

(1) Remove the aerial and earth leads from the receiver and short-circuit the appropriate sockets. Directly radiated and mains-radiated interference should then cease altogether or be greatly reduced, whereas mains-conducted interference may be hardly, if at all, affected.

(2) Test the relative strength of interference under no-signal conditions at the high- and low-frequency ends of each waveband. Directly radiated interference tends to increase as the wavelength decreases, and will usually be much increased in strength on the short wavebands. Disturbances from mains-conducted interference, on the other hand, will normally become stronger as the wavelength is increased and be most troublesome on the long waveband.

(3) If a battery-operated receiver be available, the effect of switching off the electric power at the main house switchboard should be noted. Should the interference continue it is unlikely that mains-borne radio-frequency signals are the cause. If the interference ceases, it would seem that interference is entering the premises via the mains supply wiring, or alternatively, that the source of interference is located within the premises concerned.

Identifying the Source of the Trouble

Often the interference source will be obvious or easily traced from the characteristic nature of the disturbance and from observation of the times at which it is experienced. The source of directly radiated interference is normally restricted to within a radius of a few hundred feet of the receiver.

More troublesome causes may be traced with the assistance of a portable battery receiver, preferably fitted with a frame aerial, so that rough bearings of the source of interference can be plotted. These bearings should always be taken by observing points of minimum signal strength which occur when the axis of the frame aerial is at 90° to the direction of the source; readings taken on maximum pick-up are considerably less accurate. Portable receivers which have no frame aerial can be used by simple observation of the strength of the interference, provided that it is possible to move freely within the search area.

Mains-borne and mains-radiated interference, which may originate at a considerable distance from the receiver, may prove much more difficult to trace; it will not be possible to pin-point the direction of the source until the search has been narrowed to the area within which directly radiated interference can be received. The effect of removing the fuses, one at a time, from each of the house circuits and of switching off the power at the main switchboard should be noted: this procedure will usually establish whether or not the interference is arising from defective switches, cable joints or apparatus within the building. Where mains-borne interference originates outside the premises, and the source is not obvious, it will usually be necessary to obtain assistance from the Engineering Branch of the G.P.O., who have developed specialized techniques for tracing elusive sources of interference.

Suppression at the Source

Interference should wherever possible be suppressed at the source, though in practice, suppression at the receiver is often necessary. The aim of most suppression devices is to confine any radio-frequency

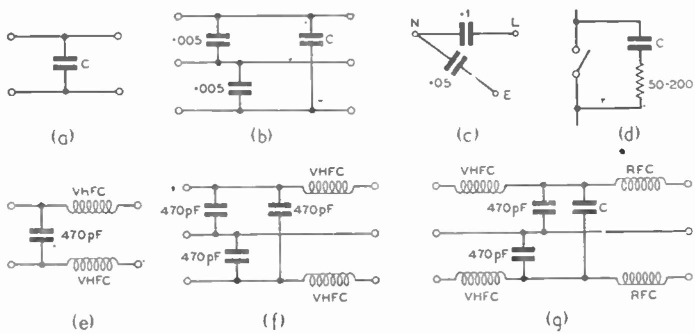


Fig. 1.—Basic Interference-suppression Devices.

Seven commonly used suppression circuits. (*a*) For two-core cable appliances; (*b*) for three-core cable appliances; (*c*) for three-pin socket; (*d*) for thermostats. Types (*e*) and (*f*) are for suppression of television interference from two- and three-core appliances. Type (*g*) is an all-wave filter for broadcast and television interference suppression. The value of C may vary between 0·01 and 0·5 μF. On type (*c*) the values given are the largest permissible.

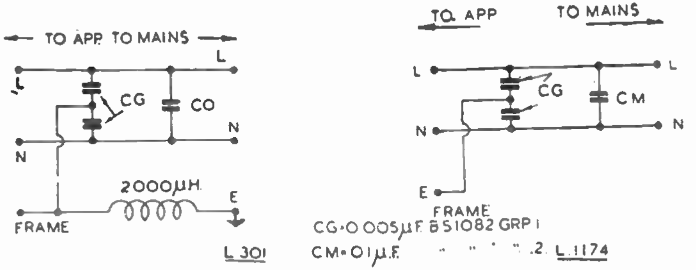


FIG. 2.—FLEX LEAD SUPPRESSOR CIRCUITS.

Capacitor suppressors should be fitted within 1 ft. of any portable appliance suspected of creating interference. The suppressor on the left is for earthed appliances (three-core flex), whilst the one on the right is intended for unearthed appliances (two-core flex).

(Belling & Lee Ltd.)

currents which may be generated to that particular circuit, or at least to the apparatus concerned, by the use of suitable isolating resistors or chokes in conjunction with by-pass capacitors. Sufficient suppression can often be achieved by the use of by-pass capacitors alone; where they alone are ineffective chokes should be included in the line circuit to offer high impedance. Since these chokes often carry considerable current, they must be of low resistance and suitably rated for the apparatus concerned.

A wide range of interference-suppression devices has been developed and marketed. Capacitors used in suppressors must always be generously rated for A.C., and normally only components specifically designed for this type of work should be used.

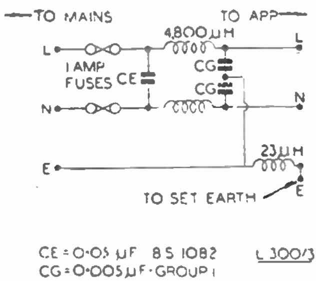
COMMUTATOR MOTORS. Capacitor or inductor capacitor filters (see Fig. 2), should be connected as close as possible to the motor. An R.F. choke should be included in the earth lead if the performance is better without the earth connection.

A.C. SYNCHRONOUS AND INDUCTION MOTORS. These do not usually cause interference; and the exceptions can generally be suppressed with a simple capacitor filter.

FIG. 3.—CIRCUIT OF ALL-WAVE MAINS FILTER.

This filter is suitable for use with 230-volt A.C. receivers and provides suppression of mains-borne interference between 10 and 2,000 metres.

(Belling & Lee Ltd.)



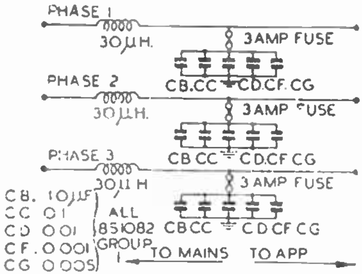


Fig. 4.—Industrial Suppressors.

A treble inductor and capacitor filter for use on AC./D.C. motors, two- or three-wire D.C. distribution circuits and three-phase power supplies. Standard range of these units is from 15 to 60 amperes, but heavy-duty units capable of carrying up to 300 amperes may be obtained.

(Belling & Lee Ltd.)

A.C. Repulsion Motors. The capacitor unit should be connected across the mains-supply feed to the motor and not across the brushes.

Small Motors. Unearthed motors in portable appliances are a prolific source of interference. The centre point of a capacitor filter of the type shown in Fig. 2 is taken to the motor frame. This tends to place the frame above earth potential, but the possibility of shock is avoided by keeping the capacitors small in value.

Flashing Electric Signs, Thermostats, etc. Capacitors should be connected across the contacts in series with resistances to introduce time-lag. R.F. chokes may be included in the mains supply lead.

Lifts. The makers of the equipment usually supply the necessary suppressors for preventing the radiation of interference from the motor control panel or trailing cables.

Electro-medical Equipment. Direct radiation on S.W. and mains-conducted disturbances may be produced over considerable distances. Complete suppression can often be achieved only by operation of the equipment in a screened room with choke filters on all wires passing out of this room.

Neon Signs. Capacitor filters should be connected across the transformer primary with a screened L.F. choke in the H.T. circuit, preferably near the centre of the sign. The casing of the choke should be earthed. All H.T. wiring should also be screened with the metal casing earthed.

Fluorescent Lighting. A small capacitor, usually of 0·02 μF, is fitted across the starter contacts to suppress radio interference which may be generated within the lamp, and also to give a cleaner break to assist starting. Occasional interference of a continuous nature, however, may be experienced on M.W. and S.W. either by direct or mains-conducted radiation. This interference may not arise immediately in a new tube, and, if the tube is handled, it may stop temporarily. Proprietary suppressors designed to eliminate this form of interference are available.

Suppression at the Receiver

Suppression at the receiver may be required where it is not possible to identify or to suppress interference at the source. Standard procedure is to connect filter circuits either at the mains input to the receiver or where the mains supply enters the premises; and/or the use of a special screened aerial to reduce radiation pick-up.

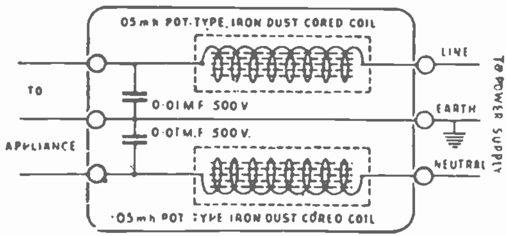


FIG. 5.—DUAL-PURPOSE MAINS SUPPRESSOR.

This can either be used to eliminate interference from vacuum cleaners and the like or alternatively as a filter in the power lead to the receiver where the source of interference is outside the control of the set-owner.

(*Aerialite Ltd.*)

It should be emphasized that interference radiation is normally confined to within a few feet of the house wiring, structural steelwork, tubing, roadway, etc. An aerial with the main part situated well clear of surrounding objects will seldom pick up any appreciable amount of unwanted noise, particularly on M.W. and L.W. At the same time it will provide a strong signal input to the receiver, which, through A.V.C. action, will thus operate at lower sensitivity and be less susceptible to conducted and induced interference than would otherwise be the case. This statement is borne out by an analysis of the G.P.O. interference investigation reports, which shows that the most frequent cause of unsatisfactory reception is the use of an inefficient aerial/earth system. In districts subject to mains-borne interference, the use of a " mains aerial " is strongly to be deprecated.

In areas subject to strong interference, the pick up of unwanted noise on the down lead and lead-in of an outdoor aerial may prove objectionable. Simple shielding of these sections of the aerial system is practicable only where comparatively short lengths of lead-in are involved. This is because of the signal losses introduced by the capacitative effects of screened cable. In such cases, it is therefore necessary to adopt transmission-line technique and to match correctly the impedance of the aerial to the feeder and the feeder to the receiver-input circuit. A low-impedance screened line may then be used without appreciable loss in signal voltage. While there are several simple arrangements suitable for use on one fixed frequency in the S.W. range, special matching transformers are usually required on M.W. and L.W.

Noise Limiters

Impulsive, or sharply peaked, interference signals, such as those produced by automobile ignition systems, are of particular importance in the range 10–100 Mc/s, and various noise-limiting circuits have been developed for use on communications receivers and also on television receivers (see Fig. 7). An important advantage of frequency modulated systems on these frequencies is the reduced susceptibility to interference of this type.

Ignition interference has exceptionally high peak amplitudes, but lasts for only short periods. Simple forms of amplitude limitation of the audio-frequency output will thus tend to " slice " off the peaks, and such devices are now included as standard practice in most television sound receivers. For communications work, however, more elaborate types of filters are often required. Basically, these permit the R.F.

noise pulses to be amplified together with the required signal, and are then rectified to provide a biasing pulse which can be used to reduce temporarily the gain of the amplifier; while this will reduce the output of the required signal, the resulting "holes of silence" will disturb the ear less than would the peaks of noise, and intelligibility is thus improved.

To be effective, the high-frequency components of the noise pulse must not be diminished, and for this reason noise limiters tend to be less efficient when used in receivers possessing extremely high selectivity. Care must also be taken in the choice of the time-constants in the noise-rectifier circuit, and stray capacitances should be avoided as far as possible.

An arrangement which provides for adjustment of the modulation percentage threshold at which limiting starts and in which the limiting is automatically adjusted to the instantaneous carrier level is shown in Fig. 6. To avoid the introduction of A.C. hum, the valve and components should preferably be screened.

AUTOMOBILE INTERFERENCE SUPPRESSION

Ignition Interference

Ignition interference is a serious menace not only to the operation of a car radio receiver but also to television and to the many other services using the H.F. and V.H.F. spectrum. The interference originates mainly at the sparking-plugs, though a certain amount can be traced to the gaps between the rotor arm and the fixed contacts in the distributor, and also to the operation of the contact-breaker in the H.T. circuit of the ignition coil. The layout of the ignition system, the length of the H.T. leads and the condition of the sparking-plugs have an important bearing on the amount of interference likely to be radiated.

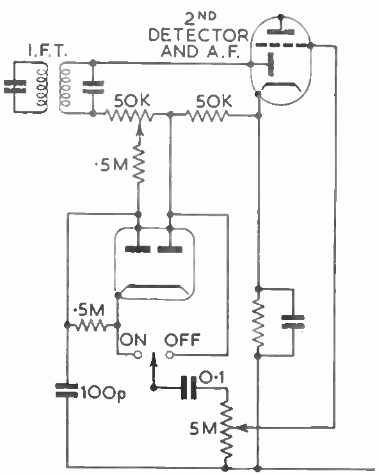


FIG. 6.—NOISE LIMITER CIRCUIT FOR COMMUNICATIONS-TYPE RECEIVERS.

Adequate suppression can be obtained both easily and cheaply by the fitting of a resistor (5,000–15,000 ohms) in the main lead from the ignition coil to the distributor; it should be placed as close as possible to the distributor. Suitable resistors are made by a number of manufacturers. Alternatively, in more difficult cases, resistors are sometimes fitted in the sparking-plug leads as close as possible to each plug. Resistors fitted in this position must be capable of withstanding high plug temperatures, and can usually be obtained from service garages. Resistors are sometimes fitted in the sparking-plug leads at the point where they leave the distributor, and occasionally some form of capacitance is arranged between the hot end of each resistor and the frame of the engine.

Where mobile communications equipment or sensitive radio receivers are installed in automobiles, more elaborate precautions to reduce interference from the ignition system and auxiliary equipment will often be necessary.

Additional methods of reducing ignition interference include: checking that the sparking-plugs are in first-class condition and that the gaps are correctly adjusted; eliminating any tendency for small sparks to occur at the " pinch fit " terminals on H.T. leads by soldering each H.T. wire to its terminating lug; increasing the distance between L.T. and H.T. wires, if unshielded, to at least 6 in. Ignition interference often reaches the receiver via the L.T. wires and any other wires and controls that pass through the bulkhead. The L.T. wire from the ignition coil to the dashboard switch is a frequent carrier of interference; this can be reduced by fitting a 0·25-μF capacitor between the switch and the chassis. Any other leads suspected of forming similar R.F. paths should be by-passed in a similar manner, the capacitors being connected with leads not more than 1 in. in length.

Other Sources of Interference

DYNAMO. Interference from the dynamo may be suppressed by the connection of a 1-μF capacitor to the live terminal (usually denoted by the connection thereto of a yellow wire); the casing of the capacitor should be bonded to the dynamo casing or to the chassis at a point within 2 or 3 in. of the dynamo.

IGNITION COIL. Interference from the ignition coil may be suppressed by the connection of a 0·5-μF capacitor to the switch terminal of the coil (i.e., not to the contact-breaker or H.T. connection). Secure the capacitor casing to the bolt holding the engine block or bulkhead with the shortest possible lead.

DISTRIBUTOR ROTOR. If the rotor in the distributor is burnt black, it should be replaced. In some cases it may also be necessary to beat out the metal contact in order to reduce the gap across which the spark has to jump, but in doing this care must be exercised to ensure that subsequently the rotor will not foul the distributor cap.

AUXILIARY APPARATUS. Auxiliary electrical equipment fitted to the vehicle which may interfere with reception includes electric windscreen wipers, electric petrol-pumps, electric clocks, thermostats, air-circulator fan motors, etc. Such interference can usually be suppressed by connecting a 0·5-μF capacitor with short leads between the live terminal of the unit concerned and chassis.

WHEEL STATIC. Interference similar in effect to natural static may be produced by static electricity generated by the rotating tyres and brake drums. This interference is normally most serious in dry climates and on paved roads. The cure is to provide some form of leakage path between chassis and earth either by the provision of specially treated tyres, the injection of special tyre powder or a little water into the inner tubes, by some form of trailing contact, or by painting the tube and/or cover with lampblack or graphite (black lead).

CRACKLES. Interference in the form of crackles may be caused by a scraping contact between metal components at different potentials. Therefore all main components of the car should be brought to chassis potential with the aid of short lengths of copper bonding braid. Parts which most often require this treatment include rubber-mounted engines, exhaust pipes and silencers and any special fittings. Care should be taken to remove all paint and rust around the braiding joints. In coupé-type cars, the hood, if possible, should be connected to chassis at several points to prevent re-radiation of interference to the aerial.

VIBRATOR HASH. The suppression of interference arising from the minute sparks that occur between the fixed contacts and moving arm of a vibrator is normally accepted as the responsibility of the manufacturer. Suitable choke/capacitor filters are usually fitted not only in the vibrator unit itself but also in any wiring which may pick up hash interference by direct radiation. Adequate screening and the careful design of the receiver input circuit also play an important role in the suppresson of this interference.

A gradual increase in hash interference, however, may be expected as a result of oxidation and mechanical wear of the vibrator contacts; excessive interference may require the replacement of the vibrator.

TELEVISION INTERFERENCE

The very high frequencies (30–300 Mc/s), on which television broadcasting at present takes place, are particularly susceptible to interference from local electrical appliances and spark-ignition systems, the interfering signals arriving at the receiver either by direct radiation from the source or by conducted radiation along power mains and overhead wiring, or by a combination of the two.

In practice, interfering signals fall into two main categories, which require entirely different treatment: the impulse type of interference producing spots or lines of peak white across the screen, and continuous wave signals on frequencies falling within the acceptance band of the receiver and producing heterodyne interference in the form of a "herring-bone" pattern of alternate dark and light bands running diagonally across the screen.

Impulse Interference

Impulse interference will cause crackling on sound and a series of white spots or bright streaks of light on the screen. In this category must also be included switching circuits, such as thermostats and dirty light switches, where slight arcing may take place. As with all forms of interference, the effect will largely depend upon the ratio of the levels of the interfering signal to the picture signal, and will thus be more severe in "fringe" areas.

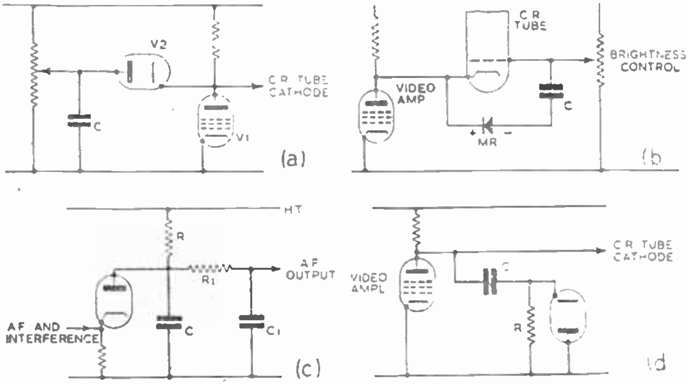


FIG. 7.—INTERFERENCE SUPPRESSION CIRCUITS.

(*a*) Basic vision-interference suppression circuit. (*b*) Automatic form of peak limiter. (*c*) Typical pulse-width limiter circuit. (*d*) Vision-interference limiter operating on a time-constant basis.

Where the source is unknown or cannot readily be suppressed, the effects can be reduced by: (1) the provision, on the sound and vision receiver units, of interference limiter circuits, designed to cut off the high amplitude peaks of the interfering signals—popular circuits for such limiters are shown in Fig. 7 (in practice the valve diodes are sometimes replaced by germanium crystal diodes)—or alternatively, by inverting the pulses so that they produce less-noticeable black instead of white spots; (2) ensuring that the most efficient aerial system is employed, with the elements as far away as possible from the source of interference or from wires and guttering that could form conduction paths; (3) where the source of the interference is known, as, for example, the cars passing along a main road, a directional aerial system may be orientated so as to provide minimum pick-up from this direction; (4) the use of interference filters in the mains supply leads to the receiver. Method (4) is unlikely to prove as effective at television frequencies as for normal broadcasting frequencies owing to the greater ease with which the leads before and after the filter can act as aerials and thus allow the interference to bridge the filter; nevertheless, a number of filters especially designed for use at television frequencies are now on the market, and will often bring about a considerable improvement, particularly when used in conjunction with methods (1)–(3).

An effect somewhat akin to ignition interference may be caused by corona discharge ("brushing") from points at E.H.T. potential, but it can easily be recognized by its more continuous nature.

Heterodyne Interference

The second and less common group of interfering signals are those in which oscillation is continuous, as opposed to trains or damped oscillation, and these are usually tunable over a comparatively narrow band; they produce heterodyne interference, against which peak limiters and suppression filters of the type so far described are ineffective. The most common causes of such interference are diathermy apparatus, adjacent-channel interference, harmonics of short-wave broadcasting, communication and amateur transmitters, and radiation from the local oscillator of other television or short-wave receivers. Susceptibility to certain forms of this interference, particularly, for example, to break-through on the intermediate frequency of the television receiver, will depend very largely upon the inherent design of the receiver and the choice of the intermediate frequency.

The interference may arise from radiation (usually harmonic) taking place at frequencies within or close to the television channel (in this case suppression at the source or careful orientation of the receiving aerial are likely to be the only effective cures); to image (second channel) response in the receiver; from "blanketing" or swamping of the receiver by very strong local signals, or cross modulation, sometimes produced by rectification in the aerial system or local metalwork; from break-through of a signal on a frequency close to the intermediate frequency; or from a combination of these. By fitting suitable traps and filters in the offending transmitter, and by careful screening, harmonic radiation can be much reduced. Rejection of fundamental or intermediate frequency signals can often be improved by screened traps or filters in the aerial feeder of the receiver, close to the receiver, or alternatively as close as possible to the control grid of the first R.F. amplifier or mixer valve. Such a wave trap resembles those used in the early days of broadcasting, and is tuned to the unwanted frequency (provided that this does not lie within the required television channel). A coil with 10 turns of 18 S.W.G. wire, spaced wire diameter, and with an internal diameter of $\frac{5}{16}$ in., tuned by a 3–30-pF trimmer will have a tuning range of about 40–50 Mc/s, and for offending signals on other frequencies the number of turns should be increased or decreased accordingly. For the rejection of signals at intermediate frequencies, high-pass filters with a pass band of about 40–60 Mc/s may be fitted in the aerial lead of the receiver, and will normally prove effective provided that the screening of the I.F. stages is adequate to prevent direct pick-up. Here again, careful positioning and orientation of aerials may prove of assistance.

The Bedford Diagram

This diagram (L. H. Bedford, *Journal of the Television Society*, March 1937) enables: (1) local oscillator frequencies for any given intermediate frequency to be rapidly determined. It shows: (2) where second-channel interference may be expected; and (3) where interference from harmonics of the I.F. will arise, these being fed back to the mixer with incoming signals to produce beats.

The diagram has become more complicated with increase in the number of active channels, and the possibilities of interference are now much greater. In Fig. 8 the bottom scale shows the I.F.s, and vertically the scale is of carrier and harmonic frequencies. Horizontally from the left-hand axis are shown the five channels. The full sloping lines represent the 2nd, 3rd, 4th and 5th harmonics of the I.F.s. Thus for an I.F. of 15 Mc/s, bottom scale, the 2nd, 3rd, 4th and 5th harmonic frequencies are respectively 30, 45, 60 and 75 Mc/s

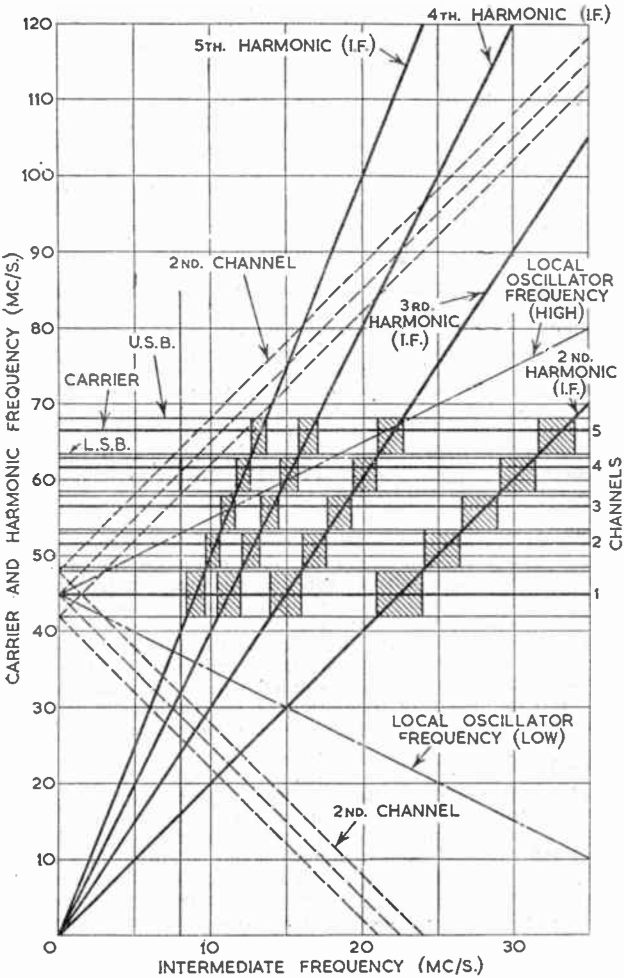


FIG. 8.—THE BEDFORD DIAGRAM.

vertical scale. The harmonic lines run through these and similarly-found points. The harmonic interference bands are shown hatched. Thus in Channel 1 the 2nd harmonic of an I.F. in the range 21·25–23·75 Mc/s would interfere, and so would the 3rd harmonic of any I.F. between 14·2 and 15·8 Mc/s. The argument is applicable to each channel and each harmonic (if necessary beyond the 5th).

The chain dotted lines show local oscillator frequencies above and below the carrier, while the broken lines indicate the second-channel interference for beating high and low. Only the set for Channel 1 is shown for greater clarity. From the diagram it can be seen that when only Channel 1 existed, a possible I.F. was 8 Mc/s; but now the second channel of this, if beating high, would be in Channel 4. Further, it will be found that if beating low were attempted, a receiver tuned to Channel 4 would have its second channel in Channel 1, with an 8-Mc/s I.F.

One of the major problems facing the designer is so to choose the I.F. and local-oscillator frequencies that with one "run" of receivers and one common I.F., any of the thirteen channels may be tuned in without I.F. harmonics, second-channel or other spurious response becoming troublesome, moreover, other bands (e.g., Amateur and Defence) must also be avoided. To clear harmonic interference, a vision I.F. of 34·65 Mc/s and a sound I.F. of 38·15 Mc/s is generally used, though much lower frequencies have been employed in the past. The local oscillator should beat high, though again the other preference may be taken. The interferences to be expected on Band III (174–216 Mc/s) can be examined if the diagram is extended to these frequencies.

Diathermy

This is a particular form of heterodyne interference, the cause in this case being the harmonic output from the relatively unsmoothed valve oscillators used in electro-medical apparatus. In addition to the herring-bone pattern, the sound channel is often affected by harsh crackling or low-pitched hum. The most satisfactory cure for this form of interference is the complete electrical screening of the offending apparatus; though in some cases relief can be obtained by slightly changing the frequency of oscillation of the equipment so that the harmonics no longer fall in the television channel concerned.

Freak Propagation

Owing to the fact that television channels are shared on a geographical basis, herring-bone interference patterns may sometimes be caused by signals being received from a distant station, normally inaudible. The most common cause of such a propagational condition is a period of "Sporadic E" when the E layer becomes highly ionized and reflects signals up to or about 70 Mc/s over distances between 200 and 1,000 miles. Pronounced temperature inversions, such as occur during summer evenings, may also cause slight interference. As such conditions seldom last for long, it is usually considered unnecessary to take precautions against this form of interference.

Camera Faults

Occasionally, a slight "herring-bone" effect may be produced by certain television cameras, and for this, of course, no remedy is possible

TABLE 1.—CAUSES OF INTERFERENCE INVESTIGATED BY THE G.P.O. DURING A TYPICAL TWELVE-MONTH PERIOD

Cause	*Television*	*Sound*
Bed-warmers	1,183	578
Calculating machines	818	342
Drills	2,492	1,177
External cross-modulation	45	280
Faulty electrical wiring of premises	494	2,194
Hair-dryers	6,954	598
Ignition systems of petrol engines	1,313	49
Radio-frequency equipment (valve)	736	91
Radio-frequency equipment (spark)	151	105
Filament type lamps	2,569	66
Fluorescent tubes	233	1,676
Street lighting	113	712
Neon signs	1,444	416
Power lines, overhead:		
Less than 650 V	230	260
650 V to 11 kV	2,052	200
More than 11 kV	1,447	222
Power lines, underground, faulty	60	132
Radiation from TV time-base circuits	3	6,805
Superhet, local oscillators	1,604	62
Radio transmitters:		
Amateurs	303	125
Others sited in U.K.	476	142
Foreign	146	533
Refrigerators	1,587	1,228
Sewing machines	8,956	1,577
Smoothing irons	198	399
Vacuum cleaners	3,269	1,043
Unknown	21,877	12,206
All other identified sources:		
Contacts type	3,356	1,978
Commutator type	7,056	2,930
Miscellaneous types	1,692	1,468
Receiver faults		
Inefficient aerial or earth installations	2,456	10,335
Faulty receivers	5,375	5,290
Maladjustment of receivers	666	227
Other conditions	2,998	3,168

at the receiving end. By noting whether such interference ceases or persists when cameras are changed, it is usually possible to determine whether the pattern is present in the transmitted image or is due to local causes.

Aircraft Flutter

Reflection of signals from aircraft may provide alternate augmentation and attenuation of the signal as the phase of the reflected signal changes in relation to that of the direct signal; this will cause the contrast of the picture to change rapidly from normal to low and then to high, the cycle being repeated in rapid succession for as long as the aircraft is in the neighbourhood.

The degree to which a particular receiver installation is susceptible to this form of interference will largely depend upon the aerial system: a system which receives only vertical polarized signals will generally be less affected than one capable of responding to the horizontal component. The effects may also be diminished by reducing the D.C. coupling to the cathode-ray tube, or by the use of some form of automatic picture control.

Interference by Television Receivers

Radiation of parasitic oscillation occurring in a television receiver may affect local broadcasting and television receivers : the cure here is to find the source of the parasitic oscillation and to improve the stability of the offending stage(s).

Radiation in the form of induced electric and magnetic fields may be set up in the neighbourhood of a television receiver, and may affect nearby broadcasting receivers, particularly on the long-wave band. The most likely sources of these fields are the line-output transformer and associated points at E.H.T., the deflector coils and the high-impedance circuits near these components. The following methods of reducing such interference have been recommended by Messrs. Mullard, Ltd. :

(1) The E.H.T. transformer, booster diode and line-output valve should be totally screened by a can which makes good contact with the chassis. Two-hole fixing of the can is not entirely satisfactory, and it is advisable to make multiple connections between can and chassis. The difference in radiation between a good and a bad connection here may amount to as much as 8 db for magnetic fields.

(2) Any width or linearity controls of the inductor type should be screened separately if they cannot be accommodated inside the line-output screening can. (The design of the line-output screening involves problems of ventilation to avoid overheating of the components enclosed by the screen. As a general guide, the maximum safe bulb temperature for the PL81 line-output pentode has been determined at 185° C.)

(3) The deflector coils should be screened as far as possible by an aluminium can or by metal foil wound co-axially around the coil and earthed to chassis. Care must be taken to ensure that there is no likelihood of voltage breakdown between the foil and the coils. This form of screening will give good reduction of electric fields, and will also reduce magnetic fields, though not to the same degree. To reduce the magnetic field still further, the deflector coil-screening can should have endplates with holes only just large enough for the tube neck to pass through. This gives a further reduction of approximately 6 db.

(4) Care should be taken in the layout of the receiver to keep circuits of high impedance well away from the worst sources of interference.

(5) The graphite coating of the cathode-ray tube should be efficiently connected to earth—preferably from two separate points on the coating.

(6) Both conductors of the mains supply should be connected to the earth terminal via 0·05-μF paper capacitors rated for 600 volts (r.m.s.) working.

(7) The use of a perforated foil screen at the back of the set will reduce radiation in that direction.

Television Interference Suppressors

The difficulties of suppressing interference at television frequencies from small universal motors have already been considered briefly. The self-inductance of normal capacitors and the self-capacitance of inductors at these frequencies are sufficient to impair their effectiveness as filter components. For these reasons it is generally necessary to use specially designed components and to pay particular attention to making all leads as short as possible and to install the filter to within

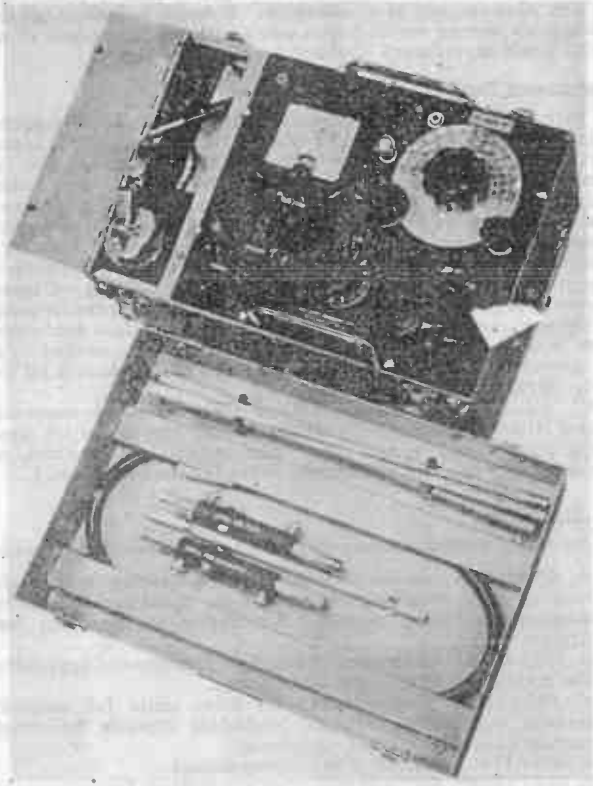


FIG. 9.—PORTABLE INTERFERENCE TRACER DEVELOPED FOR TRACING INTERFERENCE TO TELEVISION RECEPTION.

(*Murphy Radio Ltd.*)

an inch or two of the source of interference. Small moulded mica capacitors or miniature paper dielectric types with a capacitance of the order of 470 pF are usually satisfactory.

Another type of filter which has been developed recently is the inductor suppressor, in which capacitors may be completely omitted. Inductors are less affected by the by-passing action of stray fields. In practice, the self-capacitance of the inductor can be made to resonate the inductor at or near to the mid-band frequency of the television station, and thus to improve the suppression characteristics. In one simple type of filter inductors are placed in the live and neutral leads close to the source of interference. The placing of inductor filters is less critical than that of capacitor filters, but should preferably be within 6 in. of the source of interference. Suitable inductors are single-layer solenoids wound on dust-iron cores with an inductance of about 8 μH and a self-capacitance of about 1 pF.

Interference on Band III

The suppression of electrical interference at Band III frequencies presents greater difficulties than at Band I frequencies. This is mainly due to the considerably greater radiation that occurs at the higher frequencies from comparatively short lengths of wire. For example, lead suppressors fitted in the flex leads of small domestic motors that may be quite effective on Band I, may fail at Band III owing to the radiation of interference from the short length of lead between the appliance and the suppressor components; in such cases the suppressor component may have to be fitted within the motor housing. Capacitors must have extremely low inductance to be efficient at these frequencies.

For the suppression of ignition interference, individual heat-resisting resistors mounted close to each sparking-plug may be needed in place of tho one resistor in the main lead to be distributor, which is all that is generally required for Band I.

On the other hand, aerials on Band III tend to be more highly directional than those used for Band I, and can often be sited to minimize interference from a particular source. In practice, interference is usually less troublesome on Band III than on Band I.

British Standards

The following British Standards relate to interference suppression :

- B.S. 905 : 1940. Anti-interference characteristics and performance of radio receiving equipment for aural and visual reproduction (excluding receivers for motor vehicles and marine equipment).
- B.S. 727 : 1954. Characteristics and performance of apparatus for the measurement of radio interference.
- B.S. 613 : 1955. Components and filter units for radio-interference suppression devices (excluding devices for traction, marine and other special equipment).
- B.S. 800 : 1954. Limits of radio interference.
- B.S. 833 : 1953. Radio-interference suppression for motor vehicles and internal-combustion engines.
- B.S. 827 : 1939. Radio-interference suppression for trolley-buses and tramways.
- B.S. 1597 : 1949. Radio interference suppression on marine installations.

The following Codes of Practice have been prepared by joint committees of the Institution of Electrical Engineers and the British Standards Institution :

- C.P. 1001 : 1947. Abatement of radio interference caused by motor vehicles and internal-combustion engines.
- C.P. 1002 : 1947. Abatement of radio interference from electro-medical and industrial radio-frequency equipment.
- C.P. 1006 : 1955. General aspects of radio interference suppression.

34. MAGNETIC AND DISC RECORDING

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MAGNETIC RECORDING

Magnetic recording originated in the year 1899 with the invention by Valdemar Poulsen of the "Telegraphone".[1] During the period between the two World Wars, development work was carried out which resulted in improved systems, notable among which were the Blattnerphone and the Marconi-Stille.[2]

These systems employed a steel tape as the recording medium, a material which was bulky and difficult to handle. Mechanical and magnetic requirements were often mutually exclusive. The jointing of sections was also difficult.

In 1928, Dr. Pfleumer took out a patent in Germany for a new and much improved medium. This consisted of a paper or plastic base tape, having most of the required mechanical properties, on which was deposited a very thin coating of powdered magnetic material. A later patent (1935) introduced oxide of iron as the magnetic medium. This oxide, though not ideal in its magnetic properties, was a considerable improvement on the steel tapes previously available. During the 1939–45 war, considerable development work was carried out in Germany, resulting in a system of recording sound magnetically which ranked with the best on disc or film.

Concurrently, in the U.S.A., the problem of the bulk of the steel tapes had been overcome by utilizing a thin steel wire,[3] normally 0·004 in. diameter. As with the Magnetophon tape, the reduction of material available for recording upon was offset by improvements in other factors.

Up to 1946 the greater part of the development work had been carried out empirically. Since that date, however, considerable research has been undertaken into the causes underlying the observed results. Due to this research, a number of improvements has been made possible, and within the standards of current practice, magnetic tape recording shows superiority over both disc and film recording. In fact, it is now the normal practice for commercial disc and film recordings to be made through an intermediate magnetic recording process, without any noticeable loss of quality in the finished product. Even a trained ear has difficulty in distinguishing a good magnetic tape recording from the original sound reproduced over the same amplifier/loudspeaker system. Quite small and compact recorders, suitable for domestic use, while they cannot compete with the larger professional machines, are capable of a remarkably high quality of reproduction.

Wire recording has not proved itself capable of such a high standard of quality as tape, but has met a need for a very small, light and compact recording machine, suitable for the intelligible recording of speech for long runs of an hour or more. Domestic models are also available, but it would seem that generally tape recording is becoming more popular for this application.

So far as is known, magnetic recordings are permanent if properly stored. However, they have the outstanding advantage, particularly in the domestic field, that the recording can be erased at will and the material re-used as often as desired. Tape may be cut and joined with precision, on a syllable of speech or a note of music. Wire can be cut, and two pieces joined by tying in a reef knot, which will pass through the mechanism of the recording machine without mishap.

Wire

The wire used for magnetic recording is made from austenitic stainless steel, drawn to a diameter of 0·004 in., and heat treated.[4] In structure, this consists of non-magnetic crystals, embedded in a matrix of magnetic material. The non-magnetic crystals keep the magnetic elements apart, and thus reduce the tendency to mutual demagnetization. The wire is usually supplied on spools approximately 3 in. diameter and 1 in. thick, holding sufficient wire for one hour's continuous recording at intelligible speech quality. Some machines employ ready-loaded magazines of wire which have only to be slipped into the machine for the latter to be ready for use.

Tape

Recording tapes may consist of a non-magnetic base, coated with the magnetic material, or the magnetic material may be embedded in, and form the body of, the tape itself. The base of coated tapes may be of paper, cellulose acetate, specially treated polyvinyl-chloride (P.V.C.), or other plastics.[5] The impregnated tapes usually consist of equal quantities of magnetic material and vinyl plastic, calendered into a thin sheet. The material is manufactured in the form of strips, from 12 to 24 in. in width and 0·002 to 0·0025 in. thickness, and subsequently slit into tapes of $\frac{1}{4}$-in. width.

Coated tapes employ a base of similar dimensions, which is coated to a thickness of 0·0005 in. with either black oxide of iron (Fe_3O_4), red oxide of iron (Fe_2O_3) or a mixture of the two. The coercivity of the oxide may be increased considerably by special treatment. Early tapes had a coercivity of the order of 100 oersteds and a remanence of about 300 gauss. Later types have coercivities of the order of 260 oersteds and remanence of about 600 gauss.

The oxide[6] is produced in the form of a fine powder, the individual particles being less than 1 micron ($\frac{1}{1000}$ mm.) in size. These are coated on the tape surface, mixed with a lacquer which binds them to the base and provides sufficient isolation to prevent undue mutual demagnetization. The extremely small size of the particles ensures that there are always a sufficient number in the vicinity of the head gap for their performance to be considered statistically. Any appreciable irregularities in size or distribution result in imperfect contact of the tape and head surfaces, and in uneven magnetic properties, and hence lead to a high noise level and variable characteristics. A very smooth surface to the tape coating is therefore desirable, and to this end some manufacturers polish the surface after coating.

Basic Principle

The recording medium, be it wire or tape, is drawn steadily over the pole of an electromagnet, the current in the coils of which is varied in accordance with the signal waveform. As each section of the medium passes over the pole, it becomes permanently magnetized to a degree dependent upon the strength and direction of the current at that instant, and a wave of varying magnetization is therefore impressed along the length of the medium.

When the medium is again drawn, at the same speed, over the pole of a similar electromagnet which is not energized, each magnetized section in the medium causes a corresponding flux in the core, and as this flux varies, voltages are induced in the coil proportional to the rate of variation. These voltages can be applied to an amplifier, and will produce an output signal.

Immediately prior to recording, it is usual to pass the medium over an erasing magnet, which removes any previous recording and conditions the tape to receive the new signal. When replaying, both the erase and the recording head are, of course, switched off. Sometimes, in the cheaper equipments, a single head serves for both recording and replaying, being switched to one or other function alternatively.

Tape Transport

One of the most difficult problems in the design of a high-quality tape-recording system is the mechanism required for feeding the tape over the recording and reproducing heads at constant speed. A typical arrangement of such a mechanism—usually referred to as a tape-deck—is shown in Fig. 1. Tape from the left-hand reel is guided into contact with the heads by pulleys, and is passed to a take-up spool, which is driven by a lightly energized motor. The tape is carried smoothly and at constant speed over the heads by being pressed into contact with a

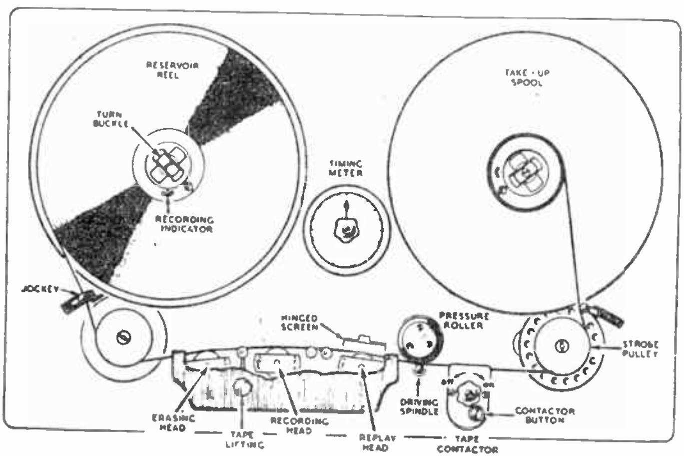

FIG. 1.—TYPICAL BROADCAST TAPE RECORDER.
(*E.M.I. Sales & Service Ltd.*)

revolving driving-spindle by means of a rubber-faced pressure roller. Even assuming constancy of rotational speed of the driving spindle, or "capstan", there is some inevitable slip between capstan and tape, and there may be variations in stretching of the tape between capstan and heads, owing to variations of the tape tension. These and other factors lead to irregularities of tape speed. Under certain conditions irregularities as small as 0·1 per cent are audible in the form of flutter and wow.

Pressure on Heads

Another important factor is the constancy of pressure of the tape on the heads. In tape-decks of the type shown in Fig. 1 this pressure is due to the tension of the tape. The left-hand spool is provided with a motor, normally used to rewind tape ready for replaying. When recording or reproducing, this motor is lightly energized to drive the spool in a clockwise direction, and the spool is therefore being pulled round by the tape against the action of the motor. The motor is arranged, under these conditions, to have a torque which is nearly inversely proportional to the speed of rotation, so that the tension in the tape remains almost constant.

Alternatively, felt pressure pads can be arranged to press the tape into contact with the heads, the tape tension being then less critical. Care is needed to see that the pressure is not too great, as the tape surface is highly abrasive and severe head wear will result.

With tape under tension, it is possible for both lateral and longitudinal vibrations to be set up as it passes over the heads, the tape behaving as a stretched string. These vibrations may result in "modulation noise", a background noise which rises and falls with signal level. It should be noted, however, that it is equally possible for pressure pads, where used, to produce damped longitudinal vibrations.

In most wire recorders the wire is drawn across the head by the rotation of the take-up spool, the rotational speed of which is maintained constant. A slowly oscillating guide distributes the wire evenly over the surface of the spool, and by maintaining even layers in the winding, prevents bunching and slipping of turns, which would result in uneven drive.

Heads

The heads may consist of straight, laminated-iron cores wound with insulated-wire coils, and tapered to an edge across which the tape is drawn. However, such construction tends to be extremely inefficient, and, in the case of the replay head, is very susceptible to hum pick-up from stray fields. This latter factor is important, as the flux induced by the tape at low frequencies is very small indeed. Much better results are obtained by making the core in the form of a nearly closed loop, the tape being carried across a small gap between the poles. A popular form is the ring-head, and this is illustrated in Fig. 2. A stack of laminations of the shape shown, to a thickness of about $\frac{5}{16}$ in., is wound with two symmetrically disposed coils, which, in the case of erase and recording heads, produce the working flux across the upper gap between the pole faces. A metallic shim is inserted in this gap, the eddy currents in which tend to force the flux out of the gap proper into the space immediately above the gap, in which the magnetic surface of the tape is situated. In the case of wire recorders the stack of laminations may be thinner, and a slot is cut in the pole tips, in the centre of the stack, and in the plane of the diagram. The wire runs in this slot, and is therefore almost completely enclosed by the poles.

It will be seen that magnetization of a particle in or opposite the gap takes place longitudinally, in a direction along the axis of the tape or wire. This method has now virtually superseded lateral and transverse magnetization, owing to the much simpler mechanical construction which it offers.

Other forms of head in current use have straight sides, joined by a yoke across the bottom. This simplifies winding of the coils.

Owing to the presence of the iron core, the relationship between flux in the gap and current in the coils is not linear. A second air-gap is

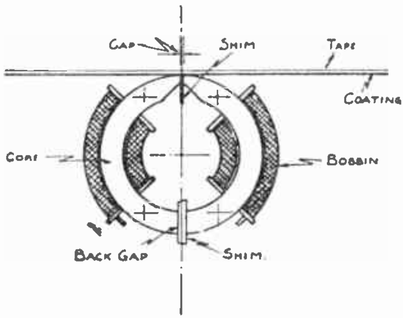

Fig. 2.—Typical Head Assembly.

therefore introduced, usually known as the back-gap, the reluctance of which is large enough to swamp variations in the reluctance of the iron at the flux densities employed, and which thus linearizes the magnetic circuit. In the case of the replay ring-head, making this gap of the same reluctance as the front gap achieves a degree of symmetry which is very helpful in reducing the effect of stray hum fields, since the flux picked up by the core will divide into two equal parts, passing through the two identical coils, and inducing equal voltages in them in phase opposition. However, this may not give adequate linearization if the gaps are small.

Modern heads are often made much smaller than the 1-in. ringhead type described. However, the smaller the pole faces, the greater the trouble which may be expected from irregular frequency response in the bass (see p. 34–15).

In the interests of good high-frequency response, short gaps of 0·00025 in. or less are frequently employed for replay heads, but the proximity of the pole tips by-passes some of the useful flux, and sensitivity, and hence signal/noise ratio may suffer.

There is no advantage obtained from reducing the length of the recording gap, since recording can be said to take place at or beyond the trailing edge of the gap.

Typical gap dimensions and core materials for 1-in.-diameter ring-heads are given in Table 1.

TABLE 1.—TYPICAL DATA FOR 1-IN.-DIAMETER RING-HEADS

Use	*Suitable Material*	*Front Gap (in.)*	*Back Gap (in.)*
Replay	Permalloy B or C, Mumetal or Radiometal	0·0002	0·008
Recording	Permalloy B or C, Mumetal or Radiometal	0·0015	0·016
Erasing	Rhometal or Permalloy D or Ferroceramic	0·015	Nil

Coil Windings

Coil windings are at the discretion of the designer, with a view to matching the associated amplifiers to best advantage. Where connections are short, high-impedance windings of many turns may be employed. Where heads and amplifiers are separated by more than a foot or two, low-impedance windings are to be preferred. In the recording head separate windings may be employed for bias (see later under " Recording ") and signal currents, or the currents may be mixed and passed through a common winding.

Double-purpose Heads

In the cheaper class of equipment erasing and recording functions are sometimes combined in a double-gapped head. The first and larger gap carries the high-frequency erasing flux. The leakage from this gap, passing across the smaller gap, serves as high-frequency bias, the signal flux in this gap being provided by a second, associated, winding.

Frequently, the same head is used for recording and replaying. Excepting for the highest possible quality, the size of gap of the recording head is not unduly critical, and adequate results can be obtained using the narrow gap of the reproducing head.

Gap Dimensions

The edge of the gap is, ideally, the intersection of two surfaces, and is therefore a line having neither breadth nor thickness. Since its area is thus zero, the flux density would be infinite, were the working flux to emanate from this edge. This is clearly impossible, so that the working flux must in fact span a gap slightly longer than the physical gap. With a view to keeping this effective gap shorter than the shortest wavelength to be reproduced, physical gap lengths of replay heads are customarily 0·0002 in. or less. Even butt-joints are sometimes used.

Recording head gaps usually lie between 0·0005 and 0·003 in., and erasing head gaps between 0·010 and 0·025 in.

In order that recorded tapes may be interchangeable from one machine to another, irregularities of the gap must be very small compared with these dimensions. Gap faces of assembled and rigidly clamped stacks of stampings must therefore be carefully ground and lapped. The lapped surfaces forming the gap must be mutually parallel, and must be set perpendicular to the direction of travel of the tape. If there is an error of alignment between the recording and replay heads, severe loss of high frequencies will result, though there will be no introduction of amplitude distortion. Heads are therefore either rigidly fixed to the backplate or, better, are made so that the angle between gap and tape axis—the "azimuth angle"—may be adjusted and set after assembly.

Surface Finish

The head surface on which the tape runs must, equally, be lapped and polished, to ensure intimate contact with the tape. An initially good surface will normally be kept in good condition by the polishing action of the tape. A surface which is bad to start with will frequently become worse with continued use, developing deep scratches and scores, which increase the distortion and the backgrouud noise level.

Screening

The heads need careful screening. A copper eddy-current screen is most effective round the erase head to prevent the erasing flux from spreading. The recording head is most effectively dealt with by means of a high-permeability alloy screen, which prevents the spread of signal flux and also effectively limits the spread of bias flux. The replay head must be protected against the influence of external fields, and here a single—or better a double—high-permeability alloy screen is effective. These screens should, as nearly as possible, completely enclose the heads.

If a head should accidentally become magnetized, it may be demagnetized by withdrawing it slowly from a strong A.C. field (say 50 c/s) to a region where the field is negligibly small or, alternatively, by passing an A.C. through the winding, and reducing the current gradually to zero.

Erasure

Except by drastic methods, such as raising to high temperatures, magnetic materials cannot, in fact, be demagnetized. The small magnetic particles, or "domains" as they are called, are always magnetized in either one direction or the other. When, in a sample of material, the directions of magnetization of the individual domains are arranged in a completely random manner, these effects will all be mutually cancelled, and there will be no external flux round the sample, which is then said to be in a demagnetized state. Conversely, if the sample has a preponderance of domains, the direction of magnetization of which has a component in any given direction. the sample is said to be magnetized along this axis.

The usual and most successful method of erasing a recorded tape takes place in two stages as the tape passes over the erasing head, the latter being fed with sufficient current at a high frequency to carry each section of tape to saturation repeatedly, first in one direction and then in the other. Every domain is thus subjected to a number of successive reversals of its direction of magnetization, after which its condition is little influenced by its initial state, the initial waveform thus being erased.

As the section moves away from the gap, in a diminishing field, this field becomes too weak to reverse first one, then another domain, the cessation of reversal depending upon a number of factors, such as the size of the domain and the direction of its axis relative to the field. The domains are thus left with their directions of magnetization almost completely random, and the section of tape is effectively demagnetized. The word " almost " is used because the action never reaches perfection, and inevitably some background noise remains, usually about 60–80 db below peak signal.

In practice, after a short interval for recovery, some tapes may again show a trace of the original recording. A second erasure will remove this, and some machines are fitted with two erasing heads in cascade. Some others have erasing heads with two successive gaps. Normally, however, the bias in the recording head successfully removes any remaining traces.

Due to losses in the iron of the erasing head, too high a frequency will cause the head to run hot. On the other hand, as high a frequency as possible is desirable, in order that as many reversals of saturation as possible shall take place in a given section of tape while it passes over the gap, and in order that the dying away of the field shall be as smooth as possible as the section moves away from the head. A frequency between 30 and 80 kc/s offers a good compromise, and a typical 1-in.-diameter ring-head would require about 40 ampere-turns for about 70-db erasure.

Distortion containing even order harmonics of the erasing current frequency must be rigidly excluded from the erasing head. Such distortion implies asymmetry of the waveform, which will leave a small majority of domains with a bias in one direction; that is, the tape is permanently magnetized, which results in increased background noise. Slight permanent magnetization of the heads will produce the same result.

The use of ferro ceramic material for erase-head cores results in greatly reduced losses, and hence greatly reduced power requirements for erasure. Magnetic saturation of the pole-tips sets a limit to the useful erasing flux, but quite small heads dissipating about 1 watt give as good erasure as alloy stampings dissipating 5 watts or more.

Recording

Signal currents are fed to the recording head, and the field formed in the space immediately above the gap causes each particle of the tape or wire to become magnetized. This magnetism is retained after the particle leaves the gap, but the residual magnetization is far from being proportional to the signal current, as is well known from the pronounced curvature of the *B–H* curve. The method now universally used to overcome this difficulty is the introduction of alternating-current bias.[6,15]

Alternating current, of carefully controlled amplitude, and frequency considerably higher than the highest audio frequency to be recorded, is

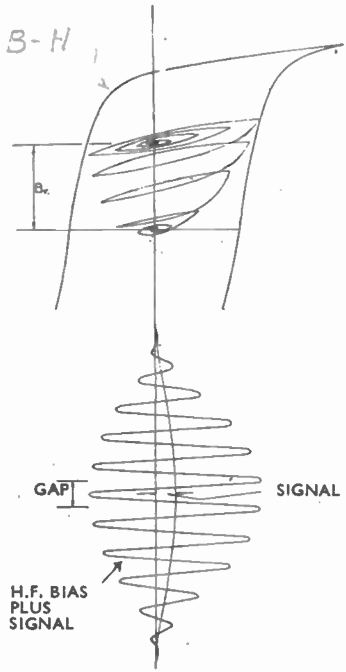

FIG. 3.—RECORDING WITH H.F. BIAS.

additively mixed with the signal current, and the combination is fed to the recording head. Suppose that the signal current, for present purposes, is substantially a D.C., say the top of a square wave or a portion of a very-low-frequency waveform. The curve marked "H.F. bias plus signal" in the lower part of Fig. 3 shows the variations of field, moving from bottom to top of the diagram, which are encountered by a particle of tape in passing over the head. The magnetic behaviour of the particle is shown in the upper part of the figure, superimposed on a *B–H* loop for the material. Starting at the origin, the particle is driven to the region of maximum excursion. Due to the high frequency employed, there are in practice perhaps five or ten complete oscillations at constant amplitude while the particle crosses the gap, and therefore several complete rotations round the largest minor hysteresis loop shown in the figure. As the particle moves into a region of weak field, this minor hysteresis loop collapses on to a point, giving a value of residual magnetization B_r, which is found to be very nearly proportional to the value of the signal current over a wide range, provided that the amplitude of the bias current has been correctly chosen.

Bias

The value of the optimum bias current is not readily calculable, as it depends upon the magnetic properties of the material, and these do not follow known mathematical laws. However, it is evident that higher coercivity materials will require higher bias currents. Typically, using 1-in.-diameter ring-heads of the form shown in Fig. 2, low-coercivity tapes may require about 2 ampere-turns, while so-called high-coercivity tapes may need up to 4 or 6 ampere-turns.

The flux density above the recording gap falls off very rapidly as the point considered departs from the surface of the head. The flux at the working surface of the 0·0005-in.-thickness tape coating is considerably greater than that at the rear surface. Hence overbiasing for the front surface will tend to set up ideal conditions at some depth into the tape coating, and imperfect contact between tape and head surfaces becomes less important. Slight overbiasing is often resorted to, as it results in a gain in steadiness and improved freedom from tape-flutter troubles, at the expense of some loss of high-frequency signal response.

The value of bias current to be used is usually taken as that which gives maximum output of a 1,000-c/s sine-wave for a total harmonic distortion not exceeding 2 per cent.

Bias frequency may be the same as that of the erasing current, or may be a multiple of it. It should be as high as can be conveniently handled by the iron of the head. Random relationships between bias and erase frequencies are undesirable, as unless the two are locked together, drift may occur, and beat frequencies may be set up between harmonics, which sometimes appear as whistles on the tape.

The use of ferro ceramic material for record heads has not proved satisfactory, as it is virtually impossible to produce or maintain a sufficiently sharp straight edge to the gap.

Signal

The signal currents which are to be recorded are fed to the recording head from the recording amplifier, and are mixed additively with the bias current. The sensitivity of a tape is defined as the recorded signal level for a given r.m.s. signal flux in the recording head, with optimum bias as given above. It is dependent upon the magnetic characteristics of the tape material, notably various aspects of the permeability. It varies some 10 db between types or brands of tape in current use.

Using a typical 1-in.-diameter ring-head, maximum signal level (defined as that which gives 2 per cent total harmonic distortion for a 1,000-c/s sine-wave signal using optimum bias) requires about 1·25–1·5 ampere turns for a medium-sensitivity tape.

Frequency Response

Assuming no losses, the magnetization of the tape would be independent of frequency for a given signal current in the head. Given a tape of constant magnetization, or surface pole strength, the method of

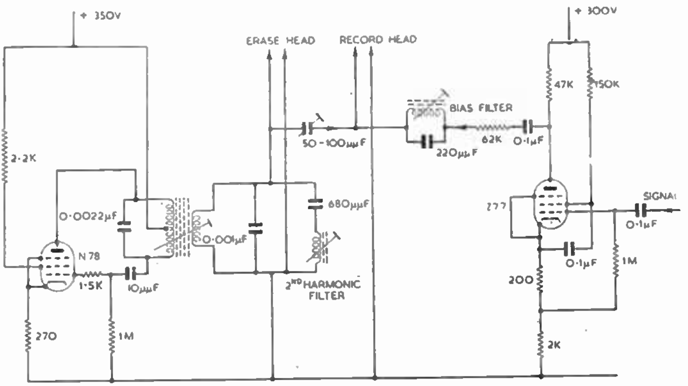

FIG. 4.—OSCILLATOR AND BIAS CIRCUIT OF THE E.M.I. TYPE TR50 TAPE RECORDER.

Circuit component values are determined by head losses and inductances.

reproduction by means of a gapped ring-head would result, in the absence of losses, in an open-circuit output voltage across the head terminals related to wavelength by the expression :

$$E_0 \propto N \sin \pi \frac{l}{\lambda} \quad . \quad . \quad . \quad . \quad (1)$$

where E_0 is the open-circuit output voltage ;
N is the number of turns in head winding ;
l is gap length
λ is recorded wavelength } in the same units ;

and the function $\pi \frac{l}{\lambda}$ is expressed in radians.

When the wavelength is long compared with the gap length the output voltage is very closely proportional to frequency for a given linear tape speed :

$$E_0 \simeq Kf \propto \frac{1}{\lambda} \quad . \quad . \quad . \quad . \quad (2)$$

giving a characteristic which rises with frequency at the rate of 6 db/octave.

Frequently the difference between expression (1) and expression (2) is considered as a "gap loss". Here, expression (1) is taken to be inherent in the reproducing system, and only deviations from this curve are considered as losses.

In practice, losses arise in both recording and reproduction, and in the tape itself. They are divisible into two main groups, those governed by frequency and those governed by recorded wavelength.

The former are easily overcome, in normal audio-frequency recording, by suitable design. The latter are a function of the linear speed of the tape, being most troublesome for low speeds.

Using the so-called high-coercivity tapes, an upper cut-off frequency of V kc/s is considered good practice, where V is the linear velocity of the tape in inches per second. With low-coercivity tapes, about half this figure is normal. Considerably higher frequencies can be recorded and reproduced, but at the expense of signal-to-noise ratio.

High-frequency Losses

The principal sources of high-frequency losses are as follows :

(1) electrical circuits of the recording amplifier and its connection to the recording head, resulting in variation of head current with frequency [F];

(2) eddy currents in the recording head [F];

(3) irregularities in the recording-head gap [W];

(4) imperfect contact between the tape and the recording head [W];

(5) part-erasure of a recorded half-wave, due to the complementary reverse half-wave occuring while the former is still in the proximity of the gap [W];

(6) self-demagnetizing action of the tape [W];

(7) imperfect contact between the tape and the replay head [W];

(8) irregularities of the replay-head gap [W];

(9) eddy currents in the replay head [F];
(10) coupling between the replay head and the replay amplifier [F];
(11) electrical circuits of replay amplifier [F].

The letter [F] indicates dependence on frequency, and the letter [W] dependence on wavelength.

Items (1) and (11) should not arise if the amplifier has been well designed. Signal currents should be fed to the recording head through a filter which prevents the bias current from getting back into the recording amplifier. Similarly, the bias supply should have a filter to prevent the signal from reaching the bias generator and modulating it. Care should be taken that these filters do not introduce loss of high signal frequencies.

Items (2) and (9) can be reduced by using thin, insulated laminations for the heads, and by choosing a form of mounting which does not behave as a short-circuited turn around the magnetic circuit.

Items (3) and (8) have already been dealt with under "Gap Dimensions".

Items (4) and (7) have also previously been mentioned. The loss of high frequencies due to poor contact between tape and head is a major factor. An empirical formula quoted[7] for this loss is:

$$A = 55 \log \frac{S}{\lambda}$$

where A is the loss in decibels;
S is the effective spacing between tape coating and head; and
λ is the recorded wavelength, in the same units.

Item (5) may have considerable effect on the frequency-response characteristic. When considering the action of the bias in linearizing the input/output characteristic, it was assumed that the signal current remained substantially constant. However, in Fig. 3, if the signal current is reduced or reverses before the medium has passed out of the leakage field, its magnetization will be stabilized at a lower value than would otherwise have been the case. This is significant at very short wavelengths.

Item (6) is of greater importance with low-coercivity tapes. At short wavelengths each recorded half-wave can be considered as a very short magnet, opposed at either end by magnets of reversed polarity. A large part of its available flux is therefore prevented from appearing as surface induction on the tape, and thus influencing the replay head.[4]

Item (10) is a matter of audio-frequency circuit design, but the problems are somewhat unusual and deserve some consideration.

Replay Head Coupling

A typical 1-in. reproducing ring-head of the type shown in Fig. 2, when excited by a typical tape sample of average remanence, recorded with a sine wave to the maximum-permissible level, to give 2 per cent distortion, has an open-circuit output voltage of the order of:

$$E_0 = 0{\cdot}006\, Nf \text{ microvolts (approximately)} \qquad (3)$$

where N is the number of turns on the replay head winding; and
f is the frequency in c/s, when wavelength $\gg$ gap length.

This can only be approximate, as it depends greatly on the head design, and, in particular, on the dimensions of the back-gap. Reducing the reluctance of the back-gap increases output, but care must be taken that the total reluctance of the iron circuit does not become non-linear.

The inductance of a head such as the above might be about:

$$L = 0{\cdot}028 \times 10^{-6} \times N^2 \text{ henrys (approximately)} \qquad (4)$$

Taking the case of direct coupling, the load into which the head has to work consists mainly of a capacitance comprising:

(a) the self-capacitance of the head winding;
(b) the capacitance of the connecting wiring;
(c) the input capacitance of the first amplifying valve.

In order not to introduce high-frequency losses, this lumped capacitance, C, should resonate with the head inductance, at or about the upper cut-off frequency of the system, f_2. The maximum permissible head inductance is then:

$$L = \frac{10^{12}}{4\pi^2 f_2^2 C} \text{ henrys} \quad . \quad . \quad . \quad (5)$$

where f_2 is in c/s, and C is in picofarads.

From (4) $N = 6{,}000\sqrt{L}$ (approximately) where L is in henrys (4a)

or, to a rough approximation,

$$N = \frac{10^9}{f_2\sqrt{C}} \text{ turns maximum} \quad . \quad . \quad . \quad (6)$$

Hence at the lower cut-off frequency f_1 the voltage to be expected on the first amplifier grid is given roughly by equation (3)

$$E_0 = 0{\cdot}006\ Nf_1\ \mu\text{V}$$

Substituting for N from equation (6) results in

$$E_0 = \frac{6 \times 10^6}{\sqrt{C}} \cdot \frac{f_1}{f_2}\ \mu\text{V} \quad . \quad . \quad . \quad (7)$$

Assuming a lower cut-off frequency of 50 c/s and an upper cut-off frequency of 10,000 c/s, and estimating a value of 40 pF for C, the value of E_0, on maximum signal, is of the order of

$$E_0 = \frac{6 \times 10^6 \times 50}{6{\cdot}3 \times 10{,}000} = 4{,}800\ \mu\text{V, or, say, 5 mV. at 50 c/s.}$$

If a 60-db signal-to-noise ratio is to be maintained, the level of hum and microphony, referred to the grid of the first valve, must be kept below 5 μV.

The gap-length might be arranged to be equal to a half-wavelength at, say, 8,000 c/s, in which case the output at this frequency, assuming no losses other than the inherent gap-loss, would be

$$E_0 = \frac{2}{\pi} \times \frac{6 \times 10^6 \times 8{,}000}{6{\cdot}3 \times 10{,}000}\ \mu\text{V, or 0·5 volts approximately.}$$

This voltage, acting in the resonant circuit formed by L and C, and controlled by such damping as may be present in the head, or added, may be used to compensate partly for losses, but care must be taken that the voltage appearing at the valve grid is not sufficient to cause appreciable distortion. The use of selective negative feedback, incorporating the greater part of the necessary basic equalization, at this point can be very helpful.

A suitable transformer may be interposed between the head and the first-stage valve. In this case, the primary reactance must be kept large compared with the impedance of the head at all frequencies, and the secondary self-capacitance must be included in the estimation of C.

Bass Response

There is an effect associated with the longer wavelengths which is often ignored, but which can become troublesome on high-quality equipment. Consider a head which is square, rather than the more usual round form. Assuming no reluctance in the iron, when the length of tape in contact with each pole face is less than an integral number of wavelengths by the length of the gap, the effective magneto-motive force acting round the iron circuit will be zero. The output would be zero at certain critical wavelengths, and doubled at intermediate wavelengths.

In practice, the tape's approach to, and departure from, the pole face is not sudden, as here indicated, but gradual, with considerable fringing of the flux, and complete cancellation does not occur. However, with most types of head, careful frequency-response measurements at intervals of a few cycles per second, up to a few hundred cycles per second, will reveal a series of peaks and hollows, sometimes amounting to several decibels in magnitude.

A similar effect is sometimes found in connection with the magnetic screen placed around the replay head. At certain wavelengths flux induced in the screen by the proximity of the tape in turn induces flux in the head, which tends to cancel or enhance the output.

These effects are much more severe in wire recorders than in tape recorders, due to the relatively sudden entry of the wire to the head and its equally sudden exit. Tapering the entry and exit tunnels alleviates the trouble.

Equalization

In the early days the equalization required to compensate for the various losses in frequency response was distributed in a rather random manner between the recording and reproducing amplifiers, at the whim of the designer. This resulted in the production of recorded tapes which were not interchangeable between one make of machine and another. Within a group, a so-called "standard" tape could be made on one machine, having bands of tone recorded at various selected frequencies, at constant electrical input level to the machine, or, perhaps, at constant signal current to the recording head. This could be used to line-up the replay equalizers of all other machines in the group, and subsequently the recording amplifier equalizers could be adjusted so that each machine had the required overall response for the particular type of tape in use.

Standard Tapes

It is obviously desirable that the levels of the tones recorded on this "standard" tape bear a known relationship to some absolute value, independent of frequency, a good starting point being to maintain constant surface pole strength on the tape. Unfortunately this has proved extremely difficult to measure, although laboratory methods have been developed which offer a fair degree of accuracy and agreement.[8]

If facilities are not available for absolute measurement, sub-standard tapes can be obtained which have been checked against the standard. These have bands of tone recorded on them at various known frequencies and levels, and using such a tape, the replay amplifier can be equalized and adjusted to give the required replay characteristic.

If the sub-standard tape has all frequencies recorded at constant surface pole strength, and the replay output characteristic is required to be "flat", the basic amplifier equalization will be of the form

$$\frac{1}{\sin \pi \frac{l}{\lambda}}$$

where l is the effective gap-length, and λ is the recorded wavelength, with the addition of a few decibels of high-frequency lift to compensate for losses in the replay head.

The C.C.I.R. has commended an international standard replay characteristic[9] which is not flat, but which includes post-emphasis in the form of an additional "top-rise" having the characteristic of an electrical circuit of 35 microseconds time-constant.*

This is equivalent to the current through a parallel combination of capacitance and resistance when fed from a constant-voltage source, the time constant of the combination, CR, being equal to 35 μS.

Standard tapes are now usually made to the C.C.I.R. recommendation, which means that the higher frequencies are de-emphasized (i.e., reduced in amplitude) in accordance with a 35-microsecond characteristic. Such a tape, replayed through a correctly equalized replay channel, will give an output which is constant with changing frequency, within the limits of the apparatus.

These tapes are available commercially. Full details of the characteristic and the method of measurement are contained in Amendment No. 1, published 1954, to B.S. 1568 : 1953.

Equalization of Recording Amplifier

Having completed the equalization of the replay amplifier, the recording-amplifier equalization can be adjusted until the overall response of the equipment is as flat as required. The basic characteristic is that of constant current through the recording-head winding, for constant input to the amplifier. In addition, some high-frequency "lift" will be needed to compensate for losses in the recording head and in the tape. The amount, and the shape of curve required, will

* This refers to tape speeds of 30 in./ and 15 in./sec. For 7½ in./sec. the corresponding figure is 100 μS.

vary according to the type of tape in use, but unless the tape speed is too low for the frequency response required, this should not be more than a few decibels at the upper cut-off. If working to C.C.I.R. standards, an additional top cut is required, having a 35-microsecond characteristic. This is the inverse of the top-lift circuit in the replay amplifier, being equivalent to the voltage across a parallel combination of capacitance and resistance having a time constant CR of 35 microseconds when fed with constant current.

The tape is much more susceptible to overload at the higher frequencies,[4] and the purpose of the C.C.I.R. 35-microsecond characteristic is to reduce the amplitudes of these high frequencies on the recording, in order to avoid distortion.

Distortion

The degree to which A.C. bias is able to linearize the transfer characteristic of a tape depends upon the magnetic properties of the material and upon the amplitude of the bias flux. Low values of bias may materially reduce the distortion at high signal levels, at the expense of high distortion for signals of medium level. Generally it is desirable to adjust the bias to a value at which the distortion rises slowly with signal level, until, at the overload point, the rise in distortion becomes rapid. The overload point is usually taken as the signal level at which the distortion reaches 2 per cent total harmonic, an amount which, on a 1,000-c/s tone, is easily heard by comparison with a pure note at the same level, but which is not very noticeable on music.

As the magnetic characteristic is symmetrical, the major distortion then takes the form of peak-clipping on both positive and negative half-cycles, resulting in the production of odd harmonics, and, if more than one frequency is present, of second-order inter-modulation products.

Owing to the self-demagnetizing effect at short wavelengths, the overload point is considerably lower in level at high than at medium and low frequencies. Typically, the overload point at 10 kc/s might be 10 db lower than that at 1 kc/s. The harmonics arising from the resultant distortion do not appear in the output, owing to factors limiting the frequency response. Second-order intermodulation products, however, arising from a two-tone signal having frequencies f_a and f_b will be found at frequencies of $(2f_a - f_b)$ and $(2f_b - f_a)$, and may be used to measure the distortion.[10] Even harmonics, and first-order intermodulation products, at frequencies $2f_a$, $2f_b$, $(f_a \pm f_b)$ do not arise, and the proposed C.C.I.R. method for measuring intermodulation distortion, which takes account only of first-order products, is unsuitable for this application.

If even harmonics or first-order intermodulation products are found in the output, they must be due to :

(a) the tone source;
(b) the amplifiers;
(c) asymmetry of the bias waveform; or
(d) accidental magnetization of one or more of the heads.

The servicing of domestic tape recorders is described in Section 39.

Modulation Noise

Modulation noise is a form of distortion which is noticeable in some types of recording. It takes the form of a hissing or grunting noise, which rises and falls with the envelope of the signal. It has been suggested, with the backing of considerable evidence, that variations of contact between tape and head, due, for example, to roughness of tape or heads, or to mechanical vibration in the machine; cause the signal level to be modulated at the interfering frequency. This modulation takes the form of sidebands on either side of the signal frequency, and, being enharmonic to the signal, they are heard as noise. Slight over-biasing tends to reduce the effect to a small degree, but its elimination must be dependent upon the reduction of all mechanical irregularities, both in the tape and in the driving mechanism.

It is exaggerated by slight magnetization of the replay head (as is normal background noise), and some types of tape are more susceptible than others.

Editing

Tapes may be cut and joined for editing purposes without noticeable effect in the form of noise. Sections of a length up to about $\frac{1}{40}$ second of playing time may be cut out completely without the loss being noticeable. It is advisable to make the cut in a note of sustained pitch, in a section of complete silence, or at the start of a sharp, transient sound, for best results.

The cut should be made at an angle of between 45° and 70° with the tape axis, so that its passage over the heads is relatively gradual. A non-magnetized cutting tool must be used.

A lap joint may be made, using suitable cement, but a better method is to make a butt joint, at an angle, the backs of the tapes being fastened with a patch of adhesive tape manufactured for the purpose, the patch being subsequently trimmed to the width of the tape.

If a jointed tape is erased and re-recorded upon, the joint should be completely unnoticeable on replaying.

Print-Through

One of the troubles which beset tape and wire recording is the problem of print-through. Any magnetic material will be magnetized by being placed in proximity to another magnet, and each particle of a recorded tape or wire is susceptible to magnetization by adjacent particles. Strong signals on a tape recording will be printed through on to adjacent layers in the reel, producing pre- and post-echoes. In extreme cases ten or twelve repetitions have been noted on either side of a strong signal, printed through as many successive layers. High-coercivity materials, recorded at low levels, are naturally less susceptible.

Print-through has been found to be a function of temperature,[11] increasing rapidly at temperatures above about 65–70° F. Tapes should therefore be stored in a cool place.

It is also a function of time, rising fairly rapidly during the first few hours of contact, and then more slowly. There is also some reduction of the printed signal during the first hour or so after separation, so that it is advantageous to rewind the tapes occasionally, and, in particular, an hour or so before they are required to be replayed.

Proximity to stray fields, for example the leakage fields from loudspeaker magnets or mains transformers, must be rigorously avoided, both on account of print-through and of distortion, particularly of the higher-frequency waveforms.

Standards

The standard width for recording tapes is ¼ in. and thickness 0·002 in.

The standard linear speed for professional recording purposes is 30 in./second, with sub-standard speeds of 15 in./second and 7½ in./second. It is recommended that, when slower speeds are required for domestic or other purposes, further sub-multiples of 30 in./second should be employed.

For tolerances, details of spools, and other details, reference should be made to B.S. 1568 : 1953.

For 30 in./sec. and 15 in./sec., the standard recording characteristic shows a "top" fall, and the standard replay characteristic a top rise, of 35 microseconds time constant. For 7½ in./sec., the standard is 100 microseconds time constant. For 3¾ in./sec. working, no standard has been laid down but the generally accepted characteristic has a time constant of 200 microseconds, in order to avoid high-frequency overload. It is evident that full frequency range can be obtained at the lower tape speeds only at the expense of reduced signal-to-noise ratio or increased distortion, with current tape-coating materials.

Standard track positions for magnetic-tape records are given below:

Monaural Tape Records

Spool. 7 in. (17·8 cm.) plastic; standard domestic fitting; dimensions within B.S. specification. Tape wound with oxide in.

Recorded Tracks (Two). Separation between tracks, 0·030 in. (0·76 mm.) Track width (each track), 0·0110 in. (2·795 mm.).

Dimension of Recommended Replay Scan. Track separation, 0·050 in. (1·27 mm.), track width, 0·090 in. (2·29 mm.): margin from edge of tape, 0·010 in. (0·25 mm.).

Winding Sense. The direction of recording will be such that with the take-off spool (magazine of tape record) unwinding in an anti-clockwise direction (active or oxide surface facing the centre of the spool) the top track is operative and called the first track. The lower or second track is recorded in the opposite direction. Leader strips will be fitted to the beginning of the tape indicating the title and record number of track one. The printing will be the right way up when track one is at the top ready for playing. A title strip will also be fitted for track two, and the printing will be so arranged that it will be necessary to turn over the spool to read and bring track two to the top. Spool turnover is unnecessary for machines fitted with track-reversal devices.

Recording Characteristic. This will be to the C.C.I.R. standard for 7½ in./sec.

Stereo Tape Records

Spool. 7 in. (17·8 cm.) plastic: standard domestic fitting; dimensions within B.S. Specification.

Magnetic Tape. Tape wound oxide in.

Recorded Tracks (Two). Separation between tracks, 0·030 in. 0·76 mm.). Track width (each track), 0·110 in. (2·795 mm.).

Dimensions of Recommended Replay Scan. Track separation, 0·050 in. (1·27 mm.); track width, 0·090 in. (2·29 mm.); margin from edge of tape—0·010 in. (0·25 mm.).

Winding Sense. The direction of recording will be such that the take-off spool (magazine of tape record) will unwind in an anti-clockwise direction during reproduction, active or oxide surface facing the centre of the spool. A leader strip will be fitted to the beginning of the tape indicating the title and record number. If the tape moves from left to right and with the active side facing away from the observer, the top track shall be designated No. 1 track and the bottom one No. 2 track. In stereophonic applications, No. 1 track shall carry the recording for the left-hand channel as viewed from the audience, and No. 2 track shall carry the recording for the right-hand channel. The tracks shall be recorded with the head gaps in line

Recording Characteristic. This will be to the C.C.I.R. standard for 7½ in./sec.

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E. W. B.-J.

DISC RECORDING AND REPRODUCTION

Although Edison held an earlier patent for the disc record, the system of recording on discs was developed by Emile Berliner, in 1887, with the object of overcoming the difficulties of duplication associated with the cylindrical records of the Edison Phonograph. Owing to the ease and economy of producing an almost unlimited number of copies, the system became—and still is—the most popular for the commercial production of recordings for domestic and other uses. Secondary advantages include easy storage and the facility of being able to select any passage at will. A fundamental disadvantage is the change of linear speed with working radius.

The original recording was done, until recent years, on wax, either in the form of a solid circular slab, about an inch in thickness, or of a coating of about 0·015 in. thickness on a circular plate of ¼-in. glass. Modern practice usually employs a so-called "lacquer" disc, this being an aluminium base, of about 0·036 in. thickness, coated to a thickness of 0·006 in. with a lacquer of cellulose nitrate, combined with suitable plasticizers.

Recording

The recording process consists of rotating the disc about its centre at a constant rotational speed, on a machine which is essentially a horizontal surface lathe, and of cutting upon its surface a spiral groove by means of a stylus mounted in the cutter head, which is moved radially over the surface of the disc.

Rotational speeds of 78, 45 and 33⅓ r.p.m. are now standardized. The pitch of the spiral may lie between about 70 and 350 grooves per inch.

At the same time, the stylus is itself oscillated in a direction radial with respect to the disc, in such a manner as to represent the waveform of the signal.

If the no-signal spiral were developed into a straight line, the waveform engraved would be, subject to certain deliberately introduced frequency distortions which are corrected in the reproducer, the integral of the input waveform.

TABLE 2.—PRINCIPAL DIMENSIONS OF COARSE AND MICROGROOVE RECORDS

Nominal speed (r.p.m.)	78	78	33⅓	33⅓	45
Nominal size (in.)	12	10	12	10	7
Outside diameter (in.)	11⅞	9⅞	11⅞	9⅞	6⅞
Diameter of outermost groove (in.)	11½	9½	11½	9½	6⅝
Diameter of innermost groove (in.)	3¾	3¾	4¾	4¾	4⅞
Centre hole diameter (in.)	0·286	0·286	0·286	0·286	1½ or 0·286

A lacquer original is suitable for direct playback using a lightweight pick-up, or it may be processed to obtain copies.

Processing

The first step in processing is to render the surface electrically conductive. In the case of wax this may be done by several methods. In one method, silver vapour, produced by heating silver to its melting point in a vacuum chamber, is permitted to condense on the surface of the wax. In another, a potential of some 2,000 volts is set up between a gold cathode and the wax, in vacuum, and a discharge takes place which causes a thin film of gold to be sputtered on to the wax.

With lacquer, a layer of silver is chemically deposited on the surface from a solution of ammoniacal silver nitrate by the action of a reducing agent.

The disc is then mounted by its edge in an electrode assembly, and becomes the cathode in a series of plating baths, being electroplated first with nickel, then copper, to a mean thickness of about 0·035 in. The whole of the electrodeposition is then stripped away from the original, which is usually destroyed in the process. This is known as the "Master", and is an inversion of the original, having ridges in the place of grooves.

To this Master, a separator coat of potassium dichromate is applied, and the electrodeposition and stripping process is repeated, producing a "Mother", which is an exact copy in metal of the original. If a very large number of pressings is expected, several "Mother" shells may be made.

After applying a separator coat, the electrodeposition and stripping process is again repeated, often a number of times, producing several "inverted" copies, or "stampers". The stampers are themselves placed in an electroplating bath, and given a very thin surface of chromium, to give added resistance to wear. They are then soldered flat on to a backing plate of copper, and after trimming the edge and locating the centre hole, are ready for the press.

Two stampers are mounted in the record press, with labels in position, one in the upper and one in the lower platen. Steam is admitted behind the stampers, which raises them above the softening temperature of the record material. A preheated "biscuit" of this thermo-plastic material is placed between the stampers, and the press is closed under pressure, which causes the material to flow into the space between the stampers, being moulded over every ridge of the waveform, and forming an exact copy of the original recording.

Admitting cold water behind the stamper cools the moulding and and causes it to set solid, when the press may be opened and the record removed. The complete pressing cycle takes less than two minutes. After trimming and smoothing the edge, the record is ready for packing and despatch.

The original waveform is maintained throughout this processing within a very close tolerance, any measurable variation being invariably due to abrasion, caused by, say, wear on the stampers, or excessive polishing of the metal shells with a view to improving the appearance of the finished product. It must be appreciated that the amplitude of a 10-kc/s wave, for example, at only 20 db below peak recording level, a readily audible signal, is less than half a micron, invisible through an optical microscope. Yet such a wave will impart to the reproducer stylus tip a peak lateral acceleration of more than 170 times the gravitational acceleration.

The biscuit which forms the pressing has been for many years compounded of shellac, with various fillers and colouring matter. Modern techniques employ, for some types of record, Vinylite, Geon or similar thermo-plastic material. These take advantage of modern pick-up design to give improved quality with lower background noise on playback.

Playback

Playback from the record is carried out by rotating it at the same speed at which it was recorded. The turntable on which the record is placed should revolve at a rate which is maintained steady within better than 0·1 per cent, and must be free from vibration, which would be reproduced as a low rumble.

A pick-up head carries a stylus having a hemispherical tip, which rests in the V-section groove cut by the recording stylus, so that it is gripped and located by the sides of the groove. The pick-up is usually mounted on an arm which is free to rotate about a centre some distance from the record centre, and is free to move only in an arc which is approximately radial with respect to the record. The pure spiral component of the groove displacement, acting through the stylus, thus causes the pick-up to track substantially radially across the surface of the disc, following the motion of the recording head.

At the same time the stylus point is made to oscillate from side to side, following the wave engraved on the disc, but, due to its inertia, the pick-up head cannot follow these rapid oscillations. There is therefore relative motion between stylus and head, which can be utilized to generate a proportional e.m.f. by electromagnetic or other means. This e.m.f. may be amplified and used to reproduce the original signal.

An alternative system, following the original Phonograph, utilizes vertical oscillations of the stylus tip, and is known, from the groove formation, as "Hill and Dale" recording. This has certain inherent disadvantages, and has become virtually obsolete.

"Hill and Dale" recording has now been revived as a component in stereo-disc recording. The disadvantage of high tracing distortion has been greatly reduced by the employment of a reproducing stylus having a tip radius of only 0·0005 in.

Recording Machine

The recording machine or lathe carries the recording blank on a turntable which must be rotated at a perfectly even speed within 0·1 per cent or better. This normally implies that the turntable shall be heavy (up to 50 lb. or more) and that the bearings shall be of impeccable quality. Since the oscillations of the groove are so extremely small, vibration of the machine due to the drive must be rigorously avoided. The actual power required to drive a recording turntable against friction and the drag of the cutter is of the order of 5–10 mechanical watts. However, a considerably more powerful motor is needed in practice, to give a reasonable run-up time, and to maintain steady speed.

In some types of machine, the cutter head is tracked radially over the disc, by means of a very accurately cut lead-screw. In others the recording head remains stationary, and the turntable is itself moved.

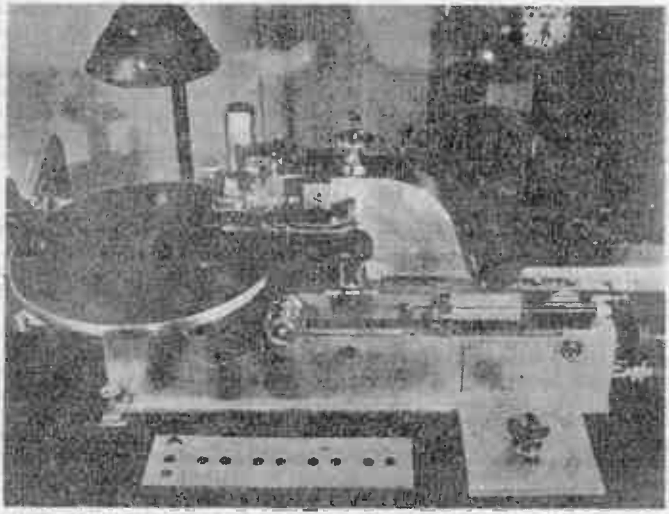


FIG. 5.—THE SCULLY RECORDING LATHE.
(Scully Manufacturing Co.)

In order to allow for imperfect flatness of the surface of the recording blank, which is often clamped to the turntable by both centre and edge, the cutter head is "floated". It is suspended on knife-edges and slightly unbalanced so that there is some small weight on the stylus, sufficient to cause it to cut into the blank to a pre-determined constant depth. Sometimes an "advance ball" is used, resting on the surface of the disc, to control the depth of cut.

In some of the latest designs a servo-mechanism is included to control the depth.

The cutting stylus has a chisel point of the shape shown in Fig. 6, and cuts a groove of the form shown in Fig. 7.

The rotational speed of 78 r.p.m. appears to have been arrived at quite arbitrarily, 75 r.p.m. being thought to be a little too slow, and 80 r.p.m. a little too fast. Groove dimensions in the early days appear to have been equally arbitrary, and grooves of widely differing sections are to be found among earlier recordings. Modern improvements in pick-ups have made possible much shallower and more closely spaced grooves, and introduced the now well-known "microgroove" techniques, requiring much smaller pick-up stylus points. This has demanded virtually a revolution in disc recording and reproducing equipment, and has cleared the way for a complete re-design of the record dimensions from an engineering standpoint, aiming at the best compromise between maximum level, background noise, distortion, and playing time. This has resulted in a rotational speed of 33⅓ r.p.m. for long-playing microgroove records, with 45 r.p.m. for a 7-in.-diameter record of medium playing time.

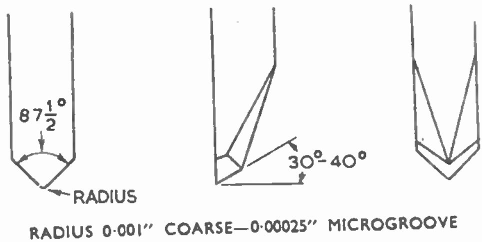


FIG. 6.—RECORDING STYLUS.

Groove pitches used on standard 78-r.p.m. records vary between 70 and 120 grooves per inch. For microgroove records the range is about 200–300 grooves per inch. In order to get even longer playing time from long-playing records, without sacrificing quality, variable groove spacing has been employed, utilizing, say, 350 grooves per inch for very quiet passages, where amplitudes of groove excursion are small, and opening out to 150 or 200 grooves per inch on forté passages with increased amplitudes. The control of groove spacing is now usually automatic, being governed by the sum of the groove width and wave peak amplitude, with a small margin for safety.

Recording Characteristics and Levels

If the peak amplitude of a record sine-wave is given by $\widehat{A}$, and ω is the angular velocity, equal to $2\pi f$ where f is the frequency, the instantaneous amplitude is given by a, where

$$a = \widehat{A} \sin \omega t \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The instantaneous velocity of the stylus tip is given by v, where

$$v = \frac{da}{dt} = \widehat{A}\,\omega \cos \omega t \quad . \quad . \quad . \quad (2)$$

and the acceleration by :

$$g = \frac{dv}{dt} = \frac{d^2a}{dt^2} = -\widehat{A}\,\omega^2 \sin \omega t \quad . \quad . \quad (3)$$

Peak values are denoted by $\widehat{A}$, $\widehat{V} = \widehat{A}\omega$, and $\widehat{G} = \widehat{V}\omega = \widehat{A}\omega^2$; r.m.s. values are denoted by A, $V = A\omega$, and $G = V\omega = A\omega^2$.

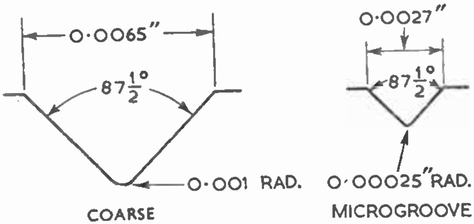


FIG. 7.—GROVE PROFILE.

If a coarse groove width of 0·0065 in. and a coarse groove spacing 0·0100 in. is assumed, there will be 0·0035 in. land between grooves, permitting safe amplitude of, say, 0·003 in., or 0·008 cm., provided two half-waves of opposite phase do not occur in juxtaposition in successive grooves. At 15 kc/s, recorded to this amplitude, the peak velocity would be about 800 cm./second. At the minimum radius, recorded on a 78-r.p.m. disc, the linear speed of the groove past the stylus tip is only about 40 cm./second, so that the maximum angle which the groove wall makes with the direction of travel is tan 800/40, or about 87°. The vertical slope of the groove wall is only 45°, so the stylus would ride over the top, rather than follow the groove. In any case, the groove waveform degenerates into a series of cusps, and the distortion would be intolerable, so that a system of "constant amplitude" recording, in which amplitude is proportional to signal input, is quite impracticable.

If the velocity is maintained proportional to input, the amplitude is inversely proportional to frequency, and if a reasonably high velocity is to be maintained at medium and high frequencies the low-frequency amplitudes will be too large for the groove spacing. A compromise is therefore effected, the characteristic laid down by Maxfield and Harrison being "constant velocity" above 250 c/s, and "constant amplitude" below. The transition is not sharp, being in fact a curve having the same shape as the curve of impedance of an electrical resistance–inductance parallel combination, with a time constant of 637 microseconds. The asymptotes to this curve intersect at 250 c/s, and this is known as the "turnover frequency".

Some manufacturers used a turnover frequency of 300 or 350 c/s, time constant 530 or 455 microseconds, while other manufacturers used 500 c/s, time constant 318 microseconds.

In 1955, an international standard was adopted, and incorporated in B.S. 1928 : 1955. This defines the recording characteristics as the frequency response relationship between the recorded waves and the electrical input to the monitor loudspeaker. The standard recording characteristics are the resultant of adding algebraically the decibel responses of three curves, having the following characteristics:

(*a*) One rising with frequency in conformity with the impedance of a series combination of an inductance and a resistance having a time constant L/R of t_1.

(*b*) One rising with frequency in conformity with the impedance of a parallel combination of inductance and resistance having a time constant L/R of t_2.

(*c*) One falling with rise of frequency in conformity with the impedance of a series combination of a capacitance and a resistance having a time constant C–R of t_3

For Coarse Groove Records	*For microgroove Records*
t_1 = 50 microseconds	t_1 = 75 microseconds
t_2 = 450 microseconds	t_2 = 318 microseconds
t_3 = 3,180 microseconds	t_3 = 3,180 microseconds

Exact equalization can be obtained in the reproducer by using the inverse of these characteristics.

Levels on the record are always measured in terms of the r.m.s. lateral velocity, expressed in cm./second, or in decibels above or below a reference of 1 cm./second r.m.s. (sometimes referred to as dBc, by analogy with dBm).

It is very desirable to equalize for the recording characteristic in one part of the replay system, and to adjust for the studio balance, for the listening conditions and for personal taste, with an entirely separate set of tone controls.

Recording Heads

The recording head must be designed to drive the stylus point at amplitudes up to, say, 0·02 cm. in the bass frequencies, and velocities up to, say, 30 cm./seconds r.m.s. above about 500 c/s, without noticeable amplitude distortion, and without undamped resonances in the frequency characteristic. The impedance offered by the disc to lateral motion of the cutting stylus varies with the linear speed, and hence with the working radius. The mechanical impedance of the stylus tip itself must be made high, in order to swamp these variations, but this impedance will be low at the natural resonance. Hence, either the natural resonance must be placed outside the recorded frequency range or heavy damping must be applied to control the resonance. The latter usually consumes power which must be supplied by the driving amplifier. In the former case the resonance can be raised by making the moving parts much smaller, but this lowers their power-handling capacity. Alternatively, the stiffness can be increased, requiring more driving power, again with danger of electrical overload.

If the resonance is to be lowered without reducing the mechanical impedance, this must be done by increasing the mass of the moving parts, requiring more driving force at high frequencies.

Very satisfactory designs have been produced using a damped intermediate resonance. One moving-iron type, having a frequency response extending from 30 to 20,000 c/s, requires only the output from a 6-watt amplifier, via a correcting equalizer, to record full commercial disc levels with less than 1 per cent distortion. Other moving-coil types, in which the electrical and mechanical parts are inherently tightly coupled, utilize the low output impedance of the amplifier to provide electrical damping without waste of power.

The usual method is to derive an electrical voltage from the recording-stylus motion, and to feed this back inversely to an early stage in the recording amplifier as " negative feedback ". Very careful control of phase-shifts is necessary.

Moving-coil type recorders operate on the well-known d'Arsonval principle, with the stylus rigidly attached to the coil to avoid spurious secondary resonances. An alternative type has the coil in the form of a single-turn loop mounted on a vertical axis and excited by transformer action from fixed speech coils. A horizontal cantilever arm connected to this coil, and carrying the stylus on its outer extremity, converts the rotational to an oscillatory motion.

Moving-iron types have the stylus rigidly fixed to an armature, one end of which is free to oscillate between the poles of a fixed magnet. A coil surrounds the armature, and currents in the coil induce a magnetic pole of alternating polarity in the free end. The latter is therefore

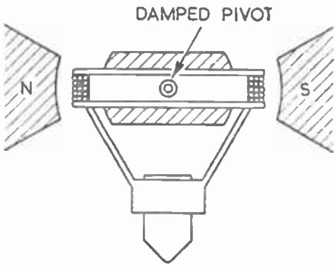


FIG. 8.—MOVING-COIL SYSTEM.

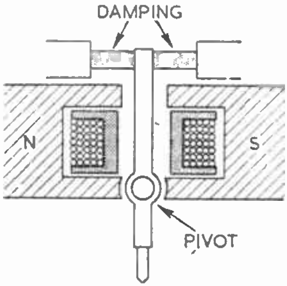


FIG. 9.—MOVING-IRON SYSTEM.

attracted first to one pole of the fixed magnet and then to the other, producing oscillatory motion of the armature. Adequate air-gaps must be provided to linearize the iron circuit with the signal currents used, and a mechanical restoring force is necessary, tending to maintain the armature tip in the centre of the gap. Adequate damping must be incorporated to control the resonance between this force and the armature inertia. Provided that the fixed magnet field is large compared with the signal flux in the air gaps, and provided that the pole tips and armature tip do not approach magnetic saturation, very low distortion can be achieved with this type of recorder, owing to the differential action.

Hot Stylus

It has been found[1] that when cutting lacquer, heating the stylus tip to a temperature in the region of 150–160° F. improves the cutting, and helps to avoid a tendency to throw up " horns ", or ridges, on each side of the groove, which interfere with stripping during processing. The working temperature of the stylus tip is very difficult to measure, and trial-and-error is the best guide. A coil of seven turns of 42 S.W.G. Eureka wire, close wound directly on the shank of a sapphire stylus and cemented in place close to the tip, requires about ¾ ampere. Care must be taken that the swarf does not ignite.

Swarf removal is best carried out by suction, a trap box being used to collect the waste, which, in the case of lacquer, is dangerously inflammable.

Unless some precautions have been taken during the manufacture of the disc, some trouble may be experienced due to the very high static charges built up on the swarf during cutting, which tend to make it adhere to the disc surface.

Pick-ups

There are three types of pick-up in common use. The moving-iron is essentially the same as the moving-iron recorder with its function inverted. The moving-coil is again based on the d'Arsonval principle, working in reverse as an electrical generator; the ribbon type, occasionally seen, is essentially a moving-coil pick-up having a single-turn coil. The third main type, which is widely used, is the crystal, in which the piezo-electric properties of a crystal such as Rochelle Salt are utilized to develop an electrical potential from a mechanical strain.

The moving-iron-type has usually the highest power output, with the moving-coil type second, and the crystal the lowest output. However, the crystal more nearly matches the impedance presented by the grid of an amplifying valve, and in the absence of matching transformers for the alternative types will usually give a higher voltage at the grid.

There is little to choose between the types as regards the transducer itself. Crystals are sometimes subject to temperature variations, but the difficulty of changes with humidity has now been overcome. Amplitude distortion due to transducer action is, in any good design, negligible compared with tracing distortion.

The principal differences lie in the linking mechanism between the record groove and the transducer proper.

The stylus assembly must have some stiffness in its fixing, to maintain it in a mean central position and enable the pick-up to be tracked across the record. There is a resonance between this stiffness and the inertia of the armature and stylus, but when the stylus rests in the groove, this resonance is no longer free, and is of no consequence. However, this same stiffness will resonate with the inertia of the whole pick-up head and arm at a low frequency, and since this resonance is free, it will cause a marked rise in response at and around this frequency.[3] More serious, the mechanical impedance at the stylus point will rise to a high value,[3] and, if the level is high enough, either the stylus will jump out of the groove, or the groove wall will be broken down. Severe wear will inevitably result, even at low and medium levels.

The record material has a resilience or compliance which resonates with the inertia of the stylus and moving parts at a high frequency. This resonance also results in a high mechanical impedance at the stylus point, which will therefore fail to follow accurately in the groove, resulting in severe record and stylus wear and distortion, accompanied by a rise in response around the resonant frequency. The compliance of the modern plastic microgroove records, taking into account the difference in stylus-tip radius, is roughly twice that of the coarse-groove shellac pressing, so the resonant frequency of a given pick-up on a microgroove record will be only about $1/\sqrt{2}$ that applying with a shellac record.

Some pick-ups employ quite complex linkages between the stylus point and the transducer proper, serving the purpose of mechanical impedance-matching transformers. These are liable to have spurious resonances, as is the pick-up tracking arm itself.

The cure for these troubles is twofold—firstly, to keep all important resonances well outside the recorded band-width, and secondly, to provide adequate mechanical damping. If this is done, the open-circuit generated voltage will automatically be linear with respect to frequency, either with reference to recorded velocity (electro-mechanical types) or to recorded amplitude (crystal types). The latter can be converted to a velocity type by differentiating the output, for example by loading the internal capacitance of the pick-up with a resistor.

Further information on pick-ups is given in Section 31.

Reproducing Stylus

The tip of the reproducing stylus which enters the record groove should be of spherical form. A radius of 0·0025 in. will fit almost all the wide range of groove shapes which have been used on different makes of record in the past.[4] For microgroove records, the tip radius should be 0·001 in. The taper from tip to shank is not critical. Any included angle from, say, 25° to 60° is satisfactory.

If the point is to slide round the waves recorded in the groove, without undue stress, it must have a low coefficient of friction, and must

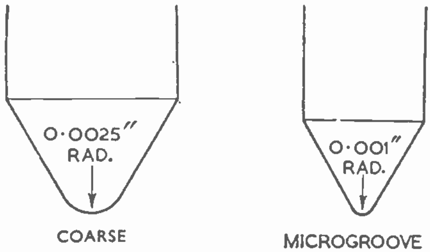


FIG. 10.—REPRODUCING STYLI.

therefore be highly polished. Diamond, sapphire and chromium are suitable materials, having resistance to wear in the order given. Wear of the point results in departure from the spherical form, which in turn leads to distortion and severe record wear. If unpolished, or of the wrong shape, the harder materials will, of course, do the more damage. However, there is no doubt whatever that a well-formed and well-preserved jewel point gives the most satisfactory overall performance.

Tracking and Tracing Distortion

The pick-up swings in an arc about the tone-arm pivot, whereas the recording head moves radially. Over most of the record, therefore, the lateral excursions of the pick-up stylus are forced to take place at an angle to those of the recording stylus, leading to even harmonic distortion. This is more serious with short arms. The effect can be very largely overcome. however, by making the radius of the arc swept by the pick-up point slightly greater than the distance between record and tone-arm pivot centres by an amount known as the " overhang ", and by offsetting the head on the arm at a slight angle, inclined towards the record centre. When playing a record, this results in a drag on the pick-up toward the centre, which offsets—or may even exceed—the friction in the tone-arm bearing. Some smooth friction at this point, in order to damp the tone arm resonance, may, therefore, also be of help in equalising the side thrust on the stylus tip.

It is desirable to keep the tracking error at a minimum toward the centre of the record, where linear speed is lowest and wavelengths therefore shortest. At the outer edge its effects are usually small. The geometry of the situation is complex, but formulæ due to Bauer[5] are given below which enable this form of distortion to be reduced to negligible proportions.

If C is the distance from record centre to tone arm pivot axis; L is the distance from stylus tip to tone arm pivot axis; $D = (L - C) =$ overhang; r_1, r_2 are the radii of minimum and maximum music grooves, respectively, all measured in the same dimensions; and B is the offset angle in radians, then :

$$D = \frac{r_1^2}{L\left[\frac{1}{4}\left(1 + \frac{r_1}{r_2}\right)^2 + \frac{r_1}{r_2}\right]} \quad . \quad . \quad . \quad (4)$$

and

$$B = \frac{r_1\left(1 + \frac{r_1}{r_2}\right)}{L\left[\frac{1}{4}\left(1 + \frac{r_1}{r_2}\right)^2 + \frac{r_1}{r_2}\right]} \quad . \quad . \quad . \quad (5)$$

Tracing distortions are those amplitude distortions which arise due to the lateral excursions of the pick-up stylus point not exactly following those of the recording stylus, with respect to time. One obvious cause is that the weight of the pick-up head, resting on the small area of the stylus tip, itself distorts the groove due to the resilience of the record material. The equally obvious cure is to counterbalance the weight to rest as lightly as possible on the record, the minimum permissible downward pressure being, in general, just in excess of the force required to overcome the maximum lateral impedance of the stylus point at the maximum groove acceleration to be expected. Any resonance must, therefore, be avoided, and, in order to allow the weight to be reduced, the moving parts must be kept extremely light, otherwise the stylus tip will not remain firmly seated in the groove.

Any motion of the needle tip in the direction of the tangent to the mean groove, i.e., " fore and aft ", will introduce phase modulation of the wave, which will produce appreciable distortion.

The axis of the sinuous groove makes an angle θ with the axis of the mean groove (i.e., the groove which would have been cut had there been no signal). The tangent of this angle, tan θ, is the ratio of the instantaneous lateral and linear velocities at that point, and becomes greater toward the centre of the record, where the linear velocity is reduced. If this angle becomes large, the forces on the stylus tip tending to produce " fore-and-aft " oscillation, and to distort the groove wall, become too great, and the distortion rises rapidly. This sets one limit on the safe maximum level which may be recorded.

At a point where the groove is at an angle to the mean groove, e.g., at a point of inflexion where a sine wave crosses the axis, the width of the groove measured radially to the record is normal, determined by the shape of the chisel-edged recording stylus. However, measured as a cross-section of the groove, perpendicular to its own instantaneous axis, it is reduced in the ratio of cos θ. Since the stylus tip is spherical, it must ride up the narrowing V groove, and a vertical component of motion is thus introduced at twice the frequency of the recorded wave. This is known as " pinch effect ". If the pick-up has any response to vertical motion, this output will appear as second-harmonic distortion. This presents a serious problem in stereo records, where the vertical component contributes to the output. Here the reduced stylus-tip radius is helpful.

Alternatively, if there is no vertical compliance of the stylus, having been once forced upwards, it will fail to drop back sufficiently rapidly, and will run round on the outer wall of the wave peak, without making contact with the inner wall, which introduces another form of distortion.

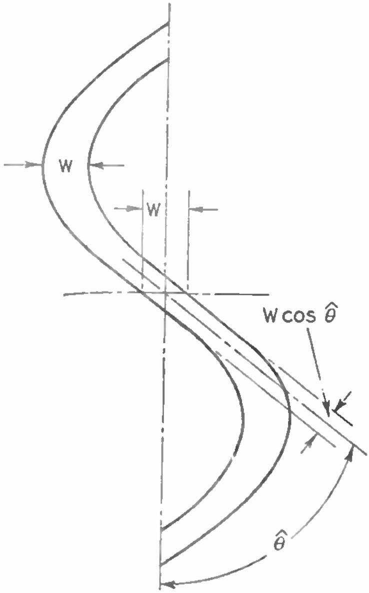


FIG. 11.—PINCH EFFECT.

On single-channel records pinch effect can be largely overcome by mounting the stylus on a short cantilever arm, which has adequate vertical compliance, but which, laterally, is rigidly fixed to the armature. Care must be taken that this does not introduce further resonances and upset the frequency response.

The stylus tip is actually controlled by two points of contact, with two identical waves in the form of the groove walls, but the two points of contact are undergoing continuous relative phase shifting, due to the sinuous motion of the groove and the rapidly changing angle. This leads to a complex distortion which has been very fully investigated.[6,7]

Roys [8,9] quotes calculated values of individual distortion and inter-modulation products. When the distortion does not exceed about 10 per cent, his expression for third-harmonic distortion may be simplified to the approximate form :

$$\text{Percentage third-harmonic distortion} \simeq 740 \frac{r^2 f^2 V^2}{S^4} \qquad (6)$$

where r is the playback stylus tip radius;
f is the recorded frequency;
V is the r.m.s. lateral velocity; and
S is the linear speed of the groove.

No reputable manufacturer of commercial records would permit appreciable amplitude distortion in the recorded waveform. Distortion is almost invariably due to the interaction between the groove and the stylus tip, and is governed by the wave shape and by the lateral impedance of the stylus over the frequency band. The upper end of the recording frequency characteristic is determined largely as a compromise between high levels, giving rise to distortion with typical commercially available pick-ups, and low levels, giving a relatively high background noise. The development of improved pick-ups has led to the use of considerable pre-emphasis on the records, with corresponding de-emphasis in the reproducer, leading to much reduced background noise. This is made full use of in microgroove records.

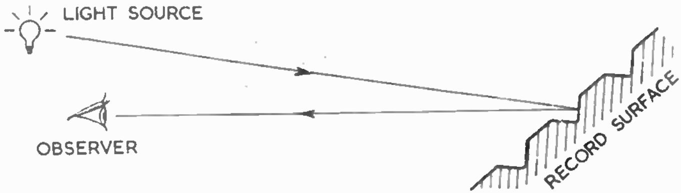


FIG. 12.—PRINCIPLE OF THE BUCHMANN AND MEYER OPTICAL SYSTEM.

Levels

For measurements of frequency response of cutter heads and pick-ups, and for distortion measurements, it is necessary to know the level recorded on the disc. At low frequencies amplitudes can be measured under a microscope, but at high frequencies they become too small, except at extremely high levels. Knowing the groove radius, however, and hence the linear velocity, the level can be determined from the angle $\hat{\theta}$, for the ratio of peak lateral to linear velocities is equal to tan $\hat{\theta}$. Buchmann and Meyer have developed an elegant optical method for measuring this angle.[10, 11]

Buchmann and Meyer Optical System

A parallel beam of light is allowed to fall on the groove walls, and will be reflected back, via a viewing telescope, to the eye of an observer, from any part of the wave which is normal to the plane through the light source, the eye, and the point of reflection, see Fig. 12.

In Fig. 13, consider a small section of sine-wave situated at a radius R, and in line with the record centre. Parallel light from a distant source will be reflected back in parallel rays from the points A and B, so the observer will see two points of light.

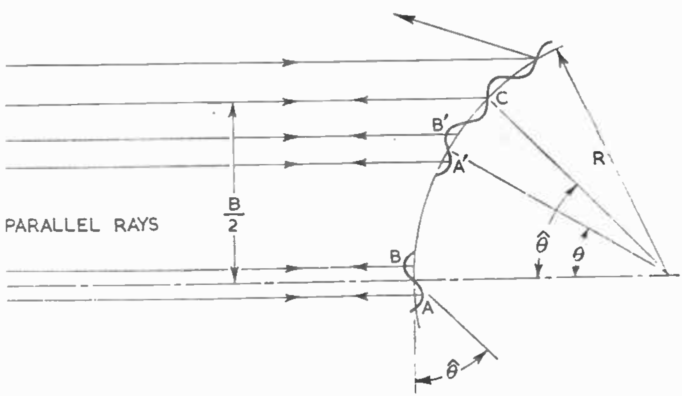


FIG. 13.—BUCHMANN AND MEYER MEASUREMENT.

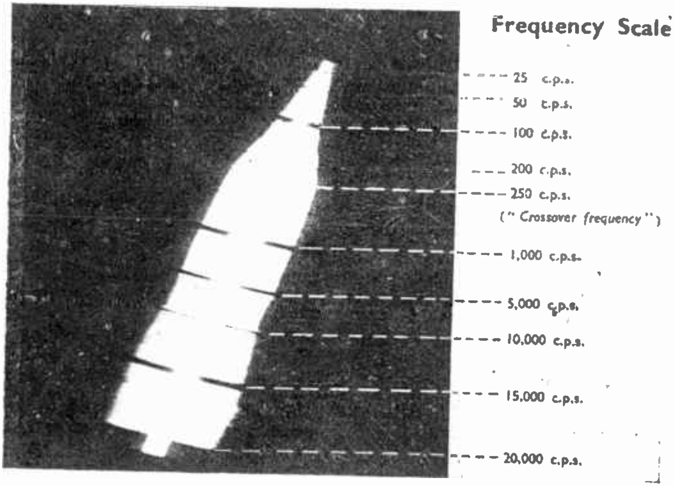


Fig. 14.—Buchmann-Meyer Pattern of Gliding Tone 20 kc/s to 25 c/s Recorded by E.M.I. Type 43 Equipment.

If the disc is now turned about its centre through a small angle, θ, the parallel reflected rays must emanate from two new points A' and B', which are closer together than A and B, and which occur at points on the wave where the groove makes an angle θ with the mean groove axis.

Turning the disc through a total angle of $\hat{\theta}$, the two points A and B will converge to a single point C, at the point of inflexion of the wave, where the angle between groove and mean groove axis is at its maximum value of $\hat{\theta}$.

Any further turning of the disc will produce no reflections to be seen by the observer, because the groove wall has no reflecting points at a suitable angle. The limiting angle of turntable rotation at which reflections can be seen is therefore equal to $\hat{\theta}$, the peak angle between the groove and the mean groove axis.

If the record has a continuous sine wave recorded on it, and if the turntable is continuously rotated, the individual reflection points will merge to show a band of light of breadth B cm. This can be measured by comparison with a scale mounted close to the record surface, or by a calibrated graticule in the telescope eyepiece.

From the geometry,

$$B = 2R \sin \hat{\theta} \qquad (7)$$

$$\text{Now } V = S \tan \hat{\theta} \qquad (8)$$

where S is equal to the linear velocity, which is $2\pi R \times \frac{\text{r.p.m.}}{60}$

(r.p.m. referring, of course, to the rotational speed at which the disc was recorded).

Hence

$$V_{\text{r.m.s.}} = \frac{\pi}{\sqrt{2}} \times \frac{\text{r.p.m.}}{60} \times \frac{B}{\cos\hat{\theta}} \text{ cm./second} \tag{9}$$

For small angles $\cos\hat{\theta}$ can be taken as equal to 1, and the error is less than 1 db for $\hat{\theta} = 25°$, $\hat{\theta}$ being half the angle subtended by the band-width at the centre of the record. The angle subtended at the centre of the disc by the light source and by the telescope objective must, however, be small compared with $\hat{\theta}$ if reasonable accuracy is to be expected.

For example, take B to equal 1 cm. on the outside of a 12-in., 78-r.p.m. record. From equation (9), the level is 2·9 cm./second r.m.s., and from equation (8), $\hat{\theta}$ is 2°. For reasonably accurate measurement the angles subtended by the light source and the telescope objective must not exceed about $\frac{2}{5}$°. If B is smaller, light source and telescope must be correspondingly smaller, for the same accuracy.

The level expressed in decibels will be 20 log $V_{\text{r.m.s.}}$, with respect to 1 cm./second.

Peak Levels

The peak levels recorded on a disc should be limited to values which can be reproduced without serious distortion by currently available pick-ups. At low frequencies the limit is set by the amplitude which can be accommodated by the groove spacing. At medium frequencies the velocity, or more exactly, the steepness of the wave front, angle $\hat{\theta}$, sets the limit. At high frequencies stylus pressure, and hence peak acceleration, is the limiting factor, and also the minimum radius of curvature of the wave must not be allowed to approach the radius of the needle tip.

Good practice indicates the following limits:

For 78 r.p.m. standard groove recordings

Maximum amplitude = 0·010 cm. peak.
Maximum velocity = 15 cm./second r.m.s.
Maximum acceleration = 5 × 10^5 cm./second^2 r.m.s.
g = 980.2 cm/sec² @ Cleveland

For microgroove recordings

Maximum amplitude = 0·004 cm. peak.
Maximum velocity = 7·5 cm./second r.m.s.
Maximum acceleration = 5 × 10^5 cm./second^2 r.m.s.

These figures are rather arbitrary, and represent a compromise between distortion and signal-to-noise ratio. They are often exceeded in practice, but at the inner groove radii they are liable to give appreciable tracing distortion. There is, however, some evidence of partial mutual cancellation between the distortions due to different causes. The best designs of pick-up, having very low mechanical stylus-point impedance, can cope with accelerations several times that indicated here, without very serious audible distortion, provided that the condition is transient.

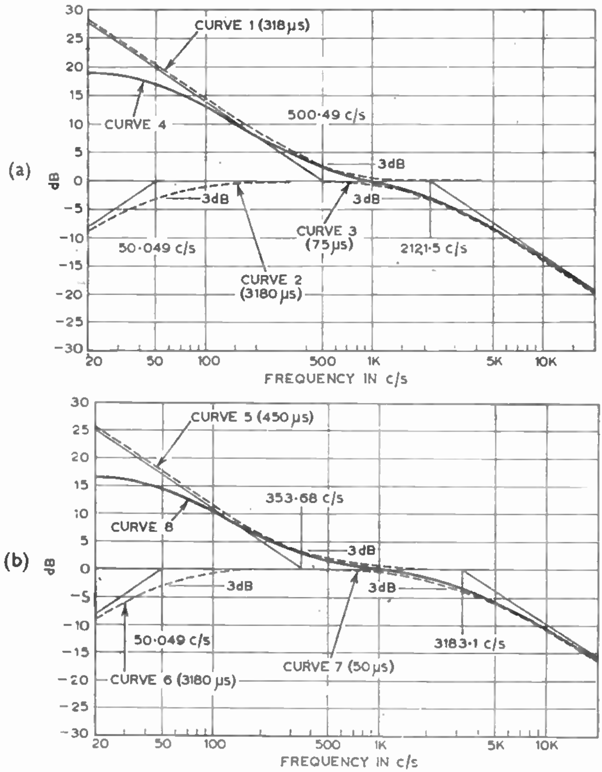


Fig. 15.—Standard Playback Response Curves.

Typical background noise levels,[12] measured on a constant-velocity basis (" flat " frequency response), with a 0·0025-in. radius point, are :

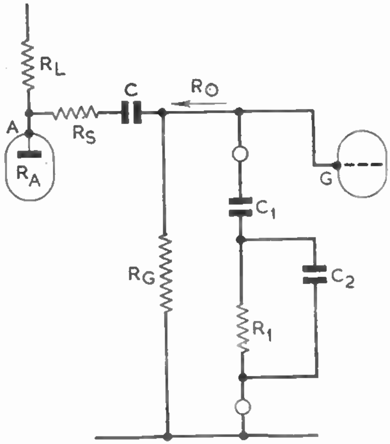
Filled shellac . . .	0·1–0·3 cm./second r.m.s.
Vinylite	0·01–0·02 cm./second r.m.s.
Cellulose lacquer . . .	0·005–0·01 cm./second r.m.s.

Pick-up Equalizers

For measuring the frequency response of pick-ups, a record may be made, and calibrated by the Buchmann–Meyer method, or suitable records of tones at various frequencies and known levels may be obtained commercially. Examples are E.M.I. No. JG881 (coarse groove) and E.M.I. No. ALP1599 and Decca LXT5346 (microgroove).

FIG. 16.—SIMPLE PLAYBACK EQUALIZER.

	Coarse Groove	*Micro-groove*
$C_1(\mu F)$	$\frac{2780}{R_0}$	$\frac{2937}{R_0}$
$C_2(\mu F)$	$\frac{4051}{R_0}$	$\frac{1007}{R_0}$
$R_1(\Omega)$	$0.141\ R_0$	$0.0804\ R_0$



The output of a well-designed pick-up when correctly loaded should be proportional to the recorded velocity. Serious departures usually indicate mechanical resonance, and mechanical damping is of more value than electrical equalization. A gradual reduction of output at high frequencies may occur if an inductive pick-up is loaded with too

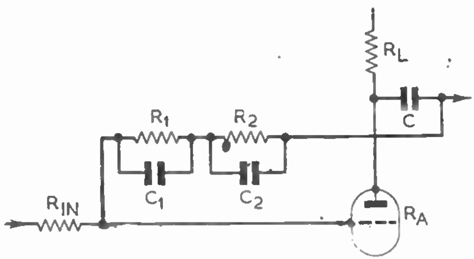


FIG. 17.—FEEDBACK EQUALIZER.

$R_0 = R_1 + R_2$ $\qquad\qquad X_C \ll R_0$

$$AR_{in} \approx KR_0 \approx K^2 \frac{R_A R_L}{R_A + R_L},$$ where K should be about 5 to 10.

A = Voltage gain without feedback.

	Coarse Groove	*Microgroove*
$C_1(\mu F)$	$\frac{3650}{R_0}$	$\frac{3450}{R_0}$
$C_2(\mu F)$	$\frac{391}{R_0}$	$\frac{959}{R_0}$
$R_1(\Omega)$	$0.872\ R_0$	$0.922\ R_0$
$R_2(\Omega)$	$0.128\ R_0$	$0.078\ R_0$

low a resistance. Similarly, a loss of bass with a capacitive crystal pick-up may also be due to too low a load resistance. Minor peaks in the curve may be reduced by a parallel inductance, capacitor and resistor, in series between pick-up and load.

Assuming the pick-up to have a substantially flat response, it remains to correct for the recording characteristic. The correct replay characteristic is the inverse of the recording characteristic described on p. 34–25. It may be obtained by using three networks, inverse to those described, and suitably separated with buffer stages; or, more conveniently, a more complex network giving the complete equalization in one step may be used.

Many circuit arrangements will produce the characteristics shown in Fig. 15. A simple arrangement, for use between amplifier valves, is shown in Fig. 16. Here R_S and R_L in parallel substantially form the load on the first valve, whose anode A.C. resistance is R_A. R_S is included to avoid too low a load on the valve when the value of Z becomes small. With a pentode, it may not be necessary.

To find the effective source impedance R_0, calculate the valve output impedance to the left of capacitor C:

$$R_{\text{out}} = R_S + \frac{R_A R_L}{R_A + R_L}$$

Then

$$R_0 = \frac{R_G R_{\text{out}}}{R_G + R_{\text{out}}}$$

The reactance of C should be made small compared with R_{out} at the lowest working frequency. Suitable component values are given.

An alternative circuit using negative feedback is shown in Fig. 17. The criterion for satisfactory operation here is that, at all significant frequencies, $R_{\text{in}} \gg \frac{Z}{A}$, where A is the amplification of the stage neglecting feedback, but not forgetting the loading effect of Z, the network impedance.

Here, the distortion due to this loading is of secondary importance, as the loading is accompanied by increased feedback. Nevertheless, Z should be made as large as possible compared with R_L, consistent with R_{in} remaining sufficiently low for the resistance noise not to become significant.

Cleanliness

Scrupulous cleanliness is most important with all types of gramophone records. Dust in the grooves will cause increase of background noise, and will also result in severe wear of both record and stylus. Finger-marks on the record surface will cause dust to adhere in patches, while the natural electrostatic attraction of the pressing material will cause the adhesion of dust over the whole surface. Shellac (78-r.p.m.) records may be sponged with a weak solution of mild detergent, rinsed and dried with a lintless cloth. Microgroove records should be cleaned with one of the special anti-static cloths available (e.g., Emitex). To avoid the accumulation of dust while playing the record, a device known as the " Dust Bug " has been found very satisfactory. When not in use, the record should always be stored in its protective sleeve, and should be kept flat, preferably under light pressure to avoid warpage.

Stereo Disc Records

The stereo disc record carries two separate recordings. These, when correctly reproduced over two identical independent channels, through two similar loudspeakers, suitably spaced, can be made to re-create at the listener's ears the original sound field, in such a way that the listener can judge the direction of the source of sound within the angle subtended by the two loudspeakers. This results in an ability to separate different sources, such as the individual instruments in an orchestra, and to distinguish between direct and reverberant sound. When the reproducing channels are of high quality a remarkable increase in realism is achieved.

In the system of recording originally patented by A. D. Blumlein, and now generally accepted, both recordings are carried in a single groove similar in form to the standard microgroove, with its two groove walls substantially at right angles. It may be said that the outer groove wall carries the recording for the right-hand channel, viewed from the listener, while the inner groove wall carries that for the left-hand channel. The two wave motions thus occur in directions mutually perpendicular, each making an angle of 45° to the record surface.

The reproducer stylus thus performs a complex motion, having both lateral and vertical components; the special dual-channel pick-up must be capable of resolving this complex motion into its original 45° components and producing two independent outputs determined by the individual waveforms.

The phasing of the two tracks is so arranged that the lateral component represents the sum of the two waves, and the vertical component the difference. Thus a normal single-channel lateral pick-up should give fair single-channel reproduction, although the niceties of balance and microphone placement will differ from the customary practice.

No official standards have yet been published, but accepted practice is to use groove dimensions similar to those of normal microgroove records, but with bottom radius on the low side, say 0·0002 in. This is to accommodate the stereo reproducer stylus tip of radius 0·0005–0·0006 in., necessary to reduce vertical tracing distortion and pinch-effect distortion. The frequency characteristic is the same as for normal microgroove.

Because of the small stylus tip, with its very small bearing area, pick-up weight must be as low as possible, and should not exceed about 5–6 gm.

For best results the two reproducer channels should be identical as regards frequency response and gain, within fairly close limits, say $\pm\frac{1}{2}$ dB between 100 and 10,000 c/s, ± 1 dB between 30 and 15,000 c/s, and their phase shifts should be the same within a few degrees over the full frequency range. Cross talk between channels should be as low as possible, preferably better than 40 dB down. The two loudspeakers should be as nearly identical as is possible: selected matched pairs show a marked improvement over random samples.

The two loudspeakers should be so placed that they subtend an included angle of about 60°–90° at the listener, and the acoustics of the listening room should be fairly " dead ", and more or less symmetrical. Theoretically the listener should be on the centre-line, equally distant from both loudspeakers, but in practice, the listening position is not usually at all critical, within fairly wide limits.[16]

It is important that the two loudspeakers should be in phase with one another, and that the overall sensitivity of both channels should be the same. This may easily be checked by playing a normal micro-groove record, when the sound source should seem to be positively located mid-way between the loudspeakers.

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E. W. B.-J.

35. GRAMOPHONE MECHANISMS

35. GRAMOPHONE MECHANISMS

For many years the only means of rotating a gramophone record at a constant speed was by means of a spring-wound motor. Gradually, however, the electric motor has superseded its spring predecessor, so that today gramophone units are driven almost exclusively by an electric motor.

The modern radio-gramophone is fitted with either a single-record player or automatic record changer. The single-record player comprises a spring or electric motor for driving the turntable, together with a pick-up and arm and automatic stop mechanism mounted together on a unit plate.

A record changer is a record player, as described above, driven by an electric motor, with means for automatically changing records. Until recently, both types of units were made to play 10- and 12-in.-diameter standard 78-r.p.m. records. However, with the introduction of micro-groove records, units to run at three speeds, 78, 45 and $33\frac{1}{3}$ r.p.m. are now in general use and capable of playing 7-, 10- and 12-in.-diameter records, with standard or micro-groove spacing.

Drives

The speed of the single-speed units, both spring and electric, has usually been controlled by a centrifugal governor. In the case of the electric motor the drive from the motor to the turntable shaft has been by means of a worm and gear.

Difficulty has been experienced in obtaining a steady drive on the lower speeds using a gear and governor, which tend to introduce wow, flutter and rumble, due to very slight unavoidable inaccuracies in the gear and unbalance in the governor.

For three-speed gramophone motors a rubber friction drive has been found to give most satisfactory results, producing a very silent and steady transmission between the motor and turntable. The principle

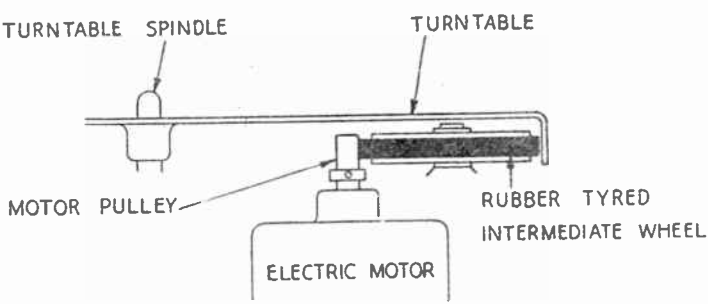


FIG. 1.—FRICTION DRIVE AS USED ON GRAMOPHONE UNITS.

of the rubber friction drive is illustrated in Fig. 1. The motor shaft is fitted with a pulley of appropriate size, and the drive from the pulley to the turntable is transmitted by means of a rubber-tyred intermediate wheel, which can drive either on to the inside of the turntable rim or on to a concentric drum attached to the centre of the turntable, according to the design of the unit.

The advantage of this type of drive is that the diameter of the driving surfaces which control the speed—the motor pulley and the turntable rim or drum—are made of metal, which does not wear appreciably with

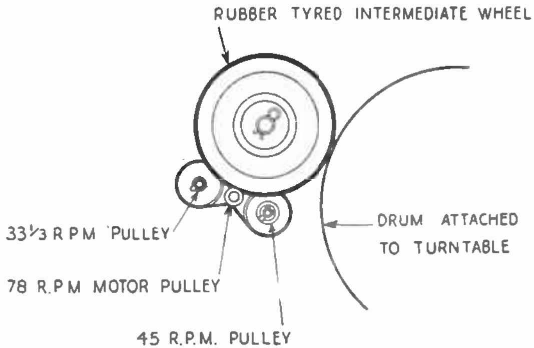


FIG. 2.—THREE-SPEED DRIVE USING BELTS.

use, whilst the size of the rubber intermediate wheel, which may wear, does not effect the speed. The ratio between the motor pulley and the turntable rim or drum remaining the same whatever the size of the intermediate wheel.

With this type of drive, it is essential for all diameters to be perfectly true in order to avoid wow or flutter in the reproduction. The driving surfaces must also be kept absolutely free from oil or grease, to prevent slip.

Three-speed Drives

Three methods are in general use for obtaining the three required turntable speeds. Fig. 2 shows the use of two small rubber belts driving from the motor pulley to two stepped pulleys. The diameters of all the pulleys are arranged to give the required turntable speed when the stepped pulleys are moved radially around the motor pulley, so that the appropriate pulley for the speed required drives the intermediate wheel. When the two stepped pulleys are in their central position relative to the intermediate wheel, the drive is then direct on the motor pulley to give a turntable speed of 78 r.p.m.

Another method, Fig. 3, uses a series of three stepped pulleys operating in an arc between the motor pulley and the intermediate wheel. One of the steps on each pulley is of rubber, and all diameters must be accurate for size, as they affect the ultimate speed. The third

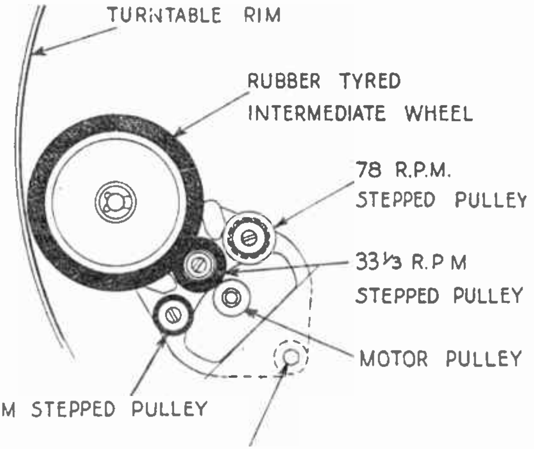


FIG. 3.—THREE-SPEED DRIVE USING THREE STEPPED PULLEYS.

method shown, Fig. 4, is now in general use, due to its simple construction. It comprises a motor pulley having three steps and an intermediate wheel which is moved on to the step having a diameter to give the desired speed. To change the speed the intermediate wheel must move vertically, and due to the length of the three-speed pulley, the driving surface on the turntable must be fairly wide. This, however, is no disadvantage; in fact by increasing the weight of the turntable it helps to reduce flutter and wow.

Electric Motors

Although a few single-record players are fitted with spring motors, the majority of gramophone units are driven by electric motors. The electric motor for use on A.C. mains supplies is usually of the shaded-pole induction type, while for D.C. mains and low voltages a series-wound commutator motor is used.

Induction Motors

The induction motor is not synchronous, but its speed is governed to a certain extent by the frequency of the A.C. mains supply, the final rotor speed being dependent on the frequency of the mains supply together with a regular applied load, maintaining a constant electrical slip of the rotor. The motor is rated to run with a slight electrical overload, so that, for the torque required to play a record, the rotor slip is substantially constant.

Induction motors for gramophone purposes usually have two or four poles, the two-pole type having a rotor speed under load of approximately 2,640 r.p.m. and a four-pole motor 1,320 r.p.m. Both types are used for gramophone work, the four-pole type being more usual, as its slower speed permits of a large-diameter motor pulley.

Fig. 5 shows a sectional view of a typical shaded-pole induction motor used for gramophones. This motor is of the four-pole type, but only two coils are used, these being placed on diametrically opposite poles, arranged so that the polarity of the poles on which they are fitted is the same, both South or North as the case may be, opposite polarity being induced in the two intermediate poles. The shading rings are of heavy gauge copper which produces the lag in the magnetic flux in the shaded portion of the pole necessary for inducing a starting torque in the rotor.

The assembly of the rotor laminations and copper bars is given a twist of about 30° in order to give an even starting torque, and reduce motor hum. As the motor is designed to run in a slightly overloaded condition, its temperature rise is higher than is usual with small electric motors, a rise of 60° C. not being considered exceptional, and this is provided for in the design. Fan blades are fitted to each end of the rotor to pass a current of air through the motor, and porous phosphor-bronze rotor-shaft bearings impregnated with a special high-temperature lubricating oil are fitted. These bearings are of the spherical, self-aligning type, and are machined to very close tolerances to prevent rattle. To avoid flutter, the rotor is very accurately balanced, and lead slugs are pressed into holes in the fan blades to dynamically balance the rotor.

The running torque of a typical induction motor, as described, is approximately 2 oz.-in. at 1,250 r.p.m., the starting torque 3·5 oz.-in., power factor 0·62 and the consumption 16 watts.

Motors are manufactured having single- or dual-voltage ranges. The dual range covers 100–130, and 200–250 volts, the coils being connected in parallel for the low range, and in series for the high range.

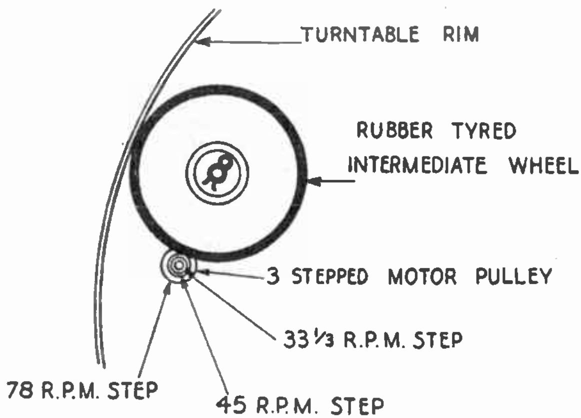
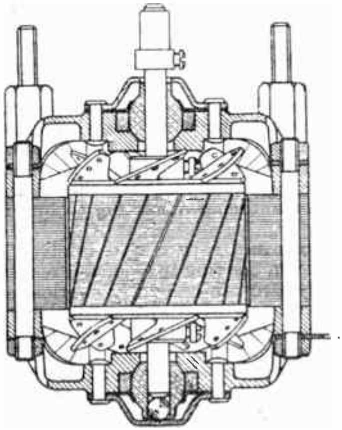


Fig. 4.—Three-speed Drive Using Stepped Motor Pulley.



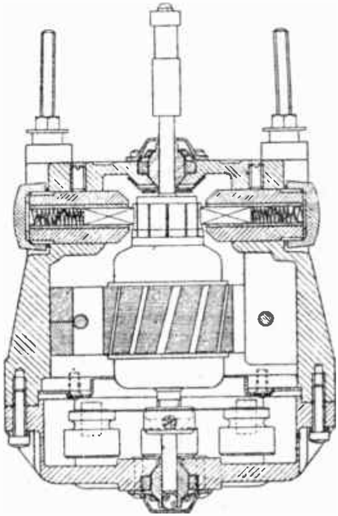


FIG. 5 (*above*).—SECTIONAL VIEW OF OF A TYPICAL INDUCTION MOTOR USED ON GRAMOPHONE UNITS.

FIG. 6 (*right*).—SECTIONAL VIEW OF A UNIVERSAL MOTOR FOR USE WITH GRAMOPHONE UNITS.

Series-wound Motors

Series-wound motors are used for D.C. mains and low-voltage supplies; and can also be made for A.C./D.C. operation, in which case they are commonly known as universal motors. Fig. 6 shows a section of a typical, series-wound universal motor. It has two field coils—on some types a permanent-magnet field is used—and the armature usually has twelve poles, the coils being double-wound connected to a twenty-four-segment commutator. In order to reduce mechanical brush noise to a minimum, the mica commutator segments are not undercut. To reduce sparking and noise, the brushes are of a special grade of carbon, a harder grade of brush being used for low-voltage supplies than for mains-voltage supplies. A pair of radio-interference-suppressor capacitors are connected across the brushes, and centre-tapped through another capacitor of 5,000-volts test to the motor frame, which should, of course, be earthed. These capacitors are usually mounted together in one housing. Television-interference-suppressor chokes may also be connected in series with each brush lead. The speed of these motors is usually controlled by a flat type of centrifugal governor which is pre-set on installation to give the correct turntable speed.

If the motor has field coils, it can be run on A.C. or D.C., a tapped resistance being connected in series with the mains supply to give approximately the same motor torque for varying voltages and frequencies. With motors having a permanent-magnet field, a tapped mains resistor is used for D.C. mains supplies, together with a metal rectifier for A.C. supplies so that the motor is always running on D.C.

For low-voltage D.C. supplies, such as 6 or 12 volts, the motor is specially wound, no resistor being required.

Transcription Motors

For professional purposes, a heavier type of gramophone motor is generally used. the principles of drive being similar to those already described. All bearings, spindles and the turntable are of heavier construction, designed to give continuous service such as may be required in broadcasting studios. These are generally referred to as transcription motors.

Speed Adjustment

A simple method of obtaining speed adjustment on gramophone units having induction motors makes use of an eddy-current brake. This applies an even load to the motor in order to reduce its speed. An aluminium disc is attached to the rotor shaft, the disc passing between the poles of a permanent magnet.

This magnet is pivoted in such a way that it can be swung across the disc, exerting a greater pull as it moved inward. The speed of the turntable is originally fixed by the size of the motor pulley with the magnet in its mid-position, so that the speed is increased due to the reduction in load as the magnet is moved away from the disc, and reduced due to increased load as the magnet is moved over the disc. The magnet is moved by a speed-control knob on the motor plate, the speed variation obtained being about 2½ per cent fast or slow.

Most gramophone units provide no adjustment for speed, the diameter of the pulleys being such that the turntable is driven at the desired speed.

Automatic Trips

Most single-record players and all record changers have an automatic trip mechanism designed to operate as the pick-up runs into the run-out groove at the end of a record. In the case of single-record players, the trip operates a switch which stops the motor. In some cases it also causes a brake to be applied to the turntable. On record changers the trip releases the operating cam to commence the changing cycle.

The mechanism for an automatic trip usually falls into one of two categories, viz., th velocity type or the ratchet type. The velocity principle is almost exclusively used on modern three-speed units, and it operates as follows : as the pick-up advances towards the centre of the record it moves at a slow, even speed and the trip does not operate, but when it runs into the record run-out groove, the speed of the pick-up arm is accelerated and the trip operates.

Single-speed units playing only 78-r.p.m. records often use the ratchet type of trip mechanism; this is operated by the reversal of the inward movement of the pick-up as it runs into the eccentric run out groove of the record. A lever attached to the pick-up arm carries a very light pawl which moves along a ratchet connected to the switch or change mechanism, the pawl riding over the ratchet teeth as the pick-up moves inward, but engaging with a tooth and moving the ratchet, which trips the mechanism, as the pick-up-arm movement reverses.

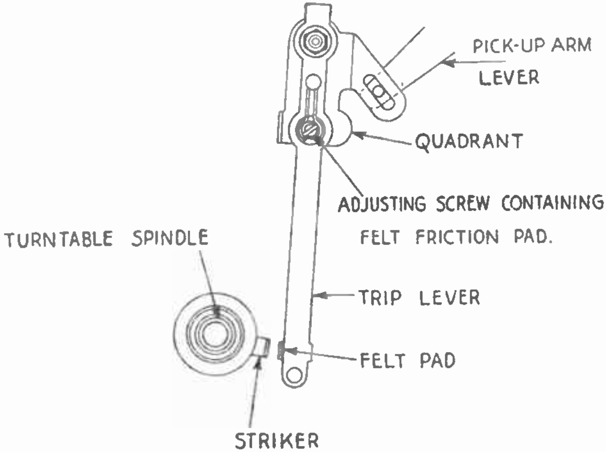


Fig. 7.—Velocity Type of Automatic Trip.

Since there is no eccentric groove on the 45-r.p.m. microgroove records, any unit intended for playing these records should have some form of velocity trip.

Velocity Trips

Fig. 7 illustrates the principle of the velocity trip mechanism. A quadrant is moved by the pick-up arm; this carries the trip lever, which is moved with it by the friction of a felt pad on the polished face of the quadrant. When the pick-up nears the centre of a record, the quadrant commences to move with it, transmitting its inward movement to the trip lever. As the end of the trip lever moves towards the turntable spindle, a felt pad, located in the end, contacts a projection—usually referred to as a striker—on the revolving turntable spindle. Each revolution of the striker pushes away the trip lever as it slowly moves inward; but when the movement of the trip lever is accelerated by the pick-up reaching the run-out groove, the trip lever moves inward too far to be pushed back by the striker, and the striker then either lifts the trip lever or moves a lever attached to it to trip the mechanism. This type of auto-trip can be made very sensitive in operation, accelerated movement of the pick-up through $\frac{3}{32}$ in. being sufficient to operate it. It should be noted that the felt friction pad on the quadrant face should not be lubricated, since the oil is liable to become gummy and also collects dust; this may increase the friction to such an extent that the action of the striker pushing the trip lever will be transmitted to the pick-up and become audible in the reproduction.

Operation of Single-record Players

To operate any three-speed single-record player, first set the speed-change knob to indicate the speed required for the record to be played,

FIG. 8.—SLOPING TYPE OF STEPPED RECORD SPINDLE, SHOWN HORIZONTALLY.

either 33⅓, 45 or 78 r.p.m. Next place a record on the turntable and see that the correct stylus is in position for the type of record to be played, one having a point radius of 0·0025 in., usually coloured green, for standard 78-r.p.m. records, and one having 0·001 in. radius point, coloured red. for microgroove records. For stereo records, radius points of about 0·0005 – 0·00075 in. are used.

All standard 78-r.p.m. and microgroove 33⅓-r.p.m. records are made with the standard 0·286 ± 0·002-in.-diameter centre hole, but the 7-in., 45-r.p.m. records have a centre hole which is 1·504 ± 0·002 in. in diameter. To play these records, adaptors to fit over the small centre spindle are obtainable. Also there are several types of clip-in adaptors, one of which is fitted into the large hole of each record. The 7-in., 45-r.p.m. records are now being made with optional centres. They normally have the standard small-centre hole, but if desired the centre can be pushed out to give the large-diameter hole.

Automatic Record Changers

The motor turntable drive and auto-trip mechanism used on automatic record changers is similar to those fitted to single-record players and already described.

Many methods have been used in the past to place records for playing one at a time on to the turntable of record changers. The method now generally adopted is to place a stack of up to eight or ten records over the top of a long, stationary centre spindle which has a step on which the records rest. They are released one at a time for playing, the records resting on top of each other on the turntable.

There are two general classes of record-dropping arrangements: the first plays batches of records of one size only, either 7, 10 or 12 in.; the other plays batches of records of mixed sizes, the mechanism automatically measuring the size of the record to be played and setting the dropping position of the pick-up accordingly.

Great care is required in making record spindles to see that all angles and gap dimensions are correct, and that the edges of the step and pushing pawl are free of sharp edges which could cause damage to the record hole.

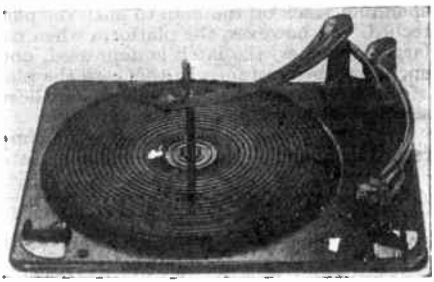


FIG. 9.—GARRARD MODEL R.C.110 THREE-SPEED RECORD CHANGER.

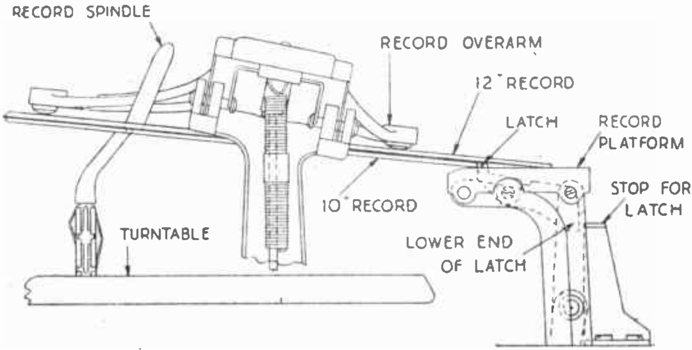


FIG. 10.—RECORD PLATFORM FOR PLAYING MIXED SIZES OF RECORDS.

Fig. 8 shows one type of record spindle. This is used in conjunction with a platform which pushes the records off the step. The distance between the face of the step on which the records rest and the underside face of the inverted step is such that only one record can pass at a time. The spindle is inclined toward a platform on which the edge of the records rest. This platform has a pawl or stud which at the correct moment engages with the edge of the lowest record and pushes it, causing the record to drop off the step; as the record falls, the slope of the spindle causes it to pull away from the platform and drop flat on to the turntable. When using a simple platform with this type of spindle, the platform position must be pre-set for the size of record to be played. Mixed sizes of records can, however, be handled with this type of spindle by using the special form of platform described below. The stack of records is held at the correct angle on the sloping record spindle by means of an overarm. This overarm is required to hold the records in position while the platform moves away to select the record size: Fig. 10 shows the mechanism in the platform. To drop a record, the platform first moves inward beneath the records to a position determined by the smallest-diameter record. Having reached its innermost position, the platform rises to touch the records; if a record is present of the appropriate size, the latch on the platform is not depressed, thus allowing its lower end to engage with a stop. The platform then remains in this position, allowing its operating lever to select the corresponding track on the cam to push the platform forward to release the record. If, however, the platform when rising encounters a record of a larger diameter, the latch is depressed, and its lower end then fails to engage with the stop. In this case the platform falls back to the edge of the record, and the latch then rises, allowing another cam track to be selected to move the platform forward.

Fig. 11 shows what is probably the simplest form of centre-spindle record-dropping. The stepped spindle is slotted; housed within this

FIG 11.—CENTRE-DROPPING TYPE OF RECORD SPINDLE SHOWN HORIZONTALLY.

FIG. 12.—PLESSEY TYPE D THREE-SPEED RECORD CHANGER.

slot is a pivoted lever so placed that the top end of the lever pushes the lowest record off the step as the lever is operated from the lower end. A loose latch allows easy unloading of the records without removing the spindle, the latch rising and moving inward to line up with the record spindle as the records are lifted. A simple type of overarm rests on the top record to hold the records level on the spindle.

The pushing lever is usually spring loaded in a vertical direction. After it has pushed the lower record off the step it will be depressed to the level of the step when the weight of the remaining records falls upon it. This prevents any damage to the hole of the lowest record being caused by the weight of the stack of records above it having to be supported only on the small area of the top of the lever. With the spring-loaded lever, the weight is thus distributed over a greater area.

There are many forms of the centre-dropping type of record spindle similar to that described, the principal differences being in the method of pushing the records off the step. Instead of a lever, as described, a small rotating cam is sometimes used; and many rather more complicated forms of the simple lever described are also found.

A special record spindle for playing 45-r.p.m. records having a $1\frac{1}{2}$-in.-diameter centre hole is made by several manufacturers to fit their own record changers. The spindle is $1\frac{1}{2}$ in. in diameter, and the mechanism to move the record-dropping blades is contained within it. This spindle fits over the existing record spindle and is operated by the pushing pawl.

The records when placed on the spindle rest on two projecting lugs. In operation, two thin blades move outward between the two lowest records. The two lugs move inward, allowing the lowest record to drop, leaving the remainder supported by the blades. The two lugs then move outward as the blades move inward, causing the records to be lowered on to the lugs ready for the next operation. The records are readily removed by lifting them upward, the lugs being spring loaded and chamfered on the underside to allow them to move into the spindle as records are moved past them.

One record spindle of this type has a projecting shaft which fits into the turntable spindle in place of the small-diameter sloping type. The shaft is rotated by mechanism underneath the changer to operate the record-dropping blades.

Using the centre-dropping type of record spindle, it is only necessary to select the pick-up dropping position. The actual record dropping, operated from the centre hole, is independent of record diameter. On some record changers the pick-up dropping position is set manually, and on others it is done automatically, the record changer then operating as a "mixer". Automatic selection of the pick-up dropping position is usually achieved by means of a projecting selector lever, which is moved by the record as it drops.

If the selector lever is not touched by a record, the pick-up automatically lands in the 7-in. position; a 10-in. record, in falling, moves the selector lever a small amount, which causes a lever to interrupt the pick-up arm in the 10-in. position as it swings inward. Similarly, a 12-in. record moves the selector lever a greater distance, thus causing the pick-up arm to lower in the 12-in. position.

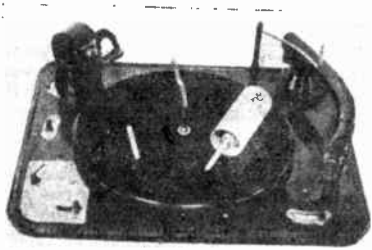


FIG. 13.—GARRARD MODEL R.C.90 THREE-SPEED RECORD CHANGER.

Another method of selecting the record size is by means of a swinging arm. This arm swings inward and is stopped by the edge of the lowest record on the record spindle. This action moves a lever which halts the inward movement of the pick-up arm at the selected place.

Record-changing Cycle

The complete record-changing cyle is usually controlled by the rotation of a cam which makes one revolution for each changing cycle and which is disconnected from its driving pinion during the playing of a record.

The cam is driven by means of a pinion on the turntable spindle driving a gear which is integral with it. Several consecutive teeth are cut from the gear, thus providing a gap in which the turntable pinion runs when a record is being played. Upon the operation of the auto-trip, the cam gear is given an impulse which causes the teeth on the gear to engage with the pinion teeth. The cam rotates until the pinion again reaches the gap, when the cam stops, leaving the pinion revolving in the gap.

To avoid the possibility of the teeth jamming as the gear is moved into engagement, the first few teeth after the gap are on a separate spring-loaded segment (see Fig. 14).

By using a number of cam tracks, the many levers that operate the motions required on a record changer can be moved in their correct sequence. The cam is geared to complete one revolution—corresponding to one complete changing cycle—in approximately 4–8 seconds. The lever which impulses the cam to start it in motion is spring loaded, the spring being reset by the cam during its cycle. The auto-trip also releases a catch which holds the impulse lever in the loaded position.

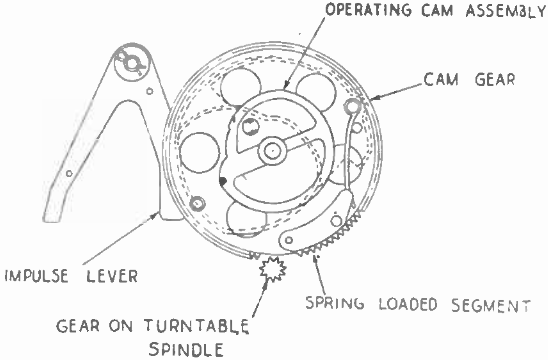


Fig. 14.—Cam Gear and Drive.

When the changer has played the last record it must stop automatically after the pick-up has returned to its rest. This is operated either by the dropping of the record overarm after the last record has dropped; by the uninterrupted inward movement of a swinging arm; or by the rising of the record spindle as the weight of the last record is removed. The usual method of operation is by means of the cam tripping the switch lever during each cycle; normally the record in position on the spindle holds the switch in the "on" position. However, when the last record has dropped, the retaining catch is moved by one of the methods already described, allowing the switch to break contact during the next rotation of the cam.

The rotation of the cam also controls the movement of the pick-up arm, which must move smoothly and descend gently on to the record with minimum impact between stylus and record. When a record has dropped, the pick-up arm lifts from its rest and moves inwards to a position ready to lower on to the commencement of the record: the correct radius is $3\frac{3}{8}$ in. for 7-in., $4\frac{13}{16}$ in. for 10-in., and $5\frac{13}{16}$ in. for 12-in. records. The pick-up should descend at a steady rate; and when the pick-up stylus lands on to the record, the mechanism controlling the movement of the pick-up arm must immediately disengage, leaving the pick-up arm free while a record is being played.

The pick-up arm movement is generally operated by a friction disc which drops out of contact with the pick-up arm immediately the pick-up lands on a record. The pick-up arm pivots must be free, but the vertical pivot for the horizontal movement of the pick-up arm must be very lightly damped, so as to prevent the stylus skidding inward as it lands on microgroove records. The damping can be applied mechanically by means of a very light spring, or alternatively a special lubricant having a high viscosity can be used in the pivot bearing.

Operation of Record Changers

Present-day record changers are designed to play automatically records conforming to British Standard 1928: 1953 "Lateral-cut Gramophone Records and Direct Recordings". A number of records which do not conform to this standard are still being made and sold, and difficulty may be experienced with such pressings. These are pre-1933 recordings, having no run-in groove on the outer edge and sometimes no run-out groove; or, where a run-out groove is provided, the diameter may be such that it will not operate the auto-trip.

To obtain the best results from records, the surface should be preserved by keeping them as dust-free as possible, and it is recommended that microgroove records be cleaned with a clean, soft, damp cloth. The turntable surface must also be kept perfectly clean. Avoid touching the music lines, keep records away from heat, and store in container when not required for playing. Do not play chipped or cracked records, as they will damage the stylus point. Badly warped records will also give trouble in dropping and playing, and should not be used. Do not leave records on the turntable when the changer is not in use.

Maintenance and Servicing

On record changers using a stationary record spindle, the turntable must revolve around it. This precludes the use of the single-ball thrust as used on single-record players. The single-ball thrust is quiet and rumble free; but the turntable thrust bearing of record changers is usually a race containing a number of steel balls which, unless special precautions are taken in manufacture, can be a source of noise, rumble and flutter. This ball race is situated either immediately under the turntable boss or at the lower end of the turntable spindle, and usually rests on a felt or plastic buffer washer. The two steel washers between which the steel balls run should be hardened, ground and lapped to a high degree of finish, and first-grade steel balls should be used.

If rumble is troublesome, first of all examine the turntable thrust race, since dirt in the race, indented washers, due to the turntable receiving a blow, or a hardened felt or plastic buffer washer may be the cause of the rumble. The race should be cleaned out and, if necessary, the washers and balls replaced.

Rumble, and often hum and flutter, may be caused by perished motor-mounting rubbers; these are rubber grommets or bushes in the unit plate. The resilience of the rubber is important in preventing unavoidable vibrations of the motor from reaching the pick-up via

Fig. 15.—Collaro Model 3RC531 Automatic, Three-speed Record Changer.

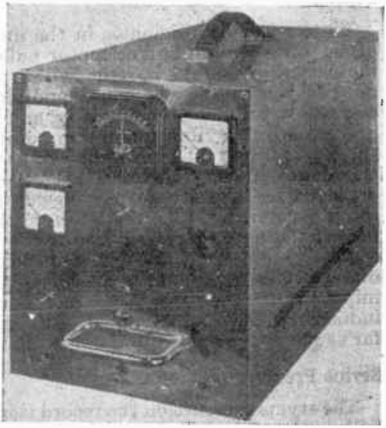


FIG. 16.—GAUMONT-KALEE WOW AND FLUTTER METER.

the unit plate. The rubbers should be replaced, if necessary, by new ones obtained from the manufacturer.

Wow

For gramophone purposes, "wow" can be briefly described as a variation in speed. Speed variations below 20 c/s being known as "wow", above this figure as "flutter".

On high-quality gramophone motors a figure of 0·2 per cent wow and 0·04 per cent flutter at 33⅓ r.p.m. is considered very good, 0·5 per cent wow being just about detectable by a keen ear. Special meters have been designed to measure wow and flutter. Fig. 16 shows such an instrument. A constant-frequency record of 3,000 c/s is played on the gramophone motor to be measured. The output from the pick-up is fed into the instrument, where it is compared with a crystal-controlled frequency of 3,000 c/s generated in the instrument. Variations between the two frequencies are measured, and indicated as percentages of wow and flutter.

Wow is more noticeable on 33⅓- and 45-r.p.m. records due to the slower speed reducing the momentum of the turntable. Turntable momentum is very important in maintaining steady running, and for this reason some manufacturers increase the weight of their turntables by loading with a metal disc or ring.

There are many causes of wow: if it is heard on a record changer the first thing to suspect is record slip. This can be checked by playing the record on which wow is suspected directly on the turntable. If it is then free from wow the trouble is most likely caused by the record slipping through bad contact with the one beneath it. To prevent slip the records should be reasonably flat and the hole free from burrs. Highly polished new records sometimes slip, and in this case a useful tip is to stick a transparent stamp hinge on the labels, thus giving sufficient key to drive the record. Incorrect stylus pressure will also cause slip, particularly on 7-in., 45-r.p.m. records, and thus should be checked: 8–10 gm. is the pressure recommended by most manufacturers.

Dirt in the turntable spindle, intermediate wheel and pulley bearings, and also slip in the rubber drive may all cause wow. The rubber tyre on the intermediate wheel must be kept absolutely free of oil to prevent slip. Slip can also occur if, in time, the motor pulley becomes highly polished with use, in which case a new pulley should be fitted. The concentricity of the intermediate wheel is important, as it sometimes wears unevenly. If the wow is very erratic, and does not occur regularly, slip in the drive should be suspected.

Flutter

Flutter usually originates in the motor driving the turntable, and is often caused by an untrue motor pulley or an unbalanced rotor. The motor pulley should be true within 0·0005 in., and any eccentricity outside this limit will cause audible flutter. The rotor should be accurately balanced dynamically, and its shaft must be true.

A mechanical vibration of the motor is produced by the frequency of the mains supply, the vibration being twice that of the mains frequency: this vibration can sometimes reach the pick-up via the motor mounting or the drive. As already mentioned, the resilience of the motor mounting is very critical: if it is too hard the motor vibration will be transmitted to the unit plate; and if too soft the motor may vibrate excessively, and transmit this vibration to the pick-up by way of the intermediate wheel and turntable. Magnetic pick-ups, especially low-impedance, low-output types, are particularly prone to pick-up hum induced by the magnetic field of the motor, which should be placed as far as possible away from the pick-up in order to minimize this trouble.

Stylus Pressure

The stylus pressure on the record is important: if too light, the stylus will not track heavily modulated grooves; and if too heavy, excessive record and stylus wear will occur.

For monophonic three-speed units the stylus pressure usually recommended is 8–10 gm. Pick-up arm counterbalance adjustment is provided on most units to facilitate this setting. It is recommended that the stylus pressure be checked on installation and that periodic checks be made as the tension of the pick-up-arm counterbalance spring may alter slightly with use. A stylus pressure gauge is shown in use in Fig. 17. For stereo pick-ups, requiring stylus pressures of the order of 3–5 gm, maker's instructions should be consulted.

If difficulty is found in setting the correct stylus pressure, it may be due to excessive friction in the cross-pivot, which should be examined: clean and lubricate, if necessary.

Should the stylus occasionally jump a groove, first check the stylus point to see that it is not excessively worn or broken. If found to be in order, see that the screened lead from the pick-up arm is not strained or twisted in such a way that it can bias the free movement of the pick-up arm. This lead should not be disconnected from the tag strip but left as supplied by the manufacturers.

Resonances of Pick-up Arms

Pick-up arms are designed so that their natural resonance is as far as possible outside the recorded range, and for this reason great attention is given to their shape, rigidity, mass and cross-section.

Earlier types of pick-up arms had resonances up to 100 c/s to help boost the low-frequency response of the pick-up, but with the latest type of pick-up the arm resonance should be as low as possible, resonances in the region of 20–30 c/s being desirable. The pivots must also be extremely free, but not loose enough to produce arm vibration, which can cause the stylus to leave the record groove.

To reduce unwanted vibrations, the pivots can be damped with a lubricant of high viscosity. The off-set angle of the arm is designed to give a minimum tracking error. On a typical pick-up arm for use on

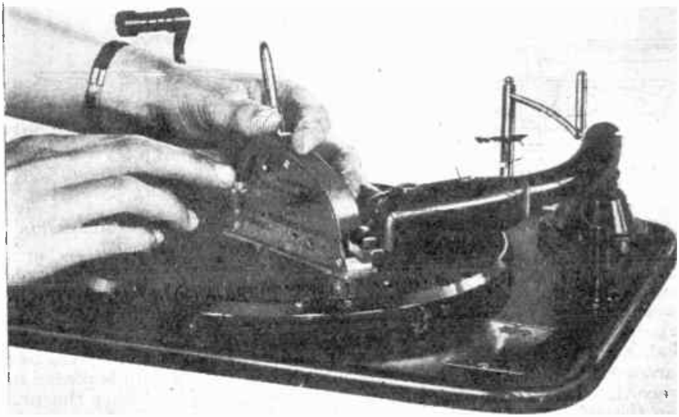


FIG. 17.—CHECKING STYLUS PRESSURE.

units having a distance of 7½ in. between the centre of the turntable and the pick-up arm pivot, the offset angle is 26° the stylus point being ½ in. in front of the turntable centre; this gives a maximum tracking error of approximately 2½ degrees.

To play the two types of record groove, a pick-up with reversible styli or two separate heads is used. If the former is preferred the pick-up can be housed in a one-piece arm, the pick-up then being integral with the arm.

When two separate pick-ups are used, they must be detachable from the arm by means of a plug-in or clip-on arrangement.

A spring is generally used to counterbalance the weight of the pick-up arm, and this is provided with means of adjustment to set the stylus pressure correctly. A spring counterbalance is normally preferable to a weight which, unless very carefully designed, may cause bad tracking on warped records.

Turntables

Turntables should run true within 0·020 in. measured by means of a flat disc played on the covering. If made of steel or some other magnetic material, the turntable should be free from holes or ribs which may produce magnetic variations which become audible when passing under some types of low-output magnetic pick-ups. On transcription motors, heavy cast turntables are generally used, having a large-section outer edge to increase the weight, these being made in a non-magnetic material, usually aluminium.

Radio-gramophones should, on installation, be checked by placing a spirit level on a record on the turntable to see that the turntable is level. The unit is levelled, if necessary, by adjusting its spring suspen-

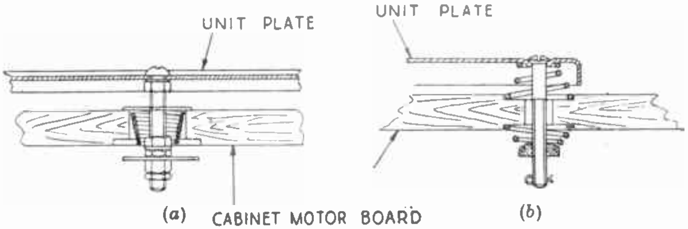


FIG. 18.—SUSPENSION SPRINGS: (a) TENSION TYPE; (b) COMPRESSION TYPE.

sions. Gramophone units are suspended in the cabinet on springs to prevent acoustic feed-back building up between the loudspeaker and pick-up. This occurs mainly in radio-gramophones with the loudspeaker in the same cabinet as the unit, and its presence is indicated by a characteristic howl from the loudspeaker when the pick-up is placed on a record. The spring suspension, by mechanically isolating the unit from the cabinet, also helps to prevent external vibrations from reaching the pick-up, which can, on the slightest vibration, jump the grooves and possibly skid across the record. This is particularly serious when playing microgroove records.

The unit should float freely on its suspension springs, and the edges of the unit plate should be well clear of the cabinet sides. Both the mains and pick-up lead connecting the unit with the chassis must be loose and not strained tight.

There are two types of spring suspension in general use on gramophone units; one (Fig. 18 (a)) uses conical springs in tension; the other (Fig. 18 (b)) has compression springs. To level the unit some types can be adjusted by means of their fixing nuts.

In order to prevent undue movement of the unit during transit, it should be rigidly clamped to the cabinet motor board, special screws being generally provided for this purpose. Care should be taken on installation to see that these screws are removed.

Single-core, screened pick-up lead terminating on a tag strip is usually fitted on record changers. The reason for this is that as the pick-up lead can bias the free movement of the pick-up arm, it must be as light and flexible as possible. It must be anchored so that it cannot be strained. Twin-screened pick-up lead is supplied on units intended for use on D.C. supplies or on A.C./D.C. chassis.

Some record changers are now fitted with a pick-up muting switch which short-circuits the pick-up during the changing cycle in order to prevent extraneous noises in the reproducer when the pick-up is not playing a record.

Switch surge suppressors can also be incorporated to suppress the "plop" in the loudspeaker as the switch breaks contact. These suppressors consist of a 220-ohm. $\frac{1}{4}$-watt resistor in series with an 0·01-μF capacitor (3,000-volt D.C. test) connected across the switch contacts.

E. W. M.

36. BATTERIES AND CONVERSION EQUIPMENT

36. BATTERIES AND CONVERSION EQUIPMENT

PRIMARY CELLS

In both the primary cell, and the storage cell or accumulator, the electrical energy is derived from chemical energy which is liberated when the external circuit is closed. But in one case—the primary cell—the process of energy conversion involves the dissolution and loss of weight of one of the electrodes forming the cell causing the ultimate exhaustion of the cell.

In the other case—the storage cell—the process of energy conversion involves only a change of chemical state of the plates and electrolyte, and when exhausted the cell is restored to its active condition by converting the plates again to their original state. This reconditioning or recharging process is effected electrochemically by passing a reverse current through the cell for a definite period.

The elements of a primary cell usually comprise two conducting plates, or electrodes, of dissimilar metals; an electrolyte, or conducting liquid, which can combine chemically with the metal of one of the plates; and a depolarizer.

In " wet " cells the inert electrodes (usually carbon or copper), together with the depolarizer, is contained in a porous pot (or " sack " if the depolarizer is a solid) which stands in a jar containing the electrolyte, in which also dips the other plate (usually zinc). The depolarizer may be either a liquid or a solid.

In " dry " cells the depolarizer is a solid, and is moulded as a stiff paste around the inert plate (which is of carbon). This element is then placed inside a cylindrical zinc cup, and the intervening space is filled with a thin paste or jelly mixed with the electrolyte. The open end of the cylinder is sealed with compound.

Why a Depolarizer is Necessary

If a primary cell is made up without a depolarizer—viz., with zinc and copper plates and an electrolyte of dilute sulphuric acid (the chemical symbol for which is H_2SO_4)—the cell will quickly " polarize " when current is taken from it. Thus the current will rapidly drop from its initial value to a very much lower value, and when the circuit is opened the e.m.f. of the cell will be much lower than its original value, to which it will ultimately return after a prolonged period of rest. Fig. 1 illustrates graphically the results of a test on such a cell.

The Cause of Polarization

The polarization is due to the collection of a film of hydrogen bubbles on the surface of the copper plate. This thin film of hydrogen sets up an e.m.f. in opposition to the normal e.m.f. of the zinc–copper–sulphuric acid combination, and in consequence the available e.m.f. of the cell is reduced.

The hydrogen bubbles accumulate on the surface of the copper plate

TABLE 1.—DATA ON PRIMARY CELLS

Name	*Cathode* —	*Anode* +	*Exciting Fluid*	*Depolarizer*	*E.m.f.*	*Internal Resistance (ohms)*	*Applications*
Leclanché (Wet)	Zinc	Carbon	Sal ammoniac	Manganese dioxide	1·5	1–3	Bells, light signal work (long periods of rest)
Daniell	Zinc	Copper	Zinc sulphate	Copper sulphate	1·1	2–6	Rough standard
Leclanché (Dry)	Zinc	Carbon	Sal ammoniac	Manganese dioxide and carbon	1·5	0·1–0·5	Radio and flashlights
Bichromate	Zinc	Carbon	Dil. sulphuric acid	Potassium bichromate	2·0	0·5–1	Signal work
Bunsen	Zinc	Carbon	Dil. sulphuric acid	Nitric acid	1·9	0·1–0·2	—
Grove	Zinc	Platinum	Dil. sulphuric acid	Nitric acid	1·9	0·1–0·2	—
Fuller	Zinc	Carbon	Dil. sulphuric acid	Potassium bichromate	2·0	0·5–4	U.S.A. telephones. Large-current experimental work
Edison	Zinc	Copper	Caustic potash	Copper oxide	1·0	0·2–0·9	Light signal work, with long periods of rest
Weston	Cadmium	Mercury	Cadmium sulphate	Mercurous sulphate	1·0183	900	B.O.T. Standard

produced by ionic action as a result of the conversion of chemical energy (due to the combination of the zinc and the sulphuric acid) into electrical energy, hydrogen being liberated in the process, and zinc passing into

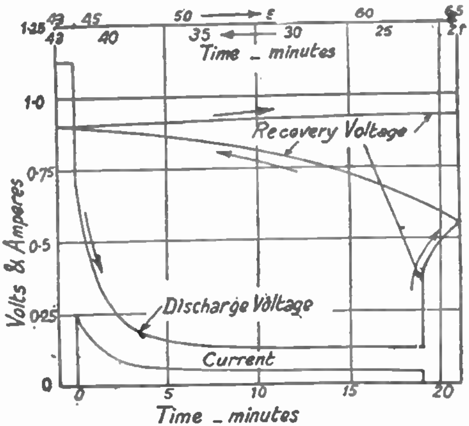


FIG. 1.—RESULTS OF POLARIZATION AND RECOVERY TEST OF ZINC–COPPER – SULPHURIC ACID CELL WITHOUT DEPOLARIZER.

It will be seen that polarization is rapid whilst recovery is very slow.

solution. The ionic action due to the current causes the hydrogen to be liberated at the copper electrode.

Function of Depolarizer

The function of the depolarizer is to prevent the deposition of hydrogen on the copper plate. If the depolarisation is complete no bubbles of hydrogen will appear on the copper plate, and therefore no back e.m.f. will be produced. The full normal e.m.f. of the cell will then always be available, and the current will remain constant when the cell is connected to a circuit of constant resistance.

Such complete absence of polarization is obtained when certain liquid depolarizers (such as copper sulphate, nitric acid) are employed.

But the use of these solutions, together with sulphuric acid, is objectionable and undesirable for the purposes for which primary cells are required at the present day, and in consequence modern primary cells have a solid depolarizer.

The use of sulphuric acid as an electrolyte in a primary cell has the great disadvantage that the zinc undergoes attack from the acid when the cell is standing (i.e., the external circuit is open and no current is required). This "open-circuit attack" of the zinc is called *local action.*

Local action may be reduced by coating the surface of the zinc with a film of mercury. In the latter case the amalgam formed on the surface of the zinc has the same contact e.m.f. with the acid at all points, and in consequence local action is minimized.

DRY BATTERIES

A dry cell has been defined as a primary cell in which the electrolyte is present in the form of a firm paste, or is retained in some absorbent material so that it does not flow if the cell is inverted. A number of cells fulfilling this definition have been put forward since the beginning of the century, but only two have been developed successfully on a commercial scale. By far the most important of these is the Leclanché type, and this is the cell normally understood by the term dry cell. It is due to the fact that the Leclanché-type cell is capable of being employed over a wide range of conditions that it has achieved its present popularity. During the past decade, however, another type of dry cell using mercuric oxide as a depolarizer has been developed commercially. This type, known as the Mallory (or mercury) cell, has special characteristics of its own, and for certain particular applications has undoubted advantages. Because of its relatively high cost and somewhat limited range of applications, in general it is likely to be complementary to the Leclanché dry cell rather than a replacement for it.

As the characteristics, construction and uses of the two types differ considerably, they will be dealt with separately here.

Leclanché-type Dry Cells and Batteries

Development of the Dry Cell

In 1868 Georges Leclanché produced the cell which bears his name. It was quickly appreciated that it had a practical advantage not possessed by other cells; virtually no action took place until the external circuit was closed, so it did not waste when not in use. Essentially it consisted of a zinc plate which was the negative pole, a solution of ammonium chloride as electrolyte, and a positive pole made by packing powdered manganese dioxide into a porous pot round a centrally-placed carbon rod. (As manganese dioxide by itself is a very poor electrical conductor, it was found necessary to admix carbon to render it conducting.) Nevertheless, it was a wet cell, and because of the obvious disadvantages of a cell containing free liquid, attempts were soon made to convert it into a dry cell.

The first "dry" cell of the modern type made its appearance in 1888, and is generally attributed to the German scientist Dr. Gassner. An important feature of his cell was that the zinc element formed into a cup was used to contain all the other constituents of the cell. Another improvement credited to Dr. Gassner is the addition of zinc chloride to the electrolyte. This considerably reduces the local action, and is one of the factors contributing to a good storage life.

Although not very efficient judged by present-day standards, it then became capable of commercial development, and soon dry cells were being produced by the Ever Ready Co. in England and other companies in Europe and America.

The impact of the First World War caused a great expansion of the industry, to meet the demand of the Armed Forces for batteries for torches, telephones and radio communication, etc.

With the advent of broadcasting in 1922 further development took place, which continued steadily through the inter-war years. The efficiency of cells was increased considerably during the same period, and the layer-type battery was developed.

The Second World War stretched the industry to the limits of its capacity, although it was now largely mechanized.

The "blackout" caused a demand without precedent for flashlight batteries. The requirements of the Services had, however, changed markedly since the First World War. Smaller, lighter and more efficient batteries for telecommunication purposes had become absolutely essential. To meet this pressing need the Ever Ready Co. began, in 1941, to produce on a large scale the Batrymax* or layer-type batteries. Since 1945 this type of battery has become generally available, and is now widely used in personal and portable receivers, hearing aids and other electronic equipment.

Construction

At the present time there are two standard methods of dry cell (and battery) construction They are: (1) the conventional round-cell construction, and (2) layer-type construction.

These are best illustrated by means of the sectional diagrams which are shown in Figs. 2 and 3.

Round-cell Batteries

These have the older conventional construction, and even today the preponderance of Leclanché-type dry batteries are still made in this way. In conditions where (*a*) a low voltage is required, (*b*) the discharge

* The word "Batrymax" is a registered trade name of The Ever Ready Co. (G.B.) Ltd., and is used in this article by permission of that Company.

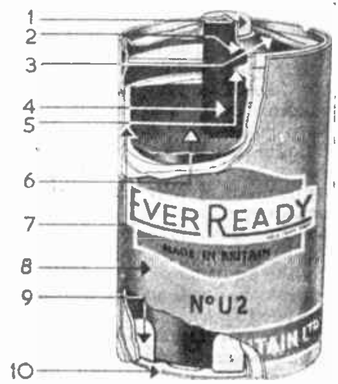


FIG. 2.—UNIT CELL (ROUND-CELL CONSTRUCTION).

1, Brass cap. 2, Insulating washer. 3. Metal seal. 4, Carbon positive electrode. This functions as a conductor, remaining unaltered by the reactions within the cell. 5, Cardwasher. 6, Depolariser. This contains manganese dioxide to act as the depolarizing agent, and carbon for conductivity. 7, Electrolyte. The principle constituents are ammonium chloride, zinc chloride, starch and wheat flour, forming a paste which sets on heating to a stiff jelly. 8, Paper tube. 9, Star-type insulating washer separating the depolarizer from the negative electrode. 10, Negative electrode of extruded zinc. When the cell is discharged, part of this is consumed to produce electrical energy, thereby becoming the negative pole.

rate tends to be heavy (say 30–300 mA per cell), and (c) the period of use per day is somewhat long (4–8 hours), this method still has advantages, partly on account of the lower internal resistance, and consequently batteries of this kind are normally employed for radio and hearing-aid L.T. purposes. Moreover, the use of round cells in parallel (or series–parallel) for the L.T. unit gives a battery of low internal resistance and relatively high efficiency to a high endpoint voltage. In the case of H.T. batteries for radio and hearing-aids or the like, where the current taken is relatively small (15 mA down to about 0·5 mA), the round-cell batteries, whilst operating reasonably effectively, do not give the *maximum* efficiency for the space available, and it is for such requirements that the layer-type batteries were developed.

Layer-type Batteries

As their name implies, layer cells are flat, and they are normally rectangular or square in section. These cells are placed one on top of the other and then bound together in good electrical contact. The bound groups of cells are known as " stacks ". The stacks can be fitted into rectangular battery cartons with a minimum waste of space. This clearly represents a considerable advance on the use of cylindrical cells which, being assembled into the square compartments created by the essential intercell insulation, results in a considerable volume of unused space. As the latter is usually filled with some inert sealing material, such as bitumen, the weight is increased without any corresponding improvement in performance. Moreover, the replacement of the carbon rod of the round cell by a thin carbon coating in the layer-type cell, and the negligible air space needed by the latter, make it the more efficient at appropriate discharge rates in comparison with the round cell.

Consequently for both technical and manufacturing reasons, this has resulted in much smaller and lighter H.T. batteries being produced for " personal " radio receivers, hearing-aids, etc.

In addition, the storage or shelf-life properties of these batteries are

FIG. 3.—HIGH-TENSION BATTERY (LAYER-TYPE CONSTRUCTION).

1, Brass positive contact. This is in electrical contact with the duplex electrode forming the positive pole of the battery. 2, Label. 3, Wax coating insulating each cell to prevent electrical leakage. 4, Binding tape. 5, Plastic cell container. 6, Depolarizer containing manganese dioxide and carbon. 7, Electrolyte soaked card separating the depolarizer from the duplex electrode. 8, Carbon-coated zinc electrode. The duplex electrode is a zinc plate covered with a layer of highly conducting carbon. 9, Brass negative contact. This is in electrical contact with the zinc plate forming the negative pole of the battery.

better than those made with the corresponding size of round cell. Consequently, for use overseas where wide variations in climatic conditions are experienced, and shelf life is an important factor, layer-type H.T. batteries are being increasingly used.

In connection with layer-type batteries the following points are of importance :

(*a*) Layer-type batteries are in the main constructed with standard stacks which are made up of ten, fifteen or twenty cells bound together. Single-layer cells (or two- or three-cell stacks) are not commercially available as yet.

(*b*) Batteries of the layer type were designed for use as radio and hearing-aid H.T. units, where the current consumption is relatively small. The advantages they possess in regard to service life will be largely, if not entirely, lost if they are called on to furnish excessive currents for fairly long periods. The suggested current range is indicated in the schedule below. Owing to the nature of its construction, the internal resistance of the layer cell is rather higher than that of the corresponding size of round cell, and in certain types of electronic apparatus (hearing aids, etc.) it is desirable that this should be taken into account.

(*c*) Because of the advantages possessed by layer-type batteries, they are being, and will continue to be, increasingly used for H.T. purposes. Consequently, when new equipment is being designed, particularly if it is portable, layer-type batteries should be used wherever practicable to provide the H.T. supply.

Physical Theory

Voltage

The initial open-circuit voltage or e.m.f. of the Leclanché cell is, in the main, determined by the respective positions of its elements (i.e., the zinc, and carbon/manganese dioxide electrodes) in the electrochemical series. It is, therefore, independent of cell size, and for practical purposes is always taken as 1·5 volts.

The exact figure depends on a number of factors, such as the nature and origin of the manganese dioxide, the type of carbon material used, the composition of the electrolyte paste, the age of the cell, etc.

With reasonably fresh cells it usually ranges from 1·5 to 1·6 volts, but more generally the e.m.f. is found to be in the region of 1·53–1·56 when measured with a high-resistance voltmeter. The open-circuit voltage of an unused cell provides no criterion of its service-life performance, i.e., a cell with an e.m.f. of 1·6 volts will not necessarily give a longer life than a similar cell measuring 1·5 volts.

Providing the previous history of the cell or battery is known (and this is usually available to the maker), the e.m.f. provides a useful guide to the condition of a battery which has undergone discharge.

Internal Resistance

The internal resistance (or I.R.) of a cell or battery depends principally on the size and construction of the cell and the method of measurement. Usually it is calculated from a measurement of the e.m.f. and the short-circuit current read instantaneously on a short-circuit ammeter (i.e.. one of which the total resistance including leads does not exceed 0·01 ohm). It is given by the formula :

$$r = \frac{E}{I}$$

where r = internal resistance of the cell (ohms) ;
E = e.m.f. (measured on a 1,000-ohms/volt voltmeter) (volts) ;
I = short-circuit current (amperes).

Another standard method of measurement is to take the e.m.f., using a high-resistance voltmeter having little or no overswing, and then shunt the cell with a resistance of 1 ohm and take an instantaneous voltage reading. The internal resistance r is then calculated using the formula :

$$r = \frac{E - E_1}{E_1} \times S$$
$$= \frac{E - E_1}{E_1} \times 1$$

where E = e.m.f. of cell ;
E_1 = instantaneous closed-circuit voltage ;
S = resistance of shunt (1 ohm).

It will be evident that these methods determine the D.C. resistance or apparent internal resistance of a cell, and in actual practice these are the only measurements made, although the shunt resistances may be varied from 1 ohm if necessary. The higher the resistance of the shunt (i.e., the lower the current taken from the cell), the higher the value obtained for the internal resistance.

The internal resistance of cells has been measured by A.C. methods, but the results obtained bear no relationship to those by the methods outlined above, and they are not normally used for practical purposes.

Contrary to popular belief, the short-circuit current is not a measure of the electrical output which may be obtained from a cell. That is, the short-circuit current in amperes is not related to the ampere-hour capacity of a cell, nor is any other measurement of the internal resistance of the cell.

The internal resistance of a cell or battery, however, merits consideration if: (*a*) the current drain is comparatively heavy, or (*b*) if the internal resistance is unusually high. To illustrate case (*a*) when a flashlight bulb, the filament of which consumes 300 mA at 3·0 volts, is connected to a small 3-volt torch battery having an internal resistance of 1 ohm, the instantaneous drop on load due to the internal resistance is $e = Ir$ or $e = 0{\cdot}3 \times 1$. That is, quite apart from polarization, immediately the bulb lights, the battery voltage will fall to 2·7. In case (*b*) if the current is less but the internal resistance is high, a similar instantaneous on-load voltage drop would be observed.

In general, if the discharge rate is low in relation to the cell size the internal resistance is usually unimportant, but if the discharge rate is relatively heavy it may be of some consequence.

Chemical Theory

Whilst the reactions occurring during discharge in the Leclanché dry cell are very complex, the chemistry of the dry cell is essentially that of the dissolution of the zinc element by the electrolyte and the collection of the current by the carbon element which conducts it to the external circuit.

A more detailed description of the principal chemical reactions occurring when a dry cell is discharged is given below. In the Leclanché cell the actual reactions which occur when the cell is discharged through an external circuit are very complicated.

In simplified form the overall reaction is as follows:

$$Zn + 2NH_4Cl + 2MnO_2 = ZnCl_2 + 2NH_3 + Mn_2O_3 + H_2O$$

This overall reaction consists of two fundamental chemical reactions viz.:

$$(a)\ Zn + 2NH_4Cl = ZnCl_2 + 2NH_3 + 2H$$

$$(b)\ 2MnO_2 + 2H = Mn_2O_3 + H_2O$$

These chemical reactions may be expressed ionically:

At the anode (negative pole)

$$(c)\ Zn + 2Cl^- = ZnCl_2 + 2e$$

At the cathode (positive pole)

$$(d)\ 2MnO_2 + 2NH_4^+ + 2e = Mn_2O_3 + 2NH_3 + H_2O$$

The ammonium and chloride ions are furnished by the dissociation of the aqueous ammonium chloride electrolyte:

$$(e)\ 2NH_4Cl = 2NH_4^+ + 2Cl^-$$

For convenience, the reactions relating to the anode and cathode are considered separately below, although in a cell during discharge they are interdependent.

Reactions Involving Anode (Negative Pole)

The zinc goes into solution forming zinc chloride as indicated in equation (*c*), and ammonia is liberated simultaneously at the cathode as shown by equation (*d*).

The ammonia diffusing through the electrolyte combines with the zinc chloride and produces the transparent needle-shaped crystals seen in slightly discharged cells (and also in cells which have been stored unused for long periods).

$$(f)\ ZnCl_2 + 2NH_3 = ZnCl_2.2NH_3 \text{ (sometimes written, } Zn(NH_3)_2Cl_2).$$

In completely exhausted cells the reactions go some stages further and give rise to opaque white deposits containing basic zinc chlorides such as Zn(OH)Cl, etc.

The ionic equation (*c*) also shows that the two electrons liberated when each zinc atom goes into solution leave the zinc element negatively charged.

Both equations (*a*) and (*c*) show that theoretically in accordance with Faraday's laws 32·69 gm. of zinc produce electricity to the extent of 96,500 coulombs. That is 1 gm. of zinc should produce $(96{,}500 \div 32{\cdot}69) \times 1/3600 = 0{\cdot}820$ ampere-hours.

In practice, this is very nearly, but not quite realized owing to local action.

Reactions Involving Cathode (Positive Pole)

The depolarizing action of the manganese dioxide is due to its oxidizing action, and as may be seen from equation (*d*) this results in the formation of water with the consequent reduction of the dioxide (MnO_2) to manganic oxide (Mn_2O_3).

The precise mechanism of the reduction of the manganese dioxide is not fully understood. Some investigators have considered its action to be ionic, whilst others more recently have regarded it as a gas-solid phase reaction. From equation (*d*) it will be seen that on reaching the carbon–manganese dioxide cathode the ammonium ions acquire electrons from it, leaving the electrode positively charged.

The way in which the current flows through the external circuit and the battery is readily seen from these reactions.

On closing the circuit, electrons flow from the zinc element through the connecting wire to the carbon rod. Inside the cell negatively charged ions carrying electrons move steadily through the electrolyte from the manganese dioxide–carbon cathode to the zinc element, whilst simultaneously positive ions migrate towards the carbon–manganese dioxide electrode. As electrons are negatively charged, this is precisely the same as a positive current flowing in the reverse direction. Conventionally, all electrical instruments show the direction of flow of positive electricity, and hence if a milliammeter is inserted in the external circuit, the current will be observed to flow from the carbon to the zinc element.

The keeping properties of cells are frequently a matter of interest. These are also connected with the chemical and physical properties of the constituents. In the main, however, the problem becomes one of reducing the extent of attack on the zinc element, and ensuring as far as possible its even distribution. It has been found that the following are beneficial :

(1) amalgamation of the zinc surface with mercury ;
(2) the use of zinc chloride in the electrolyte ;
(3) the employment of flour in the paste which acts as an inhibitor ;

(4) avoidance of soluble deleterious impurities in the materials, as for example copper or copper compounds.

The main reactions have been dealt with above, but there are a number of secondary reactions other than those described which can and frequently do occur, depending on the conditions of discharge. Owing to their complexity, they fall outside the scope of this article.

Performance and Use

The amount of electricity which can be obtained from a dry cell depends on a number of factors, of which the more important are :

(1) the physical size of the cell;
(2) the method of construction and the skill of the manufacturer;
(3) the rate at which the cell is discharged;
(4) the period of time per day for which it is used;
(5) the voltage endpoint to which it is required to operate;
(6) the temperature;
(7) the age of the cell.

The bearing which some of these have on the service life of a battery is indicated in the following sections.

Size

The effect of the size on performance is best seen by means of the curve in Fig. 4, which shows the result when two hearing-aid-type

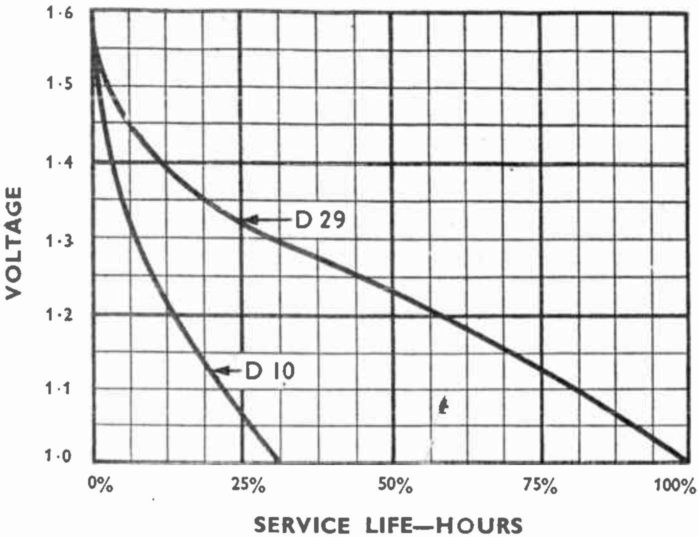


Fig. 4.—The Effect of a Battery's Size on its Performance.

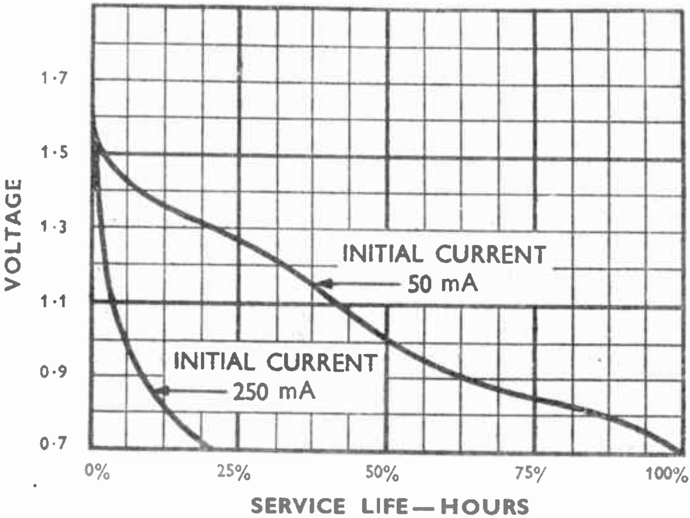


FIG. 5.—THE BEARING OF THE DISCHARGE RATE OF A BATTERY ON ITS LIFE.

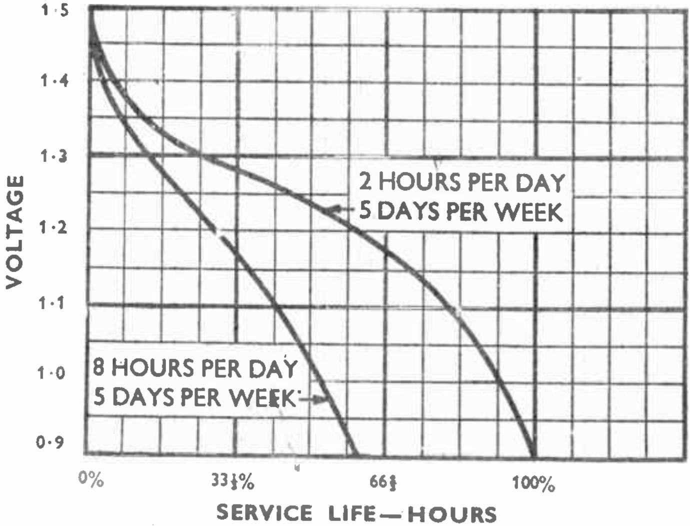


FIG. 6.—THE EFFECT OF THE TIME CYCLE ON A BATTERY'S LIFE.

batteries are discharged through a resistance of 25 ohms for 8 hours per day, five days per week, at 20° C. The D10 type is of ¾ in. diameter and 2¼ in. height, whilst the D29 type is of 1 in. diameter and 3 in. height

Discharge Rate

The bearing of the discharge rate on the life of a battery is illustrated by the discharge curve given in Fig. 5.

Period of Use

The effect of the time cycle (i.e., period of use per day on a given number of days per week) is shown in Fig. 6.

Voltage Endpoint

The special importance of this factor can best be demonstrated by means of a discharge curve as shown in Fig. 7.

Temperature

The effect of temperature on performance varies so widely, depending on the other conditions of use (discharge rate, endpoint, etc.), that only very general statements are possible. Over the temperature range from +20° to −20° C., the service life falls with decreasing temperature, and all standard Leclanché-type cells become for most practical purposes inactive in the region of −20° C.

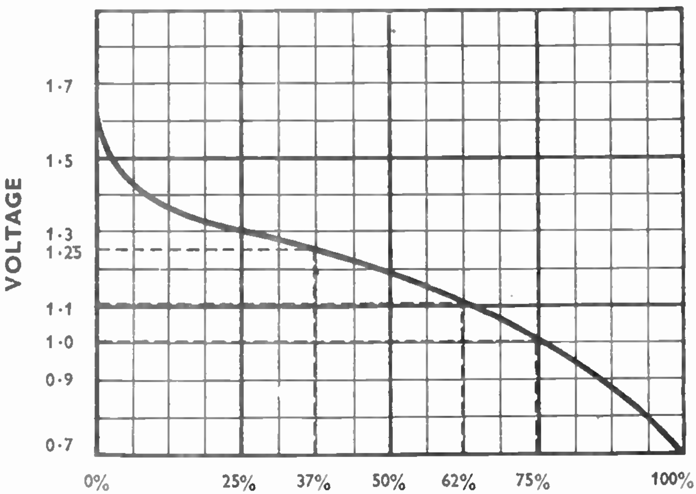


Fig. 7.—The Effect of the Voltage Endpoint on the Life of a Battery.

It may be of interest to mention that the effect of temperature on the open-circuit voltage of unused cells has been determined, and over the range of $+25°$ to $-20°$ C. has been found to be approximately 0·0004 volts/cell/1° C. This figure, however, varies somewhat with the constitution of the cell.

To obtain the best results batteries should be operated wherever possible at about 20° C.

Age of Cells

Here also it is only possible to generalize as to the effects of storage on the service lives of cells. It is an empirical but well-established fact that the larger the size of cell, the better its storage properties. An indication of the storage effect is that after keeping a cell of approximately U2 size for six months from the time of manufacture at 20° C., a loss in service life of only 5–10 per cent would be expected. This figure is, however, dependent on the conditions of use after storage.

Storage at high temperature (e.g., 50° C.) has a much more marked effect on performance, and wherever possible steps should be taken to avoid storage (or use) at such temperatures.

Choice of Battery

The selection of the appropriate dry cell or battery for a particular purpose is usually a matter of compromise. Size, weight, performance and cost, etc., all enter into it, but usually one factor is predominant.

There are, however, certain general considerations which should be studied if the best results are to be obtained. These will now be considered under the main general purposes.

H.T. Batteries

Where portability or maximum performance in minimum space are of prime importance layer-type batteries have marked advantages. They have a good storage life, and give reliable service under exacting climatic conditions.

The minimum voltage commercially obtainable with layer-type batteries is 15. They were designed for radio rates of discharge, e.g., from 0·1 to 20 mA according to battery size, and therefore are not used for radio L.T. or similar purposes, where the current drain may be in the region of 200–300 mA.

L.T. Batteries

If the current drains are relatively high (100–400 mA) and a long service life to a high endpoint voltage (1·0–1·2 volts for a 1·5-volt battery) is required, batteries containing medium-size round cells in parallel will give the best results.

When portability is a prime consideration and the discharge rate is of the order of 20–60 mA for 1–8 hours/day to an endpoint of 1·0–1·2 volts (for a 1·5-volt battery) another specially designed cell is needed. an example of which is the Ever Ready Hearing Aid L.T. Battery Series, For intermittent use and where very long storage-life is more important than maximum output to a high endpoint voltage, the larger dry cells have a unique place.

For portability and a relatively high discharge rate (200–300 mA) as required for torches, "personal" radios, etc., a further type of cell is required, and an example is the Ever Ready "Flashlight" series.

Combined H.T.–L.T. Batteries

For technical reasons connected with valve performance, and because of simplicity in handling, the combined H.T.–L.T. battery was developed in this country for radio purposes. When separate H.T. and L.T. batteries are used, the common practice has been to employ a relatively small L.T. unit, so that two or three L.T. batteries are required to equal the life of the H.T. unit.

Under-heating the valve filament is inevitably detrimental to its emissive properties but of far less consequence if the H.T. voltage is correspondingly reduced. With separate batteries, conditions arise towards the end of the life of the first, and even perhaps second, L.T. battery whereby the filament voltage is very low whilst the H.T. is still quite near the maximum rating. Using combined batteries, properly chosen for balance, the H.T. is reduced to perhaps half its maximum value by the time the L.T. reaches its minimum, thus considerably reducing the effects of under-heating. With the combined battery, the condition of low L.T. and high H.T. should not arise.

The combined H.T.–L.T. batteries are each designed for specific conditions of use, and these must be observed if the best results are to be obtained. For instance, the Ever Ready B103, 90 + 1½-volt battery is balanced for a L.T. consumption of 250 mA at 1·4 volts with an on-load endpoint of 1·0 volt, and a H.T. consumption of 12 mA at 90 volts, and an on-load endpoint of 35 volts, the period of use being 3 hours/day Under these conditions a balanced life of approximately 300 hours is to be expected when used in a set.

Makers' suggested current ranges should be considered in conjunction with the period of use per day and the endpoint voltage. If the period is long (4–8 hours/day) and the on-load endpoint high, i.e., 1·0 volt/cell or above, the current drain should if possible tend towards the minimum figure given in the schedule. Conversely, if the time period is short (½ hour per day) and the voltage endpoint normal (0·75 volt/cell) the current may reasonably approximate to the higher figure.

Batteries for Transistorized Equipment

Batteries intended primarily for use with transistorized equipment have been introduced by a number of makers. These generally have an output voltage of 6 or 9 volts, or (for use with some "single ended Class B" circuits) a twin output of 4·5 + 4·5 volts with provision for the connection of the two sections in series or parallel. Suggested current ranges for these batteries, using layer-type construction vary from 0–5 mA for the smaller units up to about 20–150 mA. Batteries with a current range of 5–50 mA are generally suitable for domestic broadcast transistorized receivers, and similar equipment.

Mallory-type Dry Cells and Batteries

Development of the Mallory Cell

Whilst the underlying principles of the Mallory cell have been well known for a considerable time, it was not until the Second World War that it was developed commercially by the Ruben–Mallory Company of America, for military communication purposes. These batteries are now generally available, and are made in this country by Mallory Batteries Ltd.

The essential elements of the Mallory (mercury) cell are: (1) a zinc anode; (2) a caustic alkali (potash) electrolyte; and (3) a cathode consisting of a compressed mercuric oxide–graphite mix-cake in firm contact with a steel container.

Construction

There are at present two standard types of Mallory construction. These are: (*a*) the rolled anode form, and (*b*) the pressed powder anode type.

The rolled-anode type is so-called because the zinc element consists of a strip of thin corrugated zinc of considerable length spirally wound or rolled to form the negative pole. In the case of the pressed powder form the zinc anode consists of a porous pellet made by compressing powdered zinc metal.

In Fig. 8 the 4R cell illustrates the rolled-anode style and the RM4 the powdered-anode type.

Theory

Physical

Voltage.—The e.m.f. is by and large determined by the positions of the elements (i.e., mercuric oxide/carbon and zinc) in the electro-chemical series. In consequence, it is independent of the size of the cells, and is 1·35 volts.

Internal Resistance.—This is usually measured by the short-circuit current method as described for Leclanché dry cells.

Chemical

With the Mallory cell the reactions which occur when the cell is discharged amount to the oxidation of the zinc anode.

The overall reaction is given by the following chemical equation:

$$Zn + H_2O + HgO = ZnO + H_2O + Hg$$

or simply

$$Zn + HgO = ZnO + Hg$$

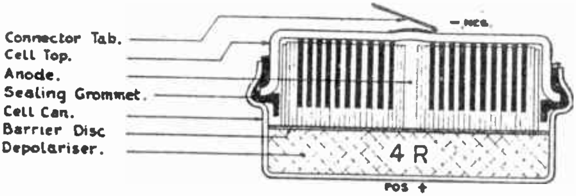


Fig. 8 (*a*).—Rolled-anode Type Cell.

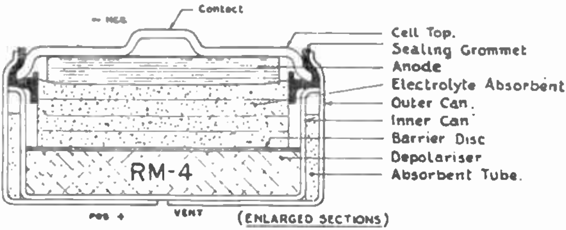


FIG. 8 (*b*).—POWDERED-ANODE-TYPE CELL.

This chemical reaction may be expressed ionically in terms of the reactions at the two poles :

At the anode (negative pole)

(1) $Zn + 2OH^- = ZnO + H_2O + 2e$

At the cathode (positive pole)

(2) $HgO + H_2O + 2e = Hg + 2OH^-$

Performance and Use

The electrical output obtainable from a " mercury " cell is determined by the following factors :

(1) the physical dimensions of the cell;

(2) the method of construction and the manufacturer's skill;

(3) the rate at which it is discharged;

(4) the on-load voltage endpoint to which it has to operate;

(5) the temperature.

It is of interest to note that in the case of Mallory cells the performance differs little whether the discharge is continuous or intermittent, provided the discharge rate is normal. That is it differs from the Leclanché cell in that no appreciable recuperation occurs in rest periods. However, if the discharge rate is abnormally heavy, some benefit is derived from intermittent service.

Size.—The effect of size is shown in Table 2 and the curves in Fig. 9.

Construction and Manufacture.—Whilst more than one kind of mercury cell exists, Messrs Mallory were the pioneers in this field, and their cells embody the results of years of experience.

Discharge Rate.—The general effect of the discharge rate is seen in the curves shown in Fig. 9. The recommended maximum discharge rates for these cells are given in Table 3.

It is undesirable that the maximum discharge rates should be materially exceeded, as this reduces the service life disproportionately.

Temperature.—Where practicable these batteries should be operated at about 20° C. (68° F.), as this is approaching the point of optimum performance. Since the ampere-hour capacity of these cells falls

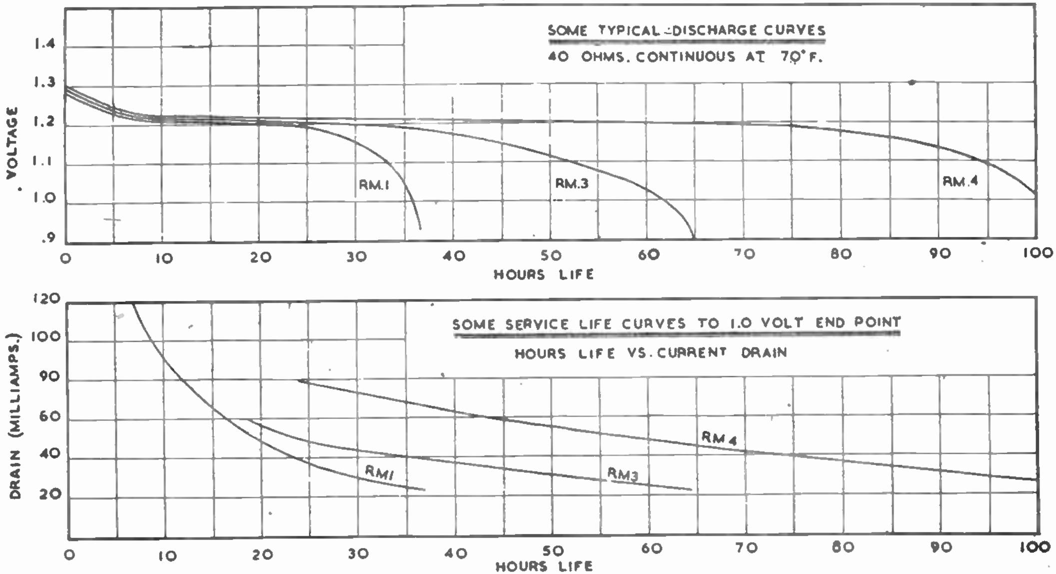


Fig. 9.—The Effect of Size on the Mallory-type Dry Cell.

TABLE 2.—EFFECT OF SIZE ON THE PERFORMANCE OF THE MALLORY-TYPE BATTERY

Type	*Nominal Voltage*	*Diameter (In.)*	*Height (In.)*	*Volume (Cu. In.)*	*Weight (Oz.)*	*Capacity (mAH)*	*S.C. (Amp.)*
RM1	1·35	0·625	0·65	0·20	0·43	1,000	0·7
RM3	1·35	0·977	0·66	0·50	1·10	2,200	0·6
RM4	1·35	1·192	0·66	0·74	1·65	3,400	1·1

TABLE 3.—MAXIMUM RECOMMENDED DISCHARGE RATES

*RM*1	*RM*3	*RM*4	1*R*	2*R*	3*R*	4*R*
50 mA	50 mA	70 mA	20 mA	30 mA	50 mA	70 mA

somewhat sharply with decrease in temperature, it is very undesirable to run them at low temperatures (i.e., around 0° C.).

At relatively high temperatures (50° C.) they behave well, and after six months' storage at 50° C. the loss in service life is only 5–10 per cent.

Choice of Battery

When selecting Mallory cells for a given application, the designer is concerned with a number of factors, i.e., size, weight, performance and operating cost. Usually, however, one is of particular importance, and this determines the type of cell required.

For L.T. purposes the steel-jacketed types RM1, RM3 and RM4 should be used, whereas for H.T. units the 1R series (which do not require the outer steel jacket because they are protected by the battery container) are employed.

At present the principal use of the RM1 series is for the L.T. unit in hearing aids, and they are used in conjunction with miniature or sub-miniature layer-type (Leclanché) H.T. batteries.

The 1R series of cells are employed principally in Government batteries for communication purposes.

E. S. B.

SECONDARY BATTERIES

Uses

The most important application of batteries to transmitting equipment is as a primary source of supply for mobile and portable transmitters. They are also used :

(*a*) Where a pure D.C. supply immune from the risk of failure of supply mains is essential.

(*b*) For intermittent use where there is no mains supply, and it would be uneconomical to run a small engine generator for short periods of time.

(*c*) For portable auxiliary equipment, such as field-strength measuring apparatus.

(*d*) For starting automatic generating plant.

(*e*) For D.C. supply to circuit-breakers with shunt trips.

For use with mobile equipment, blocks of three or six cells in sealed containers, having capacities up to 200 ampere-hours are commonly made up in multiples of 6, 12 or 24 volts to drive rotary transformers. Small Planté-type cells for fixed station work have glass containers and are installed on single- or double-tiered wood stands, liberally treated with acid-resisting paint. Cells of more than 300 ampere-hours capacity are contained in lead-lined wood boxes, erected on low stands for accessibility.

Nickel–Iron Batteries

These batteries are a suitable alternative to the lead–acid type for fixed and mobile installations, where they are likely to be subjected to rough treatment, where accommodation space is limited and the fumes from a lead–acid battery would harm nearby apparatus. or where they are required for occasional intermittent working.

They are higher in cost than lead–acid batteries and have a higher internal resistance and therefore a lower efficiency. The high internal resistance is, however, often a useful feature in limiting the discharge current and safeguarding the battery against too heavy a discharge. Against this, their advantages over lead–acid cells are :

(1) They can be charged and discharged at higher rates.

(2) They remain charged or discharged for indefinite periods without injury.

(3) Little sediment is formed.

(4) They do not give off corrosive gases.

(5) They are lighter and more robust.

(6) They have a long life.

The ampere-hour capacity is usually specified at the 7-hour charge rate and the 5-hour discharge rate. Higher charging rates than this should not be used, otherwise the iron element will be permanently damaged. The electrolyte must be renewed when the specific gravity falls below 1·16. The battery must then be discharged completely and left short-circuited for 2 hours or more, and refilled with fresh electrolyte. When it is required to take the battery out of service, it must be completely discharged and stored with its terminals short-circuited.

Battery Charging and Discharging

The ampere-hour capacity of a battery diminishes as the discharge current is increased (or the discharge time is reduced), and it is therefore usually specified at the 10-hour rate of discharge for lead–acid cells.

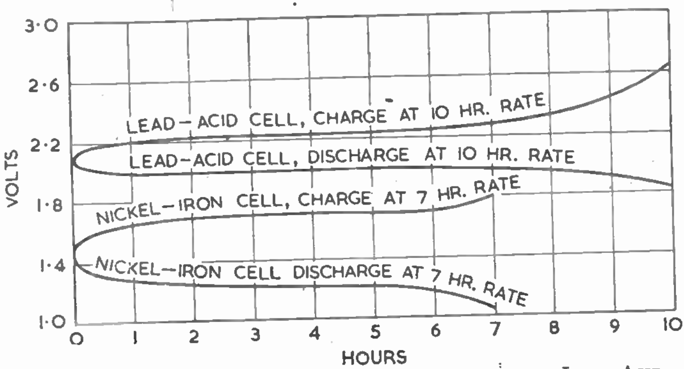


FIG. 10.—CHARGE AND DISCHARGE VOLTAGE CURVES OF LEAD–ACID AND NICKEL–IRON CELLS.

If this is taken as 100 per cent, the capacity for higher rates is approximately as follows :

TABLE 4.—BATTERY CAPACITY FOR VARIOUS DISCHARGE RATES

Duration of discharge (hrs.)	10	9	8	7	6	5	4	3	2	1
Capacity, of 10-hr. rate (%)	100	98	95	92	88	83	78	72	63	50

Battery efficiency is defined either on a current or energy basis.

$$\text{AH (or current) efficiency} = \frac{\text{AH output}}{\text{AH input}} = \begin{cases} 90\% \text{ average for lead–acid cells} \\ 82\% \quad ,, \quad ,, \text{ nickel–iron cells} \end{cases}$$

$$\text{WH (or energy) efficiency} = \text{AH efficiency} \times \frac{\text{average discharge volts}}{\text{average charge volts}}$$

$$= \begin{cases} 75\% \text{ average for lead–acid cells} \\ 68\% \quad ,, \quad ,, \text{ nickel–iron cells} \end{cases}$$

Fig. 10 shows the way in which the voltage of lead–acid and nickel–iron cells rises during charge and falls during discharge. The characteristics for both types are similar, but the voltage of the nickel–iron cells is lowor than the lead–acid cells. The voltage at any instant is given by

$$E = E_i + IR \text{ while charging}$$
$$E = E_i - IR \text{ while discharging}$$

where E_i = e.m.f. of battery off load;
I = charging or discharging current;
R = internal resistance of battery (= no. of cells × internal resistance per cell).

Thus, if the battery is switched from charge at a given rate to discharge at the same rate, the terminal voltage will fall by an amount $2IR$ volts.

The normal charge and discharge voltages for lead–acid and nickel–iron cells, for which charging facilities must be provided, are shown in Table 5.

TABLE 5.—CHARGE AND DISCHARGE VOLTAGES FOR LEAD–ACID AND NICKEL–IRON CELLS

	Lead–Acid Cells	*Nickel–Iron Cells*
Initial discharge voltage per cell	2·5	1·5
Lowest discharge voltage per cell	1·8	1·1
Charging voltage range	2·0–2·75	1·4–1·7

For example, suitable charging data for a lead–acid battery of 55 cells, 120 ampere-hour capacity, at the 10-hour rate would be :

Normal charging current	= 120/10	= 12 A
Minimum charging voltage	= 55 × 2·0	= 110 V
Maximum ,, ,,	= 55 × 2·75	= 151 V
Charging voltage range	= 110 to 151 V	
Series regulating resistance	$= (E_{max.} - E_{min.})/I$	
	$= (151 - 110)/12 = 3{\cdot}42\ \Omega$	

The rapid fall of voltage towards the end of discharge is a useful indication of the safe limit to which discharging can be carried, but the final criterion must always be the specific gravity of the electrolyte and the voltage of the individual cells while on load. It is usual with a large battery to supplement the switchboard voltmeter by charge and discharge ampere-hour meters, the discharge meter being adjusted to read about 10 per cent high to compensate for inefficiency of the battery. By logging readings of both meters, it is possible to estimate at any time the balance of charge remaining.

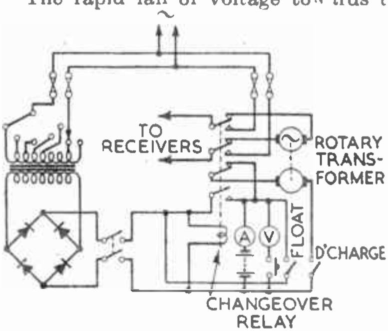


FIG. 11.—METHOD OF PROVIDING AN EMERGENCY A.C. SUPPLY WITH A FLOATING BATTERY.

Floating Batteries

The floating battery is a convenient way of providing a ripple-free supply from an A.C. rectifier, which automatically delivers current to the load if the mains supply should fail. Fig. 11 shows a typical full-wave rectifier and floating battery, with facilities for floating, charging or discharging.

W. E. P.

Secondary Cells

The trend of development for many radio and electronic applications is towards smaller and lighter accumulators. In lead–acid types considerable research into new forms of separator plates has been undertaken to overcome the effect of acid on the active plates. One development is the use of separator plates filled with diatomaceous earth under compression. Apart from enabling the active plates to be lighter and softer, this type of separator absorbs the bulk of the electrolyte, so that the cell can be regarded, for many purposes, as a dry cell. Another innovation is the use of sheets of resin-bonded glass wool and porous P.V.C. forming the separator which is also used to hold the active plates in position.

Silver–Zinc Accumulators

For applications which require an accumulator of extreme compactness while permitting relatively high charge and discharge rates over a wide range of temperature, a recent development is the silver–zinc accumulator. Silver is used for the positive electrode, zinc oxide for the negative electrode and potassium hydroxide forms the alkaline electrolyte. The plates and separators are held under pressure, using a cellulosic material. The voltage per cell is about 1·5 volts. Typical figures of overall performance are a capacity of about 32 ampere-hours per pound weight and about 2 ampere-hours per cubic inch.

SILICON SOLAR BATTERIES

It has been estimated that up to 1,000 watts of light energy reaches a square metre of the earth's surface at noon in favourable locations. In 1954 Bell Telephone Laboratories announced the development of silicon solar batteries with a conversion efficiency of the order of 5 per cent; though at a cost per cell much higher than conventional power sources. Recently solar batteries of higher efficiency and lower cost have been marketed by several firms in the U.S.A.: these cells have an efficiency of 10–11 per cent, and it is anticipated that further improvements towards the theoretical maximum of 22 per cent may occur. These efficiencies are very much higher than is possible with selenium cells.

The silicon solar cell comprises essentially a silicon *p–n* junction; light energy liberates electrons and holes within the barrier region, and the permanent electric field existing in this area causes the electrons to flow to the *n* side of the junction and the holes to the *p* side. Thus a voltage appears across the thin slice of silicon which is usually about $\frac{1}{25}$ in. thick.

A typical solar cell illuminated at 10,000 foot-candles will produce an open-circuit voltage of the order of 550 mV. The current will depend on the active area of the cell: about $\frac{3}{4}$ sq. in. being required to produce a short-circuit current of about 100 mA and a power output of the order of 30 mW.

Cells may be used: (1) for direct conversion of light energy into electric power, whenever sufficient light is available; (2) as a source of power for re-charging storage batteries to provide continuous service; (3) as control elements in photo-electric applications.

VIBRATORS AND VIBRATOR CONVERTERS

The vibrator plays an important part in modern mobile electronic equipment. It is not generally realized that vibrator units are one of the most efficient power-converting devices so far developed. The frequency stability of these units is unmatched by any other device comparable in price, size and simplicity. Although the vibrator unit has many advantages, there are, however, limitations which are not generally understood by the casual user. The various main types of vibrator unit, their applications and limitations will now be described.

Basic Vibrator Circuits

The simplest type of vibrator is shown in Fig. 12. This consists of a ow-impedance vibrator exciter coil in series with the make/break contacts and the primary of the output transformer. This circuit is very seldom used, as the output is limited to under 10 watts. The load should be a full-wave rectifier in order to reduce arcing at the contacts to a minimum. The output from this unit is not suitable for operating inductive devices, such as transformers, small motors and the like. The vibrator driving current is proportional to the load; it therefore follows that the vibrator reed "swing" amplitude will vary according to the load, resulting in poor frequency stability. As the current into the output transformer is, in effect, half-wave, the wave-form is unbalanced.

For loads ot up to 30 watts, the vibrator circuit shown in Fig. 13 is generally used. The output from this type is full-wave, and the wave-form is, therefore, balanced.

The vibrator driving coil is, however, still in series with one half of the output transformer, but, in this case, it does not carry the full input current. Nevertheless, the vibrator drive will vary according to the load. This circuit will give satisfactory results when operated at a

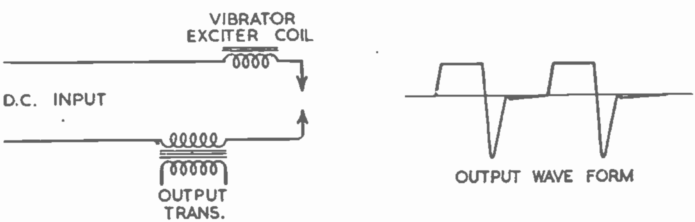


Fig. 12.—Basic Vibrator Circuit.

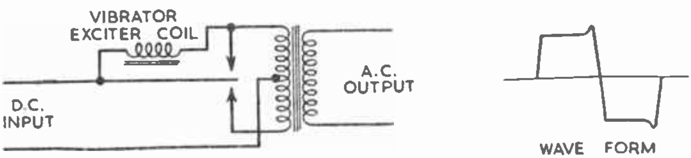


Fig. 13.—Typical Vibrator Circuit for Loads up to 30 Watts.

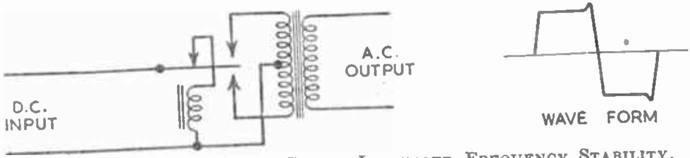


FIG. 14.—VIBRATOR CIRCUIT GIVING INCREASED FREQUENCY STABILITY.

frequency of 100–130 c/s. The output/input efficiency will be comparatively high, but the frequency stability will be poor. For this reason, the circuit is only considered suitable for feeding into a full-wave rectifier circuit.

For greater frequency stability, the circuit shown in Fig. 14 is to be recommended. Basically, this is similar to the circuit shown in Fig. 13, with the exception of the driving coil, which in this case is completely independent of the output circuit. For loads of up to 30 watts, this provides an efficient, economical circuit, coupled with good frequency stability.

All the circuits so far described depend for their operation on the use of a centre-tapped transformer, but in cases where the output voltage required is approximately the same as the input, the circuit in Fig. 15 is very satisfactory. This is in effect a double-pole, double-throw, change-over switch which reverses the polarity of the output at the periodicity determined by the natural frequency of the driving reeds. As there is no transformer winding in circuit to limit the input current, a small resistance R must be inserted in series with the input. Subject to the current-carrying capacity of the contacts, this type of vibrator circuit is capable of an output of up to 100 watts; with careful attention to the output power-factor correction, and suitable buffer capacitors, an output of up to 300 watts is possible.

All the circuits so far described provide an A.C. output with a substantially square-wave form output which, for general requirements, is quite satisfactory, especially where the output is converted back to D.C. by rectification. When a square waveform is used to provide the driving power to electronic equipment, there is the possibility that "hum" or ripple may be introduced by direct coupling or induction. For such requirements vibrator units have been developed which provide an output waveform which approximates that obtained from commercial power supplies.

The circuit shown in Fig. 16 is widely used in vibrator converters. The A.C. output is not a sine wave, but it is sufficiently close for practical

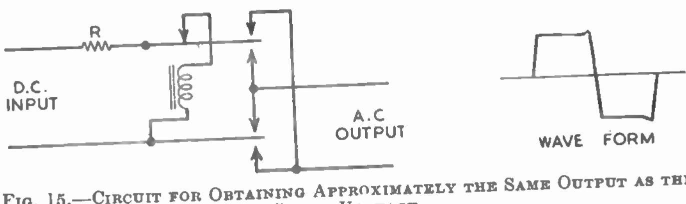


FIG. 15.—CIRCUIT FOR OBTAINING APPROXIMATELY THE SAME OUTPUT AS THE INPUT VOLTAGE.

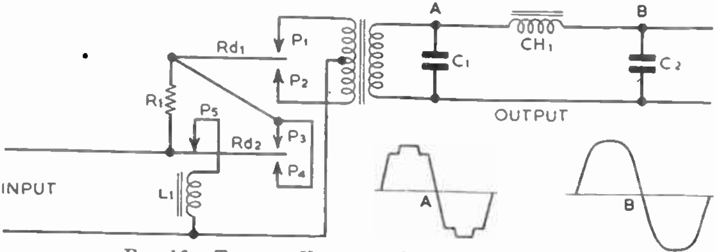


FIG. 16.—TYPICAL VIBRATOR CONVERTER CIRCUIT.

purposes to allow the operation from a D.C. source of such devices as tape recorders and dictating machines. The vibrator shown in this circuit utilizes a commutating circuit and functions in the following manner: The driving circuit is made up of the reed RD_2, the exciter coil L_1 and contact P_5. Contacts P_1 and P_2 are closer to reed RD_1 than contacts P_3 and P_4 are to reed RD_2. With a slight movement of reeds RD_1 and RD_2 (which, incidentally, are mechanically coupled) a current flows through R_1, P_1 and one half of the transformer primary: as the reeds reach approximately 50 per cent of the "swing", contact P_3 makes contact with reed RD_2, thus short-circuiting resistance R_1, and so applying the full input potential to half of the output-transformer winding. On the return half-cycle of the reeds, resistance R_1 will be re-introduced before the circuit is finally broken by P_1. The same process is repeated on the other half-cycles of the reeds with contacts P_2 and P_4. This circuit has three advantages over those previously described:

(a) As the input source is applied in steps, the current build-up in the transformer is more gradual.

(b) Resistance R_1 acts as a buffer and damper, reducing the possibility of arcing at the contacts, particularly at high input currents.

(c) The resultant output waveform shown in Fig. 16 (a) is quite easily corrected to the waveform shown in Fig. 16 (b) by a simple filter made up of capacitor C_1, iron-core choke CH_1 and capacitor C_2.

It should be noted that in all typical circuits illustrated, no buffer capacitors or power-factor correction are shown; these will, of course, be incorporated in any practical design, the actual values depending on the nature of the load and the input voltage.

Main Classes

Having described the various vibrator circuits available, it will be noted that these fall into two main classes:

Class 1.—Units for loads of up to 30 watts where the output is usually rectified to D.C. before being applied as H.T. drive to equipment such as car radio, radios and mobile amplifiers, etc. In this type of equip-

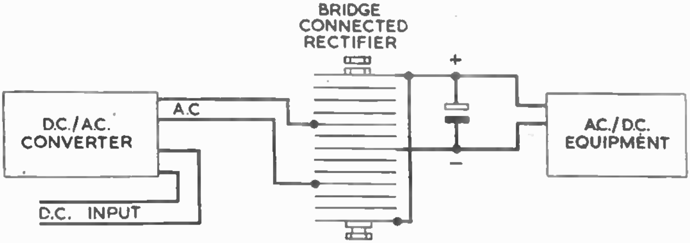


FIG. 17.—USE OF A D.C./A.C. CONVERTER WITH A.C./D.C. EQUIPMENT.

ment the operating conditions of the vibrator unit are known, simplifying the design considerably.

Class 2.—Vibrator units in this class provide a very much larger output, A.C. at 50 or 60 c/s, and in the case, for example, of Valradio units, outputs of up to 300 watts are available. The design of these D.C./A.C. converters is complicated by the fact that the characteristics of the load are not known. It is therefore necessary to incorporate constants which will fulfil most average requirements. There will inevitably be cases where attention will have to be given to the power factor of the load, and in particular to the output voltage of the converters, before the equipment is put into service. This is particularly the case when the vibrator feeds certain types of motors.

A.C. Equipment Employing Half-wave Rectification

Any load which will pass more current during one half-cycle than the other, will cause magnetic polarization of the transformer, resulting in serious arcing at the contacts; for this reason A.C. equipment incorporating half-wave rectification cannot be successfully used on vibrator D.C./A.C. converters.

Universal A.C./D.C. Equipment

When it is desired to operate A.C./D.C. equipment from a supply lower in voltage than that for which it was designed, a D.C./A.C. converter can be used, provided that a full-wave, bridge-connected, metal rectifier is interposed between the converter and the A.C./D.C. equipment, as shown in Fig. 17. This circuit arrangement is very satisfactory for operating television receivers from low-voltage supplies. The fact that the input to the television receiver is smoothed D.C. completely eliminates "hum bars" on the picture, and hum on sound.

Input Voltage

The input voltage to vibrator units should be maintained within plus or minus 10 per cent of the nominal. Where this figure is exceeded a small resistance must be connected in series with the input. A value of 1 ohm for every 2 volts in excess of the converter maximum would be suitable. For example :

The maximum input of a Valradio 32-volt converter is 36 volts. Assuming that when the batteries are being charged the voltage rises to 40 volts, the excess is 4 volts; a 2-ohm series-resistance, would therefore be suitable. In some converters a switch is provided for short-circuiting the resistance. The switch is marked "charge" and "battery".

Any rise in the input voltage will produce proportionate increase in the output voltage. To cope with change of input and variations of load, some converters are provided with three or four output tappings. These will usually be found on a panel inside the converter, and are marked "H" (High output), "M" (medium) and "L" (low output). Access to these tappings is gained by removing the base of the converter. It is always best to try the "L" tapping first and measure the output voltage under load conditions.

Interference Suppression

For most radio applications, vibrator converters must incorporate effective high-frequency suppression devices to overcome the effect of the slight sparking that occurs at the contacts, and for certain applications low-frequency filtering is also necessary.

The operation from standard types of converters of radio receivers incorporating internal frame aerials is seldom completely satisfactory, owing to the increased chance of "hash" and hum pick-up. Even where efficient suppression filters are fitted in the converter, it must be remembered that a weak field of interference will persist in the immediate neighbourhood of the converter. When installing converters, this point should be borne in mind, and the units should not be placed near the aerial section of the receiver; similarly, the aerial and earth wires should be kept clear of the converter wiring.

Servicing

Vibrator converters generally require comparatively little attention. The vibrator units, however, are subject to deterioration and have a limited life; the contact points, in time, wear or become pitted and stick, or the springs lose their tension. Repairs are possible by filing the points and adjusting the spring tension and gap, but these measures are likely to afford only a temporary relief. Towards the end of the useful life of a vibrator unit it may be found that when the receiver is switched on the vibrator fails to function, but can be induced to start by sharply striking the case; this usually indicates worn points, but may denote that the battery voltage reaching the vibrator is too low.

The main on/off switch has to carry a relatively high current at low voltage, and any resistance introduced by dirt or oxidation, caused by arcing, may reduce the efficiency of the unit or cause complete failure.

Complete failure of the power pack is sometimes caused by the breakdown of the rectifier-anode filter capacitor, connected across the secondary of the H.T. transformer. High-voltage types should always be fitted in this position. It should also be remembered that cold-cathode rectifying valves, sometimes used with vibrator packs, may if faulty give rise to a form of electrical interference resembling hash.

Where the tungsten contacts have worn, these can be replaced only by fitting new reeds and carefully re-setting the clearance gaps. It should be noted that the correct gap varies with different types of vibrators. For example, the figures in Table 6 apply to the Valradio range of vibrators.

TABLE 6.—GAP DIMENSIONS OF A TYPICAL RANGE OF VIBRATORS

Types	*Gaps*
PP6/50P, PP6/50DP, PP12/50P, PP12/50DP	All main gaps about 0·007 in.
PP24/50P, PP24/50DP, PP32/50DP, PP50/P, PP110/50P, etc.	Gaps P1 and P2 0·008 Gaps P3 and P4 0·013
PC110/50P	All gaps about 0·008 in.

When re-setting, best results can be obtained by using a cathode-ray oscilloscope to view the waveform of the output; the converter should, of course, be working into its normal load whilst the adjustments are being made.

Generally, it has been found from experience advisable to switch on the converter before switching on the load. Otherwise there is a tendency for arcing to occur at the contacts, especially if the input is high.

Tungsten contacts have a tendency to oxidize; this should be remembered when fitting a new vibrator that has been in stock for some time. To clear the oxidation, it is necessary to run the vibrator off load for a few minutes.

V. V.

ROTARY CONVERSION MACHINES

The distinction between rotary transformers, rotary converters and motor-generators is not always clearly understood. A rotary transformer or dynamotor is primarily a machine which changes the voltage of a D.C. supply, and is most commonly designed to convert a 6- or 12-volt battery supply to H.T. for receivers or low-power transmitters. It has a single armature, wound with two separate windings, each connected to a commutator at either end, and excited by a common field system. Any D.C. ratio can be obtained with a suitable ratio of turns in the windings. Machines delivering 50–500 watts have overall efficiencies ranging from 50 to 60 per cent.

It is customary to mount the machine together with a ripple filter and interference suppressors, on a common base. As a safeguard to the associated radio apparatus, the base is fitted with anti-vibration pads and the whole equipment is enclosed in a screening box lined with sound-absorbent material.

Typical connections for a rotary transformer and its auxiliary equipment are shown in Fig. 18. A starting switch S connects the battery to the motor armature M and excites the field winding F. A field rheostat R regulates the generator voltage. Commutator ripple voltage in the output is smoothed out by a ripple filter L_2C_2. Sparking at the commutators can cause severe interference at radio frequencies unless suitable precautions are taken. R.F. interference is suppressed by chokes L_1 in series with the lines and centre tap earthed condensers

C_1 connected across the input and output terminals of the machine. As a further protection the frame of the machine and the screening box are earthed.

A rotary converter converts A.C. to D.C. and vice versa. It has a single armature winding, from which tappings are connected to slip-rings at the A.C. end of the machine and to a commutator at the D.C. end. The A.C./D.C. transformation ratio is fixed by the number of phases, and is equal to $\sqrt{2}$ for a single-phase machine. For polyphase machines the ratio is given by

$$E_{ac}/E_{dc} = \{\sin(\pi/m)\}/\sqrt{2}$$

where m is the number of slip rings.

D.C. voltages differing from the fixed value can be obtained by feeding the A.C. input through a transformer of appropriate turns ratio, and

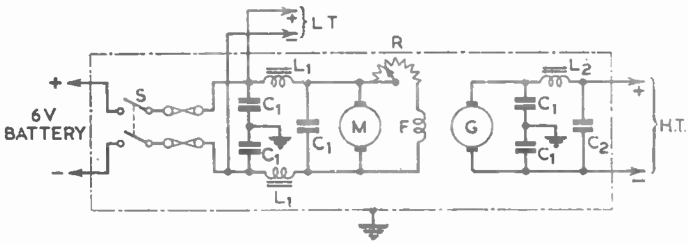


Fig. 18.—Rotary Transformer for Converting Low-voltage D.C. to High-voltage D.C.

furnished with tappings if a range of voltage is required. Conversely, different A.C. voltages are obtainable from a transformer connected to the A.C. output. Alternative methods of providing a range of voltage are the use of an A.C. voltage regulator or a variable iron core choke. It is also possible to control the voltage to a limited extent by field regulation with series reactance.

The motor-generator is the most versatile of the rotary converting machines. Consisting, as it does, of a motor driving a separate generator on a common shaft, it is possible to design machines for any voltage ratio, to convert D.C. to A.C. or A.C. to D.C., or to transform A.C. from one frequency and voltage to another. Two generators may also be flexibly coupled to a single motor on a common shaft to generate two separate supplies, such as L.T. and H.T. supplies for a small transmitter. In such combinations ripple filters must be fitted to H.T. and grid-bias generators, but are not required for an L.T. generator used for filament heating. In common with all rotary commutating machines installed adjacent to radio equipment, it is usually necessary to fit interference suppressors as close as possible to the machine terminals, and in the case of small machines to enclose all equipment in an earthed screening box.

W. E. P.

TRANSISTOR D.C. CONVERTERS

To convert a D.C. source to A.C. or to D.C. of a higher potential, it is necessary to use a converting device such as a rotary converter, vibrator unit or an electronic converter incorporating thermionic valves or transistors. Vibrators have limited life, the units tend to be bulky and their efficiency drops at low output. Valves, because of their relatively high internal resistance, can be used efficiently only on D.C. voltages exceeding 100 volts (preferably above 200 volts). Transistors, with their very low internal resistance, can be used on low-voltage D.C. supplies of up to about 50 volts (higher if necessary), retain high efficiency for power levels from about 1 mW up to hundreds of watts, and operate at relatively high frequency (usually of the order of 1,000 c/s), making possible the use of compact transformers and smoothing components. Unlike vibrator units, there are no problems of spark interference.

There can be little doubt that transistor converters will gradually supersede both the vibrator unit and the rotary converter. As far as it is possible to ascertain by practical experience to date, transistors appear to have an infinite life, and—apart from slight changes during the first few hundred hours of use–no deterioration takes place after years of continuous or intermittent use.

In order to achieve conversion, the steady direct current must be "chopped" into a repetitive series of pulses which can be stepped up by normal transformer action. In a vibrator this chopping or switching is done mechanically. In a transistor converter the transistor acts as an electronically controlled "on–off" switch. A junction transistor can provide an almost ideal switch: when the transistor is "off" only the collector cut-off current flows; when it is "on" and bottomed it behaves as a low resistance of the order of 0·2–2 ohms. By being made to oscillate, the transistor is alternately bottomed and cut-off, and can be used to interrupt a D.C. source.

Although the above action forms the basis of all transistor converters, there are a number of possible circuit configurations: the following are the most useful for practical applications:

(1) Transformer coupled, single transistor.
(2) Ringing choke, single transistor.
(3) Push–pull transformer coupled, twin transistors.
(4) Transformer coupled polarity changer, four transistors (or multiples of four).

Transistors for converter application should preferably be capable of handling high currents, have a low bottomed resistance, high maximum collector voltage rating and adequate dissipation rating. Converters can generally handle between about three and ten times as much power as the transistor dissipation rating. Conventional power transistors have been widely used for this service, although special types are being developed. Silicon transistors will withstand temperatures of the order of 100–150° C., but unfortunately this advantage is offset by the much higher internal resistance, which limits their application in D.C. converters. Germanium transistors can generally be considered electrically superior, but must not be operated at a

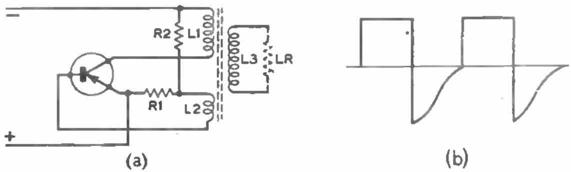


FIG. 19.—TRANSFORMER COUPLED, SINGLE TRANSISTOR D.C. CONVERTER.

junction temperature higher than about 80–90° C. For converter applications under normal ambient temperature conditions germanium transistors are generally used.

Basic D.C. Converter Circuits

Transformer Coupled, Single Transistor. This circuit is probably the simplest, and consists of a single transistor connected to a simple transformer. This circuit is shown in Fig. 19 (*a*), with its output waveform shown in Fig. 19 (*b*).

The operation of this circuit is as follows: The base of the transistor is biased slightly negative which permits an emitter to collector current of a few milliamps. The flow of this small current through L1 will induce a negative potential into L2 which will increase the base current, causing greater collector current. This process rapidly reaches a saturation point with the collector voltage bottoming; at this point no further changes in current take place; the base current rapidly turns positive, cutting off the collector current; the process is then repeated. The pulses of the interrupted D.C. are stepped up by the secondary winding L3 of the transformer. The frequency of the operation is determined mainly by the number of turns on the primary and the size and permeability of the core. The most efficient frequency is about 1,000 c/s.

Because of the severe polarization of the core, the useful output from this circuit is usually less than 10 watts, and, unless great care is taken in the transformer design, the efficiency of the circuit is not very high.

Ringing Choke. A better circuit for low-power conversion is the

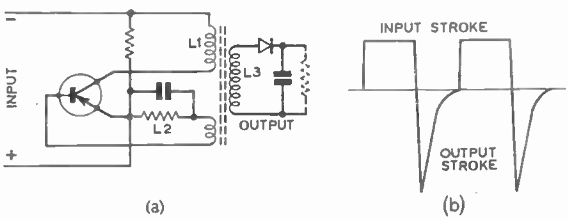


FIG. 20.—RINGING CHOKE D.C. CONVERTER.

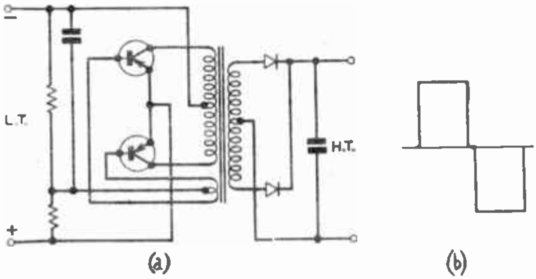


FIG. 21.—PUSH–PULL TRANSFORMER COUPLED D.C. CONVERTER.

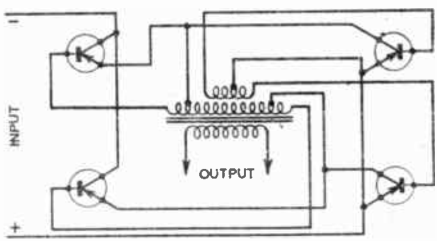
ringing-choke system shown in Fig. 20 (*a*). This functions in a similar manner to the transformer-coupled type, but the output is taken from the ringing voltage in the secondary of the transformer. The efficiency is higher than the transformer type, but it suffers from the disadvantage of poor voltage regulation.

Push–Pull Transformer-Coupled. A transistor is strictly a current-operated device, and for this reason the base circuit must be of sufficiently low resistance to allow the collector to " bottom ". When outputs greater than about 5 watts are required, transistor circuits producing a full-wave output are necessary. The most common is the push–pull type, widely used for inputs of up to 24 volts and outputs of over 100 watts.

As the description implies, the transistors operate in a push–pull manner, whereby they are switched alternatively, producing the output waveform shown in Fig. 21 (*b*). The overall efficiency is 85–90 per cent. The main reason for the very high efficiency is that the transistors are operating under conditions of minimum dissipation; that is either completely cut-off—very high internal resistance, or passing maximum current—lowest internal resistance.

Polarity Changer: The polarity-changer transistor circuit consists of four or more transistors connected in the manner of a double-pole–double-throw switch, whereby the D.C. input is reversed across the load. The basic circuit is shown in Fig. 22. The output waveform is similar to that shown in Fig. 21 (*b*). The polarity-changer circuit has the advantage that the transistors are subjected to only a slightly

FIG. 22.—POLARITY CHANGER D.C. CONVERTER.



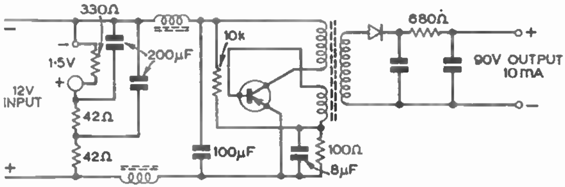


FIG. 23.—LOW POWER CONVERTER.

higher voltage than the D.C. input, whereas in the push–pull circuit the transistors are subjected to over twice the supply voltage.

Typical Units and Applications

Transistor D.C. converters have many uses, particularly for mobile and portable radio communications equipment and similar applications.

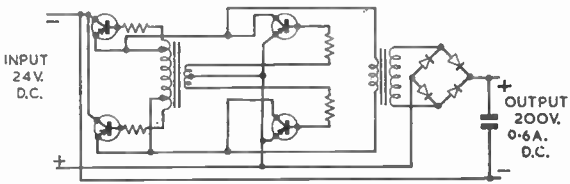


FIG. 24.—HIGH POWER CONVERTER.

They also make possible the operation of small A.C. electric motors and fluorescent lighting from D.C. supplies, etc. Since a transistor converter is practically instantaneous in starting, it can form a valuable emergency unit in the event of a mains-supply failure.

Two representative units, manufactured by Valradio Ltd., are the Types 12/1T and R24/0T.

Unit Type 12/1T (Fig. 23) is a dry-battery eliminator for operating battery receivers from a 12-volt accumulator.

Transverter Type R24/120T (Fig. 24) is a typical example of a transistor unit providing an output of over 100 watts, for operating A.C./D.C. television receivers or similar equipment from a 24-volt D.C. source. The output is 200 volts at 0·6 amp. maximum. The circuit employed is the previously described polarity changer, and the output is rectified by a bridge-connected metal rectifier. In order to increase the output, the D.C. input is connected in series with the rectified output.

V. V

37. HIGH-FIDELITY SOUND REPRODUCTION AND DISTRIBUTION

37. HIGH-FIDELITY SOUND REPRODUCTION AND DISTRIBUTION

HIGH-FIDELITY SOUND REPRODUCTION

In dealing with high-fidelity sound reproduction, it is very convenient to divide the problem into two sections. First, all those attributes of the complete system which can be objectively measured and specified; second, all those attributes which have their being in the difference between the conditions of hearing the original sound and hearing the reproduction of it. "Hi-Fi" systems sometime earn a bad reputation for themselves because they stop short at the first section and take little or no account of the second section.

Objective Attributes

During the evolution of sound reproduction, the following attributes have been isolated from the general fact that the reproduced sound is different from the original: frequency response; signal-to-noise ratio; frequency change; harmonic and cross modulation distortion; and transient distortion. The particular techniques used by different workers in the field of acoustic measurements vary somewhat, but there is fair agreement on general principles.

Frequency Response

Perhaps the most important restriction which may be set by sound-reproduction equipment is frequency coverage. At first this was synonymous with frequency range, but today it is quite possible to cover the whole of the audible frequency range; it is not so easy to cover it with uniform sensitivity to all frequencies. For the purpose of this section, the range of frequencies which can be heard by a normal adult listening to a sound pressure of 80 db above 0·0002 dynes/sq. cm. will be assumed to stretch from 25 up to 15,000 c/s. The total frequency range which is audible under strong stimulus, in the absence of any other sound, is certainly greater than this for young adults, and decidedly less for older people; but if the reproducing system has no other fault of any kind, except its inability to reproduce below 25 c/s and above 15,000 c/s, then it will be a very good system indeed.

It has been found that the maintenance of uniform sensitivity over the whole of the frequency range that is reproduced can be quite as important as having a very extended range. Many of the early high-fidelity systems suffered from this defect; and as a result, they were no more acceptable from the point of view of overall quality than those systems with a narrower frequency coverage but having greater uniformity.

The "Frequency Response Curve" is the usual way of expressing the uniformity or otherwise of the component or the whole chain of sound reproduction. It is normally plotted to show the variation in output for a constant input of various frequencies. Since the input is usually maintained constant, it can often be ignored, and the output at

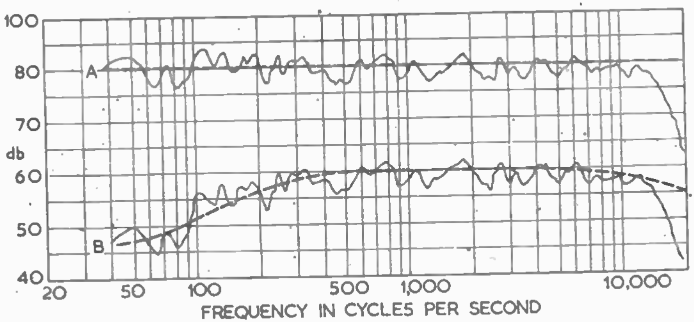


FIG. 1.—FREQUENCY RESPONSE OF HIGH-QUALITY SYSTEM MEASUREMENT AT THE EAR OF THE LISTENER IN THE LISTENING ROOM.

A. Original sound and reproduction 80 db above 0·0002 dynes/cm.².
B. Original sound 80 db, and reproduction 60 db above 0·0002 dynes/cm.².

one frequency compared directly with the output at another frequency. Do not be misled by the " shape " of the curve; what is measured and plotted is the output at one frequency compared with the output at another. The eye is used to assessing shapes, and two curves which have a different appearance may sound very similar.

There is a secondary use of the " shape " of the response curve. A normal frequency response curve is applicable to steady tones only; but if there is a rapid change in sensitivity for a small change in frequency, then the behaviour of the system to the unsteady tones of an actual programme will be less satisfactory than the steady test tones suggest. In other words, the transient response of the system will be bad, and a rough indication of the degradation can be obtained from an inspection of the rate of change of the frequency response curve. Exception should be made to those dips in the response curve which are due to sound arriving at the measuring microphone via two or more paths which have a difference in path length, so that the sound arriving from one path cancels that from another. Such cancellations will not have any effect on transient response.

Microphones and Recorded Programmes

The frequency response of a microphone is often dependent upon the direction of arrival of the sound. If only one source of sound is used, this can often be placed in the preferred direction relative to the microphone so that the required frequency response is obtained. But if there are several sources of sound, such as an orchestra, or a single source with several walls which reflect it, giving rise to reverberation, it is quite possible that some sounds will reach the microphone from angles which produce a bad frequency response. A typical example would be the direct sound from an orchestra reaching the microphone from the preferred direction, and the reverberation from the hall falling on the back of the microphone, which may have a very different frequency response.

In the case of a loudspeaker the frequency response nearly always varies with the angle at which the sound is radiated relative to the axis of the loudspeaker. There is no doubt that satisfactory reproduction could be obtained with uniform hemispherical radiation at all frequencies, but this is seldom achieved in practice; it is much more usual to find that the higher frequencies are radiated as a narrow beam, while the lower frequencies are more nearly spherical.

In the case of recorded programmes, the material used for the recording may well impose a special frequency response. This is the case for both magnetic tape and disc recording. Consider first the case of a recording made on magnetic tape. This is passed over the reproducing head, and the resulting voltage generated in it is proportional to the rate of change of flux along the magnetic tape. That is, the higher the frequency, the greater is the output voltage from the reproducing head for a given amount of magnetization. (This is true for all those frequencies where the width of the air gap in the reproducing head is considerably less than the smallest wavelength recorded on the tape.) Since the amount of magnetization permissible on the tape is set by the distortion which can be tolerated, it follows that the lower the frequency, the lower the voltage output from the reproducing head. Thus it comes about that the voltage output from a magnetic tape is proportional to frequency over most of the frequency range up to that point where various high-frequency losses come in and first flatten the response curve and finally cut it right off. The reproducing amplifier must have a frequency response which restores the output from the reproducing head to a voltage which is independent of frequency.

Considering now the case of the disc record, the early records were cut in such a way as to produce a constant transverse velocity to the needle tip from about 250 c/s upwards. It was not possible to continue this relationship below that frequency because the rising amplitude of the modulation would cause the lower-frequency grooves to run together. Thus it became the practice to record at constant amplitude below 250 c/s and at constant velocity above this frequency. Using present-day materials and groove dimensions, the lower frequencies are recorded at nearly constant amplitude, the middle frequencies at constant transverse velocity and the upper frequencies are accentuated. The amplifier used must have a frequency response which restores the voltage output from the pick-up to a voltage which is independent of frequency.

It is customary to express the frequency response of a disc record as the departure from constant transverse velocity conditions, this ratio being expressed in decibels. A good magnetic pick-up will have a voltage output which is almost exactly proportional to transverse velocity, and the reproducing amplifier then requires a frequency characteristic which is the inverse of the recording characteristic. Crystal pick-ups, on the other hand, can be made which will produce an almost ideal constant-voltage output from the present-day recording characteristic. An amplifier for use with such a crystal pick-up usually requires no corrections at all. This results in a great reduction in the total amplification necessary at very low frequencies compared with the magnetic pick-up used with the same recording characteristics.

Radio

The frequency response which can be obtained from a radio transmitter and receiver can be sufficiently good to cover the whole of the required band from 25 c/s right up to 15,000 c/s with uniform sensitivity. However, due to the closeness of the operating frequencies of adjacent radio transmitters, it is usually only possible to obtain sufficient frequency separation on the highest frequencies. In the case of the reception of medium waves, particularly after dark, it will be found that there is an audible note of about 9 kc/s due to the interaction of one radio station with another on an adjacent frequency. If a receiver is used which is capable of receiving a band of frequencies 18 kc/s wide it is quite possible that both the interfering note from the station immediately higher and the note from the station immediately lower in frequency will be heard at the same time. If the radio-frequency separation of the three stations, the upper, the wanted and the lower, are not exactly equidistant, interference beats will be produced which may be decidedly more objectionable than only one steady note of 9 kc/s. This interference has to be reduced by making the radio receiver selective, either in the radio circuits, or the audio circuits, or both, so that there is no response to a signal 9 kc/s or more from the wanted signal. Such selectivity precludes the reception of the higher frequencies beyond about 7 kc/s, and even then difficulty will be experienced with the reproduction of transients having frequencies in this neighbourhood. This problem does not exist with V.H.F./F.M. broadcasting.

Signal-to-noise Ratio

After frequency response, the ratio between the wanted signal and unwanted sounds of all kinds is probably the most important attribute to any high-quality system. The ear can survive sounds 120 db above the threshold of hearing (see Section 31, "Microphones"), but the auditory system may be damaged if a sound of this intensity is heard for a considerable time. It is not usually necessary to strive for a signal-to-noise ratio of 120 db: 100 db would be very satisfactory, but is not easy to obtain, even under the very best conditions. Often, the ratio is between 40 and 60 db only, but it cannot be said to be adequate for all occasions. The most common forms of noise are: thermal noise in resistors, various forms of valve noise, which include microphony caused by the relative movement of various components inside the valve, microphony caused by the movement of cables and components, and electrical voltages which are induced in the apparatus and connecting cables by external or internal fields, both electrostatic and electro-magnetic of almost any frequency, depending upon the particular apparatus used. With some types of apparatus it will be found that the signal-to-noise ratio is not a constant, but that the noise increases with increase of signal. This is usually termed "noise behind the signal", and badly recorded magnetic tape programmes sometimes provide an example of this fault. The noise suppression which is used in talking films and other apparatus is not necessarily quite the same as noise beyond the signal, because the value to which the noise rises when the suppression is removed may not be directly related to the strength of the wanted signal.

It will usually be found that the signal-to-noise ratio is dependent on the frequency range reproduced. If a recording and reproducing chain is added to an existing system of sound reproduction, it impairs the signal-to-noise ratio, but if it is substituted for a transmission system which has much noise, such as radio reception on medium waves, it may show a very considerable improvement in signal-to-noise ratio.

Frequency Change

If there is a change in the frequencies which are reproduced compared with the frequencies which are emitted by the original sounds, the reproduction may be rendered quite unacceptable. A change in frequency, which may be either steady or variable, is often encountered when a recording is included in the chain of reproduction equipment. It is also sometimes encountered in short-wave reception, but is usually accompanied by other forms of distortion, such as a very poor signal-to-noise ratio.

Amplitude Change

Any change in amplitude between the original sound and the reproduced sound should be considered to be a form of distortion, although in some cases the change in amplitude is deliberate and in other cases it is accidental. Volume compression, as used in radio transmission, is an example of a deliberate variation in amplitude, whilst " flutter ", which is sometimes found on tape and film records, is an example of accidental amplitude variation. The latter is obviously a form of distortion, but the former may not be suspected by the uninformed listener.

Harmonic Distortion and Inter-modulation

If harmonics of the original frequencies are generated in the reproducing chain, or if sum and difference frequencies are produced from two or more original frequencies, they will be apparent as a form of distortion. The measurement and expression of these forms of distortion is one of the hardest of all measurements to make and express. Similarly, the degree of distortion which can be tolerated by different individuals will vary to a great extent, and will even vary with the particular occasion.

Both these forms of distortion are dependent upon the size of the signal, and in all probability will have a value which varies with frequency as well. Common examples are overload distortion of a radio-frequency detector, or of an amplifier, and also intermodulation produced by a loudspeaker.

Transient Distortion

This distortion appears as an inability of the system or component to respond to sudden changes of amplitude and frequency. It is usually evaluated by measuring the rate of decay of a sudden interrupted steady frequency; it being usual to assume that the rate of decay is also an indication of the rate of growth. In cases of severe transient distortion it is quite possible for the frequency emitted during the decay period to be different from the frequency of excitation. Since transient distortion often arises from the presence of resonances, and these self-same resonances also impair the frequency response, the two forms of distortion often occur at the same frequency, and may be confused by an unskilled listener. Since the methods of objective measurement are very different, there is little chance that the measurements will be confused; but if frequency response only is measured and an attempt is made to relate it to subjective listening of a programme, the presence of transient distortion is apt to mar the objective–subjective relationship.

Subjective Differences

It is the purpose here to examine all the differences which exist between the hearing of the original sound and the sound reproduced by a single-channel system free from any of the objective distortions already mentioned. It may seem strange to find that a single-channel system, which is free from all objective distortion except the inclusion of a volume control, may not necessarily give satisfactory reproduction. In considering subjective differences it is necessary to start with the mind of the artist and to follow the chain of events from the mind of the artist right through to the mind of the listener.

Subjective Frequency Response (Pitch-loudness)

This is the most important attribute of a high-quality sound-reproducing system which undergoes very considerable modification by subjective effects. It is almost the only attribute in which some attempt at correction can be made. Consider first the human voice. The artist may be speaking particularly for one particular listener, and speaking under conditions which are ideal for the listener to hear him. But, on the other hand, he may be speaking to a number of listeners in different surroundings from himself, and differing again from one another. The announcer reading the news should read in a loud, clear voice in one home so that mother out in the kitchen washing up the dishes can hear; he should shout the news out at the top of his voice so that all those in the local pub can hear the cricket scores; at the same time he should whisper the news discreetly to the lone occupier of a small apartment, lest the neighbours complain. Clearly he cannot speak with a different " voice effort " for each person all at the same time. To overcome this defect each sound-reproducing equipment is provided with a volume control which will increase or decrease all the frequencies equally, thus preserving the original objectively faultless system.

Unfortunately, when an announcer alters his voice effort he does not maintain a constant increase or decrease of all frequencies. If normal conversation in a quiet room is taken as the standard of comparison; then for loud speech, the low frequencies are increased by about 10 db, the middle region is increased by 25 db, and the highest frequencies by about 12 db. Loud declamatory speech raises the ends of the spectrum by another 5 db, but increases the mid-frequencies by another 10 db, so that there is now a difference of 20 db between the middle and the two ends of the spectrum. Soft, intimate speech, not whispering, lowers the middle frequencies by 12 db without reducing the high frequencies at all. A " Hi-Fi " system with only an objectively flat frequency response is obviously going to be quite inadequate unless it is possible to ring up the announcer and ask him to speak in a voice appropriate to your particular occasion!

The second subjective difference occurs due to the difference between a microphone and a listen. The microphone can be made to discriminate between two sounds originating from different positions if it has suitable directional properties (see Section 31) and these properties can be varied at the will of the engineer, but they cannot be controlled by each individual listener. Take the case of three listeners in a large concert hall: one sits at the front near the orchestra and hears a loud, clear sound; one sits near the back of the hall and uses all his mental concentration to hear the orchestra as clearly as his poor seat will allow; the third also sits at the back of the hall, but he is dreaming away listening to the combined effect of the orchestra plus the reverberation of the hall. The three people require three different frequency-response curves to take account of the different amount of reverberation and distance of which each is conscious.

The third difference in subjective frequency response occurs because the ear of a normal listener has a pitch–loudness relationship which is dependent upon the loudness of the sound which is heard. This relationship is almost independent of the frequency of stimulus for sounds 70 db above the reference level of 0·0002 dynes/sq. cm., but a reduction of level of reproduction from 70 to 50 db above reference level will reduce the apparent loudness of a 100-c/s stimulus by about 10 db.

In normal domestic surroundings there are many occasions where it is not possible to reproduce a large orchestra in such a way that the loudness produced at the ear of the listener is the same as it was at the original concert when he was sitting in his preferred seat. The difference in level may easily be 30 or even 40 db, which is quite enough to give a subjective reduction in bass loudness of 15 db, in addition to the overall reduction of level. In the case of a reproducing system having a frequency response which is measured to be flat by objective means, and free from transient distortion, the reproduction will sound very thin and unrealistic. In actual present-day practice the reproducer is neither flat by objective measurement, nor is it free from transient and inter-modulation distortion. The result is that there is a tendency to hide the loss of the lower frequencies, which is brought about by subjective effects, by superimposed objective faults.

In the case of the reproduction of speech, in the home, it is customary to reproduce it louder than normal, partly because if a word is lost due to local noise, it is not possible to ask the news reader to repeat his last remark. The result of this increase in the loudness of speech is to enhance the loudness of the lower frequencies; in fact, the converse of the case with music. It has already been pointed out that there are often objective imperfections in present-day equipment, and that many of these objective imperfections increase the bass response. Thus there is a much greater tendency for the reproducer to show an excess of the lower frequencies in speech than the absence of bass in the reproduction of soft music. This is particularly true if transient distortion makes the reproduction of speech "boomy".

Subjective Signal-to-noise Ratio

The subjective signal-to-noise ratio may be considerably better than the objectively determined value might suggest. This can be due to two causes; in one case the signal may "mask" the noise to such a degree that the listener is not aware of it: and in the other case the reproduced noise is masked by local noise around the listener. The mind of the listener is able to make some discrimination between the sounds that he wants to hear and the sounds that he wants to ignore.

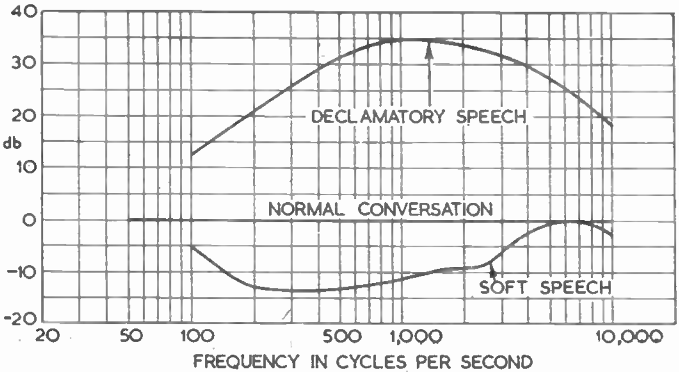


FIG. 2.—RELATIVE SOUND SPECTRA OF A MAN USING THREE DIFFERENT VOICE EFFORTS.

It may be recalled that if a particular sound is expected and the listener is waiting to hear it, it is easier to pick it out from other noises than if the listener has no knowledge of the kind of sound which may arrive.

Subjective Transient Distortion

There is little published work either on the production of subjective transient distortion or even of the subjective effects caused by objective transient distortion in the reproducing chain. The general effect is one of increasing the loudness of those frequencies near the frequency of maximum transient distortion. It is very difficult to evaluate this effect, because it is the same irregularities in the objective frequency response which give rise to the objective transient distortion.

Subjective Spacial Differences

The normal listener is presumed to be provided with two ears and a mind, which he will employ as he wishes; and let it be remembered that the wishes need not be consistent, but can be changed by a mere whim of the moment. No microphone can have this particular property, and this very limitation may often preclude the attainment of "perfect reproduction".

In the case of the loudspeaker, it is not possible, with a single channel, to change the apparent size of the source of sound to match the sounds being reproduced; neither is it possible to make the sound appear to move about. Attempts to improve this situation have been made by means of stereophonic sound; the success which these results have achieved, at least under laboratory conditions, is a direct measure of the failure of the single-channel system.

F. H. B.

AMPLIFIERS AND PRE-AMPLIFIERS

Power Amplifiers

The basic requirements of a high-fidelity amplifier are low harmonic and intermodulation distortion, linear frequency response, adequate power output, good transient response, low hum and noise level, and low output resistance.

Harmonic Distortion

Owing to the inherent curvature of valve characteristics, transformer limitations, etc., the relationship between the input and output of even the best amplifiers is to some extent non-linear. One effect of this is harmonic distortion, that is, the production of spurious harmonics to the original. For example, if the input voltage is a pure sine-wave of 100 c/s, the output may consist of a fundamental frequency of 100 c/s, plus some second harmonic of 200 c/s, third harmonic of 300 c/s and so on. The conventional push–pull output stage largely cancels second harmonics, which are usually considered comparatively innocuous, but the presence of higher-order harmonics, such as the fifth, seventh and ninth, which are not harmoniously related to the fundamental, are usually considered objectional even in very small percentages.

One method of testing an amplifier for harmonic distortion is to use a wave analyser, as shown in Fig. 3. The filter is necessary to remove any harmonics present in the output from the audio generator. The wave analyser is tuned successively to the harmonics of the frequency used for the test, usually 400 or 1,000 c/s, and the readings are expressed as follows:

$$D = \frac{\sqrt{E^2_2 + E^2_3 + E^2_4 + \ldots}}{\sqrt{E^2_1 + E^2_2 + E^2_3 + E^2_4 + \ldots}} \times 100$$

where D is the percentage of total harmonic distortion, E_1 is the amplitude of the fundamental voltage, E_2 the amplitude of the second-harmonic distortion, etc.

Since the higher-order harmonics are progressively more unpleasant, many authorities consider this analysis unrealistic, advocating instead a "weighted" distortion factor in which the harmonics are weighted in proportion to their harmonic relationship. However, as it is difficult to assess their relative unpleasantness objectively, this method is not always practicable.

Where individual harmonic figures are not required, it is more convenient to use a distortion-factor meter, which is essentially a bridge capable of filtering out a fundamental frequency, thus leaving the remaining harmonics to be read as a total percentage.

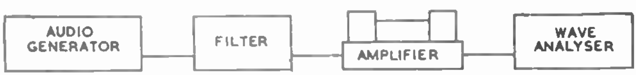


Fig. 3.—Wave Analyser Method of Testing Amplifier Distortion.

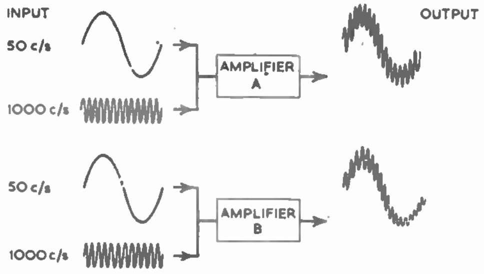


FIG. 4.—INTERMODULATION DISTORTION CAUSED BY THE NON-LINEAR AMPLIFIER "B".

Intermodulation Distortion

Intermodulation distortion is the production of spurious sum and difference frequencies when two or more frequencies are passed through a non-linear system. Fig. 4 shows two frequencies, 50 and 1,000 c/s, applied to two amplifiers. Whilst amplifier A amplifies both frequencies without interaction, amplifier B distorts the 50-c/s sine-wave in such a way that it modulates the higher frequency, forming a series of sum and difference frequencies, including 950 and 1,050 c/s, 850 and 1,100 c/s, etc.

Intermodulation distortion is much more serious than harmonic distortion because these spurious frequencies are not harmoniously related to the original tones, and are nearly always unpleasant to the ear. For this reason many authorities consider the intermodulation distortion rating of an amplifier to be more important than its percentage of harmonic distortion.

Measurements of intermodulation distortion are made by applying a composite signal of high and low frequency to the amplifier, and measur-

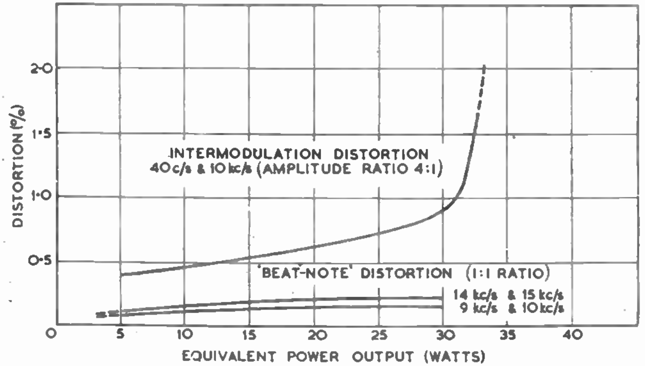


FIG. 5.—INTERMODULATION DISTORTION MEASUREMENTS ON MODERN HIGH-FIDELITY AMPLIFIER.

ing the products individually or collectively, as in the case of harmonic distortion. The usual frequencies are 40 and 10,000 c/s, 70 and 7,000 c/s, or 50 and 2,500 c/s, and the amplitude ratio is nearly always 4 : 1. The specific value of intermodulation will differ significantly with the method used, and this must be borne in mind when comparing figures. A more useful evaluation of intermodulation at high frequencies may be obtained by using two signals about 1,000 c/s apart and of equal amplitude. Fig. 5 shows the performance of a good-quality amplifier using both methods. The beat-note frequencies are 14 and 15 kc/s and 9 and 10 kc/s, whilst the standard *Society of Motion Picture and Television Engineers* (S.M.P.T.E.) frequencies of 40 c/s and 10 kc/s at a ratio of 4 : 1 were also used.

There is no simple relationship between intermodulation and harmonic distortion, although they are both due to the same cause. As a general rule, however, the proportion of 3·8 : 1, increasing as the amplifier reaches overload point, can be taken as fairly accurate.

Maximum Permissible Distortion

A high-fidelity amplifier should have a total harmonic content of not more than 0·2 per cent, and the intermodulation distortion should not exceed 0·5 per cent at 8–10 watts output. The critical ability of the human ear to distinguish distortion depends to a large extent on the volume and frequency range being reproduced, but it has been stated that an experienced listener can distinguish between an amplifier having 0·1 per cent and the one having 0·4 per cent total harmonic distortion. This may be true in certain circumstances, but it must be remembered that the distortion present in the programme material and associated equipment is generally much greater than that contributed by the average high-fidelity amplifier.

Frequency Response

The frequency response should be substantially flat from 20 to at least 20,000 c/s. Although the average adult human ear cannot hear frequencies beyond 15–20 kc/s, nevertheless, if the internal band-width of an amplifier does not extend up to at least 40 kc/s, transient response will suffer and the reproduction will lack " attack ". Further extension of the linear frequency response is unnecessary and may, indeed, be a disadvantage, as any non-linearity in the system will result in distortion, owing to the creation of audible combination frequencies. A satisfactory compromise is usually obtained in practice by " rolling off " the response gradually from about 20 kc/s, at a slope of not more than 6 dB per octave, to 40 kc/s, then increasing the slope to 12 dB per octave.

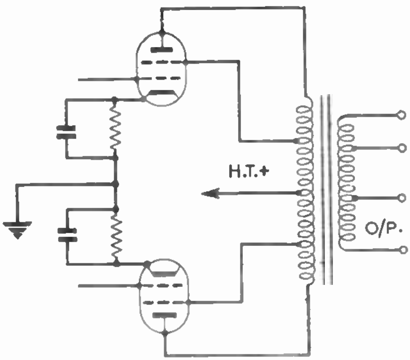
The frequency response should always be quoted with a reference figure, which is usually 1,000 c/s. A good amplifier will have a response of ± 1 dB from 20 to 20,000 c/s, reference 1,000 c/s.

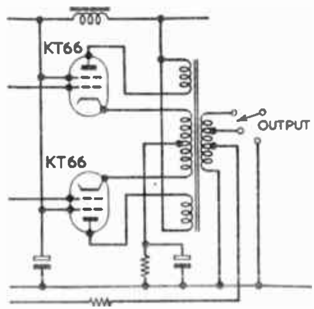
Frequency response is not necessarily identical with power response, and it is usual for the response to fall off at each end of the scale at high power outputs. If the maximum output, particularly at the bass end, is substantially less than at medium frequencies, filters must be inserted in the pre-amplifier to reduce the level of these frequencies before they reach the amplifier, otherwise severe intermodulation will occur. This is especially noticeable during the reproduction of an organ on incorrectly designed equipment, where pedal notes of the order of 16–30 c/s may cause bad distortion even though they may be inaudible in the output.

Power Output

With a loudspeaker of average efficiency, a 10–12 watt amplifier should be adequate for home listening. With an inefficient loudspeaker, or with a very large room, it is advisable to specify a larger reserve of power, perhaps 20 or 25 watts. The normal average power used under domestic conditions is of the order of 250–300 mW, but ample power must be available to handle peak transient power with an adequate factor of safety.

Output transformers capable of giving high outputs, particularly at the low frequencies, are necessarily expensive, especially if they use





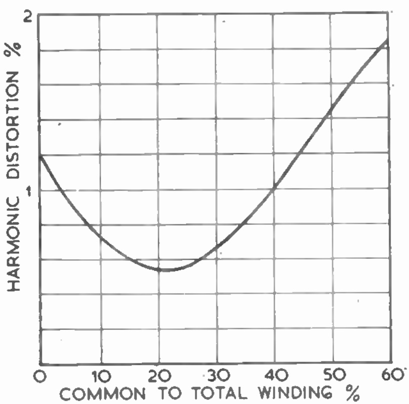


FIG. 6 (*top left*).—BASIC ULTRA-LINEAR OUTPUT STAGE. FIG. 7 (*top right*).—A VARIATION OF THE ULTRA-LINEAR PRINCIPLE.

The common portion of the output primary is connected between the cathodes of the output valves.

FIG. 8 (*bottom left*).—RATIO OF COMMON TO TOTAL WINDING FOR AN ULTRA-LINEAR CIRCUIT FOR KT66 VALVES.

Showing reduction of harmonic distortion obtained by choice of optimum tapping point for tetrode. Main feedback loop disconnected.

grain-orientated strip or " C " core; but an amplifier can be no better than its output transformer.

Output Stage

Most modern high-fidelity amplifiers use the " ultra linear " push-pull circuit. This is a modified form of the conventional push–pull output stage, the tetrode screens being connected to tappings on the output transformer to give a mode of operation intermediate between tetrodes and triodes (see Fig. 6). This distributed load has a number of advantages: less overall feedback is required for a given result, thus giving a better margin of stability; reduction of harmonic distortion (see Fig. 8); greater efficiency (36 per cent as against 27 per cent for triodes or triode-connected tetrodes); and much lower peak variations in current than are inherent with triode output stages (this can cause serious distortion at peak outputs if the power supply does not have an effectively low impedance at audio frequencies).

It can be seen from Fig. 8 that with a KT66 valve minimum harmonic distortion is achieved when the common portion of the output-transformer primary is approximately 20 per cent of the total winding. For the EL34, the common portion of the winding should be approximately 43 per cent of the total.

An alternative system is shown in Fig. 7. In this the common portion of the winding is inserted in the cathode circuit, resulting in the voltage appearing across that portion being effectively applied to the grids as negative feedback.

One possible line of development is the transformerless output stage. Although a number of circuits have been devised, practical difficulties have so far restricted the use of such systems. The circuit of a single-ended arrangement which has been used in a commercial record reproducer is shown in Fig. 9.

Transient Response

The total quality of sounds is determined not only by their harmonic content, but also by their " attack and delay " times. This is par-

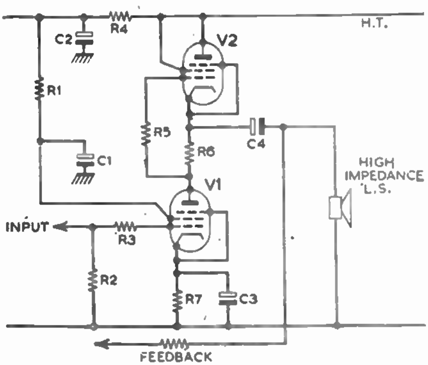


FIG. 9.—ONE PRACTICAL ARRANGEMENT FOR THE ELIMINATION OF THE OUTPUT TRANSFORMER.

This is basically similar to the simple shunt-regulated amplifier discussed in Section 10.

ticularly true in the case of percussive instruments, which produce highly complex waves of short duration. To reproduce transients satisfactorily, it is not only important that the amplifier has a wide band-width: it must also be free from supersonic oscillations or high-frequency peaks which cause "ringing". With amplifiers using a large amount of negative feedback, it is vital that the high-frequency response is controlled and correct damping applied.

The usual method of testing is to apply a square or saw-tooth wave to the amplifier, the output of which is connected to an oscilloscope and a resistive load. Fig. 10 shows typical forms of transient distortion:

FIG. 10.—(a) SQUARE WAVE INPUT, (b) OUTPUT SHOWING SEVERE "RINGING", (c) OUTPUT SHOWING HIGH FREQUENCY ATTENUATION.

(*a*) is a square wave input at 10 kc/s, (*b*) shows severe ringing due to incorrect feedback damping, and (*c*) the result of an attenuated high-frequency response.

Transient response can also be adversely affected by phase changes caused by some forms of tone control or high-pass filters having a steep cut or "roll-off".

Hum and Noise

Figures for hum and noise may be quoted separately, or as a combined figure. It is usual to quote the rated output of an amplifier as a reference figure, but a standard output of 1 watt is also used. Hum may be due to insufficient smoothing of the power supply, pick up from the heater supply (either in the valve itself or in the associated wiring) or from the mains transformer, etc. To avoid the formation of loops which would pick up hum from the stray field of the mains transformer, it is customary to use a single earthing point or bus-bar.

Internal valve noise is responsible for most of the hiss heard in a high-gain amplifier, although carbon resistors and electrolytic capacitors contribute a share. The use of high-stability "low-noise" resistors in the initial, high-gain stage, and of valves designed specifically for this application (e.g., EF86, 6BR7 and Z729) enables the noise to be reduced to negligible proportions. For example, the EF86 has the grid pins diametrically opposite to the heater pins, and incorporates internal shielding between them.

As high-frequency noise is aperiodic, i.e., not confined to any particular frequency, its effect is emphasized if the amplifier or loudspeaker has a peak or peaks. Similarly, the better the low-frequency response of the loudspeaker, the more it will reproduce hum.

A combined hum and noise level of −90 dB (relative to the full output of the amplifier) cannot be heard even quite close to the loudspeaker, and a level of −60 dB for the amplifier and pre-amplifier is satisfactory.

Damping

The ratio of load impedance to internal impedance is usually referred to as the damping factor of an amplifier. A damping factor of 30, which is a typical figure, means that at the nominal output impedance of 15 ohms the output resistance would be 0·5 ohm. A high damping factor means that a loudspeaker " sees " a very low resistance, which tends to damp any tendency of the loudspeaker diaphragm to vibrate at its natural frequency. Whereas in theory the higher the damping factor, the more efficient the acoustic " brake " becomes, practical tests show that any increase of damping factor above about 20 results in little further improvement. This is because the resistance of the loudspeaker speech coil is in series with the output.

A variable damping control, having a range from a factor of about 30 to infinity, at which point the output resistance becomes zero or a complete short-circuit, is incorporated in a number of amplifiers. Control is achieved by varying the proportions of positive and negative feedback in the main feedback loop.

Advocates of variable damping claim that every loudspeaker has a certain critical damping factor, but a good loudspeaker having a well-damped enclosure will reflect very little resonance into the electrical circuit; certainly too small a one for a critical damping point to be determined. In practice, it will be found difficult to detect any appreciable difference in reproduction as the control is turned from one extreme to the other. Nevertheless, variable damping can effect a worthwhile improvement when used with an inefficient loudspeaker system.

Negative Feedback

To say that an amplifier has 20 dB feedback means that the feedback loop has reduced the gain by 20 dB, or 10 times, and, if the amplifier is correctly designed, the distortion will have been reduced by a like amount. The overall gain of the amplifier has to be increased to compensate for the loss; but this presents no serious problem. Besides drastically reducing harmonic and intermodulation distortion, negative feedback can also be used to perform the following functions: improve frequency response; reduce the output resistance of the amplifier; improve the low-frequency characteristics of the output transformer, particularly defects due to the non-linear relation between the flux and magnetizing force; reduce the effects of noise or random changes of the parameters of the amplifier originating within the feedback loop.

The application of feedback to an amplifier having a high distortion factor will merely exchange a high proportion of lower-order harmonics, such as the second or third, for a smaller proportion of higher ones, which, in practice, may be just as unpleasant. This is due to the feedback voltage containing a high proportion of second and third harmonics, each in its turn generating its own harmonics (fourth and ninth, etc.). Great care must be taken to ensure that the feedback remains negative at all frequencies: at high and very high frequencies, reactances cause phase-angle displacements, which tend to nullify the negative feedback. In extreme cases this may even become positive, causing oscillation either inside or outside the band of audible frequencies.

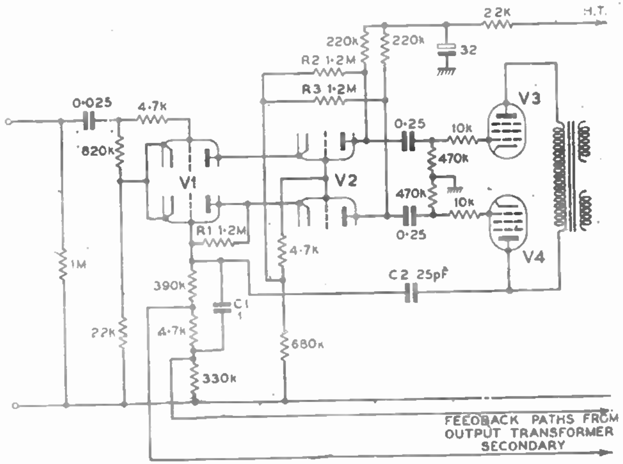


Fig. 11.—Phase-splitter and Driver Stages of a Modern Power Amplifier Showing Use of Multiple Feedback Paths.

The most common arrangement is to feed back a portion of output from the output transformer secondary to the cathode of the first stage. There is, however, a tendency to add further feedback loops between stages (see, for example, the design shown in Fig. 11). In one design, linearity is improved by feedback loops between the anode and suppressor grids of the output valves. In the design shown in Fig. 11, the main feedback is applied to the grid of the first stage. This high-impedance injection gives the amplifier an inbuilt high-pass characteristic to improve stability at sub-audio-frequencies.

It is important to ensure that the frequency range of the input signal applied to a feedback amplifier does not extend in either direction beyond the flat frequency range of the amplifier, otherwise the reduced feedback may result in an abnormal signal being applied, thus causing distortion, although the frequency itself may be inaudible.

Instability

One form of instability is an oscillation which takes place during part of a cycle only, or is triggered off by a transient. Such phenomena can best be detected by means of an oscilloscope, using a square-wave input. The stability of an amplifier should not be impaired by the connection across its output of a capacitive load of about 2 μF or an inductive load of up to 150 mH.

A tertiary feedback winding on the output transformer is sometimes used to improve the stability factor. This also simplifies the feedback network, as this remains independent of the output load.

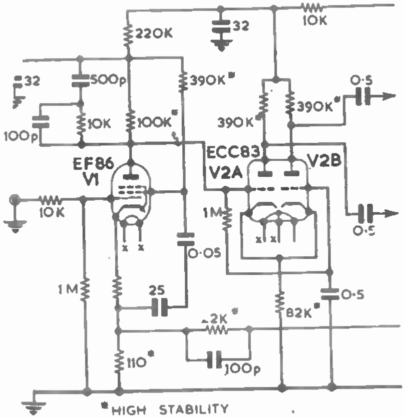


FIG. 12.—INPUT AND PHASE-SPLITTER STAGES OF A MODERN POWER AMPLIFIER.

The first stage is directly coupled to the phase splitter to reduce phase shift at low frequencies.

Phase Shift

To reduce phase shift at low frequencies, the first stage is often directly coupled to the phase-splitter stage, as shown in Fig. 12. In this circuit, there is only one low-frequency time constant (apart from the decoupling capacitors) within the feedback loop. The network in the anode circuit of the first valve forms a high-frequency " step ", and further phase correction is applied by the capacitor connected across the feedback resistor.

In the design shown in Fig. 11 a similar principle for reducing phase shift is used, V1 phase splitter stage being cascode connected to a driver stage, V2.

Phase Splitters

There are three basic phase-splitter circuits in common use: the cathode-follower, paraphase and cathode-coupled types. In the cathode-follower circuit the inputs for the push–pull output stage are derived from the anode and cathode circuits of the phase-splitter valve. One disadvantage of this circuit is the different impedance of the two outputs. The paraphase circuit is literally a phase-reversal system: the input to one output valve is taken direct from the voltage-amplifying stage, an additional valve being used to provide the out-of-phase input to the other output valve. By feeding the input to this phase-reverser valve via a potential divider, the gain of the stage is made unity. The circuit is shown and further discussed in Section 14.

In both these systems no gain is provided by the phase-splitter stage. One advantage of the cathode-coupled phase splitter is that a certain amoun. of gain is available from the stage. An ECC83 used in such a circuit will provide a stage gain of about 25. Amongst other advantages accruing from this, increased use can be made of negative feedback. Fig. 12 shows the pre-output stages of a typical high-fidelity amplifier, with an EF86 voltage amplifier followed by an ECC83 cathode-coupled phase splitter. The operation of the cathode-coupled phase splitter is as follows: A negative pulse at V2A grid produces a positive pulse at V2A anode and a negative pulse at V2B cathode (which is coupled to V2A cathode). The negative pulse at V2B cathode is equivalent to a positive pulse at V2B grid and, hence, a negative pulse at V2B anode. V2B grid is capacitively earthed, the 1M resistor connecting the two grids ensuring correct D.C. conditions in V2B. The cathode coupling resistor, anode-load resistors and grid resistors of the following stage should all be high-stability types.

Pre-amplifiers

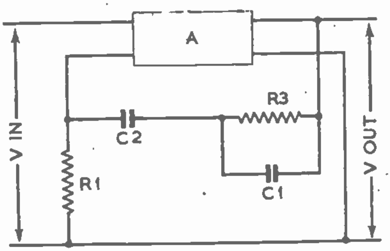
Apart from providing voltage amplification, the main functions of a control unit are: selection of inputs (i.e., tape, gramophone, radio tuner, etc.); the provision of adequate record-equalising facilities; bass- and treble-tone control; filtering; and volume control. In addition to these functions, facilities may be provided for correct impedance matching and attenuation for different pick-ups.

Equalization

With the advent of the standard R.I.A.A. recording curve, the provision of a large selection of equalizing or play-back curves has become less important: most amplifiers provide compensation for three or four characteristics only. Resistor and capacitor networks can be used to provide the necessary selective attenuation, but a feedback circuit is to be preferred, as the unwanted gain is usefully employed to reduce distortion.

Fig. 13 shows a typical equalization configuration in a feedback loop. R1 is the cathode bias resistor of the first valve, and the feedback is derived from the anode of the second valve, which in practice may be the second half of a double triode. The disadvantage of this system is that the cathode of the first valve is not by-passed, the possibilities of hum thus being increased. However, by using a balanced heater supply biased some 30–40 volts positive (conveniently obtained from the cathodes of the amplifier output valves) a hum level of −60 dB at an input of 15 mV is not difficult to achieve. If a higher gain is required, it is much more satisfactory to insert the loop around the first valve

FIG. 13.—EQUALIZING CONFIGURATION IN A NEGATIVE FEEDBACK LOOP.



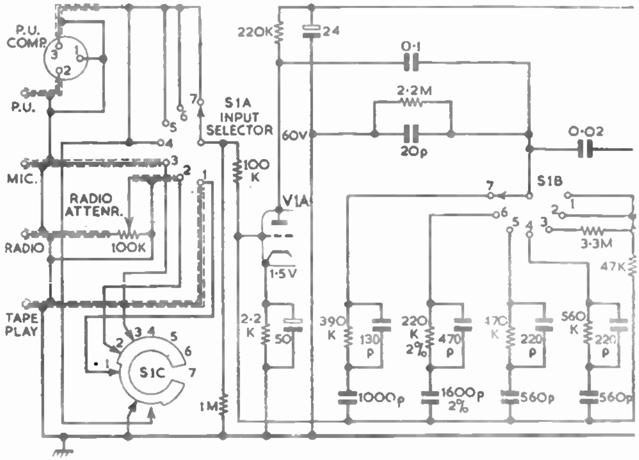


FIG. 14.—INPUT AND EQUALIZING CIRCUIT OF A MODERN PRE-AMPLIFIER. V1A is one section of an ECC40.

only. Many amplifiers use the arrangement shown in Fig. 14, and it will be seen that in this the equalizer input selector switches are combined, and that a simple feedback loop is brought into circuit to control the gain for microphone and radio inputs.

Bass and Treble Controls

Bass and treble controls are incorporated to enable the listener to compensate for room acoustics, studio or recording deficiencies, etc. For average use, a lift and cut of 12 dB at 50 c/s and 10 kc/s is satisfactory. Although switched, frequency conscious R/C networks were once common, the two control systems shown in Figs. 15 and 17 are used in most amplifiers today. Fig. 15 is a passive network, Fig. 17 a feedback circuit known as the Baxandall. The passive tone control has the effect of rotating the frequency response about a central hinge, usually

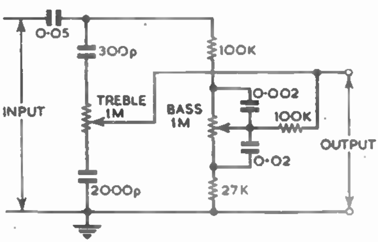


FIG. 15.—PASSIVE TONE CONTROL NETWORK.

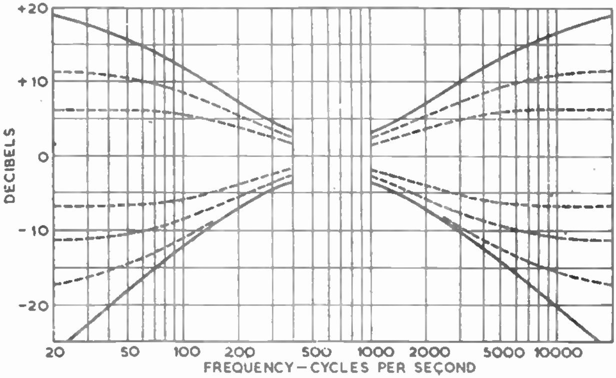
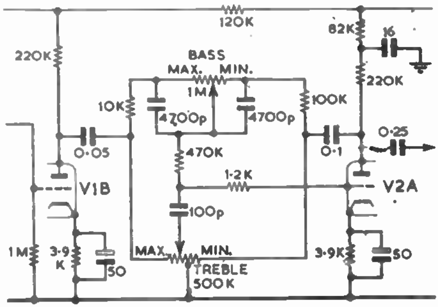
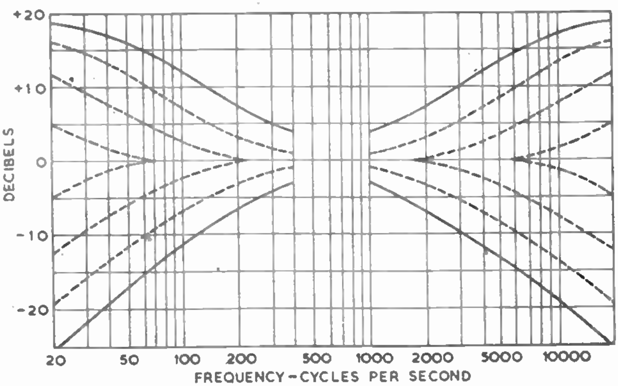


FIG. 16 (*top*).—FREQUENCY RESPONSE CURVES FOR PASSIVE TONE CONTROL.

FIG. 17 (*right*).—THE BAXANDALL TONE CONTROL CIRCUIT.

FIG. 18 (*bottom*).—FREQUENCY RESPONSE CURVES FOR BAXANDALL CIRCUIT.





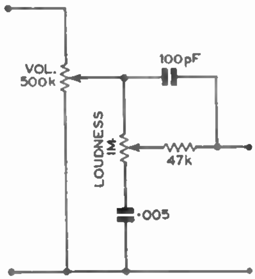


FIG. 19.—SIMPLE LOUDNESS CONTROL.

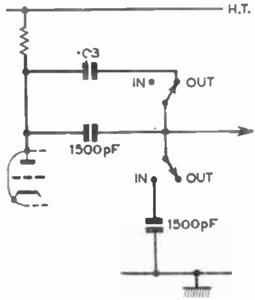


FIG. 20.—SWITCHED RUMBLE FILTER.

1,000 c/s (see Fig. 16), whilst with the Baxandall system the lift and cut is initially confined to each end of the scale (see Fig. 18). This means that with the Baxandall system the extreme low frequencies can be lifted appreciably without affecting the response in the region of 300–500 c/s. A further advantage of the Baxandall system is that, as it employs negative feedback, distortion is kept to the minimum. Nevertheless, many people prefer the shape of the contours of the treble curves in the passive network.

Loudness Control

Because the frequency response of the human ear is not only non-linear, but also varies according to sound pressure, bass boost and, to a lesser extent, treble lift are desirable when listening at low-volume levels, otherwise the reproduction tends to sound thin and unnatural. The Fletcher–Munsen and Robinson–Dadson equal-loudness curves indicate the relative intensities necessary to produce sounds of the same apparent loudness at different levels. Many pre-amplifiers incorporate a separate control to enable both the bass and treble to be increased at low levels, the basic arrangement usually adopted being shown in Fig. 19. The curves provided by such controls are somewhat of a compromise.

Filters

In order to avoid distortion produced by radio interference from adjacent transmissions, or on gramophone records due to processing defects or wear, it is often necessary to restrict the upper frequency response by means of a low-pass filter. Four switched positions, giving nominal roll-offs between 5 and 12 kc/s, are normally provided. Either a tuned inductor or a parallel-T circuit is generally used, and, provided that they are accurately designed, there is little to choose between them. A parallel-T filter circuit is shown in Fig. 21, and an ingenious circuit for a continuously variable filter circuit is shown in Fig. 22.

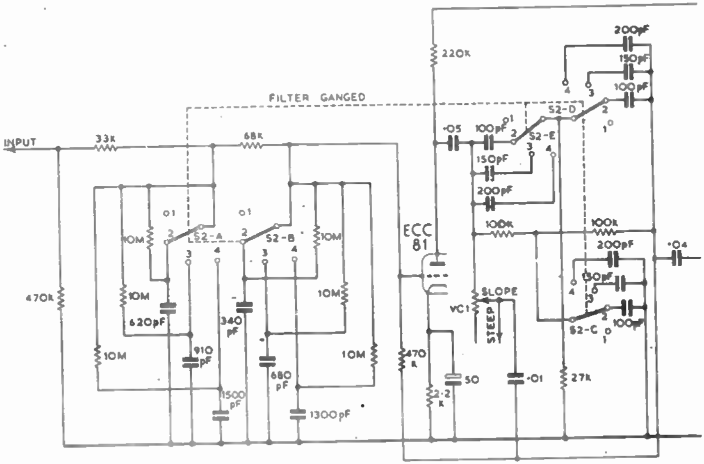


Fig. 21.—Switched Low-pass Filter Circuit Using "Parallel T" Network.

A worthwhile refinement with a low-pass filter is the inclusion of a variable-slope control permitting the rate of attenuation to be varied from a gentle slope to a steep cut. A typical control will permit variations in slope from 8 to about 35 dB per octave.

As far as the low-frequency end of the scale is concerned, it is desirable to fit a high-pass filter to reduce the effects of motor rumble and sub-audio-frequencies that may otherwise overload the amplifier. This rumble filter should attenuate frequencies below 35 c/s at not less than 12 dB per octave, and can conveniently be combined with the equalizing negative feedback loop. Sometimes a rumble filter is provided that may be switched in and out of circuit as required: a typical arrangement is shown in Fig. 20.

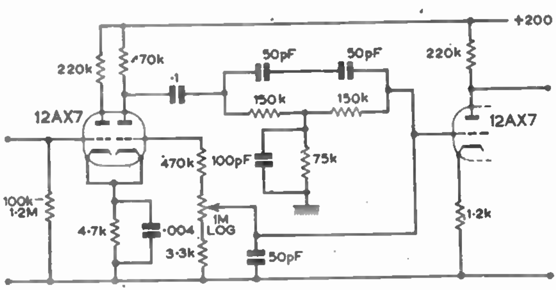


Fig. 22.—Continuously Variable Low Pass Filter.

Presence Control

A presence control is incorporated in some pre-amplifiers to give a rising mid-frequency characteristic, it being usual in this case to provide a lift of about 5 dB at 3,000 c/s. The control may take the form of an R/C network which can be switched into, or out of, a negative feedback loop. The point of this control is that the ratio of high to low frequencies is greater near an orchestra than farther away, where reverberation and absorption influence the sound heard. The effect of the presence control is to simulate the sound nearer the orchestra, providing a more accurate curve than can be had from the normal tone controls.

Output

A refinement found in some of the more elaborate pre-amplifiers is a cathode-follower output stage which, because of its low output impedance, allows long connecting cables to be used without causing undue high-frequency losses, and, incidentally, minimizes the risk of hum pick-up.

Distortion

Due to the difficulties encountered when making distortion measurements at the low levels involved in the case of pre-amplifiers, distortion figures are often quoted for the pre-amplifier and power amplifier together. A well-designed pre-amplifier should not contribute more than 0·1 per cent total harmonic distortion to the maximum rated output of the two.

STEREOPHONIC SOUND

However high the quality of a single-channel, monophonic system, as the only dimension that it can reproduce is depth, it lacks perspective. The object of stereophonic sound systems is to overcome this limitation. There is, however, still some doubt as to the exact mechanism by which our ears determine position, and experimental work on this subject continues.

Localization in Azimuth

At first sight it would appear that side-to-side location is effected by the difference in intensity at the two ears. This, however, is not entirely true of frequencies below about 700 c/s, as the difference in intensity between the signals received at the two ears at such frequencies is so small as to be negligible. Early experiments suggested that at low frequencies location in azimuth is achieved by phase differences, whilst at high frequencies intensity differences are the important factor. The conception grew out of later experiments that the important factor is time difference at the ears rather than phase difference. This theory permits the effect to be of importance at high as well as low frequencies, the difference between time and phase being of purely academic interest at low frequencies, but of major importance at high frequencies, where phase is random. The general conclusion is that at frequencies below

1,000–1,200 c/s, localization in azimuth is achieved by observation of the phase or time difference at the two ears, whilst at high frequencies the ears work on time difference assisted to some degree by intensity differences.

Localization in Zenith

There seems to be little doubt that the important factor in vertica localization is head movement. As intensity differences become quite sharp at high frequencies, they appear to help in making vertical direction finding possible with the minimum of head movement.

Judgment of Distance

Evidence suggests that, in free space conditions, previous experience of a sound is required to make even a guess at distance. In enclosed spaces the necessary indication would seem to lie in the ratio between direct and reverberated sound. However, this appears to be true only for sounds of a transient nature. A further clue to the distance between the sound source and the observer appears to be given by the quality of the received sound: judgment of this also requires previous experience of the sound at other distances.

Basic Methods of Stereophonic Reproduction

A pair of microphones clamped to either side of a dummy head at the sound source, and wired separately to a pair of headphones worn by the listener, will provide the listener with stereophonic sound. To provide the sound reproduction via loudspeakers, however, adds the complication that the listener will receive at each ear a certain amount of the sound from each loudspeaker used. Two main approaches to overcoming this difficulty have been employed in practice, the Bell "Wavefront" system and the E.M.I. "stereosonic" system. The first is the basis of the various systems used in cinemas, the second is the dual-channel system introduced for domestic tape equipment, and more recently adapted for disc recording.

"Wavefront" System

If one considers a "curtain" of an infinite number of microphones connected by an infinite number of channels to an infinite number of loudspeakers in a curtain, any sound wavefront striking the microphones will be reproduced by the loudspeakers, and a listener in front of the sound-reproducing curtain will hear a reproduction containing all the information necessary for him to have a perfect sound picture of the original sound source. Now for the reproduction of music or plays all that is really necessary is location in the horizontal plane, and therefore the "curtain" of microphones and loudspeakers may be replaced by a line of microphones and loudspeakers. Experiments using a line of three microphones feeding, via three amplifiers, a line of three loudspeakers, showed that this arrangement will give quite accurate localization in azimuth and slightly less accurate localization in depth. The stereophonic sound tracks used in present-day films are generally based on this system, usually having three channels carrying the main programme and feeding three loudspeakers behind the screen, a fourth channel carrying "effects" and feeding loudspeakers placed at intervals round the auditorium.

This "wavefront" system requires three or more channels, and is therefore not generally regarded as an economic proposition for domestic sound-reproducing equipment.

"Stereosonic" System

This system is based on the work of A. D. Blumlein, in 1929. Although devised so long ago, the system was not a commercial proposition until recently because of the lack of a suitable material on which to record the signals. The two channels used in this system may, however, be complex-cut in the groove of the modern plastics long-playing record, or recorded on the two tracks of current magnetic tape.

The recordings are made in such a manner that when reproduced on a dual-channel reproducer the loudspeakers are fed with correct relative amplitude signals at all frequencies, in order to produce at the listener's ears the same vector sound pressures as would have been heard by direct listening in a corresponding position in front of the sound source.

Under normal listening conditions the loudspeaker spacing will be less than the width of the original sound source: the angle subtended by the sound image at the listener can, however, be the same as that subtended by the original sound source at the best spot for listening in the recording studio.

The effects of listening-room reverberation can usually be ignored as the reverberation time of most domestic rooms is small compared with that of large recording studios and concert halls, and therefore the information about distance provided by reverberation is not substantially modified by the listening room.

Blumlein's work was based on his idea that the main factor in locating a source of sound was the difference in arrival time of the sound at the two ears. In the system he devised, the outputs of two microphones, placed close together, have their phase differences converted to amplitude differences, the resultant signals being fed to two loudspeakers spaced widely apart. If the relationship between phase and amplitude differences is correctly proportioned, a listener sitting in front of the loudspeakers will receive sound pressure at his ears with a similar difference in phase to those he would have received had he been in front of the original sound source.

Stereophonic Recording

One arrangement for making recordings with this system is shown in Fig. 23. It consists of a pair of velocity microphones, with their axes of maximum response at right angles; the outputs from these are passed through sum and differences transformers, the difference channel having a loss of 3 dB inserted at all frequencies above 700 c/s by the "shuffler" to allow for the different method of localization in azimuth at the higher frequencies. The resultant voltages, after passing through a further set of sum and difference transformers, are used as

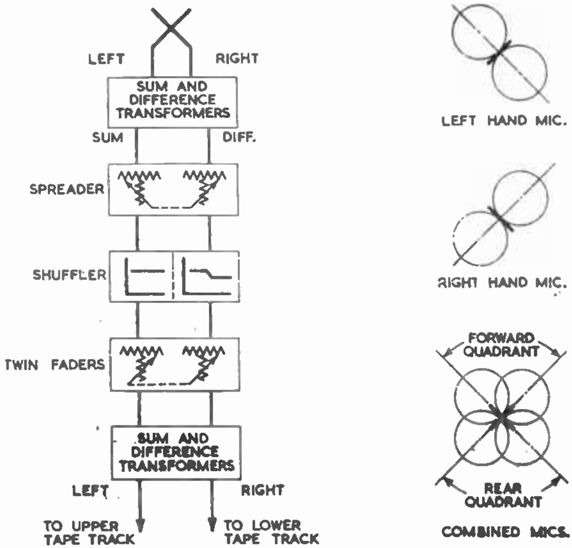


FIG. 23.—ONE ARRANGEMENT OF THE "STEREOSONIC" RECORDING SYSTEM. MICROPHONE POLAR RESPONSES ARE SHOWN ON THE RIGHT.

inputs to two identical recording amplifiers. The fader consists of two controls accurately matched to control the general level of the complete recording. The spreader is a differential attenuator used to modify the apparent angle of the reproduced sound, and is used in several ways. For example, if the microphone has to be positioned at some distance from an orchestra the result will be an apparent reduction in the stage width; this can be corrected by attenuating the sum channe in relation to the difference channel. Further information on stereo phonic recording is given in Section 34.

Stereophonic Reproduction

A block, schematic diagram of a stereophonic reproducer is shown in Fig. 24. In this, the stereophonic input is from a tape recording via a dual head, but the input may, of course, equally be from gramophone record via a stereophonic pick-up. The gains of each channel should be kept within a decibel at all settings of the volume and tone controls, and phase shifts at frequencies below 1 kc/s have to be carefully controlled. The balance control is provided to enable the gain of the two channels to be varied differentially over a small range, so that electrical correction can be made for acoustic unbalance in a room. Acoustic unbalance would cause lateral shift of the sound image, and

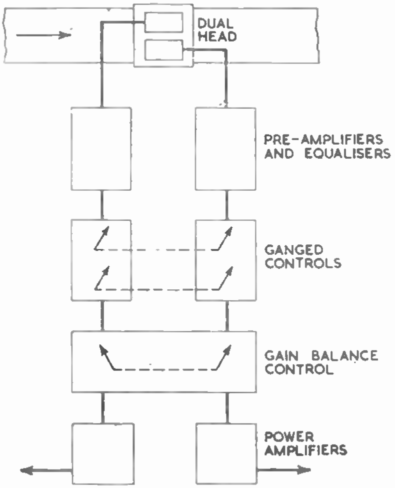


FIG. 24.—BLOCK DIAGRAM OF A "STEREOSONIC" REPRODUCER.

correction is made by raising the gain of one channel relative to the other. A balance-control circuit is shown in Fig. 25. The total differential range is 18 dB, with a loss of 1 dB when the control is at its mid-position.

The loudspeakers for reproduction should not be any closer than 5 ft. apart.

The listener should not sit closer to one loudspeaker than half the distance between the two loudspeakers.

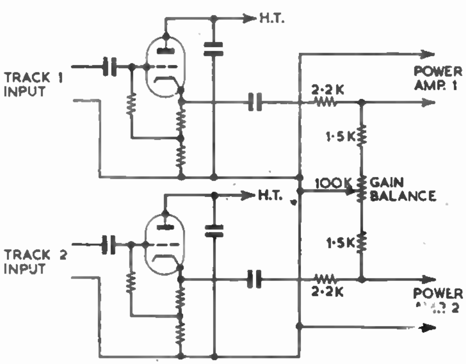


FIG. 25.—"STEREOSONIC" BALANCE CONTROL.

(Brit. Patent Application No. 9763/55)

Stereophonic Discs

A dual-channel system with the recording in a single groove may be achieved by using hill-and-dale (vertical displacements) recording for one channel and lateral (horizontal displacements) for the other. This is known as the " L/V " system. In practice, a system is used whereby the planes of the two modulations are at 45° to the surface of the disc (and 90° to one another). This is known as the " 45/45 system ". The reproduction pick-up may be arranged so that one set of movements are sensed by one transducer, the other set of movements being sensed by a second transducer. With the two channels at 90° to each other, there is little interaction between them.

The basic principle of the " 45/45 " system is shown in Fig. 26. The displacements of the recording cutter-stylus, under the influence of the two channels being recorded, results in a combined movement which changes the position of the stylus tip in both vertical and lateral planes.

If exactly the same signal (in both intensity and phase) is fed from both channels, the vertical movements will cancel out, leaving only the horizontal movement. Thus the records cut in this way can in theory be played on monophonic equipment. In practice, differences in the stylus size, etc., for monophonic and stereophonic reproduction preclude this. If, on the other hand, the phase of the signals being recorded differs by exactly 180°, and the signals are of the same amplitude, then the horizontal movement cancels out, leaving a vertical movement. In stereophonic recording there will normally be a difference in either phase or amplitude or both between the signals received from the two microphones. With the two microphones receiving the two sets of signals differing slightly in phase and amplitude, the tip of the stylus

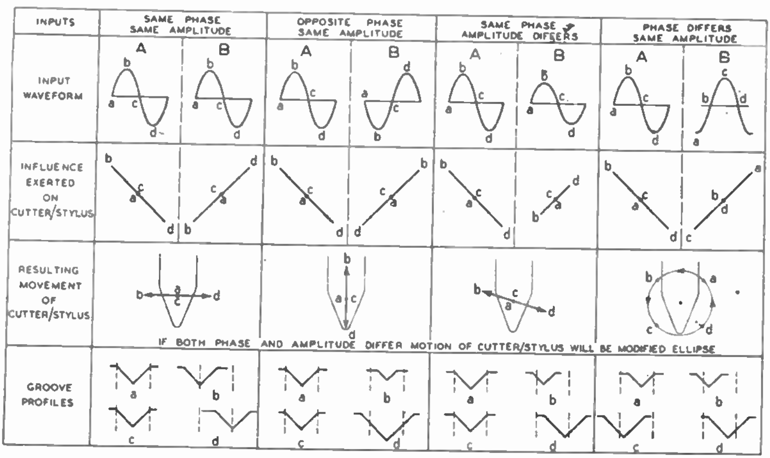


Fig. 26.—Basic Principles of the " 45/45 Out of Phase " Dual-Channel Disc Recording System.

may at any instant be to the left or right, or above or below its position of no modulation.

Advantages of the 45/45 system over the L/V system are that it is less affected by the vertical component produced by the motor rumble, as in the 45/45 system it affects both channels equally, and that each channel is similarly affected by tracing distortion.

The recommended stylus tip for stereophonic reproduction is 0·0005–0·0006 in. (12·5–15 μ, though in practice tips up to 0·00075 in. are used). A fine stylus is required because the grooves vary in width and depth (see Fig. 26), and, at the limit of vertical displacement, the groove may be about half the depth and width of a single-channel micro-groove disc.

The recording characteristic for stereophonic recording is similar to the R.I.A.A. curve.

Information on pick-ups for stereophonic reproduction is given in Section 31.

SOUND AMPLIFICATION AND DISTRIBUTION

Sound amplification and distribution equipment is used for the purpose of communicating information and entertainment either simultaneously to a large number of people or over an area greater than would otherwise be possible. This may have the purpose of sound-reinforcement—i.e., where the audience is within visual and partial audible range of the programme source—or for the convenience of distribution over a large area, as with overflow meetings, intercommunication systems, hotel paging and similar installations. But although the equipment may vary greatly in power-handling capabilities, all installations comprise the same basic elements: (1) a transducer for converting the programme source into electrical impulses i.e., microphone, gramophone pick-up, radio receiver or tone oscillator; (2) an amplifier whose gain and maximum output will be governed by the requirements of the particular installation; (3) loudspeaker(s). Also to be considered are the planning and location of the equipment, the interconnection wiring between the three units, the system of control and, for more elaborate installations, the problems of monitoring.

Although the theoretical aspects of sound distribution are receiving increasing attention, in practice empirical methods are still widely used for the determination of acoustic and electrical power requirements, and for the siting of loudspeakers. This is in part because of the wide range of acoustic conditions and defects—reverberation, echo, resonances, noise levels and dead spots—found in buildings other than those designed specifically with these problems in mind, and in part owing to the ease with which the output of an amplifier is controllable over a wide range, rendering it unnecessary to make close calculations as to power requirements.

Typical power requirements (due to Pamphonic Reproducers Ltd.) are given in Table 1. It should, however, be borne in mind that horn loudspeakers are, for example, from three to five times as efficient as cabinet-types.

TABLE 1.—TYPICAL POWER REQUIREMENTS

Location	*Size*	*Noise Level*	*Watts*
Office	20 ft. × 20 ft.	Quiet	2
Assembly shop	100 ft. × 40 ft.	Medium	25
Machine shop	100 ft. × 40 ft.	Very high	60–150
Theatre (music)	800 people	—	60
Theatre (sound effects)	800 people	—	60–250
Church	100 people	Quiet to medium	60
Railway terminal	10 platforms	High	500–1000

For many applications, an average of roughly 1 acoustic watt for each 100 sq. ft. of floor space is adequate; this is equivalent to about 2 electrical watts per 1,000 sq. ft., assuming a loudspeaker efficiency of the order of 50 per cent.

Recommendations as to the frequency range of amplifiers and loudspeakers are contained in the British Standard Code of Practice on Sound Distribution Systems (CP : 327.300 : 1952). In this, quality of aural results is divided into two categories " A " and " B ". " A " is considered desirable where the quality of reproduction should not be readily distinguishable from the original, and is likely to be achieved only where the acoustic properties of the surroundings are suitable—a frequency range of 50–7,000 c/s is suggested. " B " quality is for applications where fidelity is not all-important but where the prime consideration is intelligibility—the suggested range is 100–4,000 c/s, with the proviso that in noisy and/or reverberant locations the lower limit of 100 c/s can be raised with advantage. Some authorities consider that an acceptable range for speech reproduction is 300–3,000 c/s.

Where long runs of loudspeaker cables are required, the attenuation of the higher audio frequencies caused by cable capacitance may need consideration. For " A " quality installations capacitance losses can be reduced by limiting audio line voltage to 50 volts; this may be raised to 120 volts for " B " installations.

For permanent installations, where the operation and day-to-day maintenance may be in the hands of persons unskilled in electronics, it is highly advantageous for all components and valves to be run well within their rated values, and for adequate metering to be included to facilitate the clearing of simple faults. The selection of dry, clean sites for equipment, the provision of adequate ventilation and the establishment of a regular maintenance schedule—to be carried out by engineers with a specialized knowledge of the equipment—will also assist in securing reliable service. A running log book in which interruptions to service are entered should be kept.

J. A. R.

DELAYED SOUND REINFORCEMENT

Where a large area has to be covered, with the sound emerging from more than one loudspeaker, the listener—owing to the relatively low velocity of sound—may hear each syllable a number of times: first, from the loudspeaker nearest to him, and then as a series of echoes from other loudspeakers or from the original source of the sound. However, if a delay can be introduced so that the sound emerges from each loudspeaker at approximately the same instant that the sound from the original source (and from any loudspeakers nearer to the source than the one under consideration) reaches the loudspeaker position, then the distortion due to the receipt of multiple signals will be eliminated. For each 50 ft. between the source and a particular loudspeaker, it will be necessary to introduce a delay of approximately 0·045 second.

The practical methods of achieving delay are shown in Figs. 27 and 28. Fig. 27 shows the use of a delay tube in a theatre: the output from one loudspeaker is passed through a specially prepared tube at the far end of which is placed a microphone feeding a further amplifier; distortion introduced by the tube can be overcome by means of compensating circuits.

Fig. 28 shows a method of achieving delay by means of a continuously rotating turntable with magnetic recording material around its circumference. The output from the microphone is first amplified by a recording amplifier, the output of which passes to a recording head. A number of play-back heads pick up the recorded sound after a delay depending upon the speed of the turntable and the spacing of the heads. This spacing can be adjusted when the system is installed. After passing all play-back heads the magnetic material passes a continuously-functioning erase head, and is then ready for a further sequence of recording and play-back. The output from each play-back head is fed to a separate power amplifier feeding loudspeakers at a fixed distance from the source of the sound. For reverberant buildings, the delay mechanism may be used in conjunction with loudspeaker columns designed so that the sound is emitted in a beam which is wide in the horizontal plane (about 120°) and narrow in the vertical plane (the vertical beam angle for -6 db with reference to the axis may be as little as 10°).

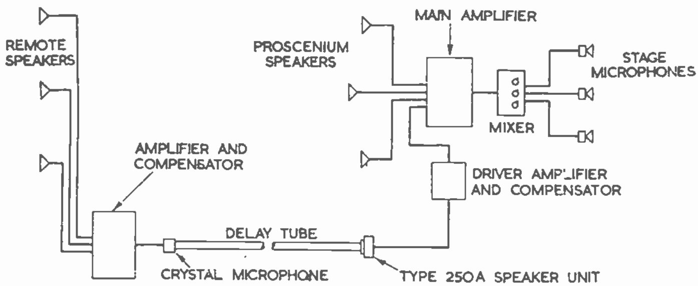


FIG. 27.—BLOCK DIAGRAM OF THE B.T.H. SOUND DELAY SYSTEM.

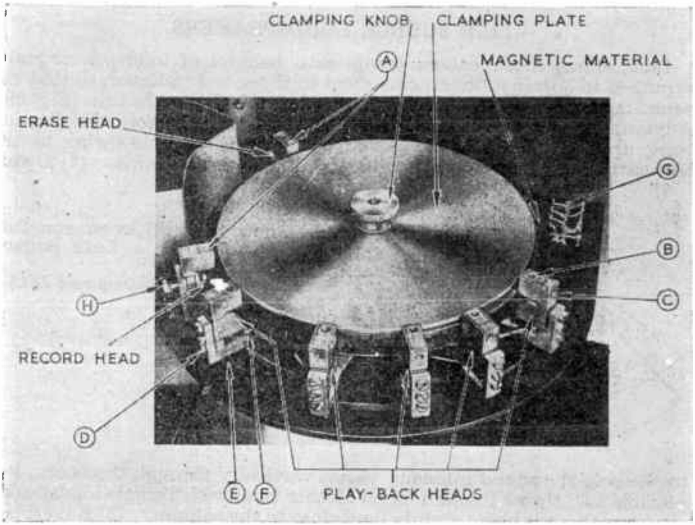


FIG. 28.—MAGNETIC DELAY MECHANISM.
Points A–E are adjustments for installation.
(*Pamphonic Reproducers, Ltd.*)

WIRING AND CABLES

For light flexible use, a good-quality twin-feeder television cable will be satisfactory for microphones.

Stout cable to withstand rougher use is made by the principle cable manufacturers, and should be twin, twisted, braided, with an outer protective covering.

For permanent installations 10 or 20 lb. cable, twin twisted lead-covered cable should be run in conduit.

Microphone and input circuits require efficient screening to avoid hum pick-up and audio-frequency feedback : it is usual to specify lead-covered twin cables. These cables should not be run within about 6 in. or so of any power, loudspeaker or telephone circuits.

Gauge of cable for loudspeaker wiring is decided by the load to be carried : V.I.R. or T.R.S. of gauge 3/0·029 or 3/0·036 will suit most purposes.

Since the output circuits customarily operate at 100 volts r.m.s., wiring contractors should treat such wiring as medium voltage power circuits, and provide insulation equivalent to 250-volt C.M.A. cables. Special cables for this purpose are available.

LINE SOURCE LOUDSPEAKERS

Line source loudspeakers comprise a column of loudspeaker units arranged to obtain a directional effect in the sound radiated, so that the sound is directed at an audience in one flat, fan-shaped beam: (Fig. 29.) Advantages of this system for public address and sound distribution work are: (1) Considerable power economy is possible owing to the concentration of the acoustic output into the desired area. (2) Owing

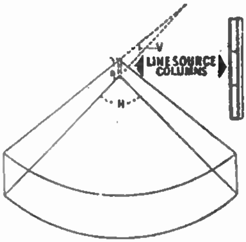


FIG. 29.—SHOWING THE FAN-SHAPED DISTRIBUTION PATTERN OF A LINE SOURCE COLUMN.
(*Pamphonic·Reproducers Ltd.*)

to the way the sound intensity varies vertically through the beam, it is possible to " throw " sound much farther than with a single loudspeaker source without it being unduly loud close to the column. (3) Reverberation effects, caused by randomly-timed echoes from boundary surfaces, are much diminished since very little sound is radiated into the reverberant upper parts of buildings.

For outdoor use, the beam angles can be made much smaller than for the corresponding indoor columns, hence the power gain and range are both greater; these columns have much greater power handling capacity. Under ideal conditions, the range of an 11 ft column could be about 500 yards.

The units of one manufacturer have a horizontal coverage of about 120°; the vertical beam angle (V) for −6 dB with reference to the axis varies from about 16° for a 6 ft. indoor column to as little as 7° for an 11 ft. outdoor column. The 6 ft. indoor column is rated at 5 watts input power and the 8 ft. 6 in. and 11 ft. columns are rated at 10 watts. Rated powers for the outdoor columns are: 6 ft. 40 watts; 8 ft. 6 in. 57 watts; 11 ft. 75 watts.

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* Journal of the British Institution of Radio Engineers, September, 1945.

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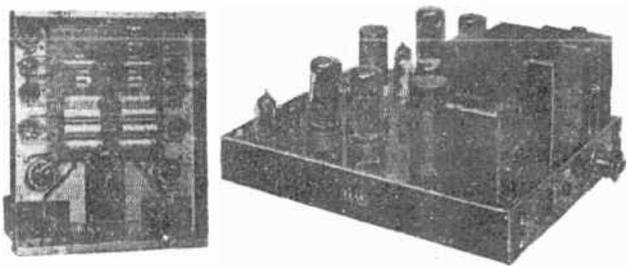

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SW/54

38. MEASURING INSTRUMENTS AND TEST EQUIPMENTS

38. MEASURING INSTRUMENTS AND TEST EQUIPMENTS

MEASURING INSTRUMENTS

General Considerations, Forces and Accuracies

Measuring instruments of the "meter" variety used in the radio profession, with the exception of those in high-powered transmitters, are generally confined to the "small normal" and "miniature" sizes. Their scale arcs are usually 100–110°, or 80° in the lower-grade instruments.

The primary consideration of an indicating instrument is the low value of electrical power it absorbs from the circuit being metered. This necessitates a high ratio of mechanical torque exercised on the spindle to weight of the moving parts. Frictional resistance of the bearings is reduced to negligible proportions by the use of hardened and highly polished steel and/or sprung jewels. Further accuracy is achieved in some instruments by the use of an "anti-parallax mirror" mounted behind the indicating pointer. In these cases the pointer and its reflection should be aligned when taking a reading.

To attain a low electrical power consumption, a low value of operating current through the basic movement is normally desirable. Hence, the lower the value of current required to effect FULL-SCALE DEFLECTION (F.S.D.) of the indicating pointer, the more versatile becomes the basic movement around which the concerned meter is constructed. Incidentally, this results in a more expensive unit. From the viewpoint of minimum power consumption there are two simple rules, these being :

(1) A current-measuring device must be wired in series with the metered circuit, and should thus have low resistance in order that it does not produce appreciable voltage-drops.

(2) A voltage-measuring device must have a high value of resistance, as it is wired in parallel with the circuit or part thereof being metered. The extra current demanded from the metered supply is thus kept to a minimum.

Both considerations involve a low value of electrical power "wasted" in the meter, these objects being attained in (1) by the low voltage-drop (E) across the current-recording instrument and in (2) by the low value of bleeder current (I) in the voltage-indicator.

Four forces operate on the spindle carrying the pointer, these being the actuating force, a damping force, mechanical friction and a restoring force.

The actuating force is provided electrically by the metered circuit in the form of ampere-turns, electrostatic attraction and the like. This actuates the pointer, tending to drive it up the scale by imparting angular movement. The damping force or "control torque" operates against the actuating force, compelling the pointer to assume its indicat-

FIG. 1.—SPRING CONTROL.

ing position of rest without violent oscillation. Two general methods of achieving this are in use, these being

(1) By generation of eddy currents in the metallic former on which the actuating coil is wound, or in an aluminium plate fixed to the pointer and operating in the field of a permanent magnet. By Lenz' Law, the magnetic flux produced by the eddy currents opposes that of the actuating current, while by Faraday's Law its magnitude increases with the rapidity of the pointer movement; hence proportional damping is effected.

(2) By the braking action of a light piston or vane acting in a practically closed air chamber.

The restoring force returns the pointer to the "zero" position on removal of the actuating force, and may also be provided by one of two methods, these being:

(1) By the action of a flat, concentric spring, as shown in Fig. 1. Sometimes two oppositely wound springs positioned on either side of the pointer-bearing shaft are used, these being self-compensating should temperature variations affect them. This is further minimized by using phosphor-bronze springs, as this material has a lower linear coefficient of expansion than spring steel. Furthermore, it is non-magnetic. The springs usually act as lead-in and lead-out wires to the actuating coil.

(2) By gravity control. In this system a small balance weight is attached to the lower extension of the pointer, this sometimes being adjustable by threaded-pendulum operation. This is usually employed only in vertically mounted instruments of the cheaper variety.

An eccentrically mounted fine wire fixed to a screwdriver-adjusted threaded bolt, and operating in an inverted U-stub on the lower extremity of the pointer, provides final zero-correction.

Voltage-indicating instruments and current-indicating instruments are graded according to their maximum permissible rated inaccuracies as shown in Table 1:

TABLE 1.—GRADING OF VOLTAGE- AND CURRENT-INDICATING INSTRUMENTS

	Grading	*Max. Permissible Inaccuracy* (%)
1	Sub-standard	Within 0·2
2	1st grade	Within 1
3	2nd grade	Within 2
4	Ungraded	—

The sub-standard limit applies to full-scale deflection values, whilst other standards are concerned with half-scale readings.

The comparable permissible inaccuracies for wattmeters are 2·5 times those shown in the table.

Galvanometers

The function of a galvanometer is to detect minute currents, the values of which may be considerably less than 1 μA. Several entirely different types may be met, these being designed chiefly for laboratory use. Fig. 2 shows the general design and construction of the more common type, which operates on the "moving-coil" (MC) principle. The magnetic field is concentrated in the air-gap between the pole faces of the permanent magnet by the soft iron core (SIC). This also provides a uniform and truly radial magnetic field in which the moving coil operates. The permanent magnet is aged by a special process to ensure

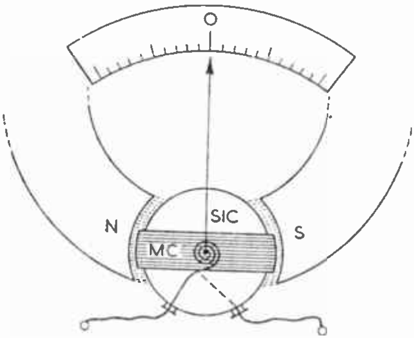


FIG. 2.—GALVANOMETER.

constancy of flux throughout its life. The moving coil is wound with fine-gauge, insulated wire on a light, aluminium former, which is pivoted in sprung jewel bearings. The large number of turns provides the necessary ampere-turns for high sensitivity, another contributory factor to this being the small air-gap, which may be of the order of 0·05 in. The energizing current is fed to the moving coil via oppositely wound control springs. The actuating force is provided by motor action between the permanent field and the magnetic field established around the moving coil by the energizing current. The resultant torque is given by $\frac{BINA}{10}$ dynes, where B = flux density, I = actuating current, N = number of turns on the moving coil and A = cross-sectional area of the moving coil.

Damping is effected by the decelerating action between the magnetic poles associated with the eddy currents induced into the coil former and those of the permanent field, this increasing with the speed of the pointer deflection.

Errors due to external magnetic fields are eliminated by enclosing the actuating section in a mumetal screen. As the torque varies with the value of actuating current, the pointer deflections are linear, hence the calibrations have equally spaced increments.

With low-grade galvanometers, such as that shown in Fig. 2, full-scale deflection values of 20 μA or so are possible. The centre-zero permits double-reading, as the pointer is permitted to swing in either direction according to the direction of the energizing current. This is useful for determining the direction of the current being detected.

Moving-coil Instruments

In general principles, a moving-coil basic movement is merely a more robust version of the galvanometer already described. Consequently, higher full-scale deflection values of current result, these ranging down to a minimum of 100 μA or so.

In spite of its somewhat high cost, this type of meter is the most used of any in radio engineering. It may be constructed double-reading or single-reading. As a reversal of the energizing current results in a

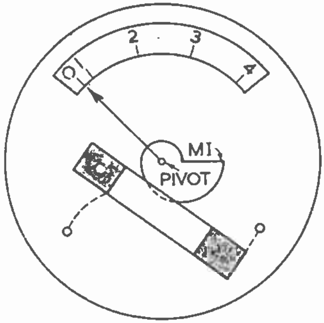


Fig. 3.—Moving-iron Instrument, Attraction Type.

reversal of the actuating torque, it can be used only on a D.C. supply, and care must be exercised to observe the current polarity to its terminals. In its favour are a high degree of accuracy, the linear scale and negligible interference from external fields.

The volt-drop developed across the basic movement of high-quality instruments at full-scale reading should be as low as 50 mV or so.

Moving-iron Instruments

These movements are used in the cheaper grades of battery testers and in certain panel-mounted instruments operating from A.C. supplies. Two general types . in use—the attraction type and the repulsion type. Both operate on the principle of magnetic induction, i.e., that a piece of soft iron becomes magnetized when subjected to a magnetic field flux, this resulting in attraction between it and any other pole of opposite polarity or repulsion between it and any other pole of similar polarity.

Fig. 3 shows the general construction of an attraction type of movement. The actuating current passes through the fixed coil C, the associated flux inducing magnetism into the lightly suspended " moving iron " (M.I.) such that its polarity is opposite to that of the energizing current. Attraction thus takes place between the two fields, this causing the moving iron to swing into the core of C and operate the pointer. As both magnetic fields are proportional to the value of actuating current, the resultant attraction, which provides the driving torque, is proportional to I^2. The scale-calibrations are therefore cramped to the left and extended to the right, this being termed SQUARE LAW. Partial alleviation is achieved in some models by designing the actual physical bulk of the moving iron so that it becomes saturated magnetically at moderately low values of energizing current. By this means, good-quality moving-iron instruments are available, with some cramping over the first 10 per cent of scale, but almost linear over the remainder.

Fig. 4 shows the general construction of a repulsion type of moving-iron meter. Two pieces of soft iron are used, one being fixed (F) and the other moving (M), this being attached to the pointer-bearing-shaft and carefully balanced. The flux associated with the energizing current flowing through the coil C induces magnetism of similar polarities into both fixed and moving irons, hence repulsion ensues. M thus moves away from F to operate the pointer. No arrangements are made in this type for swinging either of the irons near saturation.

The effect of reversing the direction of energizing current is the same in both types, i.e., both concerned fluxes become reversed in polarity, so attraction still takes place in the attraction type or repulsion in the repulsion type. Hence both types are UNIVERSAL, i.e., they will operate on A.C. or D.C. supplies alike. Actually they are more accurate on A.C supplies, as the use of special nickel–iron alloy reduces hysteresis errors to a negligible degree. Other points in favour of moving-iron instruments are cheapness and robustness.

In its disfavour are the low degrees of accuracy, high values of full-scale-deflection current and susceptibility to errors caused by external fields, as the operating field is weak. A frequency error is introduced by the reactance of the meter coil and by induced eddy currents in metalwork near the working section of the meter. In older designs hysteresis in the soft iron results in higher readings with descending values of actuating current than with ascending values.

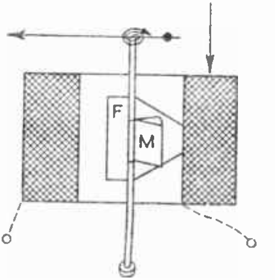


FIG. 4.—MOVING-IRON INSTRUMENT, REPULSION TYPE.

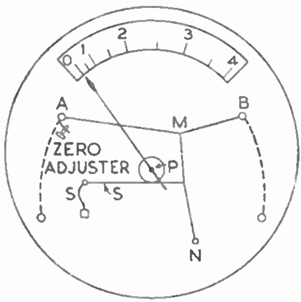


FIG. 5.—HOT-WIRE AMMETER.

Hot-wire Ammeter

This instrument operates on the expansion principle of a wire under heat treatment. Fig. 5 shows the essential components of a hot-wire assembly. AB is a "heater wire" through which the energizing current is passed. It has a high coefficient of expansion, a high melting point and a moderately high resistance (in comparison to its length). The heater wire expands due to the heating effect (I^2R) of the actuating current, the resultant sag being taken up by a silk cord SC which passes around a pulley P, this carrying the pointer. The heat developed in AB is isolated from the tautening cord SC by a phosphor-bronze wire MN. The cord thus maintains a constant length whilst its tension is held steady by the spring S. This causes the pulley to rotate and turns the pointer up the scale when the heater wire expands.

As the expansion of AB is proportional to I^2 and not to I, the scale calibrations are square law. The action is inherently sluggish, and damping is not often employed, although some of the better-class instruments are equipped with light metal vanes swinging in a magnetic field. The primary consideration of this arrangement is to avoid rapid pointer movements which would throw excessive strain on the heater wire. Variations in ambient temperature necessitate frequent zero-adjustment. Full-scale deflection current is arranged to take the heater wire almost to its melting point, so even a short overload may fuse it.

Hot-wire ammeters are "universal", but errors occur on A.C. if the waveform is not sinusoidal. At higher frequencies (radio frequencies), skin effect introduces a more serious error. Losses are high compared with moving-coil or moving-iron types, and due to the degree of heating required to effect heater expansion full-scale-deflection values below 0·5 amperes are rare.

For radio frequencies, this type of instrument has been virtually superseded by the thermo-junction type.

Rectifier Instruments

These were developed due to the need for greater accuracy on alternating supplies than that generally attained with moving-iron instruments. Step-down transformation is customary with suitable interpolation of the scalar calibrations. Small "instrument-type" metal rectifiers of the copper oxide or selenium variety are connected in a full-wave "bridge" arrangement as shown in Fig. 6. The full and dotted arrows indicate the alternate half-cycle movements of current flow, and as these both pass through the moving-coil basic movement in the same direction, D.C. operation ensues.

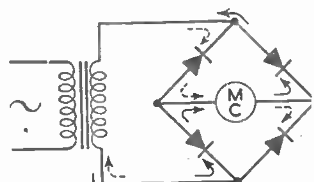


FIG. 6.—RECTIFIER INSTRUMENTS.

The Full and Dotted Arrows Indicate the Directions of the Current Flow on Alternate half-cycles.

When used at radio frequencies, care must be exercised to minimize the distributed capacitances. Errors are incurred if the applied waveform is not pure sinusoidal, as the recorded deflection current must then be multiplied by the form factor to provide the true R.M.S. value.

Rectifier instruments are chiefly used in the recording of small current-values. Shunting for higher full-scale-deflection recordings is not a practical proposition, as the rectifier resistance decreases with the input amplitude, this also producing a waveform error.

Thermo-junction Instruments

Basically, a thermo-junction instrument consists of a moving-coil millivoltmeter connected across the contact faces of a THERMO-COUPLE, the action of which is as follows: If the junction of two series-connected dissimilar metals is heated, there is a potential difference present between the contact faces. The accurate and highly-sensitive millivoltmeter is connected to the dissimilar metals and records the potential difference, the scale calibrations being suitably interpolated in terms of current or higher ranges of voltage.

The actuating current operates on a heater element positioned near the metals-junction. Suitable metals are nickel, silver, eureka, zinc, iron, bismuth, antimony, tellurium, etc. Iron and eureka are popular in view of the ease with which they are spot-welded. The thermo-metals may be housed in airtight holders, and to increase sensitivity several couples may be wired in series.

The errors in thermo-ammeters at radio frequencies are lower than in the hot-wire types, and the overload capacity is higher. The action is sluggish, and readings over the first 20 per cent of scale are unreliable, but the accuracy over the remainder of the scale is higher than in hot-wire types. Full-scale-deflection values of current range down to 0·3 ampere or so.

Electrostatic Voltmeter

This meter is essentially confined to the measurement of high voltages, although laboratory models are available which will register as low as 50 volts with fair accuracy.

The principle on which these meters function is the electrostatic

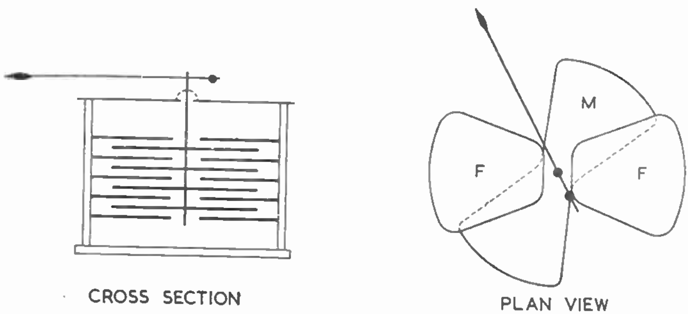
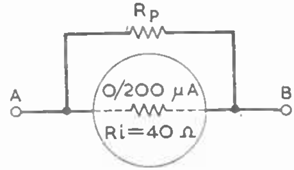


FIG. 7.—ELECTROSTATIC VOLTMETER.

Fig. 8.—Meter Shunting.



attraction existing between two oppositely charged metal plates in near proximity. To increase the sensitivity, it is customary to use two sets of interleaving plates carefully insulated from each other and from the frame. The assembly is somewhat similar in construction to a variable capacitor, as can be seen from the sectionalized and plan views shown in Fig. 7. One set of plates (F) is rigidly fixed, whilst the other (M) is lightly balanced and rides in jewelled bearings, this set carrying the pointer. Universal operation (A.C. or D.C.) ensues.

Theoretically, it has a linear scale, as the deflecting force provided by the electrostatic attraction between the plates is directly proportional to the actuating voltage, but a cramped effect is noticeable at low readings, due to the torque necessary for overcoming the initial friction of the bearings.

The chief merit of the system is the extremely low value of operating power, this being virtually zero on D.C. supplies and negligible on A.C. supplies. External fields cause no ill effects on the meter action.

Over-readings can be catered for by tapping the meter into a capacitance potentiometer. Excessive power consumption rules out resistance potentiometers.

Meter Conversion

Current-reading Conversion

The actuating coil of a sensitive basic movement, such as a moving-coil unit, is wound with fine-gauge enamelled wire to provide the necessary ampere-turns. As the temperature coefficient of expansion is moderately high in the case of copper, errors occur with varying ambient temperatures. These may be as high as 0·4 per cent per degree rise in room temperature. This is minimized in many designs by the incorporation of a Swamping Resistor of about four times the coil resistance and wired permanently in series with it. The swamping resistor is constructed from low-temperature-coefficient metal, such as eureka or manganin, so that the total percentage change in meter resistance with temperature variations is considerably reduced. This precaution is not necessary in voltmeters, due to the stabilizing action of the multipliers.

Current-recording instruments may be modified to handle higher values of current at full-scale deflection, although the reverse does not apply, i.e., they cannot be modified to record lower values at full-scale deflection. Suppose a 0–200 micro-ammeter with a meter resistance of 40 ohms has to be converted into a 0–10 milliammeter. Fig. 8 shows how this is effected. A " meter shunt " Rp has been connected in parallel with the original basic movement. The terminals A and B now become those of the new " composite meter ". The current passing from A to B at full-scale deflection will now be 10 mA. As only 200 μA

(0·2 mA) is permitted to pass through the basic movement, the excess current (9·8 mA) must be shunted via Rp. Hence the value of Rp must be

$$\frac{0{\cdot}2}{9{\cdot}8} \times 40 = 0{\cdot}816 \text{ ohms}$$

$$\text{i.e., Rp} = \frac{R_t}{N - 1}$$

where R_t is the resistance of the basic movement and N is the required multiplying factor. The calibration at full-scale deflection (200 μA) must now be changed to read 10 mA *et seq.*

Meter shunts consist of very short strips of manganin sheet soldered into massive terminal blocks. The manganin achieves a low-temperature coefficient. When used for heavy current work, shunts must be well ventilated, and are sometimes mounted in a separate case to the basic movement.

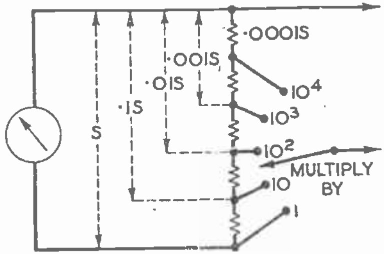


FIG. 9.—UNIVERSAL SHUNT.

A current-conversion switch sometimes operates on a UNIVERSAL SHUNT, as shown in Fig. 9, where the relevant resistor values and switch-position multiplying factors are self-explanatory.

Voltage-reading Conversion

Two operations are concerned in voltage-reading conversion : (*a*) the process by which a low-value current-reading instrument is transformed into a voltmeter, and (*b*) the system by which an existent voltmeter is modified to record a higher value at full-scale deflection.

Case (*a*) is dealt with by simple application of Ohm's Law. Consider a basic 0–500 microammeter with a meter resistance of 80 ohms, which has to be converted into a voltmeter recording 200 volts at full-scale deflection. The required modification is shown in Fig. 10. This merely consists of wiring a resistor " R " in series with the original meter to limit the full-scale deflection current to the correct value of 500 μA (= 0·5 mA) through the basic movement when the maximum voltage of 200 volts is connected to the terminals A and B. The total resistance of the new " composite meter " is thus given by E/I = 200/0·0005 = 400 kΩ, E being the new full-scale deflection voltage recorded and I the basic full-scale deflection current. From this the theoretical value of R is determined, it being 399,920 ohms for the example in Fig. 10. In practice, the tolerance of even a sub-standard resistor would cover the small error caused by ignoring the 80-ohm resistance of the basic movement. The resistor R is termed a MULTIPLIER, and academically its value is given by

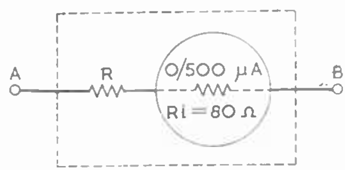
$$R = \left(\frac{E}{I} - R_i\right)$$

where R_i is the resistance of the basic movement.

Multipliers should be non-inductive and have a constant resistance. Their power ratings should be adequate, and if their dissipations are high, cooling arrangements should be made.

Case (*b*) is dealt with in a similar manner to case (*a*). An existent voltmeter can be modified to register higher values of voltage at full-scale deflection, provided that arrangements are made to limit the full-scale deflection current to its normal value. This means that the total resistance of the new meter must be increased in the same proportion as the voltage to be recorded at full-scale deflection, e.g., if the new voltmeter has to record seven times the old voltage at full-scale deflection, then the resistance of the meter must be increased to seven times

FIG. 10.—METER MULTIPLYING (CONVERSION TO VOLTMETER).



the old value by the addition of a series multiplier of six times the original value. Hence the value of additional multiplier R_m is given by

$$R_m = R_i(N - 1)$$

where R_i is the resistance of the original meter and N is the required multiplying factor.

Determination of the Resistance of a Voltmeter

The resistance of a voltmeter is determined as follows: Connect the voltmeter across a normal supply and take the reading. Call this V_1. Next, connect it in series with a known resistor R across the same supply. Call the second reading V_2. The resistance of the voltmeter is given by $R_i = \dfrac{V_2.R}{V_1 - V_2}$.

Ohmmeters

An ohmmeter is a resistance-measuring arrangement involving a current-reading meter of low full-scale deflection value, a battery, a limiting resistor and the component under test. The effect of the latter is to modify the value of current passing through the meter, thus compelling the pointer to take up an intermediate recording position. The scale is suitably calibrated in ohms, kilohms or megohms.

Two basic types are in use, the series type and the shunt type, there being two variants of the former. Basic circuitry of all three systems is shown in Fig. 11. In all cases the variable resistor R_2 adjusts the zero position for resistance-recording, this being full-scale deflection where the basic movement is concerned.

Figs. 11 (*a*) and 11 (*b*) both show series types, but the zero-adjustments are different. In Fig. 11 (*a*) R_1 is a safety limiting resistor to avoid a deflecting current considerably in excess of full-scale deflection value should R_2 be inadvertently left at a low value. R_2 is comparable in value to that of R_1 to allow for the normal fall in the cell voltage. The half-scale reading (in ohms) is $R_1 + R_2$.

In Fig. 11 (*b*) the resistance connected to the test prods is again connected in series with the meter, which is shunted by $R_2 + r$, R_1 acting as limiting resistor. To maintain full-scale deflection current through the meter when the test prods are shorted (resistance under

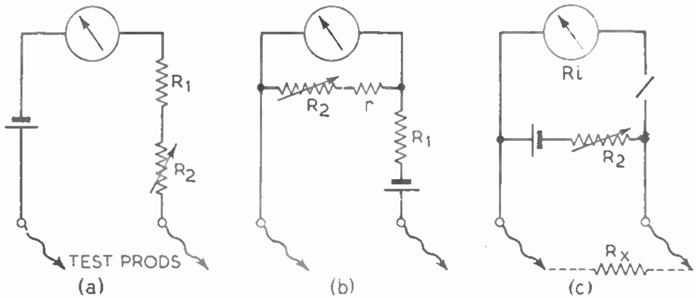


FIG. 11.—OHMMETER ARRANGEMENTS.

A and B are series types, C is a shunt type.

test = zero), R_2 must be increased in value as the cell voltage falls. This is necessary to shunt less current from the basic movement. Greater accuracy results from this arrangement than with the circuit in Fig. 11 (*a*).

The calibrated zero (for resistance values) in both Figs. 11 (*a*) and 11 (*b*) is at full-scale deflection of the pointer, i.e., to the right. Scale-cramping increases progressively to the left, i.e., at the higher resistance readings. Greater accuracy results at the higher readings if a battery of higher voltage is used. This is invariably connected externally, and suitable interpolation must be employed. The zero-adjuster R_2 must be manipulated with the test prods shorted.

Fig. 11 (*c*) shows the basic circuitry of the shunt type. R_2 is adjusted with the test prods at insulation (open-circuited), and the switch must be closed. The resistance under test R_x is connected in parallel with the basic movement by the test prods. If this is equal in value to the internal resistance of the movement R_i, half the battery current passes through the meter, so a half-scale reading results. A lower value for R_x results in more current being shunted from the movement, hence the lower resistance-values are recorded to the left and higher values near full-scale deflection. This type of ohmmeter is used for the recording of low resistance values, and the greatest accuracy again occurs at the lower readings. The lower limit can be extended by shunting the movement with a resistor comparable in value to its own resistance and interpolating the readings.

Universal Test-meters

A universal test-meter is essential in even the smallest radio workshop. It is a switchable A.C./D.C. combined voltmeter, ammeter and ohmmeter built around a basic sensitive moving-coil movement, the full-scale deflection of which may vary from 50 μA to 4 mA. Knife-edged pointers and anti-parallax mirrors are common.

Multiple-position switches introduce the necessary shunts or multipliers, which extend the current range from 50 μA up to several amperes, and the voltage range of 0·2 or 0·3 volts up to 2,000 volts or so. The meter resistance in the case of the latter should be at least 1,000 ohms/volt, some of the higher-grade instruments having sensitivities of 20,000 ohms/volt on the D.C. ranges and 2,000 ohms/volt on A.C. ranges.

Alternating current and alternating voltage are catered for by switching iron-cored transformers, metal rectifiers and rectifier shunts into circuit at the "A.C." switch-positions. On the "A.C. Volts" ranges, switched shunts are essential. In the above cases the calibrations are interpolated in terms of R.M.S. values, and the tolerance accuracy is guaranteed only if the metered waveform is sinusoidal. Due to the non-linearity of copper oxide rectifiers in the conductive direction, the tolerance accuracy is less than on D.C. ranges.

Ohmmeter services are catered for at other switch-positions when dry cells, limiting resistors and zero-adjusting resistors are brought into circuit. Shunt-connection of the last-named is usual, and according to the sensitivity of the basic movement, reliable resistance-measurement up to several megohms is possible. A self-contained buzzer is supplied in some models for quick continuity testing.

A mechanical spring-operated cut-out is preferred generally to a fuse. This is operated by the pointer when it swings above full-scale deflection, thus tripping a lightly balanced relay. A small overload is sufficient to actuate this safety device.

Wattmeters

The wattmeters encountered in radio installations are principally of two types—dynamometer and induction. The torque exercised on the indicating pointer must be proportional to the phase angle existent between the voltage across the metered circuit and the current through it.

The dynamometer arrangement is shown in Fig. 12 (*a*). A pair of low-resistance, astatically (oppositely) wound current coils C are wired in series with the metered circuit. They are fixed to the pointer and rotate inside a pair of high-resistance, astatically wound pressure coils P, which are wired in series with an anti-phasing resistor R across the metered circuit. The fluxes due to the currents in P and C are thus in phase with the voltage and current respectively, so the torque on the pointer is proportional to $EI \cos \theta$. E.m.f.s induced into the astatic

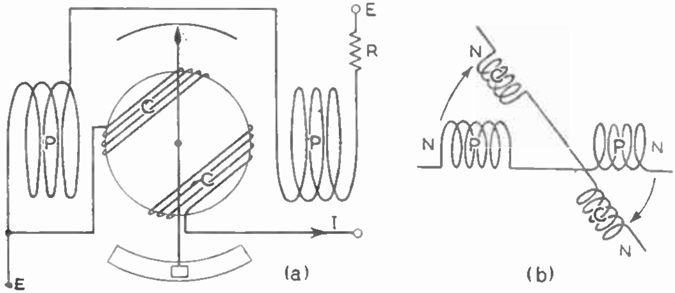


FIG. 12.—DYNAMOMETER TYPE WATTMETER.

pairs by external fields annul due to the winding-reversals, but the torque due to the concerned currents is unaffected, as shown in Fig. 12 (*b*). This type is accurate but usually somewhat fragile, and so is chiefly confined to laboratory use.

The induction type functions on the principle that the rotating fields produced by the current coils C and the pressure coils P shown in Fig. 13 induce eddy currents into a disc or drum, thus making it rotate. The P coils have many turns wound on a laminated iron magnetic circuit, so the current through them lags by 90°. The C coils have a relatively small number of turns of heavier gauge wire wound on an open iron circuit, so E and I are virtually in phase with the fluxes. The drum D is of special alloy, and rotates in the field produced by the quadrature fluxes, so the resultant torque is proportional to EI cos θ.

This type is not usually so accurate as the dynamometer, but is cheaper and more robust. Its scale is non-linear, and variations in temperature or frequency may produce errors.

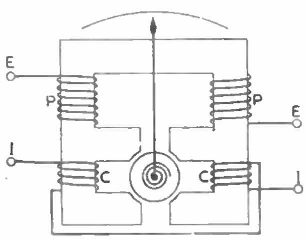


FIG. 13 (*above*).—INDUCTION-TYPE WATTMETER.

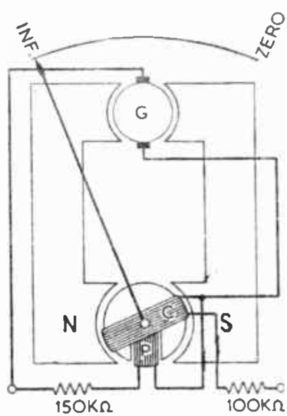


FIG. 14 (*right*).—"MEGGER" INSULATION TESTER.

Megohmmeter

"Megger" is the trade name adopted by Messrs. Evershed and Vignoles for their insulation testers designed for the measurement of very high resistances such as the leakages of insulators. These involve values of the order of hundreds of megohms.

A built-in hand-driven generator (or two series-connected generators), fitted with a folding handle and a slipping clutch, is provided to develop a constant voltage of 500–2,000 volts according to the model, this supplying the actuating currents to the meter. In the small, WEE MEGGER Testers the generator voltage may be as low as 100 volts.

The indicating system is shown in Fig. 14, and operates on the "bicoil" principle, in which two coils are fixed in relation to each other, but the composite unit so comprised acts as a lightly balanced moving-coil system rotatable in the field of a permanent magnet as shown. The pressure coil P is connected across the D.C. generator, and the flux associated with its current acts on the permanent field to produce a torque which tends to rotate the pointer towards "infinity". The current coil C is connected in series with the external circuit, and the flux developed in it acts on the permanent field to produce an opposite torque, i.e., tending to rotate the pointer towards zero. According to the value of resistance connected to the terminals, the pointer assumes an intermediate reading. The resistors shown in Fig. 14 are safety limiters.

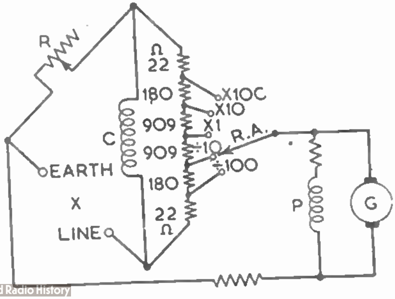
Full-scale calibrations range from 10 megohms up to 2,000 megohms according to the model, and from a practical viewpoint, the accuracy of the readings is sufficient for normal requirements.

Low-voltage testers registering resistance values of 2 ohms, 10 ohms or 100 ohms at full-scale deflection are also available. Their ohm-meter actions incorporate shunt connection of the resistors under test, and the instruments are usually referred to as CONTINUITY TESTERS.

Bridge-megohmmeter

This instrument consists of a high-voltage megger tester and a sub-standard DECADE resistance box, the value of which is indicated numerically through four "windows", the maximum being 9,999 ohms. Fig. 15 shows the associated circuitry. R is the decade box and R.A. is a built-in "ratio arm" with five positions calibrated ÷ 100, ÷ 10, × 1,

FIG. 15.—BRIDGE-"MEGGER" TESTER.



$\times 10$ and $\times 100$. Resistance-values of 0·01 ohm (1 ohm $\div$ 100) to 999,900 ohms (9,999 $\times$ 100) are thus measurable with very good accuracy. The resistor under test " x " is connected to the " Line " and " Earth " terminals, and the Wheatstone Bridge is balanced when the pointer remains at " infinity " on turning the hand-generator. In this case no current flows through the current coil C in Fig. 15, " x " then having a value of R times the multiplying factor indicated by R.A. At the " dividing positions "of this switch the decade box and " x " exchange legs in the Wheatstone Bridge.

The words " increase " and " decrease " are printed opposite the " infinity " calibration. These direct the operator how to manipulate the decade box for bringing the pointer to " infinity " when the hand-generator is turned.

A three-position switch provides services as follows : " Meg " for normal high-resistance megger operation, " Bridge " for accurate bridge-megger measurement and " Varley ", this involving the circuitry for locating an earth on a line by the VARLEY LOOP TEST. This is effected as follows :

(1) Connect the " Varley Earth " terminal to the " Earth " terminal.

(2) Short-circuit the remote ends of the line under test and measure the resistance of the loop so formed by bridge operation. Call this R_l.

(3) Connect the free (near) ends of the line to the " Line " and " Earth " terminals.

(4) Switch to " Varley " and balance again. Call the recorded value R.

The distance of the earth down the line measured in ohms is now given by

$$\frac{R_l - R}{2} \text{ if the ratio arm is at } \times 1;$$

or

$$\frac{10R_l - R}{11} \text{ if the ratio arm is at } \div 10;$$

or

$$\frac{100R_l - R}{101} \text{ if the ratio arm is at } \div 100.$$

TEST EQUIPMENTS

The test equipments which follow include such meters and instruments as are found in a fully-equipped radio and communications workshop. The descriptions apply, as far as possible, to instruments in a particular class, rather than to individual models.

Power-supply circuitry has been substantially ignored, as this is not pertinent to the action of the equipment concerned. Actually, such power supplies are not wholly conventional to normal receiver or transmitter techniques. Both full-wave and half-wave systems of rectification are used. Filtering arrangements of the mains ripple may be conventional choke/capacitor, resistance/capacitor or completely omitted, i.e., rectified raw A.C. Errors are inevitable in some equipments if the supply waveform is not pure sinusoidal. Voltage-regulated transformers are used in some equipments to provide a constant output from any supply mains between 200 and 250 volts. Radio-frequency filtering is incorporated in the mains-supply leads of most equipments.

Overlapping of services takes place in some equipments, e.g., a universal test-meter may be convertible to an output power meter or to a valve voltmeter with considerable limitations. Such instruments find an appeal in small workshops, where expense is a major item.

Valve Voltmeter

An "A.C. voltmeter" of the conventional type is useless for measuring radio-frequency voltages, as these are usually developed across a parallel-tuned circuit, the dynamic resistance of which invariably exceeds that of the voltmeter. The resultant damping imposed on the tuned circuit and the capacitance effects of the voltmeter thus render such an instrument totally unsuitable for radio-frequency voltage measurements.

The valve voltmeter (V.V.) presents an impedance of the order of 2 or 3 MΩ to the tuned circuit under test, this probably having a dynamic resistance of 50–150 kΩ, so the shunting effect of the valve voltmeter introduces minor errors. Furthermore, the parallel capacitance it presents is of the order of 4–9 pF, this merely involving slight retuning of the test circuit when the valve voltmeter is connected across it.

In addition to being a test equipment in itself, a valve voltmeter is also used as an integral part of other test equipments, such as a Q-meter, an Impedance Bridge, etc.

Fig. 16 shows a generalized circuit of a valve voltmeter. The radio-frequency voltage to be measured is applied to the "H" and "L" terminals, these notations referring to the "high" potential and "low" potential ends of the test circuit. The diode D rectifies the input voltage, developing an anode bias which is almost the peak value of the input, due to the long time-constant of the 500 pF/50 MΩ load combination. This is applied to the grid of a D.C. amplifying valve V2 wired in one arm of a Wheatstone Bridge arrangement. Radio-frequency ripple is filtered by the 10 MΩ/0·02 μF/0·02 μF/100 pF network. The bridge is "no-signally" balanced by adjustable positive bias on the grid of V2 tapped by the " Set Zero " control. The negative potential from the load combination due to the applied signal decreases the anode current of V2, thus increasing its *Ra*-value. This throws the bridge out of balance, and the 0–200 microammeter records the out-of-balance current. The scale of this meter is accordingly calibrated severally in R.M.S. input voltages according to the ranges covered (five in the case shown). Typical full-scale deflection coverages are 1·5, 5, 15, 50 and 150 volts. In practice, the range of frequencies over which such a valve voltmeter will give good accuracy is about 20 c/s to 120 Mc/s.

At frequencies above some 10 kc/s or so it is not politic to use long input leads to the diode, due to their excessive self-capacitance, hence D and its associated radio-frequency components are built into a small **Probe** unit on the end of a flexible multi-core cable in order that it may be brought right up to the measured job, connecting leads thus being kept to a sheer minimum. The flexible cable carries the diode

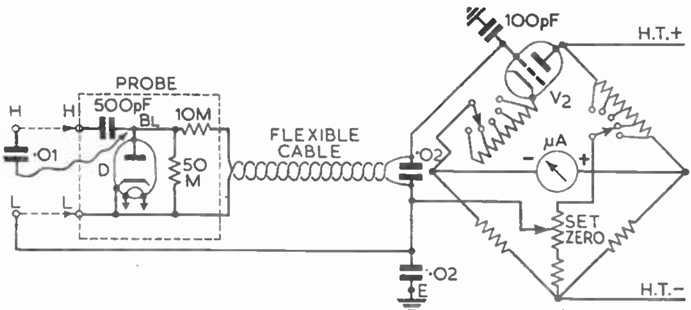


Fig. 16.—Typical Valve Voltmeter and Probe Unit.

heater leads and the output rectified voltage. D is a miniature valve of the D1 or EA50 types.

Below frequencies of 10 kc/s, the foregoing precautions are unnecessary, so the probe unit is secured inside the instrument proper with its " H " and " L " prongs making contact to the inner terminations of the H and L terminals on the front panel. The flexible lead from the 0·01-μF capacitor must be connected to the " B*l* " terminal. This parallels the 0·01-μF and the 500-pF capacitors to decrease the input reactance at the lower frequencies.

D.C. supplies can be measured by connecting the negative supply to the " B*l* " terminal and the positive to the L terminal.

The five-position switch shown in Fig. 16 varies the cathode bias of V2. this necessitating higher inputs at higher bias values to effect comparable out-of-balance currents. At the 50- and 150-volt positions, meter multipliers are switched to the microammeter, as the bias involved would be excessive on these ranges.

Operations should be commenced at the highest voltage range and switched downwards if the reading is too low for accurate measurement.

Balanced Valve Voltmeter

This type of valve-voltmeter is used as a monitor when aligning the discriminator stage of a frequency-modulated receiver. Fig. 17 shows typical circuitry.

The discriminator output is applied to the H and L terminals, and unmodulated input at the intermediate frequency is injected into the intermediate-frequency chain. If discriminator output exists, the positive-going terminal produces an increase in the anode current of its associated valve via the G1 connection, whilst the negative-going terminal produces a corresponding decrease through the other valve. The cathodes now assume different potentials, so the centre-reading microammeter records the out-of-balance current in the concerned direction, and is thus suitably calibrated in terms of ± input volts. Discriminator adjustment must be made for zero reading on this meter. If its tuning is now offset in either direction, opposite deflections take place.

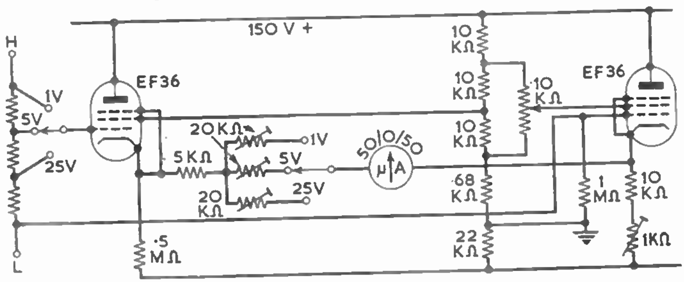


FIG. 17.—CIRCUIT OF TYPICAL BALANCED VALVE VOLTMETER.

The 20-kΩ pre-set resistors are adjusted on the concerned ranges with the input terminals shorted, the 1 kΩ cathode resistor acting as an auxiliary control on the highest range (25 volts) after the lowest range (1 volt) pre-set 20 kΩ has been adjusted. The screen potentiometer to V2 acts as the general "Set Zero" control by balancing the cathode potentials.

Method of Use.

(1) Inject unmodulated input at the intermediate frequency into the intermediate-frequency chain.

(2) Switch to the 25-volt range and roughly align for zero.

(3) Switch to the 5-volt range and re-align.

(4) Switch to the 1-volt range and make the final adjustment with great care.

Beat-frequency Oscillator (B.F.O.)

A Beat-frequency Oscillator is an instrument designed to generate an audio-frequency signal of specified frequency and amplitude. This should be variable from about 20 c/s to 10 kc/s or 12 kc/s with reasonable accuracy, and should be free from amplitude distortion. The output is used to test the audio-frequency sections of receivers and transmitters, i.e., the headphones, loudspeaker, phone jack, output transformer, modulation transformer, speech amplifier and audio-frequency equipment of transmitters generally, audio-frequency amplifier and associated components of receivers as far back as the detector diode anode. It may be used to plot audio-frequency response curves, to determine harmonic content in output stages and to trace phase-shift through an audio-frequency amplifier. It may also be used to determine the resonant frequency of an audio-frequency filter or to centre the speech coil of a moving-coil loudspeaker.

Suppose a frequency coverage of 50 c/s to 10 kc/s is required. A straight L/C oscillator would need a capacitance ratio of 200^2 to 1, this being impracticable. Hence the technique shown in Fig. 18 is adopted. Two high-stability oscillators are used. a fixed one operating at 100 kc/s and a variable-frequency oscillator (V.F.O.) with a frequency coverage of either 90–100 kc/s or 100–110 kc/s. Coupling between these oscil-

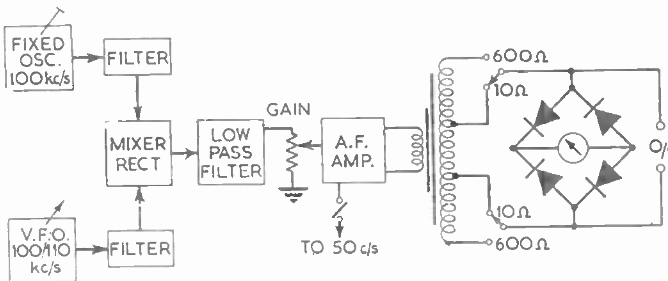


Fig. 18.—Block Schematic Layout of a B.F.O.

lators is kept low by adequate decoupling and screening. The outputs from both oscillators are filtered to remove harmonics and then applied to an electronic mixer. The resultant difference-frequency becomes the desired audio frequency (0–10 kc/s), the original oscillator frequencies and their sum frequency being filtered off by a simple π-section low-pass filter.

The output amplitude is controlled by a conventional gain control. Switched tappings on the output transformer effect approximate matching to high (600 ohms, 2,500 ohms) or low (2–10 ohms) impedance loads. A high output voltage (50 volts) is developed across the former or a low value (5 volts) across the latter. The output may be matched to any other desired impedance by an appropriate matching attenuation pad.

The output indicator may be a moving-coil milliammeter as shown in Fig. 18 with a full-scale deflection current of 1, 2 or 5 mA, or it may be a " magic eye " tuning indicator operated from the output secondary.

For " Set Zero " operation, the " 50-c/s check " switch is made, this applying a small signal from the mains to the grid of the amplifier. The " Set Zero " control must now be operated to produce the slowest heterodyne flicks on the output indicator (1 or 2 per second) with the frequency control set at 50 c/s. The "Set Zero" control may be a mechanical adjuster of the fixed oscillator inductor or a small pre-set capacitor across it.

The frequency control may be a small capacitor across the variable-frequency oscillator or an adjustable brass damping plate operating on the variable-frequency oscillator inductor. Sometimes a vernier frequency control is fitted, this covering a restricted frequency range. In this case the output frequency is the sum of the two frequency calibrations.

The oscillators may be straight positive feedback resistance-stabilized types, Hartleys, Colpitts or cross-connected push–pull types. The last-named use the triodes of triode-hexodes with the mixer sections commoned.

Efficient smoothing of the mains ripple (100 c/s) is particularly necessary, separate filter chokes and capacitors sometimes being used for the output stage. Ample audio-frequency output is catered for (about 2 watts) to minimize distortion, push–pull operation being used in some models, negative feedback in others, with triodes general. These precautions result in a tolerably good sinusoidal waveform when viewed on a cathode-ray oscilloscope.

For accurate work, such as stage-gain measurements or overall audio-frequency sensitivity checks, the instrument can be used in conjunction with an attenuator to provide small outputs at the required impedances.

Audio Oscillator

The function of this instrument is virtually the same as that of a beat-frequency oscillator. The usual circuitry is conventional resistance–capacitance technique using two triodes and negative feedback and an output pentode. A frequency coverage of 20 c/s to 20 kc/s is obtained in three bands by variable resistors. An almost pure sinusoidal waveform with negligible amplitude variation is produced in common with normal resistance–capacitance oscillator technique. Output power may be 1 watt or so, matched to two or three load impedances.

Signal Generators

Modulated Test Oscillator

A modulated test oscillator is virtually a low-grade signal generator intended for rough bench operation, in which rapid fault diagnosis of a receiver under test may be carried out. It is thus essentially a "time-saver" to the repair man. In the absence of a standard signal generator, if the frequency inaccuracies are known, the instrument is sometimes used for rough alignment.

Compared with a standard signal generator, a lower degree of accuracy, both in calibrated frequency and output voltage, is usual. Normal frequency coverage is from about 100 kc/s or so up to 50 or 60 Mc/s.

Conventional circuitry is employed. Fig. 19 being a typical block schematic arrangement. A Hartley audio-frequency oscillator functions at 400 c/s and acts as internal modulator to a radio-frequency oscillator. Its maximum output may be 10–15 volts, this being variable in some models and taken to the output audio-frequency terminals. This is useful in small workshops to obviate the necessity for a beat-frequency oscillator when it is desirable to carry out checks to the audio-frequency section.

The internal audio-frequency oscillator is switched into circuit when the master switch shown in Fig. 19 is at the "INT" (internal) position. The depth of modulation so effected is about 30 per cent.

The radio-frequency oscillator may also be modulated by an external source of audio frequency, such as a beat-frequency oscillator, if the master switch is set to the "EXT" (external) position. In this case there is no guarantee of the modulation depth.

The radio-frequency oscillator may be a conventional Colpitts circuit tuned by a turret-operated range switch. At the "R.F." or "C.W." position of the master switch it is disconnected from all sources of audio frequency; hence the output is unmodulated C.W. fed via a coaxial cable to a simple dummy aerial. The R.M.S. value of radio frequency delivered to the output attenuator is about 1 volt, this varying somewhat with frequency.

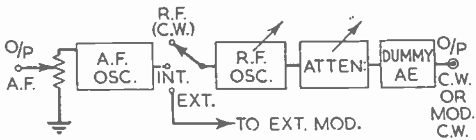


FIG. 19.—BLOCK SCHEMATIC LAYOUT OF A TEST OSCILLATOR.

The attenuator delivers the full radio-frequency output from the oscillator to the dummy aerial for "force" fault-finding purposes or lower variable values of 50 mV downwards. Correct feed to an unbalanced input circuit is provided by the coaxial line, and the dummy aerial imposes normal aerial loading on the receiver input circuit.

Except when operated under "force" conditions, the radio-frequency output should be restricted to the minimum required for the tests concerned. This avoids operation of the automatic-gain-control system with the concomitant misleading results. The serious demerits of a test oscillator are: (1) the poor screening, this resulting in excessive stray fields which make low-level work impossible; and (2) the "makeshift" attenuator.

Standard Signal Generator (S.S.G.)

A "signal generator" may be classified with a modulated test oscillator, whereas a "standard signal generator" is a precision signal generator designed to develop highly accurate voltages by means of a precision attenuator which is conveniently calibrated.

Its functions are to align receivers accurately, to calibrate receivers, transmitters or oscillators (in conjunction with a standard frequency-meter), to determine stage-gain, sensitivity, bandwidth, signal-to-noise ratio, selectivities, overall response curves (in conjunction with

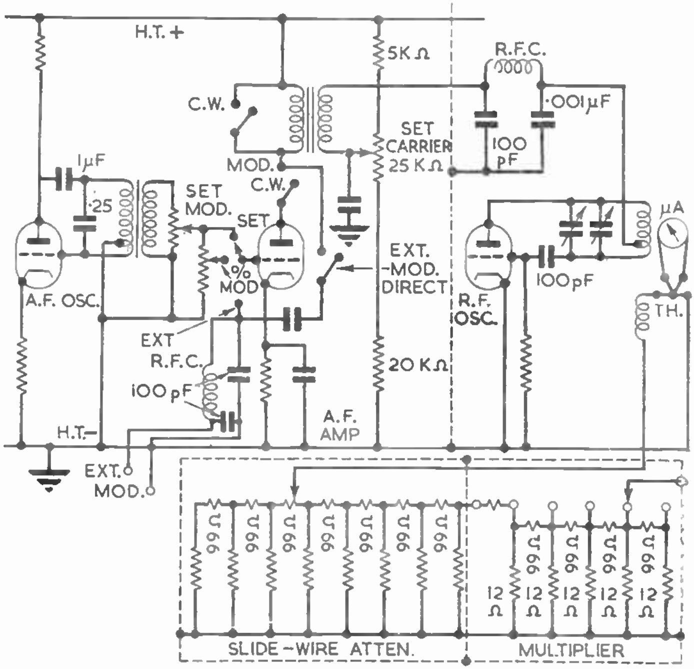


FIG. 20.—TYPICAL CIRCUIT OF A SIGNAL GENERATOR.

a beat-frequency oscillator) and to check automatic-gain-control efficiency.

The circuit of a popular design is shown in Fig. 20, this being basically similar to that of a modulated test oscillator with certain refinements, these being :

(1) " Set Carrier " control. This enables the amplitude of the derived unmodulated radio-frequency oscillations to be accurately adjusted such that when the radio-frequency output monitor records a pre-determined value, the radio-frequency output delivered to the attenuator is exactly 1 volt R.M.S.

(2) " Set Modulation " control and " Percentage Modulation " control. These enable the depth of modulation to be accurately adjusted to any desired value up to 75 per cent or so by use of a second calibration on the output meter.

(3) The radio-frequency output monitor ensures that a constant output is delivered to the attenuator. It may be a valve voltmeter, measuring the radio-frequency voltage developed across the attenu-

ator or a thermo-couple meter recording the current passing through it.

(4) A higher degree of radio-frequency tuning results from a larger tuning dial and the provision of vernier band-spreading ("Fine Frequency Control" or "Incremental Tuning"). The radio-frequency oscillator may be calibrated direct in kc/s and Mc/s or used in conjunction with calibrated charts provided with the instrument.

(5) A high degree of accuracy in the calibration of the radio-frequency attenuator, this being denoted in μV, mV, db and multiplying ratios. This is effected by a resistance ladder-network, the elements of which are sometimes screened from each other to avoid capacitance effects which would vary with frequency.

(6) Switching arrangements whereby external audio frequency may modulate the radio-frequency oscillator direct, or may be amplified by an audio-frequency stage prior to application to the radio-frequency oscillator.

(7) Each section of the instrument is separately screened and housed in a heavy, rigid screening case. This minimizes radiation of stray fields which, if picked up by the receiver under test, would despoil the measurements.

As the instrument has a fixed internal impedance, an external matching pad is necessary to match a particular input load.

V.H.F. Signal Generator

The primary function of a V.H.F. signal generator is to generate signal voltages across a known impedance at frequencies between some 100 and 300 Mc/s. These may be amplitude modulated sinusoidally or by square-wave signals for radar applications. The radio-frequency tuning is calibrated direct for this frequency band, and also carries a blind 0–100° scale for the interpolation of calibration charts when secondary bands over the 10–60-Mc/s range may be involved.

Fig. 21 shows the basic circuitry of a typical instrument which uses a conventional audio-frequency oscillator for anode modulation of the radio-frequency oscillator. The latter employs a short-path V.H.F. triode (RL18) in a Colpitts circuit formed by the inter-electrode capacitances. The dual top contacts to this valve permit very short leads to the tuned circuit, which comprises a "series-gap" type of tuning capacitor (0–270° sweep) and an inductor, which is merely a thick aluminium bar plugged directly across the capacitor. Other plug-in coils wound on low-loss material cater for the lower-frequency bands. All radio-frequency components are heavily silver-plated.

The output "piston" attenuator is magnetically coupled to the tuned circuit, and is calibrated from +10 db to −100 db. If very small outputs are required, a further 26 db attenuation may be achieved by use of an external T-pad provided with the instrument.

A piston attenuator consists of a brass piston mechanically moved up and down inside a brass cylinder which acts as a waveguide and so has application at V.H.F. and U.H.F. only. Its outstanding points are that it attenuates along its length exponentially, i.e., its attenuation is "linear in db", and that the resultant attenuation is independent of frequency up to tube lengths of $\lambda/10$.

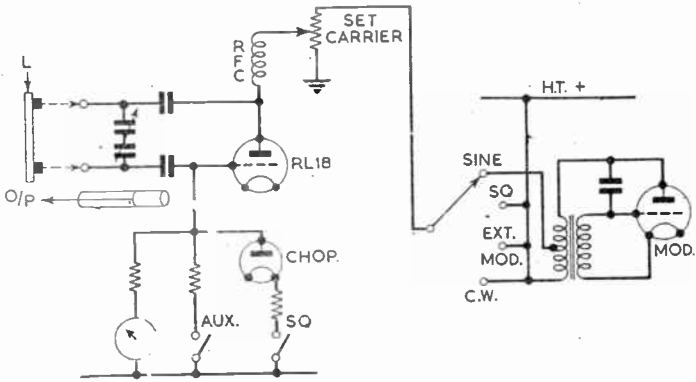


FIG. 21.—V.H.F. SIGNAL GENERATOR.

The carrier amplitude is adjusted by the " Set Carrier " H.T. potentiometer, and is monitored by a 0–100-microammeter which registers the value of oscillator grid current. This has a " normal " calibration mark for initial setting-up purposes. On the lower-frequency bands the radio-frequency oscillator develops a greater output, so an extra shunt is switched across the microammeter by switching to the " Auxiliary " position.

At the " square-wave " switch-position, a saturating chopper diode (EA50) is switched into the oscillator grid circuit to limit its output for 50/50 square-wave operation.

Frequency-modulated/Amplitude-modulated Signal Generator

This type of signal generator is more comprehensive than usual in that it provides an output which may be (*a*) unmodulated C.W., (*b*) sinusoidally-modulated frequency modulation or (*c*) sinusoidally or square-wave modulated amplitude modulation from either an external or internal source.

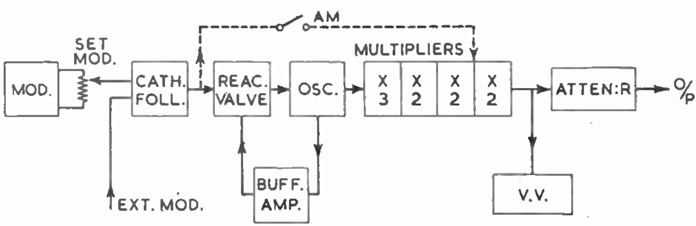


FIG. 22.—BLOCK SCHEMATIC LAYOUT OF AN F.M./A.M. SIGNAL GENERATOR.

Frequency coverage may be between 10 and 200 Mc/s at output voltages from 1 μV to 100 mV into 75-ohm loads. The internal modulating frequencies may vary between 300 c/s and 3 kc/s. The resultant frequency deviation may range up to 300 kc/s on the lower bands or up to 600 kc/s on the higher bands. Fundamental oscillator frequencies between 5 and 10 Mc/s are usual.

A limiting chopper diode converts sine-wave input to square-wave modulation. When operative on the amplitude-modulation system, modulation depths of up to 80 per cent are possible. Fig. 22 shows a general schematic arrangement.

Variation in deviation is attained by control of the audio-frequency input to the reactance stage, which is usually operated in push–pull. The output frequencies are monitored by a small loudspeaker when an internal oscillator heterodynes the output of the crystal oscillator. This provides twelve to sixteen spot frequencies per band for check purposes, intermediate values being determined by interpolation.

The built-in balanced valve voltmeter records the deviation in terms of frequencies, the modulation depth also being recorded.

Test Bridges

A test bridge is a " bench " class of instrument capable of carrying out the following functions with a reasonable degree of accuracy, i.e., satisfactory for normal purposes but not usually intended for laboratory use :

(1) Measurement of unknown resistance-values between 5 ohms and 50 MΩ by comparison with internal sub-standards.

(2) Measurement of unknown capacitances between 5 pF and 50 μF by comparison with internal sub-standards.

(3) Measurement of the power factors of capacitors with values of 0·05 μF upwards.

(4) Measurement of any resistor, capacitor or inductor over 0·1 henry by comparison with a suitable external component of similar phase-angle (switch position " E ").

(5) Determination of the approximate leakage-resistance of a capacitor (switch position L).

(6) Measurement of alternating voltages from 0·1 to 15 volts R.M.S. at frequencies up to 20 kc/s by the use of a built-in valve voltmeter (switch position " V "), the 50-volt A.C. supply then being disconnected.

A Wheatstone or De Sauty bridge arrangement is used as shown in Fig. 23. Its energizing voltage is usually 50 volts at 50 c/s, the output being fed via a triode amplifier to a metal rectifier or via a pentode amplifier to a diode rectifier. The rectified load current is recorded by a sensitive microammeter. The potentiometer P is adjusted for null output from the bridge, i.e., for minimum reading on the diode monitor. It is calibrated in terms of " multiplying factor ' (from 0·05 to 50), this being the factor by which the concerned internal or external substandard component value must be multiplied to provide the value of component connected to the " x " terminals. As the reactance of a capacitor is inversely proportional to the value of C, the multiplying factor for capacitors is the reverse of that for R. On similar reasoning, the relevant multiplying scale for inductors is the R calibrations.

At the C3 position the power-factor control R1 may be manipulated for a true null reading on the monitor after the bridge is balanced. This provides the same phase-angle as that across the capacitor under test, and so is calibrated in terms of power factor.

External sub-standard components must be connected to the " EXT " terminals. Matching of two components may be effected by connecting them to the " x " and " EXT " terminals.

An electrolytic capacitor may be measured if a 120-volt dry battery and a 50-kΩ resistor are connected in series across it for polarization purposes. The battery voltage should exceed the peak A.C. input to the bridge by 25 volts, this allowing for a polarizing current of 0·5 mA.

When the nine-way switch is at the " L " position (L = " leakage " NOT " inductance "), the subsidiary circuit shown dotted in Fig. 23 becomes operative. The capacitor under test, a neon lamp and a 50-kΩ resistor are then connected across the H.T. supply. The neon

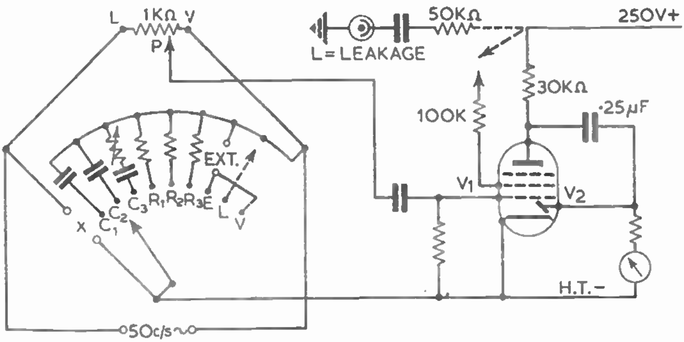


FIG. 23.—SIMPLIFIED CIRCUIT OF A TYPICAL TEST BRIDGE.

strikes as the voltage across it builds up and reaches the striking value. If the capacitor has a relatively low leakage resistance, self-discharging takes place at a frequency decided by the self-time-constant. The frequency of striking thus gives an indication of the leakage resistance.

The low bridge-energizing frequency (50 c/s) precludes accurate measurement of inductors below about 0·1 henry in value.

A break in a line can be located by measuring the capacitance between it and another in the same cable and comparing this to the capacitance between the other broken section and the same sound line. The capacitances concerned will be proportional to the line lengths.

Universal Impedance Bridge

This instrument is virtually a " precision test bridge " and measures resistance-values of 0·05 ohm to 1 MΩ, capacitance-values of 5 pF to 100 µF and inductance-values of 5 µH to 100 henrys. The circuits are mains-operated bridges using Wheatstone, De Sauty or Maxwell arrangements respectively over five ranges in each case. Other services available are measurements of coil-magnification between values of 0·1 and 100 at a frequency of 1 kc/s or so, and capacitor losses (ωCR) between values of 0·0001 and unity.

The Wheatstone bridge is energized by a D.C. supply, the degree of unbalance being monitored by a centre-zero galvanometer.

The A.C. bridges are energized by a conventional audio-frequency oscillator functioning at 800 or 1,000 c/s. In these cases the unbalance voltages are monitored by a built-in valve voltmeter after suitable amplification, or an aural null indication may be found with telephones. A " separator valve " is used to isolate the test bridge from the audio-frequency oscillator.

The bridges are adjusted by " coarse " and " fine " controls in all cases, the two calibrated readings being summated to supply the true measurements.

The " Phase Balance " control provides the Q-value of the inductor concerned at the frequency of the audio-frequency bridge input.

Electrolytic Capacitor Bridge

The function of this instrument is to carry out more extensive and accurate measurements on electrolytic capacitors than that provided by

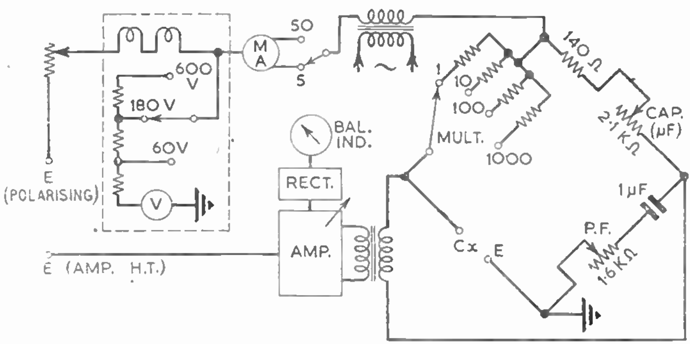


FIG. 24.—TYPICAL ELECTROLYTIC CAPACITOR BRIDGE.

an impedance bridge, and to measure their power factors and leakage currents. Fig. 24 shows the circuit arrangements of a popular model which caters for capacitor values of 0·2–2,200 μF.

A conventional bridge circuit energized from the mains is employed, its output being amplified, rectified and monitored by a " Balance Indicator ".

Two independent power rectifiers are used, one to supply D.C. to the bridge amplifier, the other providing the D.C. polarizing voltage to the capacitor under test, this being continuously variable over three ranges. It should be applied progressively up to 90 per cent of the working voltage and returned to zero slowly on completion of test This ensures that the capacitor is fully discharged.

A double-range milliammeter (F.S.D. = 5 or 50 mA) in series with the polarizing D.C. and the capacitor under test records the leakage current. In the event of a capacitor-short, the milliammeter is protected by the lamps. The leakage current of "dry" electrolytics should not exceed $0{\cdot}15 \times C \times V$ μA, where C is the capacitance in microfarads and V is the working voltage.

The capacitor under test is connected to the "Cx" terminals with its negative lead to earth (E). The capacitance control (Cap) must be adjusted for minimum reading on the Balance Indicator. The former is calibrated in terms of microfarads (0·2–2·2) and, when read in conjunction with the setting of the multiplier, provides the required capacitance. To measure the percentage power factor, the "P.F." control must be manipulated until the residual reading on the Balance Indicator reduces to absolute zero. "P.F." now shows the value of power factor.

It is noteworthy that electrolytic capacitors usually record higher values than the rated capacitances.

Valve Tester

A valve tester is an essential equipment in any radio/television workshop. The measurements which can usually be effected with it are the mutual conductance (slope) of three-electrode and multi-grid valves, the emission of all types of diode, the heater/cathode insulation under working conditions (hot) and an approximation of the degree of evacuation (gas-test). Short-circuits between electrodes may also be detected.

The D.C. supplies to the valve under test are usually rectified but unsmoothed A.C. On some models a "Set ~" control provides slight variation in the input mains voltage. The L.T. supplies are variable from 1·1 to 117 volts.

About twenty valve-holders are required to cater for all types of valves not yet completely obsolete, and vacant positions are generally provided for fitment of new types when introduced. The valve-contacts are usually connected to their relevant D.C. supplies by a multi-way, independently-operated roller switch. Discreet use of stopper resistors prevents parasitic oscillations when the valve under test is a modern "high-slope" type. A cut-out safeguards the valve from overloads in some models.

Special switching arrangements involve the following considerations:

(1) A variable *Ea* supply provides a very low value when signal diodes are under test.

(2) "Other anodes", such as mixer anodes, D.D.T. anodes, D.D.D. anodes, etc., are switched-selected in turn.

(3) Thyratrons, cold cathode rectifiers (C.C.R.) and neon indicators need an external load resistor connected to the terminals provided. The *Ea* switch in this case must be set to a value just above the striking voltage of the valve.

(4) The "slope-test" switch increases the value of *Eg* by 1 volt for direct reading of the *Gm*-value after zero-setting has been completed. Full-scale deflection may be switched from 15 mA/volt to 3 mA/volt to provide a more accurate reading with low-slope valves.

(5) The screening grid switch inserts a load between H.T. positive and G2 of output A.F. multi-grid valves to prevent "trioding".

(6) The "cathode insulation" switch converts the valve under test into a rectifier. A capacitor and the cathode/heater leakage resistance now become the load combination. The meter records the load current, which is inversely proportional to the leakage resistance, and the scale is suitably calibrated in megohms for this purpose.

(7) The "gas-test" switch short-circuits a 100-kΩ resistor in the grid circuit of the tested valve when *Eg*1 is about −1 volt. If reversed grid current caused by positive ions striking the grid exists, the removal of the volts-drop by the shorting action produces a slight fall in anode current. The reversed *Ig* may be as high as 10 μA.

A booklet is generally provided by the manufacturers with each meter, listing the roller-switch positions and other relevant data for

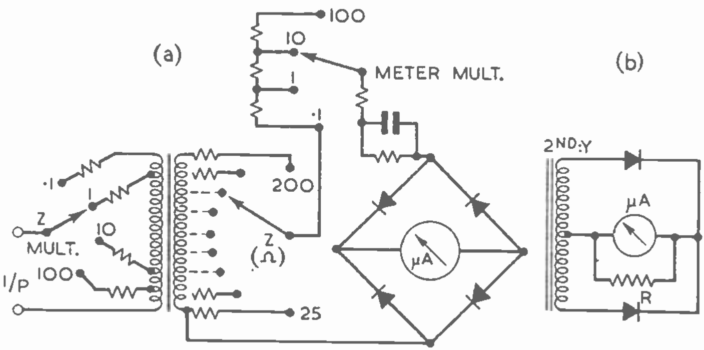


FIG. 25.—OUTPUT METER ARRANGEMENTS.

some 3,000 valves in use. In the case of the Mullard High-speed Tester correct potentials are applied automatically by the insertion of a punched card.

Output-power Meter

The function of this instrument is to measure the output audio-frequency signal power or the output noise power of a receiver.

Basically, it merely consists of a tapped audio-frequency transformer for effecting a match between the meter and the receiver output valve together with a suitable monitor. Fig. 25 (*a*) shows typical circuitry of the more superior types, whilst Fig. 25 (*b*) shows the monitoring arrangement of other types, R being a calibrating resistor. The bridge rectifiers shown in (*a*) and the straight full-wave rectifiers in (*b*) are small "meter types" such as WX1.

A fixed 150-ohm impedance is provided by lower-grade meters, whilst that of Fig. 25 (*a*) is variable between 2·5 ohms and 20 kΩ. The output rectified D.C. is recorded by the microammeter, which has a full-scale-deflection current of the order of 50 μA and is calibrated in terms of milliwatts and decibels. The "Meter Multiplier Switch" (0·1–100, four-step) varies the calibrated output between values of 5 milliwatts and 5 watts at full-scale deflection.

The useful frequency range should be from 50 to 8,000 c/s, over which the meter response should be level within. ± 1 db.

A.F. Absorption Wattmeter

This instrument is virtually a "modified output power meter", as it usually registers audio-frequency output powers between some 10 μW and 6 watts by a ten-step switch. A tapped matching transformer provides about twelve impedance-values over the range of 2·5 ohms to 20 kΩ. It functions satisfactorily over an audio-frequency range of 50 c/s to 20 kc/s. Temperature compensation is incorporated.

R.F. Absorption Wattmeter

This type of instrument is the transmitter-counterpart of an output-power meter in that it registers the radio-frequency output of a transmitter.

In a typical instrument, two impedance-values are presented by the instrument to the transmitter under test, these being monitored by a double-range meter and are 0–1 watt working into a 52-ohm load and 0–25 watts working into a 70-ohm load. The frequency coverage is from 1 to 100 Mc/s.

Decibel Meter

These instruments frequently provide four ranges of power measurement to a reference power level of 1 mW in a 600-ohm load.

The impedances presented by a switched audio-frequency transformer are usually of the order of 75, 150 and 600 ohms. The power ranges covered may be from -15 db to $+30$ db over a frequency range of 200 c/s to 120 kc/s. The output-rectifying arrangement uses four germanium crystals in a bridge-circuit.

This meter has its chief applications in high-power transmitters, line and V.F. level measurements

Circuit Magnification Meter (*Q*-Meter)

The primary functions of a *Q*-meter are :

(1) Measuring the *Q*-values of coils within a frequency range of about 50 kc/s to 75 Mc/s.

(2) Measuring the values of small capacitors (less than 450 pF).

(3) Measuring the self-capacitances of inductors.

The tuning capacitor and its associated vernier of the test circuit (together denoted "T.C." in Fig. 26) are calibrated in picafarads for the capacitance measurements. The versatility of this instrument enables several secondary measurements to be effected.

Fig. 26 shows the basic arrangement generally adopted. A conventional positive feedback tuned grid or a cross-connected push–pull type of radio-frequency oscillator with high-frequency stability passes radio-frequency current through the heater element of a sensitive thermo-couple and a 0·04-ohm manganin strip "R" The D.C. millivoltmeter "*Q*-range" records the potential difference produced across the thermo-junction, any radio-frequency energy being removed from the meter by a radio-frequency filter. The radio-frequency voltage developed across R by the oscillator acts as the input to the series-tuned circuit comprising the inductor under test, R and the calibrated capacitor(s) T.C. When the radio-frequency oscillator develops a *Q*-range value of "X1", 20 mV are developed across R, this being magnified "*Q* times" to appear across L or T.C. The voltage across T.C. is applied to the valve voltmeter, and has a value of 5 volts R.M.S. when the output "*Q*-value meter" records full-scale deflection. The latter is calibrated in *Q*-values of 10–250. If the radio-frequency oscillator output is halved by manipulation of the "Magnification" control, the *Q*-range indicator records "X2", 10 mV then being produced across R. If F.S.D. is again effected on the *Q*-value meter, the input to the valve voltmeter must still be 5 volts, hence the magnification of the circuit under test has obviously doubled, and all *Q*-values recorded must be doubled (20–500).

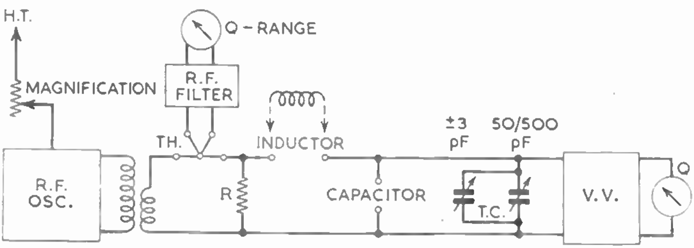


FIG. 26.—BLOCK LAYOUT DIAGRAM OF TYPICAL *Q*-METER.

An "External Meter" jack provides for the connection of an external meter. The "Set Zero" control is that normally associated with valve voltmeters.

When setting-up the instrument it is essential that the magnification controls are fully anti-clockwise before switching. This protects the thermo-couple. To protect the microammeter in the valve voltmeter the instrument should not be switched on without an inductor being connected across the concerned terminals. Before commencing operations, a half-hour warming-up period should be allowed.

Measurements with a *Q*-Meter

By virtue of the calibration accuracy in both oscillator frequency and in test-circuit tuning capacitor T.C., several useful measurements can be effected with a *Q*-meter as follows:

Determination of the *Q* of an Inductor

(1) Tune the radio-frequency oscillator to the required frequency.
(2) Connect the coil to the "Inductor" terminals.
(3) Adjust "Magnification" until the "*Q*-range" meter shows "X1".

(4) Adjust T.C. for resonance, i.e., for maximum on the "Q-range" meter. If the pointer swings above F.S.D. (Q = over 250), adjust magnification until "X2" is recorded on the Q-range meter and re-resonate.

The value of Q is that shown on the magnification (Q-value) meter if the Q-range was set up on X1, or double this if on X2.

At frequencies below 15 Mc/s, the self-capacitance of the coil Cs throws an error, in which case the true Q-value is $Q_0(1 + Cs/C1)$, where Q_0 is the recorded Q-value and $C1$ is the value of T.C.

Another method of determining the Q of a coil is as follows :

(1) Resonate both circuits to the required frequency.

(2) Adjust the oscillator output until the magnification meter indicates a convenient Q-value (141 or 100). Note the oscillator frequency.

(3) Detune the oscillator until half-power output results, i.e., until the recorded Q-value has fallen to 0·707 times that of sub-para. 2 (100 or 71). Call the change in oscillator frequency δf.

The Q-value is now given by $fo/2\delta f$, where fo is the resonant frequency. As δf is only a small change in frequency, accurate results are not usual with this method. Greater accuracy is obtained with the following system :

(1) Resonate both circuits to the required frequency and adjust the oscillator output for the same Q-value as before. Note the oscillator frequency and the calibrated value of T.C. Call this C.

(2) Detune T.C. until half-power output results (0·707 × original Q-value). Call the required change in capacitance δC.

The Q-value is now given by $C/\delta C$. Actually, the value of δC taken should be the average of the capacitance changes on either side of resonance necessary to reduce Q to 0·707Q.

As both of the foregoing methods involve only small percentage detuning, inaccuracies are inevitable. Still greater accuracy results if the following system is adopted :

(1) Resonate both circuits to the required frequency.

(2) Adjust the oscillator output for a convenient value of Q such as 224, 158 or 103. Call this $Q1$. Note the oscillator frequency fo.

(3) Detune the oscillator until the Q-value shown falls to a convenient value $Q2$ such as 100, 50 and 25 respectively for the examples suggested in sub-para. 2. Call the change in oscillator frequency δf.

The Q-value is now given by $\frac{fo}{2\delta f}\sqrt{\left(\frac{Q1}{Q2}\right)^2 - 1}$, this simplifying to $\frac{fo}{\delta f}, \frac{3fo}{2\delta f}$ or $\frac{2fo}{\delta f}$ for the three examples previously suggested.

Comparable accuracy results if detuning is effected on T.C. instead of varying the oscillator frequency. In this case the Q-value is given by $\frac{C}{\delta C}\sqrt{\left(\frac{Q1}{Q2}\right)^2 - 1}$, where C, δC, $Q1$ and $Q2$ have the previous significations. For the three cases quoted, this expression simplifies to $2C/\delta C$, $3C/\delta C$ and $4C/\delta C$ respectively.

Determination of the Self-capacitance of a Coil

The self-capacitance of an inductor may be determined by one of the following methods with varying degrees of accuracy :

(1) Connect a coil of $\frac{1}{5}$–$\frac{1}{10}$ the inductance of the inductor under test to the "Inductor" terminals.
(2) Resonate to a frequency requiring not more than 100 pF on T.C. Note this value of tuning capacitance.
(3) Connect the inductor under test to the "Capacitor" terminals.
(4) Reduce the value of T.C. to obtain resonance again. Note the new value of T.C.

The self-capacitance of the test inductor is now given by the difference in the T.C. values of sub-paras. 2 and 4. Greater accuracy devolves from the following method :

(1) Connect the coil under test to the "inductor" terminals.
(2) Set T.C. to about 70 pF and tune the oscillator to the test circuit. Note its frequency.
(3) Reset the oscillator frequency to exactly half that of sub-para. 2 and re-resonate on T.C.

If the values of T.C. in sub-paras. 2 and 3 are denoted by Ca and Cb respectively, then the self-capacitance of the inductor is given by $\frac{Cb - 4Ca}{3}$. Alternatively, the inductor may be resonated at any two frequencies f_1 and f_2, when the self-capacitance of the inductor under test is given by $\frac{f_1^2Cb - f_2^2Ca}{f_2^2 - f_1^2}$ pF. If $f_2 = nf_1$, this simplifies to $\frac{Cb - n^2Ca}{n^2 - 1}$.

Capacitance Measurement

Capacitors with values up to about 400 pF may be measured with good accuracy as follows :

(1) Connect a suitable coil to the "Inductor" terminals and resonate it with a high value of T.C. (450–500 pF). A suitable frequency is of the order of 2 or 3 Mc/s. Note the calibrated value of T.C.
(2) Connect the capacitor under test to the "Capacitor" terminals.
(3) Reduce the value of T.C. until resonance once more obtains. Note the new value of T.C.

The value of the capacitor is the difference between the values of T.C. in sub-paras. 1 and 3.

Inductance Determination

The inductance of a coil can be determined as follows: Connect it to the "Inductor" terminals and resonate the oscillator to the test circuit. From the values of T.C. (C) and resonant frequency (f), the value of L is given by $\frac{1}{(2\pi f)^2 C}$. Strictly speaking, this involves an error due to the self-capacitance of the inductor. This should be determined and added to the value of C if greater accuracy is required. The error may be avoided by the following procedure:

(1) Connect the coil to the "Inductor" terminals.

(2) Resonate it to a frequency requiring about 100 pF on T.C. Note the oscillator frequency (f_1).

(3) Tune the oscillator to exactly half this frequency (f_2).

(4) Tune T.C. for resonance once more and note its new capacitance. Call the change in capacitance from sub-para. 2 to sub-para. 4, C_d.

The true inductance-value of the coil is now given by

$$\frac{3}{\omega_2{}^2 C_d} \quad \text{or} \quad \frac{3}{4\omega_1{}^2 C_d}$$

where ω_1 and ω_2 refer to f_1 and f_2.

Determination of Radio-frequency Resistance

The radio-frequency resistance of an inductor at a given frequency is determined as follows:

(1) Connect the coil to the "Inductor" terminals.

(2) Tune the oscillator to the required frequency.

(3) Adjust T.C. for resonance. Note the Q-value recorded (Q_1).

(4) Connect a sub-standard non-inductive resistor R in series with the inductor and re-resonate. Call the new Q-value Q_2.

The radio-frequency resistance of the inductor at the specified frequency is now given by $\frac{Q_2 . R}{Q_1 - Q_2}$.

Determination of Dynamic Resistance

The dynamic resistance Rd of a parallel-tuned circuit may be determined as follows:

(1) Connect a coil comparable in inductance to that of the test inductor to the "Inductor" terminals and resonate to the required frequency. Call the value of T.C. C and the resultant Q-value Q_1.

(2) Connect the test circuit across the "Capacitor" terminals and tune it to the desired frequency (from previously noted calibration of its tuning capacitor). The recorded value of Q will fall. Call the new value Q_2.

The dynamic resistance of the test circuit is now given by

$$Rd = \frac{0{\cdot}159 \times Q_1 \times Q_2}{fC(Q_1 - Q_2)} \text{ M}\Omega$$

where C is quoted in pF and f in Mc/s.

Measurements of Small Impedances

These are concerned with the radio-frequency resistances of components having low D.C. resistances, e.g., thermoammeters, small inductors, the inductances of large capacitors, etc. The method is as follows :

(1) Connect a small inductor to the " Inductor " terminals and resonate at a fairly high frequency (20 Mc/s). Call the recorded Q-value Q_1 and the value of T.C. C_1.

(2) Connect the test impedance in series with the earthy end of this inductor, ensuring that no coupling exists between them.

(3) Re-resonate to the original frequency. Call the recorded Q-value Q_2 and the new value of T.C. C_2.

At the selected frequency the required values of radio-frequency resistance of inductance or capacitance respectively, denoted Rx, Lx or Cx, are now given by

$$Rx = \frac{1}{\omega}\left(\frac{1}{C_2Q_2} - \frac{1}{C_1Q_1}\right), \qquad Lx = \frac{C_1 - C_2}{\omega^2 C_1 C_2}, \qquad Cx = \frac{C_1C_2}{C_2 - C_1}$$

Greatest accuracy results in the case of Lx and Cx when $C_1 = 2C_2$ or $C_2 = 2C_1$ respectively.

Determination of the Characteristic Impedance (Zo) and the Velocity Propagation Constant (Vc) of a Feeder Line

The characteristic impedance of a feeder line may be determined as follows :

(1) Select a length of feeder less than $\lambda/4$ long.

(2) Connect a suitable coil to the " Inductor " terminals and resonate it to the oscillator at the frequency corresponding to λ in sub-para. 1. Note the value of T.C.

(3) Connect the feeder length to the " Capacitor " terminals.

(4) Reduce the value of T.C. until resonance once more obtains (the feeder is now acting as a capacitor). Note the value of T.C. Call the required decrease Coc.

(5) Short the far end of the feeder length and retune for resonance. As it now acts as a parallel inductor, more capacitance than that of sub-para. 2 is required. Call this increase Csc. The value of Zo is now given by $\dfrac{1}{\omega\sqrt{Coc.Csc}}$, where the capacitance values are quoted in farads and f in c/s.

The velocity propagation constant may be determined with the same arrangement as follows :

(1) Proceed as in sub-paras. 1 and 2 above.

(2) Connect the feeder length to the " Inductor " terminals and short the far end.

(3) Increase the oscillator frequency. Re-resonate and note the value of T.C.

(4) Connect the feeder length to the " Capacitor " terminals and

re-resonate. Note the value of T.C. The difference capacitance will have decreased.

(5) Repeat sub-paras. 3 and 4 until this difference capacitance has fallen to zero. Note the oscillator frequency when this point is reached and evaluate the corresponding $\lambda/4$ free space value ($= V/4f$).

(6) Measure the physical length of the sample feeder.

$$\text{The value of } Vc \text{ is now given by } \frac{\text{Feeder Length}}{\text{Free Space } \lambda/4 \text{ Value}}.$$

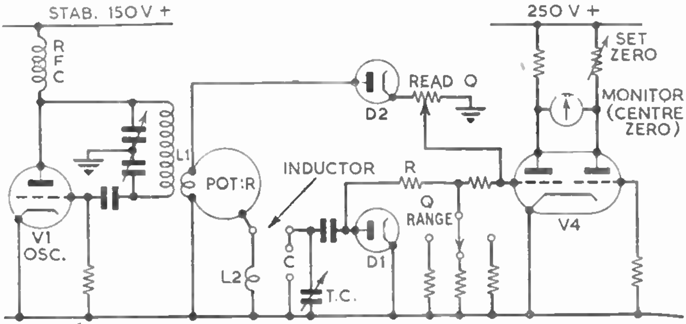


FIG. 27.—V.H.F. *Q*-METER CIRCUITRY (POWER SUPPLIES OMITTED).

V.H.F. *Q*-Meter

The V.H.F. version of the *Q*-meter is functionally different to the normal medium-frequency/high-frequency model, as the frequency range extends higher into the V.H.F. spectrum (15–170 Mc/s).

A different monitoring circuit is usually employed, as can be seen from Fig. 27. In this instrument the radio-frequency output developed across the coupling coil L1 is rectified by the diode D2, attenuated by the "Read *Q*" potentiometer and applied as D.C. input to the D.C. amplifier V4. A fraction of the oscillator output is simultaneously injected into the test circuit via the common coupling inductor L2. This becomes magnified "*Q* times", rectified by the diode D1 and applied as D.C. to V4 via the potential divider formed by R and the "*Q*-range" switch. This voltage is opposite in polarity to that from D2, and if equal in magnitude to it, the anode voltages of V4 may be equalized by the "Set Zero" control. The monitoring meter then reads zero (centre). This system avoids the necessity for accurate oscillator monitoring by a thermo-couple, etc. The "Read *Q*" control is calibrated in terms of *Q*, and must be adjusted for a centre-reading (zero) on the monitor.

Zero-setting is actually carried out with the radio-frequency oscillator out of action, this being effected when the Set Zero control is depressed.

The test circuit tuning capacitor "T.C." has a maximum capacitance of the order of 90–100 pF.

Circuit Analyser (Signal Tracer)

This instrument simulates and detects signals at all stages of a receiver for rapid fault-finding purposes. Hence it is a combined test oscillator, valve voltmeter and audio oscillator. A built-in audio-frequency amplifier boosts the receiver audio-frequency signal for operation of a magic-eye indicator or the receiver loudspeaker.

The instrument often has its own loudspeaker and associated matching transformer, or it may be fed direct from the receiver output transformer. External headphones may be connected if the signal is very weak.

The probe unit carries a high-efficiency pentode detector which operates the magic eye direct or is amplified and applied to the loudspeaker. A high input impedance will permit detection of D.C. potentials associated with the automatic-gain-control circuit.

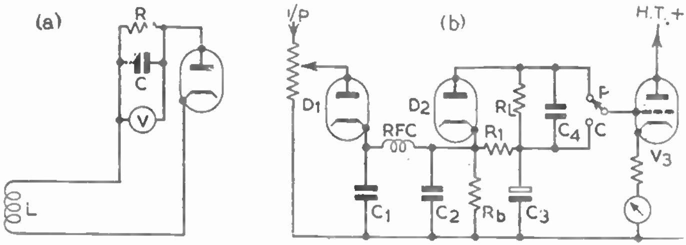


Fig. 28.—Modulation Meter Arrangements.

Modulation Meter

A modulation meter is a device giving a direct indication of the percentage modulation effected by a high-power transmitter. Two arrangements are shown in Fig. 28. The coupling coil L of Fig. 28 (*a*) is coupled to the tank circuit of the final power-amplifier stage. The pulses of current rectified by the diode charge the capacitor C at a short time-constant, i.e., C.Ra. The voltmeter (V) is an electrostatic type, and due to the long discharge time-constant of the load combination (R = 50 MΩ) it registers the voltage across C. The full-scale deflection value of the voltmeter may be 1,000 volts or so, and due to the cramped scale associated with this type of instrument, the unmodulated calibration is arranged to coincide with half full-scale deflection, i.e., 500 volts. 100 per cent modulation increases this to 1,000 volts. Hence the open extent of the scale which covers most of the dial (500–1,000 volts) can be accurately calibrated in terms of "percentage modulation".

The arrangement in Fig. 28 i(*b*) compares the amplitude of the audio-frequency modulating component with the mean carrier amplitude. The input is applied to the dode D1, the resultant radio-frequency current being smoothed by the C1/R.F.C./C2 filter. Hence the common cathode current produced through Rb is proportional to the peak carrier input i.e., it carries the modulating audio-frequency component. This is filtered by R1/C3, leaving only the D.C. component of voltage across

C3, this being proportional to the "mean" or unmodulated radio-frequency input. The voltage developed across R1 is thus the difference between that across Rb and C3, i.e., it is proportional to the modulation. This biases D2 anode negatively during modulation peaks and positively during the modulation troughs. The latter produces rectified current through the load combination Rl.C4 of D2; hence the voltage across C4 is proportional to the peak (negative) modulation. This voltage is of opposite polarity to that across C3. With 100 per cent modulation they are equal in magnitude, so the voltage between the grid of V3 and earth when the switch is at P (peak) is zero. With the switch at C (carrier only) the V3 cathode meter is arranged to read full-scale deflection; this corresponding to 0 per cent modulation. The action of the modulating component thus produces an annulling negative potential on the grid of V4, which takes this valve virtually to cut-off with 100 per cent modulation. The V3 cathode meter is thus calibrated "0 per cent

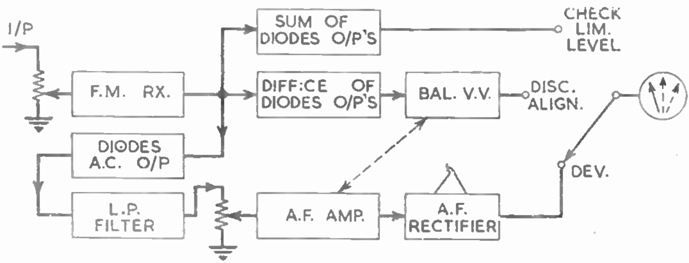


Fig. 29.—Deviation Meter.

modulation" at full-scale deflection and "100 per cent modulation" at zero deflection. The percentage modulation is read direct when the switch is made to the P position.

Frequency-modulation Deviation Meter

This instrument records the frequency deviation of a frequency-modulation transmitter. Monitoring may also be effected aurally with headphones or by visual inspection of the modulation waveform with the aid of a cathode-ray oscilloscope.

It comprises a calibrated low-sensitivity frequency-modulation receiver up to and including the discriminator stage, a two-stage audio-frequency amplifier (which is switch-convertible to a balanced-valve voltmeter) and an audio-frequency rectifier. The monitor is a sensitive microammeter switchable to one of three positions, as shown in the block schematic layout of Fig. 29. These are:

(1) *Check Limiter Level.* At this position the gain of the monitoring receiver is adjusted such that the combined output of the discriminator diodes (one diode output reversed in polarity) is to some pre-determined level such as the right-hand position of the monitor pointer in Fig. 29.

(2) *Discriminator Alignment.* The monitor now becomes a

centre-zero balanced-valve voltmeter in conjunction with the audio-frequency amplifier valves, so normal discriminator alignment may be effected.

(3) *Deviation.* At this position the amplifier valves assume normal functions. The audio-frequency output from the discriminator is amplified, rectified and recorded by the monitor. The reading on the monitor is thus proportional to the peak deviation in the frequency of the transmitter carrier. The monitor is calibrated in terms of " kc/s deviation ".

The frequency coverages of the instrument vary with the transmitter for which it is designed, and so may be between 10 and 200 Mc/s. The deviation recorded may proceed in steps of 5 kc/s, or may be continuously variable up to ± 300 kc/s.

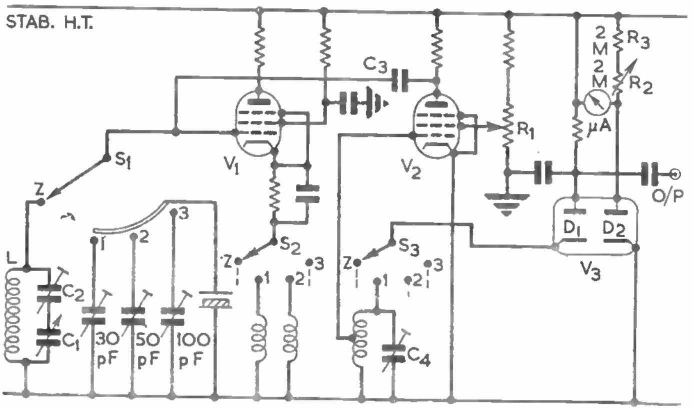


FIG. 30.—CRYSTAL ACTIVITY TESTER.

Crystal Activity Tester

With this type of instrument a check can be made on the activity of a quartz crystal having a fundamental frequency between some 100 kc/s and 8 Mc/s. This is usually accomplished by comparing the equivalent parallel resistance of the crystal under normal working conditions with that of an internal LC circuit tuned to a pre-determined frequency (250 kc/s). Sockets are provided to take the standard crystal-holders.

Fig. 30 shows the essential circuitry. The V1/V2 oscillator comprises a cathode follower stage (V1) which is cathode coupled via a radio-frequency transformer to a conventional resistance–capacitance amplifier (V2), from the anode of which positive feedback is applied to V1 grid via C3 (10 pF). The inputs to V1 grid and V2 grid are arranged to be anti-phase to maintain oscillations. The amplitude of these depends on the dynamic resistance of the circuit selected by S1. At position " Z " this is the test circuit L–C1–C2. At other positions it is the crystal under test shunted by different capacitances (20 pF, 50 pF, etc.). The frequency of the derived oscillations is decided by the cathode secondary, and C4, S1, S2 and S3 are ganged.

The amplitude of the radio-frequency oscillations is controlled by R1 and monitored by the microammeter. Output is adjusted for a convenient reading on the monitor with the crystal in circuit and shunted by the appropriate capacitance. The LC test circuit is now switched in and tuned for a similar output by manipulation of C1 (100 pF), this being calibrated in kΩ. C2 (50 pF) sets the scale calibration. L–C1–C2 is a high-*Q* circuit, and is negligibly damped by the cathode follower.

V3 provides limiting action to the meter when the oscillations become excessive. Rectified current from D1 due to radio-frequency voltages from S3 produces volt-drops across R2 + R3 which back off Ea and decrease the standing anode current. This limits the movement of the microammeter pointer.

The parallel resistance recorded in kΩ on C1 should be up to the standard specified by the crystal manufacturer for the particular shunting capacitance concerned.

Output may be taken from the instrument to check frequency meters.

Frequency-measuring Instruments

A great variety of frequency-measuring devices is available. Generally, these may be segregated into four distinct groups according to their modes of operation.

(1) The heterodyne type of frequency-meter is basically a very low-power transmitter convertible by exchange of telephone points into a very low-sensitivity receiver. In both cases the degrees of frequency-calibration and frequency-stability are of a high order. When a transmitter is under frequency check, the instrument becomes a receiver very loosely coupled to it. When a receiver is under frequency check, the telephones are plugged into it and the instrument becomes a transmitter very loosely coupled to it. In both cases heterodyne action takes place, so tuning is adjusted for zero-beat note, the two frequencies concerned then being virtually the same. The required frequency is always read from the meter. This method is the most accurate, errors of within 0·02 per cent being common.

(2) The absorption type of frequency-meter relies for its action on the resultant absorption of radio-frequency power from the metered circuit when the meter is coupled to it. This power is used to actuate a resonance indicator such as a microammeter, a small pea-lamp or a neon tube. These meters are only of use with transmitters or oscillators, and accuracy is of a low order.

(3) A crystal calibrator is basically a heterodyne type of frequency-meter, which does not provide continuously variable tuning but a series of "spot" frequencies, intermediate frequencies being located on the equipment under test by careful interpolation.

(4) The grid-dip oscillator (G.D.O.) functions on a somewhat reverse principle to the absorption-type frequency-meter. It is coupled to a "dead" (non-oscillatory) circuit under check, and the energy abstracted affects the action of the grid-dip oscillator. A resonance indicator shows when this is a maximum.

None of the first three types can be used to determine the frequencies of dead circuits.

Heterodyne-type of Frequency-meter

This type is generally used at all frequencies up to 40 or 50 Mc/s. The principle of operation adopted is usually based on the production of sideband-component frequencies when a carrier wave is modulated. The modulating frequency is arranged to have a waveform rich in harmonic content in order that multiple sidebands are produced, e.g., if a carrier frequency *fo* becomes modulated by a non-sinusoïdal frequency *fm*, then the spot frequencies usable for check purposes may be $fo - 3fm$, $fo - 2fm$, $fo - fm$ (lower sidebands), *fo* (the carrier), $fo + fm$, $fo + 2fm$, $fo + 3fm$ (upper sidebands), etc. To aid this action, the carrier oscillator is deliberately over-modulated.

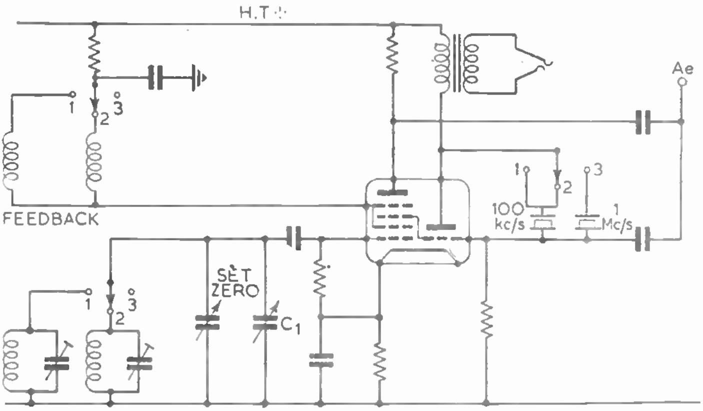


FIG. 31.—HETERODYNE TYPE OF FREQUENCY-METER.

Fig. 31 shows the essential circuitry of a popular instrument. The hexode regenerative tuned-grid oscillator is tuned to a pre-determined spot frequency at each of the switch positions 1 and 2 such as 4 and 7 Mc/s. These act as the carrier frequencies. C1 is a vernier tuning capacitor (20 pF) capable of varying the carrier frequency by 100 kc/s, and is suitably calibrated 0–100 kc/s. At these switch-positions the carrier is modulated electronically by the 100-kc/s triode crystal oscillator. This produces sideband components of the carrier at intervals of 100 kc/s, which together with the frequency variation provided by C1, give continuously variable coverage from 1·5 to 10 Mc/s. The "Set Zero" control is adjusted for zero beat note with C1 at 0 (zero). To use the instrument intelligently, it is essential that the frequency of the receiver or transmitter under check is known within 100 kc/s.

At switch-position 3 the hexode oscillator is out of circuit, and the triode functions at 1 Mc/s. Beat notes will be heard at 1-Mc/s intervals if the transmitter frequency is varied.

Calibration charts are provided in some models when an accurate slow-motion drive is used on full-range frequency coverage. Crystal-check frequencies at frequent intervals enable accuracies within 0·01 per cent to be realized.

Loose coupling must be used between the instrument and the equipment under check, or the mutual inductance will throw an error in the developed frequency.

Absorption Type of Frequency-meter

Few frequency-meters of the absorption type are commercially available for use on medium-frequency and high-frequency bands. The usual monitoring arrangement is a microammeter of some 50–250 μA full-scale-deflection current rectified by a germanium crystal. The tuning inductor usually plugs into the end of the instrument case for ease in coupling to the metered transmitter. This type of instrument is convenient for tracing parasitic oscillations in a transmitting installation.

Fig. 32 shows the circuit of a popular V.H.F. arrangement, as the design of heterodyne types becomes difficult at these frequencies. The primary frequency coverage is 150–300 Mc/s, but secondary

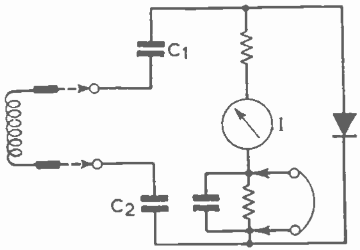


FIG. 32.—V.H.F. ABSORPTION FREQUENCY-METER.

coverages of 20–80 Mc/s are available by using other plug-in tuning inductors. Calibration charts are provided for each range. The plug-in inductors are large-diameter, air-spaced, high-Q coils very lightly loaded by the germanium crystal. This is further aided by the light coupling to the monitoring circuit, C1 being only a few picafarads and C2 merely the capacitance between the front panel and the tuning capacitor. I is a 0–50 microammeter, which may become damaged if the coupling between transmitter and instrument is too great. Tuning is remarkably keen in view of the high-Q inductor and the light loading.

Crystal Calibrator

Basically, a crystal calibrator is a heterodyne type of frequency-meter which provides spot frequencies between about 1 and 40 Mc/s at intervals of 10 kc/s (or 20 kc/s). These may be modulated at 400 c/s if necessary, when tho output valve becomes converted to an audio-frequency oscillator.

Fig. 33 shows a typical block schematic layout. The 1-Mc/s crystal oscillator generates copious harmonics, its output being fed direct to the aerial or to a multi-vibrator the frequency of which it locks at 100 kc/s. The output of this in turn locks a 10-kc/s (or 20-kc/s) multi-vibrator. The outputs of both multi-vibrators and of the crystal oscillator are

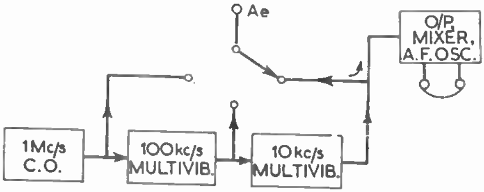


FIG. 33.—BLOCK SCHEMATIC LAYOUT OF A CRYSTAL CALIBRATOR.

developed across a common load and fed to the aerial. As multi-vibrators are noted for their impure waveforms, spot frequencies are available for check purposes at intervals of 10 kc/s (or 20 kc/s) over the whole band. The free-running frequencies of the multi-vibrators are panel-adjusted by variable grid resistors.

An external signal may be injected via the aerial terminal to develop voltages across the common load. These are mixed with the instrument oscillations, and the beat note is heard in the telephones.

Grid-dip Oscillator

Very few models of this type of frequency-meter are commercially available, although it meets with considerable favour in amateur circles.

Fig. 34 shows typical circuitry of a grid-dip oscillator which is a conventional Hartley oscillator except that the value of the grid resistor is considerably lower than normal. This results in greater values of grid current recorded on the microammeter, the full-scale-deflection value of which may be 100 μA. The oscillator must be loosely coupled to the "cold circuit" being metered and the tuning capacitor adjusted for maximum dip in the microammeter. Under these conditions the resonant frequency of the test circuit is the same as the grid-dip-oscillator frequency. The tuning capacitor may thus be accurately calibrated in terms of frequency.

If this instrument has a small battery valve, it may used as a "field-strength meter".

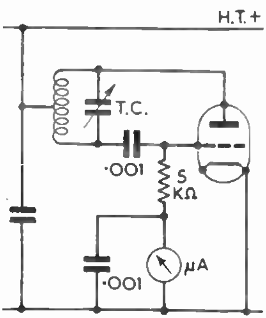


FIG. 34.—GRID-DIP OSCILLATOR.

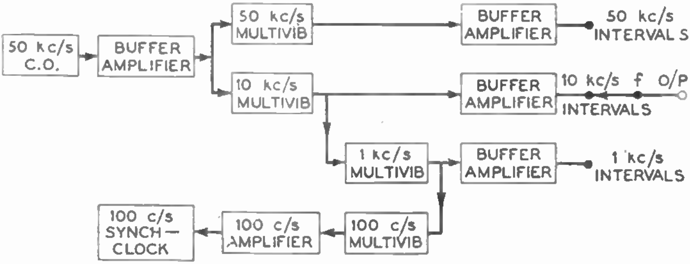


FIG. 35.—BLOCK ARRANGEMENT OF A STANDARD FREQUENCY-METER.

Standard Frequency-meter

This instrument is sometimes termed a synthesizer. It is a primary "master" frequency-meter, against which secondary frequency-meters may be checked for accuracy.

A very stable crystal oscillator using a specially ground crystal functions at a frequency of 50 or 100 kc/s. This triggers a train of multivibrators as shown in Fig. 35. The output from the final multi-vibrator (100 c/s) operates a synchronous clock. As the input to this clock is exactly $\frac{1}{500}$ or $\frac{1}{1000}$ of the crystal frequency, the stability of the crystal oscillator can be verified by checking the time recorded by the clock over a period (say 3 hours) against G.M.T. The series of harmonics developed by the multi-vibrators also have comparable accuracies, so the errors over long working periods may be within one part in 10^7.

To avoid confusion over the multiple frequencies involved, calibrations not in use are masked. Sometimes an interpolating oscillator with a frequency variation of 1 kc/s is used to provide an output which is continuously variable in frequency.

W. C. R.

CATHODE-RAY OSCILLOSCOPES

The cathode-ray oscilloscope has come to play a most important role, both in the laboratory and in day-to-day servicing and adjustment.

The invisible beam or pencil of electrons which is produced in a cathode-ray tube is focused to its maximum concentration at the point where the beam encounters the fluorescent screen on the end of the tube.

The brilliance is controlled by the voltage on the modulator of the tube, while the application of suitable deflection fields will cause the beam to deviate from its normal position and consequently the spot of light to move across the face of the screen.

This movement of the spot is called a *trace*, and the pattern produced by the trace is termed a *display*.

Rectangular Displays

There are many forms of display, the simplest and most common being the rectangular display, in which there are two distinct deflecting forces at right angles. One of these will cause the spot to move horizontally across the screen, while the other will cause it to move vertically up and down.

FIG. 36.—MODEL 1035 CATHODE-RAY OSCILLOGRAPH

(Cossor Instruments Ltd.)

The deflecting plates are carefully aligned in manufacture so that the two deflection directions are accurately at right angles, but the movement of the spot relative to the observer depends upon the orientation of the tube itself. It is necessary to mount the tube in such a position that when voltage is applied to the X plates, the movement of the spot is in a horizontal direction. It is customary to mount the tube in a holder which permits a small rotation so that the trace can be accurately lined up. This is particularly necessary if the tube is provided with a ruled transparent graticule or cover plate to facilitate measurement of the co-ordinates of the trace.

It is customary to make the horizontal movement of the spot proportional to the primary variable, e.g., time, with the vertical deflection proportional to the dependent variable.

The necessary deflecting voltages can also be produced by electromagnetic means, though this is less frequently used in oscillograph work.

The most common application of a rectangular display is the presentation of waveforms. Here the horizontal movement of the spot is made proportional to time, and the vertical deflection is proportional to the instantaneous value of the voltage to be examined. The movement of the spot then traces out the waveform.

Usually the time occupied by a single waveform is only a small fraction of a second. The time-base is thus synchronized with the "work" so that it generates a succession of horizontal movements at accurately repeated intervals. The successive traces then become exactly superposed, giving the impression of a stationary image. It is possible by this method to examine waveforms having frequencies ranging from a few cycles up to many megacycles per second.

Waveform examination, however, is only one of the examples of the rectangular display. Actually the horizontal movement can be made proportional to any desired primary variable, such as position, frequency, resistance, pressure, etc.

Frequency Bases

It may be more convenient to employ the usual time-base to sweep the spot across the screen, and either to use this voltage as the primary variable or to synchronize the time-base with the external source.

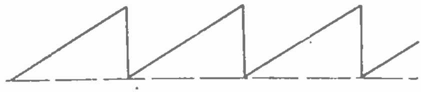


Fig. 37.—Sawtooth Waveform.

The plotting of frequency response curves provides an example of this usage. The conventional time-base used in an oscilloscope generates a "sawtooth" voltage which increases at a uniform rate to a certain predetermined value and then falls rapidly to zero as shown in Fig. 37. This causes the spot to move horizontally across the screen and then to fly back and repeat the cycle indefinitely.

There are various forms of circuit, known as frequency modulators, in which the frequency generated can be controlled by the voltage applied to the grid of a valve. If such a circuit is fed with sawtooth voltage from the oscilloscope time-base, the frequency will change uniformly over a certain range. This varying frequency can be supplied to the circuit under test, and the output can be applied to the vertical plates of the cathode-ray tube. The resulting display on the tube will portray the frequency-response curve of the circuit.

The same result could be achieved by using a mechanically rotating capacitor to produce the required variation of frequency. In this case the oscilloscope time-base would be synchronized with the rotation of the capacitor shaft so that the movement of the spot corresponded exactly with the shaft rotation and hence with the frequency generated.

Time-base Circuits

It is desirable to consider more closely the behaviour of the time-bases and amplifiers used with rectangular displays. This consideration will be related more particularly to the examination of waveforms, but it will be clear that the remarks apply to any of the many possible applications.

As already mentioned, the time-base generates a sawtooth voltage, causing the spot to move uniformly across the screen and then fly back and repeat the performance. Essentially a time-base consists of a capacitor charged through a high resistance. The voltage on the capacitor then increases gradually and more or less uniformly, and as it does so the spot is caused to travel across the screen. Across the capacitor is connected a valve or combination of valves so arranged that at a certain pre-determined voltage the capacitor is discharged, thus

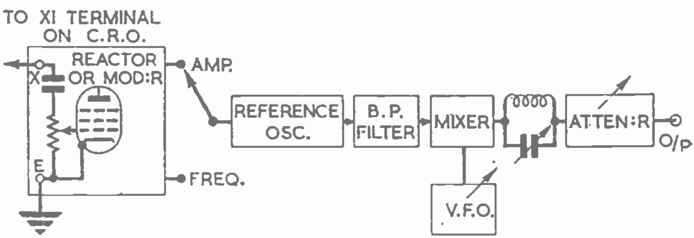


Fig. 38.—Use of Frequency Modulated Oscillator (Known also as "Ganging Oscillator", "Sweep Generator" or "Wobbulator") for Tracing Frequency Response Curves.

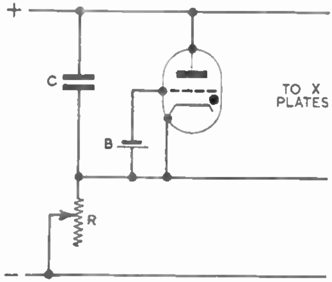


FIG. 39.—SIMPLE TIME-BASE CIRCUIT.

The capacitor C is charged through the variable resistor R to permit the charging rate (and hence the speed of movement of the spot) to be altered. The triode across the capacitor conducts, thus discharging the capacitor when the voltage reaches a pre-determined value (derived here from the battery B) controlled by the voltage on the grid.

producing the fly-back and permitting the operation to recommence. Fig. 39 shows a simple time-base circuit.

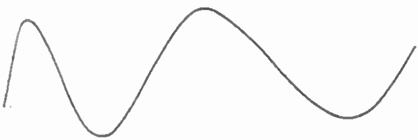
Various requirements have to be met in a satisfactory time-base. Simple resistance charging is only satisfactory in certain limited conditions because as the capacitor charges, the current through the circuit will gradually decrease (actually in an exponential fashion) so that the rate of increase of voltage, and hence the rate of movement of the spot across the screen, will not be uniform but will gradually slow down. This will effectively change the scale of the display on the screen, giving rise to the effect known as non-linearity, as illustrated in Fig. 40. It is more usual, therefore, to charge the capacitor through a pentode valve which can be so arranged that it delivers a constant current over the range of operation required, and with such an arrangement the rise in voltage is strictly linear. Fig. 41 illustrates a pentode-charged arrangement.

Effect of Trace Speed on Brilliance

It is important also that the time occupied in the discharge of the capacitor at the end of the sweep shall be small. Otherwise a distinct trace will be visible on the return stroke, and this may confuse the observation. If the spot is moving rapidly, however, it will not leave any appreciable trace, and no confusion will result. In circumstances where a slow fly-back cannot be avoided, it is possible to arrange that during this portion of the trace a suitable negative voltage is applied to the modulator of the tube, causing the trace to be " blacked out ".

Attention should perhaps be directed to this relationship between the speed of the spot and the apparent brilliance of the trace. Since with a given setting the light energy in the spot is fixed, it is clear that the greater the distance moved by the spot in a given time, the less will be the apparent brilliance of the trace which it leaves. There are very few waveforms in which the spot is moving at a uniform rate throughout the whole of the trace, and consequently there is always some variation in brilliance. For simpler waveforms this variation is barely noticeable.

FIG. 40.—DISTORTION OF WAVEFORM OWING TO NON-LINEAR TIME-BASE.



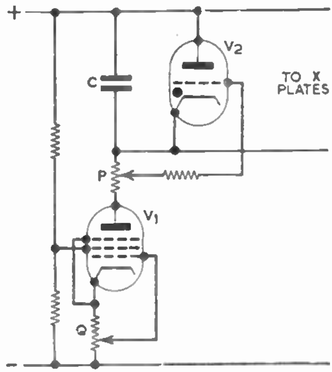


FIG. 41.—PENTODE CHARGED TIME-BASE.

. Here the resistor R in Fig. 39 is replaced by the pentode V1, the effective resistance of which is controlled by the voltage on its grid determined by the setting of Q. (This resistance does not vary with the anode voltage of V1 as the anode current, i.e., the charging current of C remains nearly constant for varying anode voltage.) In the place of B in Fig. 39, P controls the bias on the grid of V2.

but in waveforms in which there is a very rapid rise or fall in voltage the spot may be required to move so rapidly that the trace is apparently missing. Fig. 42 gives an example of a trace of this type. The missing portion of the trace can be made visible by increasing the intensity of the spot, but this usually means that the image is too brilliant in the slower-moving portions of the trace and, in any case, if the brilliance control is set too high, the focus may suffer. Circuits are sometimes used whereby the modulator of the tube is supplied with a voltage proportional to the rate of movement of the spot, so that the brilliance is automatically adjusted over the whole trace, but for all normal work this refinement is not necessary.

Synchronism

In the majority of applications the waveforms under examination are recurring. Consequently the time-base is arranged to operate at an exact sub-multiple of the frequency of the waveform under test, which results in the successive traces lying exactly on top of one another, giving the impression of a stationary image. With a 50-c/s wave, for example, the time-base could be set to run at exactly 25 c/s, which would result in exactly two complete waves appearing on the screen. If the time-

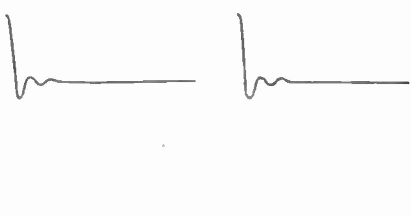


FIG. 42.—EXAMPLE OF WAVEFORM WITH PARTS "MISSING" OWING TO RAPIDITY OF SPOT MOVEMENT.

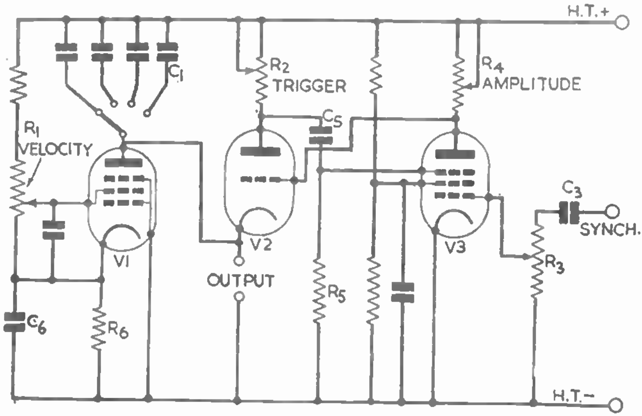


FIG. 43.—THE PUCKLE TIME-BASE, EMPLOYING "HARD" VALVES: A POPULAR ARRANGEMENT FOR GENERAL-PURPOSE OSCILLOSCOPES.

R2, 2 k (max.); R3, 0·5 M; R4, 0·25 M; R5, 1 M; C3, 0·1 μF; C5, 0·1 μF; V1, screen potential, 20–150 V.

base frequency is not an exact sub-multiple, the successive traces will lie slightly displaced from one another so that the image will appear to drift through the screen in one direction or the other, while, of course, if the relationship is nowhere near correct, the result will only be a confused pattern of traces.

It is impracticable to adjust a time-base to operate so precisely that its frequency remains an exact sub-multiple of the work frequency, even if the work frequency itself were constant, which does not necessarily follow. Arrangements are therefore made to synchronize the time-base with the work. This is done by injecting a small proportion of the work voltage into the time-base discharge valve. The circuits used are such that the point at which the discharge valve becomes conducting is controlled by the voltage on the grid of the valve, and the setting is normally chosen such that the time-base discharges when the spot has moved approximately across the whole screen. If the time-base is already in a condition where it is approaching the point at which it will discharge, the injection of a small pulse of voltage will in fact cause the discharge to take place. All that is necessary, therefore, is to arrange the time-base to operate at a frequency just slightly slower than the correct frequency and to superpose on the grid of the discharge valve a small voltage derived from the waveform under examination.

This will have the effect of causing the time-base valve to discharge at the correct point (once every two cycles in the example previously given), so that the time-base is automatically maintained in operation at a correct sub-multiple.

In practice, this is accomplished very easily by a mere alteration of the setting of the time-base control. Provided a suitable synchronizing voltage has been fed from the work into the time-base, it will then be

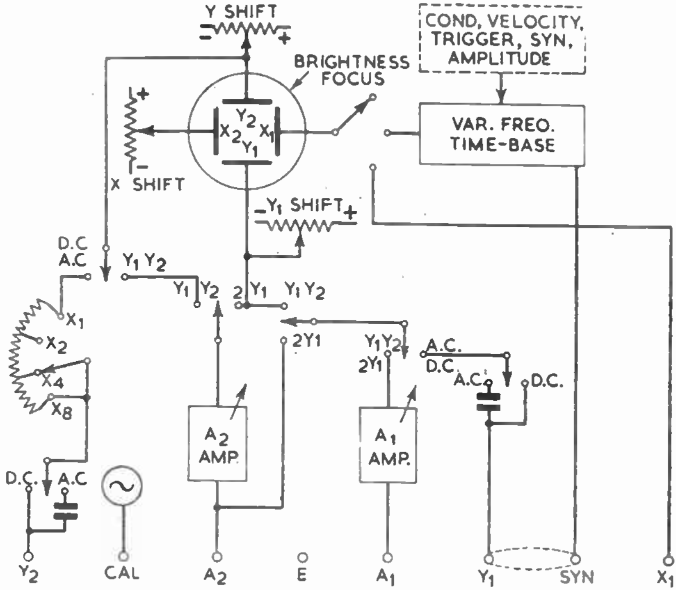


FIG. 44.—SWITCHING AND INPUT CIRCUIT OF A TYPICAL OSCILLOSCOPE.

found that as the frequency control is altered the time-base locks at a series of sub-multiples giving two, three, four, etc., complete waveforms, on the screen

Control of Signal Voltage

Turning now to the vertical deflection, the requirement here is that the Y plates shall be supplied with a voltage proportional to the effect to be examined. If this is already an electrical voltage the problem is simply a matter of ensuring that the voltage is of the right order. If it is too large, so that it would deflect the spot off the screen, it must be attenuated by a suitable network. If it is too small to produce a satisfactory image it must be amplified.

If the effect under examination is not electrical, it must be converted into a voltage by means of some suitable signal converter.

Currents can be measured in two ways. One is to pass the current through a small non-inductive resistance and to observe the voltage developed across it. This is the most common arrangement, and is quite convenient if the oscilloscope incorporates an amplifier. The alternative is to use magnetic deflection, the current being passed through deflecting coils suitably located around the neck of the tube. Some instruments are provided with magnetic deflecting coils for this purpose.

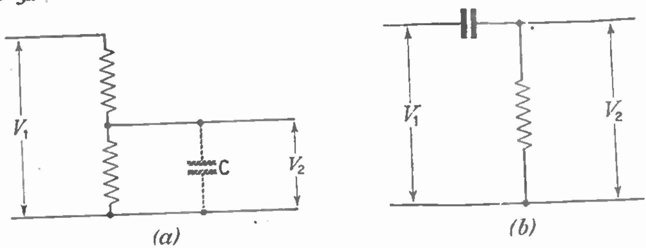


FIG. 45.—ATTENUATING NETWORKS SUBJECT TO FREQUENCY DISTORTION.

Avoidance of Distortion

It is impracticable here to discuss in detail the design of deflection amplifiers. Most oscilloscopes are provided with amplifiers which have been very carefully designed to give faithful reproduction over the range for which they are designed, both as regards amplitude and phase.

It may be noted here that in oscilloscope technique frequency response, in the accepted sense, is of secondary importance. It is the phase displacement which the signal suffers in its transit through the amplifier which matters, and a negligible phase displacement implies a much higher frequency response than would be required from a normal amplifier.

It should be remembered, moreover, that it is quite possible to introduce distortion in the method of connecting the oscilloscope to the work, and the user should always satisfy himself that his method of connection is such as to obviate any error of this sort, and also make sure that he is really looking at what he thinks he is. It is always useful, having connected the circuit, to switch off the signal itself, leaving the circuit otherwise undisturbed. The trace on the screen should then be a horizontal line (assuming the time-base is still running). It is sometimes even advisable to repeat this process on the vertical axis by switching off the time-base and making sure that the trace on the screen reduces to a single spot. It is quite possible to have some spurious signal injected into the horizontal deflecting system which would, of course, introduce queer distortions into the waveform which are not in fact actually present.

Coupling or attentuating networks can also introduce distortion. For example, if a potentiometer is being used to feed the voltages to the oscillograph as shown in Fig. 45 (*a*) the ratio of V_2 to V_1 will fall as the frequency increases because of the inevitable stray capacitance across the input to the oscilloscope. This would have the effect of suppressing harmonics and making a waveform appear smoother than it actually is. Conversely, if the network of Fig. 45 (*b*) is used, the lower-frequency components will be attenuated. These effects will indeed be familiar to the circuit engineer, but it is as well to mention them, because their effect may be overlooked.

Mains hum or other spurious signals can be picked up on the leads to the oscilloscope, particularly if these are long and appreciably separated. Hum voltages will be superimposed on the work, the effect produced

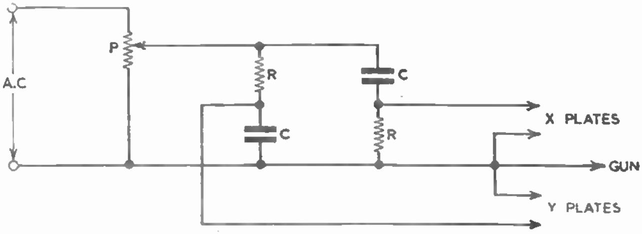


FIG. 46.—ESSENTIALS OF A CIRCULAR TRACE CIRCUIT.

If $R = 1/\omega C$, the voltage on the X plates leads the input voltage by 45°, whilst that on the Y plates lags by the same amount, giving the required 90° phase difference. The diameter of the circle is controlled by P. Since it is difficult to make the capacitances exactly equal, the resistors should be made variable over a small range.

depending on the relative frequencies of the work and hum voltages. The image may wobble up and down, or there may be a confusing multiplicity of traces. Cutting off the signal but leaving the remainder of the circuit in operation may disclose the presence of hum, but there is no hard-and-fast rule, and it is often a good plan, if any doubt exists, to try to observe the effect required in more than one way to see if similar answers are given.

Transient Phenomena

The applications so far considered have all been recurrent, but the use of the oscilloscope is by no means limited to such phenomena. The same technique can be used to depict transient impulses. In this case a "single-sweep" time-base is used, in which the spot is moved across the screen but then remains deflected until it is restored to normal. The actual trace may be too swift to be observed visually, in which case it must be recorded photographically, or by using a delay screen which retains the image for an appreciable time.

For photographing the trace, cameras are available which focus the image on the screen on to a film or plate. The shutter is opened just prior to the arrival of the transient and closed afterwards. The film or plate is then developed in the normal way.

Radial Displays

The rectangular display is not the only possible arrangement. Frequent use is made of the circular trace or polar display. A magnetic deflecting system can be so arranged as to move the spot outwards from the centre to an extent dependent upon the magnitude of some signal. The actual direction of the movement will depend upon the position of the deflecting system, and if this is caused to rotate, it becomes possible to arrange a display in polar co-ordinates such that the angular position of the spot is proportional to the primary variable, while its distance from the centre is determined by the dependent variable. This is a form of display much used in radar, but it has industrial applications.

It is not necessary to have a mechanically rotating magnetic deflecting system. The same result can be produced by suitably related voltages applied to two pairs of deflecting plates at right angles. If two alternating voltages of the same frequency are connected to the two pairs of deflecting plates the result will be a diagonal line. If, however, there is a phase difference between the voltages, the go and return traces will not coincide, and the resulting trace will be an ellipse. If the phase difference is made exactly 90°, and if the maximum deflections in both horizontal and vertical directions are made equal, the result is a circle. Fig. 46 shows the essential features of a circular-trace circuit.

Control of Spot Intensity

A circular display can be used in various ways.

One method is to control the intensity of the spot by varying the grid voltage. The brightness control may be set so that the spot is normally just not visible. The arrival of the signal to be investigated is caused to increase the brilliance momentarily, so that the spot becomes visible for a brief instant at a position depending upon the relative time. Suppose, for example, that it were desired to examine the relative regularity of a series of impulses arriving approximately fifty times a second. The circular trace would be arranged so that the spot rotated exactly fifty times a second. Every time a signal arrived the spot would become momentarily visible at some position around the circle, and provided the successive impulses all arrived at exactly $\frac{1}{50}$-second intervals, the position at which the spot appeared would always be the same, so that the spot would appear to be stationary. If the signal frequency were slightly greater or less than 50 c/s, the spot would appear to move around a circle in one direction or the other, while if the impulses were arriving in an erratic manner it would appear to dance from side to side.

Radial Line Display

Alternatively, instead of causing the spot brilliance to be varied, it is possible to change the radius of the circular trace momentarily, by altering the voltage applied to the deflector plates. For example, the voltage could momentarily be cut off altogether, in which case the spot would return to the centre and then immediately fly outward to its original position when the voltage was restored. The effect of this would be to produce a radial line, the relative position of which could provide the information required.

* * * *

Information on test equipments specially developed for the servicing of television receivers will be found in Section 40.

39. RADIO RECEIVER INSTALLATION AND SERVICING

39. RADIC RECEIVER INSTALLATION AND SERVICING

INSTALLATION NOTES

While the positioning of the receiver will usually be claimed as the privilege of the owner, the installation engineer should be prepared to point out, tactfully, the disadvantages of unsuitable locations. In particular, the receiver should not be installed in small, unventilated recesses or in unduly warm or damp positions. Whenever possible, positioning should be made to allow freedom of air circulation on all four sides of the cabinet. Attention should also be paid to the accessibility of the manual controls and to the ease with which the tuning dial can be read without parallax errors. High shelves and low tables are unsuitable in normal circumstances.

The mains-supply socket should be fused for not more than 2 amperes; if the only suitable point is of a higher rating than this, the use of a 2-ampere fuse plug is recommended, unless mains fuses are incorporated in the receiver. Mains leads should be so arranged that the minimum of movement will be necessary in the course of domestic cleaning operations.

Before connecting the receiver to the mains, check the type of supply, whether A.C. or D.C. and the voltage under normal load : the word of the resident should never be accepted in lieu of such a test. Supply voltages even in the same district often vary considerably, as does the type of supply.

The plug of an A.C. mains-operated receiver should always be tried in both positions to see which gives the least hum. If three-pin sockets are not fitted in the room, a mark should be made on the plug and socket and its significance explained to the user. With D.C. mains, of course, only one position is possible. Where no results are obtained on first switching on, the plug should be reversed in the mains supply socket.

Make sure that all valves and scale lamps are firmly seated in their sockets, the valve clamps, if fitted, are properly in place and that all top-cap connections are securely fastened. Each valve should be tapped lightly to ensure that there are no internal loose connections or intermittent short-circuits.

For general coverage in areas not subject to bad radiated interference, an outdoor aerial of between 50 and 100 ft. in length (including lead-in), erected at least 20 ft. from the ground, will generally be satisfactory for reception throughout the range 10–2,000 m.

Information on the types of aerial and their erection is provided in Section 21.

While the importance of the earth connection has, to some extent, decreased with the advent of high-sensitivity receivers and better smoothing filters, it should be remembered that, particularly on long wave, it forms a valuable part of the aerial matching system. as well as affecting the interference and hum levels of mains-operated receivers. An efficient earth is also an important safety factor for all mains-operated receivers.

A rising-main water-pipe is suitable, provided that its surface is thoroughly cleaned, after which the connection should be made by means of a tightly fitting clamp. Gas, hot-water or drain pipes should not be used, nor should any connection be made to an existing telephone earth.

More detailed information on earths is given in the section on " Earthing ", Section 21, pages 30–32.

SERVICING METHODS AND EQUIPMENT

The speedy tracing of faults in broadcast receivers and allied electronic equipment depends basically upon the systematic location of, first, the section or stage of the receiver that is defective, and then the particular valve(s) and/or component(s) concerned. This is often accomplished by the progressive elimination of the unaffected sections of the circuit, involving—in some instances—many separate tests. With experience, however, many of the preliminary checks can be carried out by the intelligent use of eyes, ears, nose and touch.

In these tests the service engineer must call into play his knowledge of radio fundamentals, the test equipment at his disposal, his past experience of similar faults and whatever data on the particular model that is to hand. His procedure will be largely governed by his test equipment, but will, in almost all cases, be compounded of voltage, current and resistance measurements, the substitution of suspected valves and components by others of known goodness, and the use of a substitute signal such as that provided by a signal generator, or, as it is often called, a service oscillator.

The limitations, as well as the uses, of test equipment should be borne in mind when trouble tracing. For example, for the reasons shown in Fig. 1, a low-sensitivity testmeter may give misleading voltage readings. Again, although valve testers are of undoubted assistance in the showing up of defective valves, they should not be considered

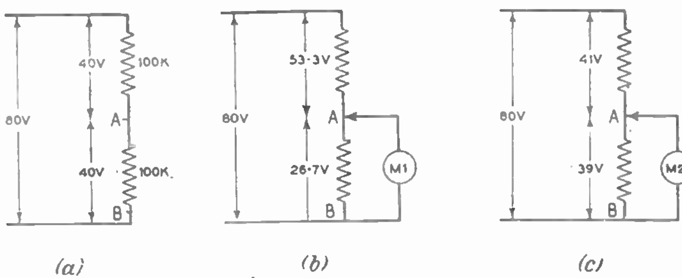


Fig. 1.—Effect of Meter Resistance.

The above series of illustrations shows that when measuring high-resistance circuits, a voltmeter of higher sensitivity gives more accurate readings. (*a*) The actual voltage conditions existing in a simple potential divider circuit when no voltmeter is present, showing 40 volts between points A and B. (*b*) A 1,000-ohms/volt meter (M1) on a 100-volt range (total resistance 100,000 ohms) would indicate 26·7 volts when connected between points A and B. (*c*) A 20,000-ohms/volt meter (M2) on a 100-volt range (total resistance 2,000,000 ohms) would indicate 39 volts between A and B.

infallible : valves that pass muster when tested on an instrument, may nevertheless be unsuitable for particular applications ; a fact that will be found only by substitution of a good valve while measuring the output of the receiver.

Faults

Before considering practical servicing procedures, it is first necessary to review briefly some of the faults which most frequently occur in standard broadcast receivers and the effects which they give rise to.

Defects in radio equipment can be roughly classified into two overlapping groups : (1) mechanical faults ; and (2) electrical faults.

Purely mechanical faults, which are often self-evident, most frequently occur in the tuning circuits ; cord-drives, for instance, sometimes break, and other manual controls and switches are subject to considerable mechanical wear. Loosening of the cement used to fasten the valve envelopes to their bases or to the top caps is a common occurrence in old receivers. Extensive physical damage to a chassis or cabinet, caused, for example, by the accidental dropping of a receiver may also be met with.

More numerous are the mechanical/electrical faults where mechanical failures adversely affect the electrical circuits. These are frequently not self-evident, and can often be traced only with the aid of correct servicing procedure. Short-circuiting of variable capacitors, either over their entire span or in one or more positions ; open-circuits in battery, power, aerial, earth and loudspeaker leads ; short-circuits caused by the accidental bending of wires, the presence of metallic dust or other foreign bodies ; physical damage to the cones or speech coils of loudspeakers ; noisy potentiometers caused by wear or dirt ; and deterioration or damage to valve sockets and valve pins ; these are frequent causes of failure.

There are also many possible electrical and mechanical/electrical defects in the basic components. These include :

Valves : heater failures ; loss of emission ; ionization (producing " soft " valves) ; breakdown of inter-electrode insulation, including heater/cathode leakage ; fracture or disconnection of the leads between electrodes and valve pins ; microphony (sensitivity to mechanical vibration).

Resistors : open-circuits caused by excessive power dissipation or deterioration ; development of high or low resistances at wide variance from rated values.

Capacitors : complete or partial failure of insulation producing short-circuits or D.C. " leakage " ; fracture of internal leads ; loss of insulating wax as the result of overheating ; " ageing " of electrolytic types producing effective capacities well below rated values and an excessively high power factor ; poor contacts between the rotary plates and outer sections of variable types.

Inductors, Transformers : disconnection or breakage of windings by excessive power dissipation or faults in the wire ; short-circuiting of adjacent turns or sections ; loose turns causing the effective inductance to vary.

Switch Contacts : high resistance or open-circuit caused by dirt, corrosion or distorted contact springs. The on/off switch is a frequent source of trouble with battery models.

Insulating Materials : D.C. or radio-frequency leakage across " insulating " materials may be caused by moisture, metallic dust or the decomposition of electrolytic capacitors, or the use of unsuitable materials.

The deterioration of radio components takes place comparatively rapidly in hot and humid atmospheres or where the receiver is kept in a damp place, and for this reason equipment intended for use overseas often requires special treatment. Salt-laden atmospheres may also increase the rate of deterioration owing to electrolysis occurring. Components thus affected often show characteristic green spots, as do soldered contacts. The widespread adoption of " tropicalization " during and after the Second World War has led to much greater reliability of certain radio components.

There are occasions when certain inherent defects or shortcomings in the original design of a receiver may only become noticeable, for example, where the set is used in particular localities or when the wavelength of the local transmitter is changed. Poor " image " rejection; intermediate-frequency breakthrough; inefficient shielding; lack of selectivity and sensitivity may be met with in some areas and under certain conditions. Such faults often require modification of the original design.

Incorrectly " coded " resistors and capacitors sometimes slip through manufacturing tests and give rise to trouble in the receiver after it has been installed.

Effects of Faults

It has been shown in the preceding paragraphs that almost all components and valves which together go to make up a receiver are liable to develop faults. Their effect on receiver performance may take the form of :

(1) the clearly apparent fault which cannot be ignored;
(2) a gradual falling off in the standard of performance;
(3) the intermittent fault which may re-occur only at widely spaced intervals;
(4) the type of fault which disappears immediately any component is touched.

Symptoms also vary considerably and fall, roughly, into the following classifications :

(1) no output;
(2) distorted output;
(3) excessive hum;
(4) whistles, " motor-boating ", " ringing ", " howling " and excessive interference between stations;
(5) crackles, noise, excessive fading, " blasting ";
(6) overheating;
(7) no signals on certain wave-ranges, or portions of wave-ranges, " blind spots ";
(8) lack of output;
(9) lack of sensitivity.

The radio service engineer should have a thorough grasp of the more common origins of these symptoms; though, at the same time, he should constantly be on the alert for unexpected and " rare " faults.

NO OUTPUT. Complete absence of signals may be due to almost any of the basic faults already listed. Systematic checking to localize such faults is described later under "servicing procedures". The more frequent causes include the failure of: one or more valves; electrolytic smoothing capacitors; power and output transformers; smoothing resistors and chokes; power supplies, power transmission leads and open-circuit fuses (not necessarily blown); voltage-dropping resistors (particularly where the dissipation of power is high); "line cords", barreters and pilot lamps in A.C./D.C. ("universal") receivers. The choking of valve control-grids due to over-biasing, damage to loud-speaker speech coils, fracture of aerial leads and switch faults are also common causes of a complete absence of signals.

DISTORTED OUTPUT. Distorted output may be due to wrong bias voltages; faulty anode and screen resistors; short- or open-circuit by-pass capacitors; loss of emission in valves; "soft" valves; leaky A.V.C. lines; misalignment of push-button and other tuning devices, misalignment of tuned circuits; parasitic oscillation, particularly in audio-frequency stages; and leaky inter-stage coupling capacitors causing positive biasing.

EXCESSIVE HUM. Most A.C. and "universal" receivers possess a residual mains hum, but in a quiet room this should not be audible at approximately more than 1 ft. from the loudspeaker. The level may, however, be considerably increased by: open-circuit or "ageing" smoothing capacitors; transposition of mains-input plug ("universal" types); external pick-up of stray mains hum; absence or failure of earthing system; short-circuit of windings or leakage of power transformers; defective valves, especially where the cathode/heater insulation has completely or partially broken down; failure of grid resistors; open-circuit or resistance of switch contacts; open-circuit of grid inductors; and the transposition of connections to the loudspeaker hum-bucking coil.

WHISTLES, MOTOR-BOATING, RINGING, ETC. A tuneable whistle on all signals usually indicates intermediate-frequency instability, which may be due to: failure of intermediate-frequency screen or anode by-pass capacitors; excessive gain of intermediate-frequency stage resulting from failure of the intermediate-frequency bias resistor or capacitor, or from a too sharply peaked alignment of the intermediate-frequency transformers; batteries which have developed a high internal resistance; poor screening; stray capacity or inductance coupling; and poor lay-out of components.

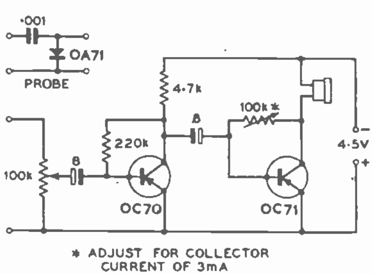


FIG. 2.—TRANSISTOR SIGNAL TRACER.

This is a two stage amplifier for A.F. signal tracing. With the crystal diode probe it can be used for I.F. and R.F. circuits.

Constant howling or "motor-boating" usually denotes audio-frequency instability, although blocking of the local oscillator, loose valve bases, valve microphony and poor voltage regulation can also cause this trouble. Audio-frequency instability may arise from: failure of anode or screen by-pass capacitors; batteries with high internal resistance; stray coupling.

"Ringing" usually denotes valve microphony, but occasionally it may be due to vibration of the oscillator tuning capacitor.

Excessive interference may be caused by: misalignment of the tuned circuits or poor tracking of the gang capacitor; incorrect adjustment of the intermediate-frequency wave-trap; faulty intermediate-frequency transformers; "spurious" responses, denoting incorrect alignment. poor screening or excessive radio-frequency output from the local oscillator; high resistance of switch contacts. Reception of Morse signals from ship stations operating in the 600-m. band usually denotes intermediate-frequency break-through. Reception of medium-wave stations on the long-wave band may be due to insufficient "image" rejection.

CRACKLES, NOISE, EXCESSIVE FADING. Crackles and noise may be caused by external sources of interference or faults within the receiver. Partial breakdown of insulation in capacitors, inductances, intermediate-frequency and audio-frequency transformers; faulty resistors; dry joints; loose or high-resistance switch contacts; loose valve pins and sockets; faulty power transformers and smoothing chokes; dirty or faulty potentiometers can all give rise to this type of symptom.

Excessive fading or "blasting" can usually be traced to faults in the A.V.C. network. The breakdown of resistors and capacitors should be suspected, although the possibility of faults in the aerial or aerial-input circuit should not be completely ignored.

OVERHEATING. Possible causes include: short-circuit or partial failure of capacitors, especially those employed for smoothing; shorted turns or sections of windings in power transformers; short-circuits in heater wiring; faulty bias resistors or capacitors, particularly in the output stage.

NO SIGNALS ON SOME WAVELENGTHS. May be caused by the rotary vanes of the tuning capacitor touching the stationary ones; loss of emission in the local oscillator valve; fractures or short-circuits in inductor windings; poor contacts on wave-change switches; misalignment or poor tracking of the tuning gang capacitor; excessive aerial coupling producing "blind spots" (straight receivers).

LACK OF OUTPUT. May be caused by loss of emission in one or more valves, especially in the output stage; damaged loudspeakers (distorted speech coils, loss in field strength); faulty resistors or capacitors in valve inter-coupling networks; short-circuits producing an excessive drain on the power supply; partial failure of metal rectifiers.

LACK OF SENSITIVITY. May result from loss of emission in one or more valves; faults in the radio-frequency or mixer stage; misalignment or poor tracking of tuned circuits; faulty resistors or capacitors; leaky capacitors in the A.V.C. line, for example, can cause flat tuning and general loss of sensitivity.

INTERMITTENT FAULTS. Where any of the above symptoms occur intermittently, particular attention should be given to components subject to mechanical movement or vibration, such as potentiometers, switch contacts, aerial and power-supply leads. The effect of heat on valves, resistors and power transformers is another likely source of trouble. When testing a receiver known to suffer from intermittent faults, the symptom may sometimes be induced by temporary application of voltages about 10 per cent above normal, either by adjustment of the mains-input tapping or, in the case of A.C. models, with the aid of an auto-transformer. It is important not to remove or disturb the chassis until the fault has been heard, otherwise the trouble may be temporarily cleared, only to return after the receiver has been restored to its owner.

Electrolytics

When electrolytic capacitors have been out of service for an appreciable period of time, the insulation resistance falls sharply. For example, when a faulty electrolytic is replaced from stock, the initial insulation resistance on first switching on the receiver may be only a fraction of its normal value. This fact may give rise to a heavy initial current through the capacitor, heating the electrolyte, and this in turn will further decrease the insulation resistance; thus starting a chain of events which may lead to the complete breakdown of the capacitor, the rectifier or the power transformer.

For this reason, unless capacitors rated well above the circuit peak and surge maximum voltage are used, it may be advisable to "re-age" (re-form) any electrolytic capacitor that has been out of service for more than, say, three months. This may be done by placing the capacitor in series with two 15-watt electric lamps (series-connected) and a D.C. supply—not necessarily smoothed—about 20 volts above the working voltage of the capacitor. When first connected, the lamps will probably light, but after some little time—and certainly within 1 hour—they should cease to glow. A milliammeter may then be inserted into the circuit and the re-forming process continued for about 30 minutes after the current has dropped to below 1 mA.

A service department should arrange that electrolytic capacitors are used strictly in the order in which they have been stocked. It is also preferable to err on the generous side in regard to ratings and to use the largest type, physically, that will fit the space available, which should not be subjected to undue heat. It should be noted, however, that where an electrolytic capacitor has been used for any length of time at voltages lower than those at which it is rated, it may no longer be safe to apply the full rated value.

Servicing Procedures

Rigid servicing procedures, to be adopted on receipt of faulty receivers, cannot usefully be formulated, since so much will depend upon the symptoms reported by the owner of the receiver and the servicing equipment available. Investigated in a systematic manner, the majority of faults can be traced in a reasonably short time with a minimum of tools and instruments, although a full range of servicing equipment will save much time even during routine checking.

The following equipment should be regarded as essential:

A reliable general-purpose multi-range "universal" testmeter with an internal resistance preferably of not less than 1,000 ohms/volt.

A modulated signal generator.
An output meter.
A valve testmeter.
A supply of suitable meter and test leads terminated with prods and clips.
A non-metallic trimming tool.
An electric soldering-iron.
An assortment of screwdrivers in various sizes (lengths and blades).
A pair of long-nosed pliers (pendulums).
A pair of combination pliers.
A pair of side-cutters.
An insulated prod.

Before work on the receiver is started, all available service data and workshop records on the particular model concerned should be consulted. It is most important that the circuit diagram be carefully examined for unusual features before any power is applied to the receiver or any tests are made.

Preliminary Tests

Receivers reported simply as " not working " require a brief visual examination for obvious defects; for instance, fracture of valve envelopes, disarrangement of the tuning device, faulty manual controls. The presence of any considerable quantity of wax oozing from transformers, chokes and capacitors, or of badly burned resistors should be noted. It will also be advisable at this stage to remove at least the outer layer of dust and dirt from the chassis.

Unless the owner has reported the fusing of the mains power supply, the receiver may now be connected to the appropriate power supply and switched on, care being taken to ensure that it can be immediately switched off again should there be any sign of overheating or valve ionization (soft valves usually glow a bright pale blue or pink towards the base of the electrode assembly; overloaded valves often arc between anode and cathode). With battery receivers a quick check should always be made to ensure that there is no short-circuit across the H.T. supply.

When servicing A.C./D.C. receivers (or A.C. models usin A.C./D.C. technique), precautions should be taken against the risks arising from " live " chassis. A double-wound isolating transformer with a 1 : 1 voltage ratio should be inserted in the mains-supply leads where practicable. In other cases the output from instruments such as signal generators should be isolated by means of 1 : 1 ratio transformers or good-quality 0/1-μF capacitors. Unless a mains-isolating transformer is fitted, the receiver should never be tested where it is necessary to stand on a damp floor; it is also advisable to stand on a rubber mat whilst working on the receiver.

With the majority of mains receivers, the pilot or dial lamp should light immediately the receiver is switched on and the glow of the valve heaters appear within a few seconds. On A.C. receivers, if no light or heater glow is visible, the fault is probably in the mains-plug or leads, receiver fuses (if any), the primary of the mains transformer or a faulty voltage-tapping connection. But it may be that the heaters of all the valves have been burnt out through operating the receiver from incorrect power supplies. With A.C./D.C. receivers the valve heaters are usually wired in series, so that should one heater become open-circuited or a faulty connection develop in the valve socket, the other heaters will cease to function. In this case it is usual to check the heaters of each valve separately with the aid of a valve tester, ohmmeter or continuity meter. Pilot lamps, ballast valves (barretters) and mains voltage-dropping resistors also require checking, as they are usually in series with the valve heaters. The construction of certain valves

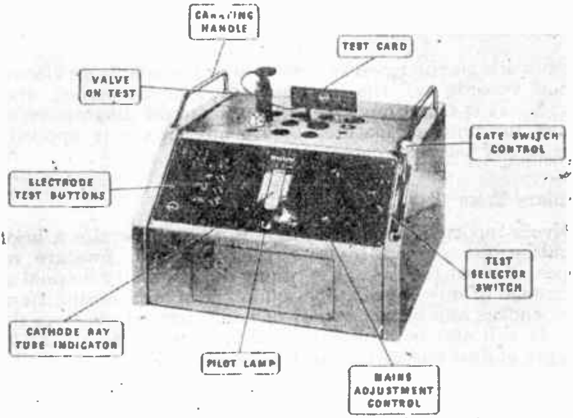


FIG. 3.—HIGH-SPEED ELECTRONIC VALVE TESTER.

All electrode potentials are automatically applied by the insertion of the appropriate perforated test card in a multiple-gate switch. The tests include: filament or heater continuity; electrode insulation with and without H.T. applied; heater-cathode insulation; grid current; emission.

(*Mullard Ltd.*)

(e.g., metal or shield types) prevents the observation of heater glow, but in mains sets at least, the valve envelopes become warm or hot when the heaters are on, and can be checked by lightly touching them. Where a valve is suspected of being faulty, another equivalent type should be tied in its place. An important exception, however, is the rectifier. Before substituting another valve in this position it is essential to check that no short-circuit exists across the main H.T. line. In A.C. models this can quickly be done by disconnecting the mains-supply leads, removing the rectifier valve and measuring the resistance between chassis and filament (or cathode in the case of indirectly heated rectifiers). This should be not less than 10,000 ohms, and will probably be considerably more; the actual value depends upon the screen voltage networks and any bleeder resistors. This precaution is necessary, since rectifier failures are often caused by a short-circuit on the H.T. line, and the substitution of a fresh valve before the original fault is cleared would lead to further damage. Similarly, it is a wise precaution to check the bias voltage and inter-stage coupling capacitor of any valve that has to be replaced, since the life of a new valve may be sensibly reduced if run with no bias or with a positive voltage applied to the control grid. Where a short-circuit exists, the electrolytic or paper capacitors used for smoothing should be examined; the presence of foreign bodies, such as small pieces of solder, stray snippets of wire, metal washers and so on, can also be a source of trouble.

Should the screen grid of the output valve become red hot, this will usually indicate that there is a failure in the anode circuit of this valve; likely faults include fracture of the primary winding of the output transformer or damage to the anode pin of the valve or its socket.

When testing components or valves by removal or substitution it is most important that only one change or test should be made at a time and that the circuit should be restored to its original condition at the completion of each test. Serious damage can be done by haphazard alterations.

Aural Tests

A valuable guide to the fault is sometimes provided by the kind of "noise" coming from the loudspeaker. Complete silence may also prove useful: for instance, it would suggest that a fault exists in the receiver power supply (H.T. secondary of power transformer, smoothing choke, smoothing resistor or smoothing capacitors); the loudspeaker or its associated leads; or the output stage. To test the loudspeaker, disconnect the receiver from the mains supply and measure the resistance of the primary of the output transformer with a multi-range testmeter; the reading should be about 400 ohms, and a faint click may be heard as the meter is connected. The power supply can most easily be checked by measuring the voltage across the main H.T. line and between chassis and the anode of the output valve.

A slight mains ripple or valve noise would suggest that the power supply, the output valve and the loudspeaker are all functioning, though not necessarily correctly. Rustling, crackling and an increase in noise when the volume control is turned towards maximum output would suggest that the audio section is operating. In the case of A.C. or battery receivers, this can easily be confirmed by touching, with a finger, the grids of the output and audio-amplifier valves. The audio-amplifier valve will often be the triode section of a double-diode-triode or the pentode section of a double-diode-pentode. A loud hum or howl will probably result from contact with the grid of this valve, while a "plop" or dull hum can be expected from touching the grid of the output valve. This test is not advisable with D.C. or A.C./D.C. receivers, where the grids may be at high potential to earth. A similar, though reduced, response can be obtained by touching the valve-grid pins with the metal end of an insulated screwdriver.

Touching the grids of the intermediate-frequency and mixer valves may also provide a rough indication as to whether or not those stages are working; but the hum thus produced may not be very loud. The audible effects, if any, of rotating the wave-change switch and of manipulating the tuning control should also be noted; clicks or rustling noises would tend to suggest that the fault lay in the radio-frequency section. A quick check of the frequency-changer valve by substitution is usually worth-while, since loss of emission and consequent failure of the local oscillator is common.

Measurements

All the tests so far described can be quickly carried out and require only the minimum number of instruments; and in a surprisingly large number of instances will result in the location of the fault. But a more detailed examination of valves and components may prove necessary. This normally means: (1) valve testing and the measurement of voltages, currents and resistor and capacitor values at selected points in the circuit; and (2) the tracing of broadcast or locally injected signals through the various stages.

FIG. 4.—"Avo" ELECTRONIC TESTMETER.

For the speedy performance of these operations, reliable test gear is of prime importance. Routine measurement of voltages, currents and resistances should be carried out with a high-grade multi-testmeter possessing an internal resistance of not less than 1,000 ohms/volt. Meters with a lower rating may give misleading readings. Where no loading of the circuit can be tolerated a valve-voltmeter or electronic testmeter must be used.

Check voltages under specified conditions are, in most cases, indicated in the appropriate section of this manual and in manufacturers' service literature. Deviation of up to about 20 per cent from published figures may be expected, and would not necessarily indicate a fault.

The first checks usually include the voltage of the main H.T. line and the voltage readings at the anode and screen-grid pins of the valves. A quick check of the oscillator section of a superheterodyne receiver can be made by connecting a voltmeter across the cathode resistor of the mixer valve and short-circuiting the oscillator grid to chassis: the voltage drop across the resistor should change appreciably as the valve stops oscillating.

The use of valve testmeters and valve analysers will enable the valves themselves and the currents at the electrodes to be checked.

The windings of transformers and inductors can be checked for open-circuit either with an ohmmeter or a continuity meter. Manufacturers' literature usually specifies the approximate D.C. resistance of the main components, but even where this information is not available, the experienced service engineer should be able to estimate whether the reading obtained is approximately correct. Resistance measurements may also reveal faults in wave-change switches, although occasionally misleading readings may result from the meter battery "breaking down" appreciable radio-frequency resistance. Faults in tuning circuits can usually be detected by means of a grid-dip oscillator, an instrument which indicates circuit resonances.

Combined capacitor analysers and resistance bridges which are simple to operate and which can be used to measure the full range of resistor and capacitor values likely to be encountered in routine servicing are available from a number of manufacturers. The majority of these instruments can additionally be used to ascertain the power factor and leakage of electrolytic and other capacitors. In general terms, capacitors used for filter and by-pass applications should not be less than about 66 per cent of their rated values, or between 90 and 110 per cent where tuned filter circuits are involved; for untuned filter and by-pass applications a considerable degree of excess capacitance is usually unimportant. Tuned radio-frequency circuits will usually require capacitors with close tolerances (better than ± 5 per cent).

The power factor of electrolytic smoothing and bias capacitors should not exceed about 40–50 per cent; where such a capacitor is also being used to by-pass radio-frequency signals, this figure must be reduced to about 10 per cent unless additional mica or paper capacitors are wired in parallel with the electrolytic capacitor to prevent regenerative coupling between the stages.

When testing capacitors of less than about 1,000 pF it is important to use short leads between the component and the bridge, otherwise inaccurate results may be obtained owing to the stray capacitance of the leads.

A simple check to ascertain whether the A.V.C. is functioning correctly consists of measuring the screen voltage of the intermediate-frequency valve under "no-signal" conditions, tuning in a strong carrier and noting the change in reading. With a standard mains-operated receiver the screen voltage should show an increase of the order of 10 volts.

Signal Tracing

Signal tracing is probably the most satisfactory method of locating the less-obvious faults, since it enables each stage or section of the receiver to be tested under working conditions. Simple forms include the injection of an audio signal (e.g., from a pick-up) into the audio stages, working backwards from the output stage; and the use of a crystal rectifier and earphones to trace an incoming signal from the aerial through the early stages of the receiver. Elaborate equipment, however, will permit of more accurate and speedy checking than is possible with the simple systems outlined above. For example, a signal generator can be of invaluable service, since it enables a modulated signal of either the intermediate or radio frequency to be injected into each radio-frequency stage in turn (starting with the demodulator), thus rapidly localizing a fault to one particular stage.

Alignment

It is not usual to check the receiver alignment until other routine tests have failed, since the tuned circuits seldom become so far out of adjust-

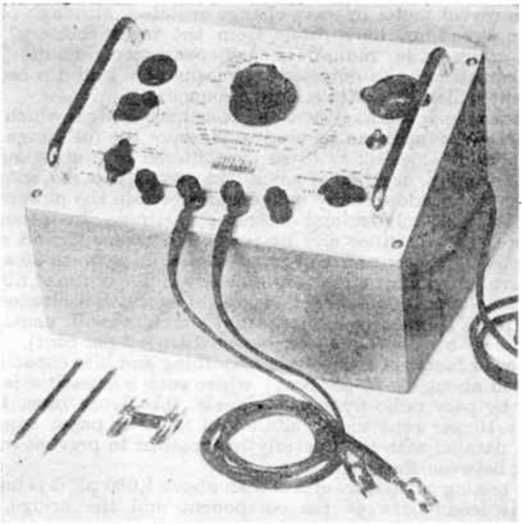


FIG. 5.—CAPACITOR ANALYSER AND RESISTANCE BRIDGE, TYPE CRB3.

This test bridge covers ranges 20 pF to 500 μF; 50 ohms to 100 MΩ. It measures capacitance by means of a Wien bridge; the resistance of all types of carbon and wire-wound resistors; the leakage resistance of paper and electrolytic capacitors and all types of insulation, 25–500 volts, by flashing neon. It directly indicates leaky, shorted, low-capacitance, high-capacitance and high-power-factor capacitors of both usual and intermittent types. Measurements are made directly, and no calculations are necessary.

(A. H. Hunt (Capacitors) Ltd.)

ment as to render a receiver completely inoperative. One exception, however, should be noted: over-enthusiastic owners do sometimes tighten up all "loose screws" in the hope of improving performance!

The alignment of short-wave ranges can easily be impaired by movement of any leads connected to the coils or other components at high radio-frequency potential; care should therefore be taken when handling the chassis or making tests to avoid disturbing such wiring.

Alignment faults can usually be detected most readily with the assistance of a signal generator. The procedure for re-alignment is described elsewhere in this manual. Where any of the radio-frequency or intermediate-frequency components or valves are changed, manufacturing tolerances and/or the disarrangement of wires usually make it necessary to re-align the receiver. This should also be done if sensitivity is below normal.

Partial Failures

Where it is suspected that capacitors or resistors have become open-circuited the temporary connection of a similar value component in parallel often permits a quick check to be made; particularly in cases of excessive mains hum. Complete removal of suspected components should, where possible, be avoided until the suspicion is confirmed. A sudden but not violent increase in mains-hum level may be due to a faulty earthing system, though modern receivers do not often suffer from inherent mains hum.

Leaky capacitors—a common source of "noise" and distortion—can be tested most satisfactorily with a "Megger" insulation tester or a component bridge, though it will usually be necessary to disconnect at least one connecting wire.

In cases of apparent modulation hum (i.e., tunable hum which appears as a background on stations but disappears when no station is being received) or crackles, the first step should be to ascertain whether the interference is being picked up externally by the aerial and mains-supply leads or whether there is an internal fault. Decrease of noise with the removal of the aerial will usually indicate an external source of interference. Internal sources of crackles and the like can often be located by the careful movement of components with an insulated prod. A plastic crochet hook is a useful accessory, since the hook enables wires to be firmly gripped in otherwise inaccessible spots. Each stage, beginning with the radio-frequency section, can also be examined in turn by temporarily short-circuiting the grid pins of the valve-holders to chassis, provided, of course, that reference to the circuit diagram shows that no damage could result. For example, removal of bias voltage from an output valve could seriously impair its emission.

A pronounced buzzing in the loudspeaker may be due either to the speech coil being out of centre or to loose turns in the coil. A similar effect can also be caused by distortion in the audio circuit.

Parasitic oscillation at frequencies above audible range may occur in the audio stages of a receiver and produce distortion. Where this fault is suspected the effect on the anode current of touching the grid should be observed; parasitic oscillation being denoted by a change in current.

"Ringing" and microphony can usually be traced by gently tapping each valve with an insulated prod. An elastic band wrapped round the end of the prod will make it easier to detect microphony. If the valves prove faultless, the oscillator section of the tuning capacitor should be similarly tested.

Radio-frequency switches, particularly where ganged wafer types are employed, are a frequent cause of noisiness and lack of sensitivity. Carbon tetrachloride or one of the commercial cleaning materials intended specifically for this purpose will usually cure the trouble. The cleansing agent can be applied with a small brush while rotating the switch. The repair of distorted contact springs requires dexterity; but can often be accomplished with the aid of a small pair of pliers or tweezers or the tools made for this purpose. A useful switch lubricant consists of two drops of fine oil in a teaspoonful of lighter fuel. The lubricant can be applied with a clean feather.

When the work of repair is finished, a routine check of valves and voltages should be made. This may prevent further and possibly more serious trouble later on. Resistors should also be carefully examined for signs of overheating and, if necessary, replaced by higher-wattage types. The chassis, cabinet and tuning dial should be thoroughly cleaned, and a little fine oil applied to any rotating mechanism such as the tuning drive.

After the receiver has been re-assembled in its cabinet, it should be subjected to a thorough test under normal operating conditions. Where possible, A.C./D.C. models should be operated from both A.C. and D.C. power sources, since certain faults (e.g., poor smoothing) may not otherwise be apparent.

Details of the fault should always be entered in workshops records. Carefully compiled records can save many hours of work by making clear the faults from which particular models are prone to suffer.

FIG. 6.—MODEL 130A INSULATION TESTER.

(Taylor Electrical Instruments Ltd.)

Performance Tests

It is, perhaps, after a receiver has been restored to something like normal working order that the need for the more elaborate test equipment is most keenly felt. In such circumstances the ear is an unreliable instrument for quantitative—and even qualitative—tests. Since the aim of the service engineer should always be to hand back the receiver with a performance fully equivalent to that for which it was designed, no receiver can be considered as completely serviced until instruments have confirmed beyond doubt that its calibration, sensitivity, selectivity and audio-frequency response are in no way inferior to those ruling on the day it left the factory; this is especially true of the receiver which has not received attention for a considerable period, and in which gradual deterioration may have passed unnoticed by the owner until a complete breakdown forces him to take action. Though, unfortunately, it must be admitted that, owing to the time factor involved, comprehensive servicing and maintenance is not always feasible.

For full performance tests, essential items of equipment are a signal generator with calibrated attenuator and an audio-frequency output meter of an impedance suitable for matching to the receiver under test; in addition, a calibrated audio-frequency generator with near sine-wave characteristics, and an oscilloscope together with a sweep oscillator (" wobbulator ") suitable for the signal generator in use, will facilitate the carrying out of more stringent receiver tests.

Sensitivity

With the aid of the signal generator and output meter it is possible to check the sensitivity and calibration of the receiver at various points throughout its wave-range. The need for frequent checking of the calibration of the signal generator itself, preferably against a crystal substandard, or transmissions of known frequency, should not be overlooked. Where sensitivity figures are supplied by the manufacturer these are normally based on an audio output of 50 mW and a 30 per cent modulated test signal. Should any general lack of sensitivity be discovered with the signal generator connected, via a standard dummy aerial, to the aerial socket of the receiver, the cause may rapidly be traced to the offending stage by injecting the signal in turn to: (1) the control grid of the frequency changer, at signal frequency; (2) the control grid of the frequency changer, at the intermediate frequency; (3) the control grid of the intermediate-frequency amplifier, at the intermediate frequency; (4) the control grid of the 1st audio-frequency amplifier, at the audio modulation frequency; (5) the control grid of the audio power amplifier, at the audio modulation frequency. For operations (2) to (5) the signal should be injected via a 0·1-μF capacitor. To obtain the standard output of 50 mW from a typical A.C. receiver it would be necessary to increase the input from approximately 40 mV to some 1·5 volts for operation (5), the intermediate steps depending upon the voltage gain of each stage. In practice, of course, it is unlikely that there will be sufficient output from the generator to permit the standard output to be obtained in the later operations, but after a little experience with his instruments on various types of receivers the service engineer soon learns to determine immediately whether or not a particular stage is " pulling its weight ".

Selectivity

The recognized method of defining the selectivity of a broadcast receiver is the amount off tune, in kilocycles, that a 30-per-cent-modulated carrier must be moved in order to reduce the power output of the receiver by one-half (3 db) from the figure attained when the test signal is exactly tuned in. When making tests with a signal generator and output meter to determine the selectivity of a receiver, the signal input must be kept below the level at which the A.V.C. circuits become operative, or alternatively the A.V.C. must be disconnected.

Incidentally, in checking the performance of a stage, the limitations of the normal valve tester become apparent. While such instruments are valuable for detecting a serious loss of emission or changes in mutual conductance, performance checks frequently show that valves which pass these tests without difficulty are not always satisfactory under actual working conditions: and where stage gain is found to be low, or distortion persists, it is as well to test the valve concerned by substitution while measuring the output of the receiver. It should be noted that where similar types of valve are used for different applications in a receiver—and more particularly in a television receiver—their positions should not be changed indiscriminately, since the condition of a valve may make it unsuited for certain applications, and inter-electrode capacitances may differ slightly.

Audio-frequency Response

Where an audio-frequency generator is available, and can be employed to modulate the signal generator, the overall audio-frequency characteristics of the receiver can be checked. Should there be no facilities for applying an external source of modulation to the signal generator in use, then the audio-frequency generator must be connected directly to the first audio-frequency amplifier, and the tests will then cover only the audio section of the receiver. If the receiver under test has a high degree of selectivity, the overall audio-frequency response will probably be considerably more restricted in frequency range than where the audio section alone is concerned.

Useful aural tests of the loudspeaker can be carried out with an audio-frequency generator by varying the generator frequency while listening and watching the loudspeaker for indications of excessive resonance effects. In the absence of an audio-frequency generator, the special frequency test records marketed by several of the leading gramophone companies may be employed, provided that this is done in conjunction with a pick-up of known characteristics. Common loudspeaker faults giving rise to uneven response include : off-centre speech coil, torn cone, loose speech-coil wires, dust between the speech coil and the field poles, loose parts and sympathetic vibration of nearby components and cabinet fittings.

Oscilloscopes

A well-designed oscilloscope, incorporating an efficient vertical amplifier, and linear, wide-range time-base, is an invaluable asset to speedy servicing and, when used in conjunction with other test instruments, in the adjustment of almost any piece of electronic equipment. Instead of laboriously having to plot graphs based on a series of individual measurements, the service engineer can see at a glance the waveforms present in a receiver and observe immediately the precise effects of adjustments. For radio work, in order to obtain maximum benefit from an oscilloscope, a swept oscillator (" wobbulator ") and a source of audio tone with good sine-wave characteristics are also required.

The main uses of an oscilloscope in radio servicing are : (1) for visual alignment, facilitating the setting up of any desired band-width; (2) as a means of detecting the presence and tracing the source of audio-frequency and intermediate-frequency distortion, parasitic oscillations and hum; (3) for the measurement of A.C., audio-frequency and radio-frequency peak voltages. Taken together, these facilities make it possible to trace the progress of a signal through each stage and to determine the exact point at which any unwanted characteristics are introduced.

To the service engineer not experienced in the use of an oscilloscope, the array of control knobs may at first sight appear formidable, but their use can rapidly be mastered. " Brilliance " and " focus " have the same significance as in television practice, though it should be noted that electrostatic focusing and deflection are normally used. "X-shift" and "Y-shift" are for positioning the trace on the screen of the cathode-ray tube by altering the bias voltages applied to the deflecting plates, "X" referring to the horizontal axis, and "Y" to the vertical axis. The "coarse" and "fine" controls govern the setting of the time-base frequency, and the "amplitude" control alters the output, and hence the length of the horizontal sweep. The "sync" control is to enable the time-base to be synchronized with the wave-form under examination, so that the trace remains in a stationary position on the screen. In addition, there will generally be either a master switch or alternative input sockets, so that the signal may be connected to the "X" and "Y" plates, directly for D.C., and via isolating capacitors for A.C. (to remove any D.C. component); additional positions will normally provide for attenuation or amplification of the signal by fixed amounts.

FIG. 7.—MODEL 1039M PORTABLE OSCILLOGRAPH.

The single-stage amplifier covers a frequency range of 25 c/s to 1·5 Mc/s (30 per cent down) at a gain of 20, or from 25 c/s to 120 kc/s (30 per cent down) at a gain of 75.

(*Cossor Instruments Ltd.*)

For radio servicing, the time-base should cover the range of approximately 10–100,000 c/s or more, and the vertical amplifier should provide gains in known steps of up to about 500 times; a more limited range of amplification to external signals applied to the horizontal plates is also useful for frequency comparisons and phase diagrams. The frequency response of the vertical amplifier is of considerable importance, particularly if it desired to use the oscilloscope for television work, where a good response characteristic up to about 3 Mc/s is desirable. The amplifier should also be as free as possible from phase-shift and amplitude distortion, and have a low input capacitance and high input impedance, so that the oscilloscope may be connected across high-impedance grid circuits without unduly affecting their operation. Screened, low-capacitance cable, kept as short as possible, should be used between the connection to the receiver and the oscilloscope. It is common practice to connect either a small capacitor (about 5 pF or less) or a large resistor (1 MΩ or more) in the tip of the test prod or crocodile connector, so as to reduce to a minimum the disturbance caused to the circuits under test. As with almost all test equipment, poorly designed or constructed oscilloscopes may prove more of a hindrance than a help, and when choosing a commercial model it is as well to pay more attention to the quality and facilities incorporated than to the price of the instrument.

For visual alignment of the intermediate-frequency stages, a swept oscillator is required : when used in conjunction with a signal generator, this should produce a signal varying regularly about 15–25 kc/s above and below the intermediate frequency, with the repetition rate—usually about 25 c/s—synchronized with the cathode-ray-oscillator time-base. This frequency-modulated signal is injected into the grid of the frequency-changer valve, as in standard alignment practice, and either an intermediate-frequency signal taken—via a small capacitance—from a convenient point immediately prior to the demodulator valve, or, more commonly, an audio-frequency signal tapped-off via a large coupling capacitor from the intermediate-frequency filter capacitor or across the volume control and fed to the Y-amplifier. The local oscillator of the receiver should be rendered inoperative during alignment. Where the frequency sweep of the " wobbulator " is known, the exact band-width of the intermediate-frequency stages may be directly observed and any necessary adjustments made to the intermediate-frequency transformer trimmers to obtain the desired band-pass characteristics.

Drive Cords

The mechanical failure of tuning and control-drive systems is still a common fault, particularly where the receiver has received rough handling. Whilst their repair is generally a fairly straightforward task, occasionally this calls for a considerable degree of mechanical dexterity.

If full details of the drive system are not available, it is worth spending a little time in carefully tracing out the purposes of the various pulleys and springs before attempting to re-thread the system.

A useful accessory, when threading cords in inaccessible corners, takes the form of a steel rod with a hook at one end and a small two-pronged fork at the other; but care should be taken that all rough edges are removed.

Always use a good-quality cord that will not chafe or stretch easily : it will usually be found advisable to pre-stretch the cord by suspending it for several hours from a hook with a weight attached to the lower end.

A common fault, when repairing drives, is to wind too many turns around the drive spindle, causing the turns to ride up and bind, making the drive stiff and increasing back-lash. The correct figure usually lies between 1½ to 2½ turns, depending upon the particular layout of the drive system.

Where one end only of the cord is terminated in a spring, it will usually prove more satisfactory to begin threading from the fixed end. When tying off a cord, or making a loop, the use of a small tubular aluminium rivet, through which the cord is doubled back, and the rivet then securely pinched, has much to recommend it. The difficulty of tying a knot when the cord is on the short side is too well known to require emphasis.

ALIGNMENT

The performance of a modern superheterodyne receiver is, to a considerable extent, determined by the accuracy of the alignment of its various tuned circuits, and by its ability to hold such alignment throughout the track of its ganged tuning capacitors or permeability tuned inductors. In a standard superheterodyne circuit this involves the tracking of the local oscillator on each waveband with one or more tuned circuits at the signal frequency; the adjustment of the intermediate-frequency circuits to give the required response characteristics; and the adjustment, where fitted, of the intermediate-frequency wave-trap and image-rejector circuits.

It cannot be too strongly emphasized that optimum results are unlikely to be achieved by persons not possessing a sound understanding of the fundamental principles of superheterodyne reception: nor by those unwilling to recognize that the complete re-alignment of a receiver is a process that cannot be hurried. Unless adequate time is available for a thorough and complete re-alignment, involving the adjustment and re-adjustment of each circuit, it is usually better to leave all trimmers and cores severely alone; injudicious and haphazard "peaking" of circuits is more likely to result in a general deterioration of the quality of reproduction and "patchy" sensitivity than in any overall benefit.

Alignment of tuned circuits at radio frequencies is normally achieved by adjustment, at the higher-frequency end of the tuning span, of small variable capacitors ("trimmers"); and, at the lower frequency end, of dust-iron or brass cores in the inductors or variable capacitors ("padders") connected in series with the inductors. Present-day trends favour a combination of variable cores and fixed padders. Alignment and tracking by adjustment of the vanes of the ganged tuning capacitors and by the use of specially shaped vanes in the oscillator section is now seldom met with, except in older models.

Alignment of tuned circuits at the intermediate frequency is achieved by adjustment of either variable capacitors or variable cores.

Why Re-alignment is Necessary

Although alignment of the receiver is carried out by the manufacturer before it leaves the factory, re-alignment may be necessary:

(1) When certain components or valves are replaced.

(2) With the ageing of inductors and capacitors or the gradual accumulation of metallic dust.

(3) Where vibration or rough handling affects the adjustment setting of trimmers and cores, and the spacing of coil windings.

(4) Through the mishandling of manual controls, which may affect the tracking of the ganged tuning capacitors or the setting of the tuning indicator.

(5) Owing to changes in the relative position of the wiring or components.

(6) Where changes in the aerial system affect the loading of the radio-frequency input circuits.

(7) Where the alignment controls have been tampered with.

Of these, (1) is the most likely cause of serious misalignment, owing to the accepted tolerances in the manufacture of components—particularly inductances—and valves. The components which most directly affect the resonant frequencies of the tuned circuits include: grid and tuning capacitors; inductances and intermediate-frequency transformers; wave-change switches. Changes in the inter-electrode capacities of the following valves are also likely to affect alignment: radio-frequency amplifier, frequency changer, intermediate-frequency amplifier and demodulator. The lower the frequency, the less marked will be the effects of small changes of capacity. Small changes in the positioning of wires, however, are sufficient to destroy the accuracy of alignment on short wavebands.

Effects of Misalignment

Misalignment of radio-frequency and intermediate-frequency circuits can be recognized by:

(1) Lack of sensitivity over whole or part of the waveband ranges, often accompanied by an unusually high level of background noise.
(2) Inaccurate calibration of the tuning dial.
(3) Flat or "double-humped" tuning.
(4) Deterioration of quality due to uneven response of the intermediate-frequency and radio-frequency circuits.
(5) Prevalence of heterodyne whistles and poor selectivity.
(6) Excessive fading due to lack of signal voltage.
(7) Tunable whistles on all signals due to radio-frequency or intermediate-frequency instability.

Equipment

The equipment required for correct and accurate alignment of tuned circuits is as follows:

A modulated signal generator with a frequency coverage of 100 kc/s–30 Mc/s and fitted with a calibrated attenuator. A good instrument should be free of frequency drift, be unaffected by mechanical vibration (within reason), have a good range of attenuation, be adequately screened, possess a clear and accurately calibrated scale and a smooth tuning mechanism free of backlash and slip. The accuracy of calibration should be checked periodically against signals of known frequency.

A dummy aerial suitable for use with the signal generator.

An audio power-output meter (or sensitive A.C. voltmeter) with a range extending from a few milliwatts to several watts and affording matching to standard output impedances. Methods of alignment not requiring the use of an output meter are reviewed on pages 32–33.

An adequate supply of non-metallic trimming tools of various shapes and sizes.

A supply of soft wax should be to hand for the sealing of trimmers and variable cores; sealing-wax and quick-drying cellulose paint may also be necessary.

A 4 B.A. box spanner may be required for the adjustment of certain types of trimmers.

If a greater degree of calibration accuracy than is obtainable with a normal signal generator (about 1–2 per cent) be required, a crystal-controlled substandard or frequency meter should be used. Such instruments usually provide 1,000- and 100-kc/s check points with an accuracy of 0·01–0·02 per cent, and may also incorporate a multi-vibrator circuit for the provision of intermediate 10-kc/s points.

The use of a single- or double-beam oscillograph and a frequency-modulated ganging oscillator ("wobbulator"), although not essential, greatly facilitates alignment, particularly where bandpass circuits are involved, as they enable the effect of adjustments to be seen immediately.

Alignment Tests

Where misalignment is suspected, the signal generator should be connected through the dummy aerial to the aerial and earth sockets of the receiver. A suitably matched output meter is connected across the primary (via a 0·1-μF isolating capacitor) or the secondary windings

FIG. 8.—"AVO" SIGNAL GENERATOR.

A turret-switched signal generator giving continuous coverage, with optional modulation, from 50 kc/s to 80 Mc/s.

of the output transformer. It is usually more convenient to leave the loudspeaker connected. The sensitivity and calibration of the receiver may then be checked at various frequencies by injecting modulated signals, tuning the receiver to them and noting the outputs obtained for similar settings of the attenuator control on the signal generator. The input should be kept at a minimum, otherwise the A.V.C. circuits will come into operation. Three check points on each waveband will usually prove sufficient; that is, one each at the top, bottom and middle of the band. Marked differences in peak amplitude at these points will usually mean that re-alignment is necessary, as will any appreciable variation in the calibration accuracy. It should be noted, however, that where the calibration error is the same or nearly so at all points, adjustment of the tuning mechanism is indicated rather than re-alignment.

The intermediate-frequency response curve can be roughly checked in the following manner:

The modulated signal generator should be connected to the control grid of the frequency-changer valve and tuned over a range of about

± 25 kc/s from the intermediate frequency specified for the particular receiver under test until maximum output is obtained. The frequency of the signal generator should be found to be within about 5 kc/s of the recommended figure. Next the frequency of the signal generator is varied on either side of the maximum point until the output falls to one-half. When the intermediate-frequency circuits are correctly aligned the two points at which this occurs should each be the same number of kilocycles from the original setting.

Re-alignment

Re-alignment should never be attempted unless suitable service equipment is available; adjustment by ear should not be attempted, since only relatively large variations of output can be so detected.

Adjustment of intermediate-frequency transformer cores or trimmers should always be followed by complete re-alignment of the radio-frequency section. Intermediate-frequency circuits should be aligned to a frequency as close as possible (within 5 kc/s) to that specified by the manufacturer, otherwise considerable difficulties may be experienced in the correct tracking of the radio-frequency circuits. An exception may occur where local conditions make it essential to avoid particular frequencies.

Where possible, the re-alignment procedure recommended by the manufacturer should be carefully followed; particularly with regard to the order in which the various wavebands are adjusted. The trimmers are often common to more than one waveband, and accurate alignment will then be impossible unless the circuits are adjusted in the correct order.

Before re-alignment is attempted, both the receiver and the signal generator should be allowed to reach normal working temperatures. Ten to fifteen minutes are usually sufficient, but a longer period may be necessary where accurate S.W. calibration is required.

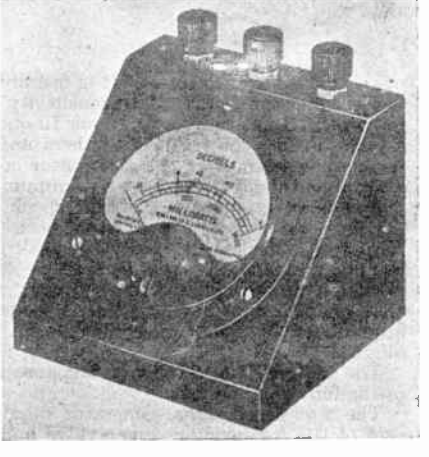


FIG. 9.—OUTPUT METER.

The use of a separate output meter for alignment and output measurements generally results in the freeing of a more expensive multi-testmeter for other purposes. This E.M.I. instrument provides a low range of 0–500 mW and a high range of 0–5 watts with decibel calibration zero levels at 50 and 500 mW respectively. The impedance is 5,000 ohms at 400 c/s; the meter is thus suitable for matching to the primary winding of most output transformers.

Standard Intermediate-frequency Alignment Procedure

With standard superheterodyne circuits the following procedure can normally be adopted.

Connect the output meter to the receiver as detailed under " alignment tests " (page 23).

Connect the output leads from the audio-modulated signal generator to the control grid of the frequency-changer valve (via a 0·01-μF isolating capacitor) and chassis, the outer (screened) lead being connected to chassis. Where the wiring to the control grid has to be disconnected, care should be taken to ensure that a D.C. continuity exists between grid and chassis; if necessary an additional 1-MΩ resistor may be connected between these points.

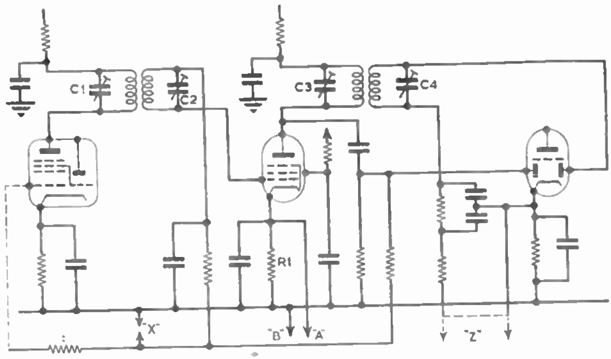


FIG. 10.—INTERMEDIATE-FREQUENCY STAGES OF A TYPICAL BROADCAST RECEIVER.

Short-circuit the oscillator section of the ganged tuning capacitors and tune the receiver to 550 or 2,000 m. (i.e., vanes of the tuning capacitors fully in mesh). Turn the volume control to maximum gain and the tone control to maximum high-frequency response.

Set the signal generator to the required intermediate frequency and inject a modulated signal just sufficient to give a reading on the output meter. The output from the signal generator should be kept at a minimum at all times during intermediate-frequency and radio-frequency alignment and progressively reduced as the circuits are brought into alignment.

Adjust the intermediate-frequency trimmers or variable cores of the intermediate-frequency transformers for maximum response in the following order: second intermediate-frequency transformer (secondary); second intermediate-frequency transformer (primary); first intermediate-frequency transformer (secondary); first intermediate-frequency transformer (primary). Unless otherwise stated in the manufacturer's instructions, repeat these adjustments to ensure utmost accuracy.

Where possible, re-seal the trimmers or cores after adjustment.

Finally, remove the short-circuit from the oscillator section of the ganged tuning capacitors.

Most modern British superheterodyne broadcast receivers are designed for an intermediate frequency of about 465 kc/s; the manufacturer's specification, however, should always be consulted. Where the correct intermediate-frequency is unknown, the point of maximum response should be found (see " Alignment Tests ", page 24) and the receiver aligned to that frequency. Some older models, mainly of American origin, use an intermediate-frequency of 110 kc/s; short-wave receivers, in order to minimize second-channel (" image ") reception, may employ a much higher frequency, a common value being 1,600 kc/s.

Intermediate-frequency transformers are designed for optimum, over-optimum or sub-optimum coupling. The optimum-coupled ones provide maximum sensitivity and good selectivity, but some cutting of side-band response and consequent attenuation of the higher audio frequencies may be experienced. The over-optimum type, if correctly adjusted, gives a band-pass effect, permitting reception of the full side-band range with, however, some loss in sensitivity and selectivity. Sub-optimum coupling is used where high selectivity is required; receivers employing such transformers are usually provided with tone-correction circuits to compensate for the loss of the higher audio frequencies. A response curve similar to that associated with over-optimum coupling is sometimes obtained by stagger tuning optimum-coupled transformers (e.g., primaries tuned to 467 kc/s, secondaries tuned to 463 kc/s). Manufacturer's literature will normally provide full information in such instances.

Occasionally it may prove advisable to render the A.V.C. circuits inoperative during re-alignment. One method is to short-circuit the points marked *X* in Fig. 10. Where muting circuits are used to reduce receiver noise between stations, these should always be disconnected during alignment.

Intermediate-frequency Wave-trap

Many receivers are fitted with either a series-tuned (acceptor) or parallel-tuned (rejector) wave-trap adjusted to the intermediate frequency. Its function is to reduce break-through of transmissions at or near this frequency. Alignment should be made after the intermediate-frequency circuits have been adjusted and before the final alignment of the radio-frequency circuits. Standard procedure is as follows:

Connect the output leads from the signal generator to the aerial and earth sockets of the receiver through a dummy aerial.

Tune the receiver to 550 m.

Set the volume control at maximum gain and the tone control at maximum high-frequency response.

Inject a strong modulated signal at the frequency to which the intermediate-frequency stages have been aligned.

Adjust the core or trimmer of the intermediate-frequency wave-trap for minimum response on the output meter.

Radio-frequency Alignment

Alignment is usually carried out at two points on each waveband; one near the high- and the other near the low-frequency limits. Typical alignment points are: Medium waveband, 200 m. (1,500 kc/s) and 500 m. (600 kc/s); Long waveband, 1,000 m. (300 kc/s) and 2,000 m. (150 kc/s); Short waveband, 16 m. (18·8 mc/s) and 50 m. (6 mc/s). Convenient frequencies are usually given by manufacturers in their service literature, although the frequencies chosen are not critical, except, perhaps, where special alignment calibration marks are provided on the dial for the convenience of the service engineer.

An example of the radio-frequency alignment on the medium waveband of a standard superheterodyne receiver is given below. For special instructions relating to A.C./D.C. models and those with internal-frame aerials see pages 30–31.

Check that the tuning cursor or drum, or both, are correctly adjusted in relation to the position of the rotary vanes of the ganged tuning capacitors. Normally when the cursor is at the lowest reading on the scale the rotary vanes of the ganged capacitors should be fully open, and when the cursor is at the highest reading they should be fully enmeshed.

Test the tuning mechanism for slip and backlash, and make any necessary adjustments prior to circuit alignment.

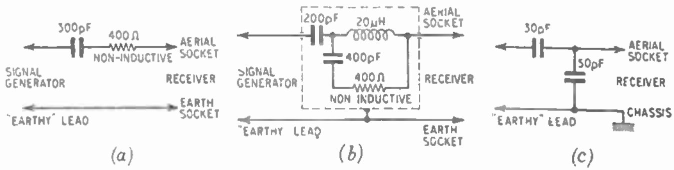


FIG. 11.—DUMMY AERIAL ARRANGEMENTS FOR RADIO-FREQUENCY ALIGNMENT.

(*a*) Simple system for general coverage; (*b*) standard dummy aerial particularly recommended for medium-wave alignment; (*c*) suitable arrangement for the alignment of automobile radio receivers.

Connect the output leads from the modulated signal generator to the aerial and earth sockets of the receiver through a dummy aerial, taking care that the " earthy " lead (usually the outer conductor) is connected to the earth socket.

Connect the output meter to the receiver as detailed under " Alignment Tests ", page 23.

Set the signal generator to the higher of the two alignment frequencies (e.g., 1,500 kc/s, 200 m.).

Switch the receiver to the medium waveband.

Set the tuning cursor accurately to the selected wavelength and adjust the medium-wave oscillator trimmer until the signal is heard. The correct adjustment position will be that which gives a maximum response on the output meter.

Adjust the signal-frequency circuit of the frequency-changer valve (usually the aerial trimmer) for maximum output, reducing the signal-generator output as receiver sensitivity increases.

Adjust the tuned circuit of the radio-frequency amplifier stage (where such a stage is provided) in a similar manner; i.e., set the trimmer to give maximum response on the output meter.

Where only slight misalignment of the receiver circuits is suspected, make a quick test of the sensitivity at the centre and low-frequency end of the band, since adjustment of the trimmers may be sufficient to restore the accuracy of alignment. Where the performance of the receiver is still unsatisfactory, however, continue re-alignment as follows :

Set the signal generator to the lower alignment frequency (e.g., 600 kc/s, 500 m.).

Set the tuning cursor accurately to the selected wavelength and adjust the medium-wave oscillator padder or medium-wave oscillator dust-iron core until the signal is heard. The correct adjustment position will again be that which gives a maximum reading on the output meter.

Adjust the tuned signal frequency or " aerial " circuit of the frequency-changer valve by means of the padder (often omitted) or adjustable dust-iron core to give maximum output, rocking the ganged tuning capacitors slightly during each adjustment and reducing the signal-generator output as the receiver's sensitivity increases.

In a similar manner, adjust the tuned circuit of the radio-frequency amplifier valve (where provided).

These adjustments at the higher wavelength will probably affect those carried out earlier at the lower wavelength; therefore all adjustments should be repeated at both alignment points until no further improvement can be obtained. Where it is impossible to eliminate completely this fault, which is known as " pulling ", the final adjustment should be made at the higher-frequency position (e.g., 1,500 kc/s).

The alignment procedure outlined above is equally applicable to the long and short wavebands, in each case adjustment to the trimmers being made at the higher frequency and adjustment to the cores or padders being made at the lower alignment frequency. In some receivers, however, certain of the adjustment points are omitted on one or more of the wavebands, and the procedure must be modified accordingly. On portable receivers, for example, only trimming capacitors may be provided; and in such cases the receiver is usually aligned at 1,500 m. (200 kc/s).

It is difficult to secure accurate alignment over an entire waveband, especially where the frequency ratio is high. Practical alignment is therefore often a matter of compromise, and some variation in sensitivity will usually remain, no matter how carefully the circuits are adjusted. For normal broadcast reception such discrepancies are usually of little practical importance. Tracking errors on any one band can be corrected by adjustment to the outer rotary vanes of the ganged tuning capacitors; but this correction may not hold good on other wavebands.

Where the layout of trimmers and inductors is unknown, this must be traced before commencing re-alignment. Most manufacturers supply such information, but failing this, examination of the circuit wiring and comparison of the relative sizes of the various inductors must be resorted to. By lightly touching adjustment screws with a metal screwdriver while the receiver is successively switched to each waveband, the strength of the resulting clicks will also prove a valuable guide as to which components belong to which circuits.

Short-wave Alignment

Although the re-alignment of short wavebands is fundamentally similar to that already described for the medium waveband, certain additional precautions are necessary if maximum performance is to be obtained.

Standard dummy aerials are usually unsatisfactory for short-wave operation, and should therefore be replaced by a 400-ohm non-inductive resistor connected between the signal generator and the aerial socket of the receiver.

As the frequency is increased, the second channel (" image ") response will become more prominent. This is due to the decreasing percentage difference between signal and oscillator frequencies. Care should therefore be taken to ensure that the receiver is adjusted to the correct frequency and not to that of the second channel. The oscillator circuit is usually designed to track at a frequency higher than that of the signal frequency, though this arrangement is sometimes reversed. Generally speaking, there are two positions of the oscillator trimmer which give a satisfactory response, and the one requiring the lesser capacitance should be chosen.

Adjustment of communications-type receivers or receivers incorporating electrical bandspread circuits on short wavebands should not be lightly undertaken. It should be attempted only where tests clearly indicate that this is necessary and a full range of service equipment (including an accurate frequency sub-standard) is available. Manufacturers' instructions should be carefully followed at all times.

Image Rejector

The image rejector, where fitted, should be adjusted after the alignment of the long waveband. A popular arrangement consists of a series-tuned (acceptor) wave-trap connected across the long-wave input transformer, and adjusted to the wavelength of the offending medium-wave station. For example, the 247-m. B.B.C. transmitter may appear on approximately 1,056 m. where an intermediate frequency of 465 kc/s is used. To minimize such interference, procedure would be as follows :

Connect the modulated signal generator to the aerial and earth sockets through a dummy aerial, as for radio-frequency alignment, and connect a suitably matched output meter.

Inject a strong 1,214-kc/s (247-m.) signal and tune the receiver on the long waveband until this is heard.

Adjust the trimmer or dust-iron core of the image-rejector circuit until minimum response on the output meter is obtained.

A.C./D.C. Receivers

With A.C./D.C. receivers, A.C./battery receivers and some A.C. receivers using A.C./D.C. technique, particular care must be taken during re-alignment to minimize the risk of shock and damage to equipment, which arises from a high potential (" live ") chassis. Before commencing work on a receiver always examine the circuit diagram (or receiver itself) to ascertain whether there is a D.C. continuity between chassis and power-supply leads.

The most satisfactory and complete safeguard is to supply power for the receiver from A.C. mains through a double-wound (1 : 1 ratio) isolating transformer. No other special precautions will then be necessary.

Where such a transformer is not available, it is advisable to make sure that the chassis is at earth potential before starting to re-align the receiver. This can be done by connecting the receiver to A.C. supply mains, switching on and checking by means of a sensitive A.C. voltmeter or neon bulb in series with an earth lead, and reversing the mains-input plug where the chassis is found to be above earth potential.

Where the receiver cannot be operated with the chassis at earth potential, as may be the case where power is obtained from D.C. supply mains, great care should be taken not to touch the chassis unless standing on a rubber mat, nor to allow any piece of earthed equipment (e.g., signal generator, electric soldering-iron) to come into contact with the chassis. The screened output lead from the signal generator should be connected to chassis via a good-quality $0{\cdot}1$-μF isolating capacitor for all alignment processes. As a further precaution direct earth connections should be removed from any piece of equipment which may come into contact with the chassis.

Portable Receivers

Where receivers are normally operated from internal frame aerials (e.g., portable and "personal" models) re-alignment must be carried out with these aerials in circuit.

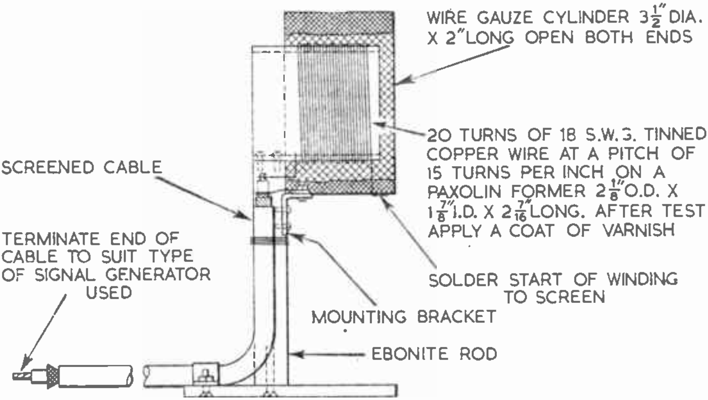


Fig. 12.—Construction of a Shielded Coil.

Details of the directional dummy aerial recommended for the re-alignment of models incorporating frame aerials.

(*Philco*)

For radio-frequency alignment the signal should preferably be injected via an R.M.A. standard shielded coil spaced about 1 ft. from the frame aerial. This coil is defined as follows: "This shall be a cylindrical coil, 5 cm. in radius and 6 cm. deep, wound with 20 turns to an approximate inductance of 40 microhenrys. The whole coil shall be shielded by a wire cage arranged to avoid magnetic screening (i.e., there shall be no complete circuits whose planes are normal to the axis of the coil). The connecting leads shall also be screened."

Where a coil of this type is not available, a small loop aerial (about four turns) of approximately the same area as the frame aerial of the receiver should be connected to the output leads or socket of the signal generator and set up at a minimum distance of 2 ft. from the receiver. For preliminary adjustments, it may prove necessary to loosely couple the signal-generator output lead to the grid of the frequency-changer valve by laying the lead near to this point.

The receiver should always be aligned with the batteries and loud-speaker in the same position relative to the frame aerial as would be the case under normal operating conditions, otherwise the inductance of the frame aerial may be affected.

Straight Receivers

The alignment of straight (T.R.F.) receivers is comparatively simple, except perhaps where band-pass circuits are involved, since the difficulty of tracking radio-frequency circuits at different frequencies does not arise.

Alignment is usually achieved by means of trimmers and split vanes on the ganged tuning capacitors. In practice, adjustment of the vanes is seldom necessary. Medium-wave trimmers usually remain in circuit when the receiver is operated on the long waveband, and in such cases the medium waveband must be aligned first.

The equipment is set up as for the radio-frequency alignment of super-heterodyne circuits. The tuned circuits in the demodulator stage are trimmed at a point near the high-frequency end of each band until the dial calibration is correct, and the tuned circuits of any radio-frequency stages are adjusted for maximum output. The split vanes of the ganged

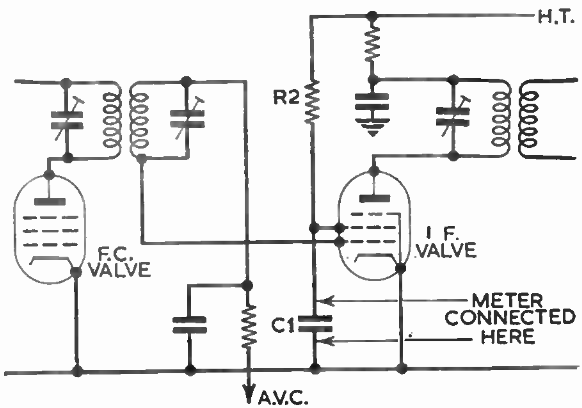
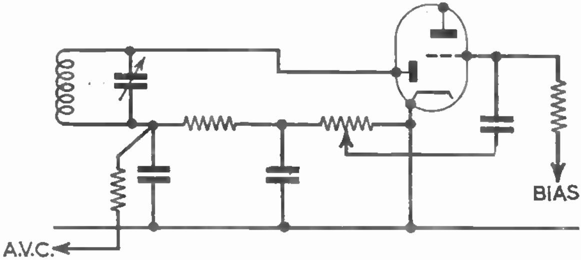
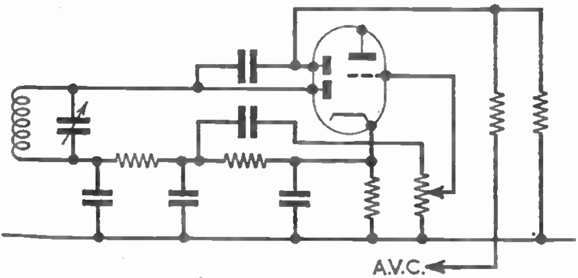


Fig. 13.—Intermediate-frequency Stage of a Battery-operated Receiver.

The voltage drop across R2 is varied by A.V.C. action, and the screen voltage is measured by a voltmeter connected across the screen by-pass capacitor C1. Resonance will be indicated by an increase in screen voltage.



(a)



(b)

FIG. 14.—TYPICAL A.V.C. CIRCUITS

(a) Method of obtaining A.V.C. from a signal diode; (b) A.V.C. obtained from a separate diode section, fed from the secondary of the intermediate-frequency transformer. How resonance indications are affected by the type of circuit is explained in the text.

capacitors may then be adjusted near the low-frequency end of the band on which maximum sensitivity is required.

Band-pass coils in the aerial circuits should be adjusted in a manner similar to that already described for band-pass intermediate-frequency circuits (see page 25), the signal-generator control being slowly rotated until the receiver output falls to half its peak value, and the number of kilocycles off resonance noted. Should the figures obtained differ by more than about 1 kc/s, the trimmers must be adjusted to givo more output on the side which showed the greater difference. Alignment of bandpass circuits, however, can more easily be carried out with the aid of an oscillograph and a frequency-modulated oscillator.

Alternative Methods of Indicating Resonance

Should neither an output meter nor sensitive A.C. voltmeter (0–5 or 0–150 volts) be available, alternative methods of indicating circuit resonances can be employed.

Where a " magic eye " or similar type of visual tuning indicator is fitted, this provides a rough indication of the signal voltage reaching the demodulator valve, and can therefore be used in much the same way as an output meter : an unmodulated signal generator can then be used. This form of indicator tends to be sluggish in response, and is not recommended.

A.V.C. circuits permit reasonably accurate alignment to be made with a sensitive D.C. voltmeter, indication of circuit adjustment being obtained by noting the effect of the A.V.C. voltage on one of the controlled stages. With this system, the injected signal should be unmodulated and of adequate strength to overcome the initial delay of the A.V.C. circuit. The D.C. voltmeter may be connected in any of several possible ways, two of which are shown in Fig. 10 (points " A " and " B ") and in Fig. 13. That shown in Fig. 10 is recommended for mains-operated receivers, a voltmeter measuring 0–10 volts being connected across the bias resistor of the intermediate-frequency valve. As the circuits are brought into resonance, the A.V.C. applies additional bias to the grid of this valve, reducing the anode current and hence the voltage drop across the bias resistor. Resonance is thus indicated by a *decrease* in voltage. In Fig. 13—recommended for battery-operated models—a D.C. voltmeter measuring 0–100 volts is connected between the screen grid of the intermediate-frequency valve and chassis. The action is similar, but resonance is indicated by an *increase* in reading. With these systems it is possible to align a receiver on incoming broadcast signals, since the varying modulation depth will not affect the reading.

The details given in the preceding paragraph apply to the adjustment of all circuits up to and including the primary of the final intermediate-frequency transformer, but the type of indication to be expected when adjusting the secondary of this transformer depends upon the design of the receiver. As this circuit is brought into resonance it tends to extract power from the primary. Thus when the A.V.C. voltage is derived from the anode of the intermediate-frequency valve, as in Fig. 10, a voltmeter connected across R1 would show a slight increase in reading at resonance. Where, however, the A.V.C. is derived from the signal diode (Fig. 14 (*a*)) or from a separate diode connected to the *secondary* of the final intermediate-frequency transformer (Fig. 14 (*b*)), the voltmeter reading will decrease at resonance.

One further method of obtaining an indication of resonance is to insert a sensitive micro-ammeter in series with the signal diode load at a point of low potential, such as " Z " in Fig. 10. Though reliable and positive, this method is seldom convenient to use. As with a power-output meter, the signal input must be kept low in order to prevent the A.V.C. from coming into play, and adjustment is made for maximum current on all circuits.

Tuning Wands

When dealing with receivers possessing readily accessible, air-cored inductors, a most useful accessory takes the form of a " tuning wand ", consisting of an insulated rod some 6 in. long and ¼ in. in diameter, having a piece of ferromould (dust-iron) core material about ½ in. long and ¼ in. in diameter attached to one end, and a piece of brass of similar dimensions attached to the other. The effect of introducing the iron end into a coil will be to increase the inductance, while inserting the brass end will lower it. The rod can thus be used to check alignment.

SERVICING V.H.F. RECEIVERS

The introduction of V.H.F. broadcasting in Band II (87·5–100 Mc/s) has presented the service engineer with a number of new problems. The design of combined A.M./F.M. receivers is discussed in Section 14, and it is important that the basic principles of the circuits used and the problems involved should be fully understood when dealing with these receivers.

While much of the trouble-tracing will be similar to that already described for standard medium or all-wave receivers, it must be remembered that the circuits at V.H.F. may easily be affected by small changes in the positioning of wires or by any deterioration of valves and insulation materials that might pass unnoticed on the lower frequencies. The radio-frequency circuits and the 10·7-Mc/s intermediate-frequency circuits demand the same careful treatment as in television-receiver servicing.

The most fundamental change in servicing procedure, however, is in the re-alignment of F.M. receivers, particularly of the discriminator stage. Since there is considerable variation in the finer details with different models, it is strongly recommended that manufacturers' recommendations be followed wherever possible. Procedure will also be affected by the type of servicing equipment available, and for this reason the following notes are intended as a general guide only, and are concerned mainly with cases where only a minimum of test equipment is available.

Alignment

The quality of an F.M. receiver depends very largely upon the accuracy of alignment of the ratio detector transformer and the I.F.

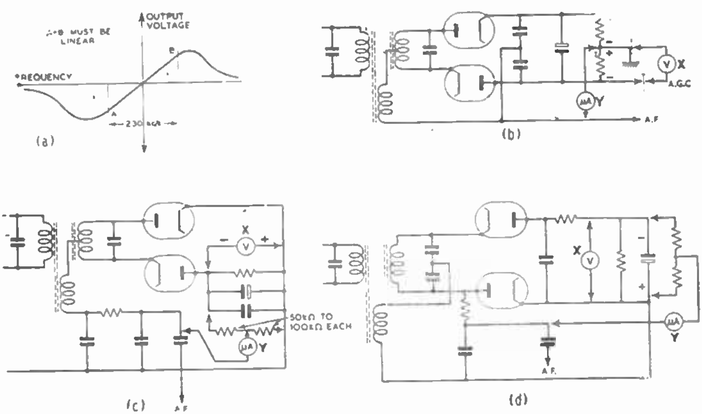


FIG. 15.—ALIGNMENT OF BALANCED AND UNBALANCED RATIO DETECTORS.

In practice the ideal linear bandwidth of 230 kc/s cannot be achieved on many receivers, about 180 kc/s being fairly normal.

stages, as the response curve must be linear over the full band-width and balanced about the nominal I.F. (see Fig. 15 (*a*)). The most accurate method of alignment of these circuits is by means of frequency-modulated oscillator and full visual display equipment. Where this is not available, the procedure given in Table 1 will give good results, provided care is taken. The equipment required is a signal generator covering 10·7 Mc/s; a valve voltmeter or high-resistance voltmeter (5–10 volts F.S.D., about 20,000 ohms/volt or higher); a micro-ammeter (200–250 μA F.S.D.); and, for unbalanced ratio detectors, a pair of matched carbon resistors between 50k and 100k. In practice, the two meters may be available in a single multi-tester. The 10·7-Mc/s signal is injected into the I.F. strip, and the procedure is then as outlined in the table.

TABLE 1.—ALIGNMENT OF RATIO DETECTOR AND F.M./I.F. CIRCUITS

Recommended order of procedure when using unmodulated signal generator

Circuit(s)	*Meter and Adjustment*
1. Primary of ratio detector transformer	Connect voltmeter as X in appropriate section of Fig. 15. Adjust core for maximum reading
2. I.F. transformers (working backwards from last I.F. stage to mixer output)	As 1, reducing input as necessary to maintain reading of about 4 volts
3. Secondary of ratio detector transformer	Micro-ammeter as Y in Fig. 15. Adjust for zero reading centrally between two peaks of opposite polarity
4. Readjust primary of ratio detector transformer	As 1
5. Readjust I.F. transformers	As 2
6. Finally, readjust ratio detector transformer secondary	As 3

7. Check response curve to ensure that the peaks (Fig. 15 (*a*)) are centrally balanced about zero point. To check this, measure Y readings at 10·6 Mc/s and 10·8 Mc/s (or as appropriate if I.F. is other than 10·7 Mc/s). Readings should be within a few micro-amperes of each other

When adjusting intermediate-frequency transformers at 10·7 Mc/s or above, it will be necessary to consider—as in television practice—the affect of trimming tools, which should therefore be of non-metallic, low-capacitance type with a long handle. It is also more than ever important to allow plenty of time for the receiver to reach its settled operating temperature before starting adjustments.

SERVICING PRINTED WIRING PANELS

From the servicing viewpoint the introduction of printed wiring panels may be regarded as a mixed blessing. For instance, one advantage is that the component identification "R" and "C" numbers can be printed alongside the actual component for rapid identification, and with such sets all components are usually readily accessible without having to dig down through several layers of resistors and wires. On the other hand, tracing out an unknown circuit can be more difficult, even though the panels are often translucent so that the position of components can be observed when looking at the wiring side by bringing a 60-watt electric lamp close to the component side of the panel.

The main difference in the servicing of printed-circuit receivers is in the replacement of faulty components, calling as it does for greater skill and care in soldering. Too large or too hot a soldering-iron may cause blistering and damage to the laminated board. Very sparing use of solder is also necessary, as otherwise it may run between adjacent copper "wires" and be difficult to clear.

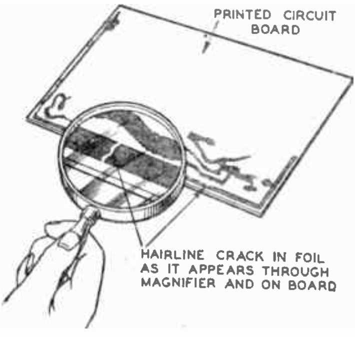
Tools

The following tools and aids are recommended:

(1) A low-wattage soldering-iron with a small point or wedge bit. The ratings should be less than 50 watts, and preferable less than 35 watts.
(2) Supply of 60/40 resin-cored solder.
(3) Soft wire brush (such as suede-shoe brush).
(4) Pair of diagonal wire cutters.
(5) Pair of long-nosed pliers.
(6) Small wire pick or soldering aid.
(7) Needle-point probe for circuit testing.
(8) Magnifying glass for detection of small cracks.

General Precautions

(1) Avoid damaging the copper foil. Be careful when removing components not to cause small breaks in the foil. Should a small break



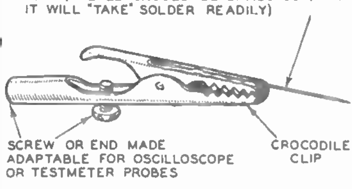


FIG. 16 (*left*).—DETECTION OF HAIRLINE CRACKS.

FIG. 17 (*right*).—CONSTRUCTION OF NEEDLE POINT PROBE.

occur, this can usually be "jumped" with molten solder. Larger breaks should be repaired with single strand connecting wire.

(2) Never apply excessive pressure to the wiring board, as this can easily cause cracks or breaks in the foil.

(3) Excessive heating from large soldering-irons or due to overlong application of a hot iron may cause the bond between the board and the copper foil to break or blister.

(4) When replacing components avoid large deposits of solder. These can easily cause a short-circuit or intermittent fault by bridging adjacent copper foils.

(5) When brushing off molten solder small particles may be left sticking to the board. Before installing a new component remove these particles with a cloth dipped in solvent.

Replacing Components

It is best to clean and tip with solder any new components before inserting them through the holes in the laminated base. The wired ends of resistors and capacitors should be carefully trimmed and bent over so that when soldered into position there is no tendency for the component to force the copper foil away from the base. A method of overcoming this with simpler and non-critical circuits is to clip away the defective component with wire cutters, leaving as much as possible of the original connecting leads in place, and then to shape the leads of the replacement component into small loops which can be slipped over the original leads and soldered into place; if the original legs are very short it may prove advisable to cut the old component in half with wire cutters and then strip the component away from its internal leads to provide slightly longer connecting wires. It should be noted, however, that when components are changed in more critical circuits, such as where parasitic or spurious oscillation is liable to occur with a change of stray capacitance or slight change in position of a component, this method of clipping off the lead and soldering the new component to the wire ends can result in instability. In such circuits it is better to remove the old component completely and to solder the new component in its place. In all cases where it is necessary to unsolder wires and lugs on the wiring side of the board, the following procedure should be used:

(1) Heat the connection on the foil side of the panel with a small soldering-iron. When the solder becomes molten brush it away with a soft wire brush, taking care not to overheat the connection, and remove the iron while brushing away the solder. It may require several sequences of heating and brushing to remove all the solder.

(2) Insert the blade of a knife between the copper foil and the bent-over component lead and bend the latter perpendicular to the board. It will sometimes be necessary to apply the iron to the joint while doing this where the connection has not been broken by the brushing.

(3) While applying the iron to the joint, gently "wiggle" the component until it comes away from the panel.

(4) Remove any small particles of solder which may be sticking to the protective coating of the circuit board.

(5) If there is a thin layer of solder left over the hole, pierce this with the new component wires after heating the solder.

(6) Place the new component in position and cut the connecting leads as necessary. Bend over the ends against the copper foil and resolder the joint.

(7) Finally, recoat the affected area with a protective coating such as polystyrene dope.

Measurements

Resistance and component measurements are usually possible from the component side of the board, but, should it be necessary to work on the wiring side, it should be remembered that the protective coating over the foil forms an insulator: this can easily be penetrated by using a probe consisting of a brass needle soldered to a crocodile clip.

SERVICING TRANSISTORIZED EQUIPMENT

Servicing transistorized equipment presents a number of pitfalls for the unwary, including the possibility of destroying or damaging the transistors. These dangers arise mainly from unfamiliarity with transistor practice, as once this is understood there is little that is inherently difficult in dealing with transistor equipment. Just as it soon becomes second nature to avoid putting H.T. across the filament pins of a battery valve, or high positive voltages on a grid, so similar semi-automatic reflexes develop with transistors. But until considerable experience has been gained of handling transistors, the service engineer and constructor should keep continuously in mind the ways in which transistors can be damaged.

First, correct battery polarity must be carefully observed. Connection of the wrong battery polarity to the collector, even momentarily, will almost certainly destroy a transistor. Since, for both point-contact and *pnp* junction transistors, the correct polarity—with the " anode " (collector) negative— is the direct opposite to what service engineers have been accustomed when handling valve circuits, there is a real danger of connecting the battery wrongly unless vigilance is exercised. Particular care should also be taken not to allow voltages to develop accidentally across the various electrodes. The internal resistance may be very low, particularly with power transistors, so that even comparatively low voltages may cause currents sufficient to burn out or damage a transistor. One way in which such voltages may be introduced into a circuit is by the use of the resistance-measuring or continuity testing ranges of a conventional multi-test meter with an internal battery. Never use a low-sensitivity continuity tester, and make sure that any testmeter has an internal resistance of several thousand ohms; when in any doubt connect an external resistor into circuit or disconnect the transistors from the circuit under test. Do not overlook the possibility that some servicing instruments and even soldering-irons may have a small potential between the instrument and earth, and that this can represent a source of danger. It is usually safe to use a voltmeter for checking potentials, but even here it is better to connect the meter into circuit before the receiver is switched on and not withdraw it until after the set is switched off; the insertion of meters for current tests is not usually recommended, as this may alter the operating conditions of the transistor. It is safest to use isolating capacitors. Do not forget that the sudden discharge of a high-value capacitor through a transistor may cause damage Excessive signal inputs during alignment are also dangerous.

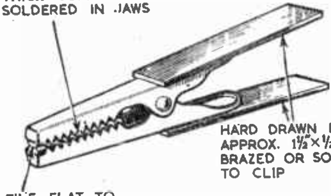


FIG. 18.—A USEFUL HEAT SINK.

A convenient heat sink that leaves both hands free for soldering can be made from a standard crocodile clip. The thick copper wires conduct heat away from the wire on which it is clipped while the brass plates assist dissipation of the heat to the air.

A further important matter is that of temperature. It is essential that the soldering of transistors (and, incidentally, crystal diodes) into and out of circuit be carried out as quickly as possible with low-wattage irons to prevent heat travelling along the connecting wires into the body of the transistor. The use of a " heat sink " is always to be recommended when soldering these components; this usually takes the form of gripping the wire between the point of soldering and the body of the transistor with a cool pair of pliers, so that there is good thermal contact between the wire and the pliers which will then absorb most of the heat. Transistor wires are tinned and plated to enable soldering to be carried out as quickly as practicable. Generally, solder as far away as possible from the body of the transistor, but do not leave very long leads or this may cause vibration of the transistor. Incidentally, when connecting transistors into circuit take care not to bend the joining wires too close to the body of the transistor. Never remove or disconnect transistors without making sure that the equipment is " off ", and all potentials removed.

It is not only while soldering that the transistor may be affected by heat and humidity: the designer must take care that all transistors are sited well away from heat-producing components and with adequate ventilation, and the service engineer should be careful not to make any alterations that would permit the transistors to become overheated. Another aspect of the same problem is that of not allowing, by component changes, the rated collector dissipation to be exceeded. Transistors should not be stored in places where there is any danger of extremes of temperature, or dampness.

One point which should not be overlooked is that transistor material is sensitive to light. To overcome this, transistors normally have an opaque coating, and care should be taken not to damage this. Should the coating be removed, there is always the danger that a transistor may be affected by light, and this could, for instance, be the source of hum pick-up from an A.C.-operated light bulb.

A useful tool is a non-metallic crochet hook with a notch suitable for gripping wires at one end, and a fine point for probing or cleaning away solder at the other.

SERVICING COMMUNICATIONS RECEIVERS

The number of valves and stages used in communications receivers may be considerable, and therefore the servicing of such instruments is sometimes regarded as a formidable task. A typical communications receiver might comprise : one or two stages of radio-frequency amplification; frequency changer stage, often with a separate local oscillator valve; one, two or three stages of intermediate-frequency amplification; beat-frequency oscillator; detector and A.V.C. diodes; noise limiter; audio-frequency voltage amplifier; audio-frequency power amplifier; power rectifier. A neon voltage stabilizer is commonly fitted to control the voltage applied to the anode of the local oscillator valve.

With an intermediate frequency of 465 kc/s, it is often considered necessary to employ two tuned stages of radio-frequency amplification in order to reduce image interference on the higher frequencies. This arrangement necessitates four ganged tuning circuits: first radio-frequency, second radio-frequency, mixer and oscillator. Sometimes the difficulty is overcome by raising the intermediate frequency to, for example, 1,600 kc/s, or by the use of double conversion. In the case of double conversion, incoming signals are converted, for example, to 1,600 or 2,075 kc/s and then, by means of a second frequency-changer valve, to 465 kc/s or, in some instances, to a much lower frequency such as 85 or 110 kc/s, the main amplification taking place at the second, i.e., lower, intermediate frequency. The double-conversion method is becoming increasingly popular, for it enables satisfactory image rejection to be achieved with one radio-frequency stage; at the same time the high selectivity and high gain of low intermediate-frequency circuits are retained. A typical double-conversion arrangement consists of: radio-frequency amplifier; frequency changer and separate triode oscillator (2,075 kc/s); triode-hexode frequency changer (455 kc/s); three stages of amplification at 455 kc/s; double-diode noise limiter and A.V.C.; diode detector; double-triode B.F.O., and audio-frequency amplifier; audio-frequency output; neon stabilizer; power rectifier. When double conversion is employed, efficient screening and careful layout of the local oscillator circuits are necessary if spurious responses to harmonics of the lower-frequency oscillator are to be avoided. For details of a typical receiver see Section 22.

The tuning mechanism of communications receivers is usually calibrated in kilocycles. Owing to the sharpness of the tuning, smooth-action reduction gear as free as possible from backlash and with good re-set accuracy is essential. Elaborate precautions may also be necessary to avoid loss of sensitivity and blind spots resulting from stray coupling at the wave-change switch; several manufacturers use rotary turrets or plug-in coil assemblies to overcome this difficulty.

Alignment

While the alignment procedure for communications receivers is basically similar to that employed for standard broadcast receivers, the large number of tuned circuits and the high standard of performance required will naturally tend to make the task not only more lengthy, but greater accuracy of adjustment will also be required. On the other hand, the air dielectric trimmers often used and the sturdy construction of many models diminish the likelihood of radical misalignment occurring in normal circumstances. The apparent complexity of re-alignment can best be resolved by working systematically through the various wavebands, resisting any temptation to make haphazard adjustments. Careful attention should be given to manufacturers' recommendations, and it should also be borne in mind that selectivity and sensitivity are likely to be regarded by the user as being of far greater importance than fidelity of reproduction. All intermediate-frequency circuits should therefore be peaked and no attempt made to obtain band-pass characteristics.

Any of the methods for obtaining indication of resonance which are specified in the section on the alignment of broadcast receivers (see page 32) may be used. In addition, a high impedance A.C. voltmeter may be connected to the headphones-jack. Where an S-meter is fitted, this in itself forms a suitable indicator, and the circuits should be adjusted for maximum deflection of the needle. It should be noted that A.V.C. is not always applied to the first radio-frequency amplifier, the frequency changer or the later intermediate-frequency stages; the first intermediate-frequency stage is, however, almost invariably A.V.C. controlled.

Intermediate-frequency Alignment

Where no crystal filter is fitted, there is little difference between the intermediate-frequency alignment of a communications receiver and that of a broadcast receiver. If separate intermediate-frequency and radio-frequency controls are incorporated, the intermediate-frequency control should be set at the maximum-gain position and the radio-frequency control at minimum. Where these controls are combined—as usually they are—the control should be set for maximum gain. If any form of audio-output meter is used, the audio-frequency gain control should be set at maximum gain and the A.V.C. switched off. Where A.V.C. action is used to obtain indication of resonance, the audio-frequency gain may be set to a convenient level. Readings should be

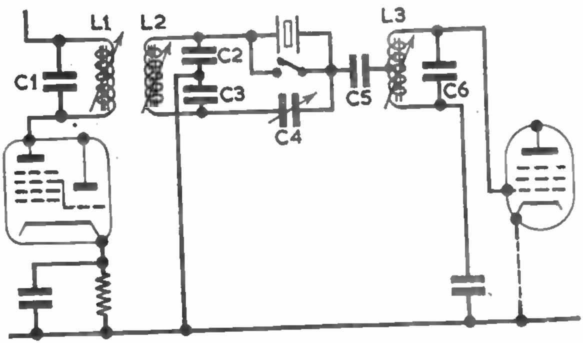


FIG. 19.—CRYSTAL FILTER.

In this circuit L3/C6 is tuned to the exact frequency of the crystal, while C4 is the phasing control. C5 may be made variable, and then becomes a selectivity control.

kept within reasonable limits by adjustment of the attenuator on the signal generator.

Should the receiver incorporate a crystal filter, the intermediate frequency must be made to coincide with the resonant frequency of the crystal, even where this differs slightly—owing to tolerances in crystal etching and lapping—from the nominal intermediate frequency specified by the manufacturers. It can be seen from the circuit of a typical crystal filter in Fig. 19 that an additional tuned circuit L3–C6 is involved.

To find the resonant frequency of the crystal, the procedure is as follows :

Switch the crystal into circuit and set the phasing capacitor at approximately half-mesh. Inject a fairly strong modulated signal to the control grid of the frequency-changer valve. Vary the tuning control of the signal generator very slowly over a range of a few kilocycles either side of the nominal intermediate frequency of the receiver. At one point a very definite sharply peaked response should be obtained. The frequency of the signal generator at this setting corresponds to the frequency of the crystal; normal intermediate-frequency alignment procedure should then be carried out, peaking all intermediate-frequency circuits to this frequency. While making intermediate-frequency adjustments it will be necessary to advance progressively the attenuator on the signal generator to prevent over-loading. Adjustment of the secondary circuit of the first intermediate-frequency transformer L2 will probably affect the resonance of the crystal coil L3 and vice versa; repeated trimming of the various circuits is therefore essential.

With the phasing control at half-mesh, a sharply peaked symmetrical response curve should be obtained. At other settings a clearly defined rejection " notch " should appear on one or the other side of the curve, the frequency difference between the peak and the " notch " varying with the setting of the phase control. If no oscilloscope is available, this effect may be tested by slightly altering the frequency of the signal generator until a noticeable drop in output is obtained, and then slowly rotating the phasing control; at one setting the output should drop practically to zero. The signal generator should then be mistuned by an equal amount the other side of the intermediate frequency and the test repeated. The difference between the peak amplitude and the rejection minimum will depend largely upon the " goodness " of the particular crystal. and usually varies slightly from model to model.

Although rare, communications receivers are sometimes fitted with a band-pass crystal filter, involving the use of two matched crystals with a frequency separation of between 300 and 3,000 c/s, depending upon the design and whether the filter is intended for the reception of telegraphy or telephony transmissions. Varying the signal generator, in this case, will produce two points of maximum output, corresponding to the two crystal frequencies. The signal generator should be set to a frequency mid-way between these two points and the intermediate-frequency circuits aligned to this frequency. With a band-pass filter the phasing control is usually pre-set and should be adjusted only if it is found necessary to equalize the response of the two crystal peaks.

In appearance the B.F.O. coil unit will generally resemble an intermediate-frequency transformer, with a single aperture for adjustment of the trimmer or iron-dust core, fine adjustment being by means of the pitch-control knob. To check the alignment of the B.F.O., the pitch-control capacitor should be set at half-mesh. Then, with the B.F.O.

switched off, a steady unmodulated signal should be accurately tuned in —the frequency of the signal being unimportant—and the B.F.O. switched on. If the unit be correctly adjusted, rotation of the pitch control should result in a heterodyne beat note becoming audible, variable up to about 4,000 c/s each side of the centre null—" zero beat "—point, which should coincide with the half-mesh position. If this effect is not obtained, the adjustment in the B.F.O. coil unit should be altered as required.

The user of a communications receiver generally requires exact dial calibration and, therefore, a normal signal generator alone will seldom be accurate enough for the setting of the oscillator circuits. But wide-coverage frequency meters, which can be set with reference to a built-in crystal oscillator, are ideal for this purpose, typical examples being the American Service types BC221 and LM7, a number of which have become available in the United Kingdom as Service Disposals. Where such an instrument is not available, a crystal-controlled sub-standard may be used. This device is not calibrated. It is, however, designed to give a strong harmonic output at intervals of either 100 or 1,000 kc/s throughout the high-frequency spectrum. Since neither frequency meters nor sub-standards are generally designed to give a modulated output, they must be used in conjunction with the B.F.O. on the receiver, which should always be set at half-mesh and signals tuned to zero beat. For adjustment of the oscillator circuits it is usually better to rely on setting by ear rather than on meter readings, since the zero-beat method provides an extremely accurate indication of resonance.

Adjustments to the oscillator circuit are made on each range in turn, moving the trimmer at the high-frequency end and the dust-iron core or variable padder capacitor at the low-frequency end. Calibration checks should always be carried out at convenient intermediate points in addition to those made towards the high- and low-frequency ends. Where dust-iron cores are used, the padder or tracking capacitor will almost certainly be fixed on the higher-frequency ranges, but may be variable on the lower-frequency ranges: if the calibration is still incorrect after repeated adjustments of the core, further trials should be made, this time varying the padder capacitor.

After the oscillator circuits have been adjusted, the sub-standard or frequency meter may be replaced by a modulated signal generator connected to the input terminals of the receiver via a short-wave dummy aerial, the B.F.O. being switched off and the radio-frequency gain control set to the maximum-gain position. If a panel-controlled aerial trimmer is fitted, the vanes should be set at half-mesh throughout the alignment of the signal-frequency circuits. This operation is then carried out in the normal manner, adjusting trimmers and cores for maximum response at frequencies near to but not quite at the end of each range. Adjustments should always be repeated to ensure optimum gain, and the final adjustment made at the higher-frequency point.

Manufacturers' literature often gives detailed figures of sensitivity, both intermediate frequency and overall, selectivity and similar data. Whenever possible a check should be made after re-alignment to ascertain whether or not the performance of the receiver is up to standard.

A final test should always be carried out with a standard aerial, as this may show up small defects likely to be overlooked when testing with instruments.

AUTOMOBILE RADIO RECEIVERS

Automobile radio receivers are liable to develop the same faults as standard domestic receivers, and the servicing procedure differs little from those already described. However, owing to the more rigorous conditions under which automobile models are operated, there is a tendency for certain types of faults to assume greater importance than with domestic receivers.

It should be remembered that these receivers are subjected to considerable vibration, and their mechanical construction must of necessity be sound : particular attention should therefore be paid to all soldered joints, and all variable cores and trimmers should be sealed after adjustment, otherwise their setting may change. There is also an increased risk of moisture entering the receiver housing, either by absorption or by condensation, resulting in the lowering of the insulation resistances of capacitors, transformers and other components, leading to inefficient operation and premature breakdown. The best safeguard against faults developing from this cause is the use of high-grade, tropicalized components for all replacement purposes.

The vibrator units are also subject to considerable deterioration, and have a limited life; the contact points wear or become pitted and stick, or the springs lose their tension.

The main on/off switch has to carry a relatively high current at low voltages, and any resistance introduced by dirt or oxidation, caused by arcing, may reduce the efficiency of the receiver, or even cause complete failure.

Complete failure of the power pack is often caused by the breakdown of the rectifier-anode filter capacitor, connected across the secondary of the H.T. transformer. High-voltage types of exact capacitance should always be fitted in this position. Cold-cathode rectifiers are sometimes fitted: a faulty valve in this position may give rise to a form of electrical interference resembling hash.

When re-aligning the radio-frequency circuits the dummy aerial employed should be of a high-capacitance type (see page 28). Final adjustment should preferably be carried out with the aerial and receiver in position. Most car aerials are tightly coupled to the input circuits of the receiver, so that any change in the aerial capacitance is liable to detune the input circuit, and thus reduce sensitivity. In some vehicles, generally of American manufacture, the aerial consists of a section of the bodywork, such as a door, or part of the hood, which is completely insulated from the remainder of the bodywork. This type of aerial generally possesses considerable capacitance to chassis, and may thus " pull " the input circuit unless a suitable dummy aerial has been used.

When installing a receiver in a vehicle not previously fitted with radio, care should be taken to ascertain that the car batteries and charging generator are capable of handling the extra drain. The average receiver requires about 3 amperes from a 12-volt source, and up to twice this figure from 6-volt sources. It is desirable that it should be connected in the vehicle electrical circuit so that the generator is protected from overload. This could be accomplished, for example, by taking the feed for the receiver from the A2 terminal on a Lucas control box. Heavy cable will, of course, be required to prevent undue voltage drop.

The polarity of the car chassis in regard to the engine battery should be checked: many receivers are fitted with electrolytic capacitors which will be damaged if the battery is wrongly connected.

Car aerials are discussed in Sections 14 and 21 and the suppression of interference in Section 33.

TAPE RECORDER SERVICING

The principles of domestic and professional magnetic recording are described in Section 34. In this section information is given on the routine adjustment and servicing of domestic tape recorders.

The electronic section of a conventional domestic tape recorder consists of a power supply; a record amplifier; a playback amplifier (usually the record and playback amplifier is a combined dual-purpose unit); a recording-level indicator; and an H.F. bias generator. A representative circuit is shown in Fig. 20. While a similar arrangement to this is used in the majority of currently available commercial tape recorders, common variations are: (*a*) to use one valve as H.F. bias generator on record and output stage on playback, in which case the record signal is taken off at V2B anode; and (*b*) to take the record output from V2B anode circuit so that V3 can be used for monitoring.

H.F. Bias Generator

The purpose of the H.F. bias is to improve the quality, reduce the background noise and increase the amplitude of the recording. Any fault or maladjustment in this stage therefore has a marked effect on the quality of the recording.

Bias-level Adjustment

Some recorders are fitted with a pre-set H.F. bias control enabling the bias applied to the recording head to be adjusted for optimum results. It is good practice, when servicing a recorder, to check the generator output against the manufacturer's figure. This is usually given in terms of bias voltage measured across the head or of bias current through the head, and may be checked using the appropriate A.C. range of a test meter. If no manufacturer's recommended figure is available, the optimum level of H.F. bias can be found approximately as follows: make a trial recording while progressively increasing the level of the bias in steps, at the same time measuring the bias voltage across the head and reading this off into the microphone with each adjustment; then play back the recording; and decide which level gives the maximum recorded signal. The correct level of bias for optimum results will then be slightly in excess of this figure.

The graph shown in Fig. 21 (*a*) shows how the bias level affects the amplitude of the recording. Above a certain bias level, the amplitude falls, owing to partial erasure of the recording by the bias. The bias level required to give the maximum recorded signal decreases, however, as the frequency of the recording increases, as shown in Fig. 21 (*b*). This is because the higher signal frequencies are more easily erased than the lower ones, and for optimum results therefore require less bias. The bias setting is, therefore, a compromise between the high-frequency and low-frequency optimum levels.

As previously described, one method of achieving this compromise is

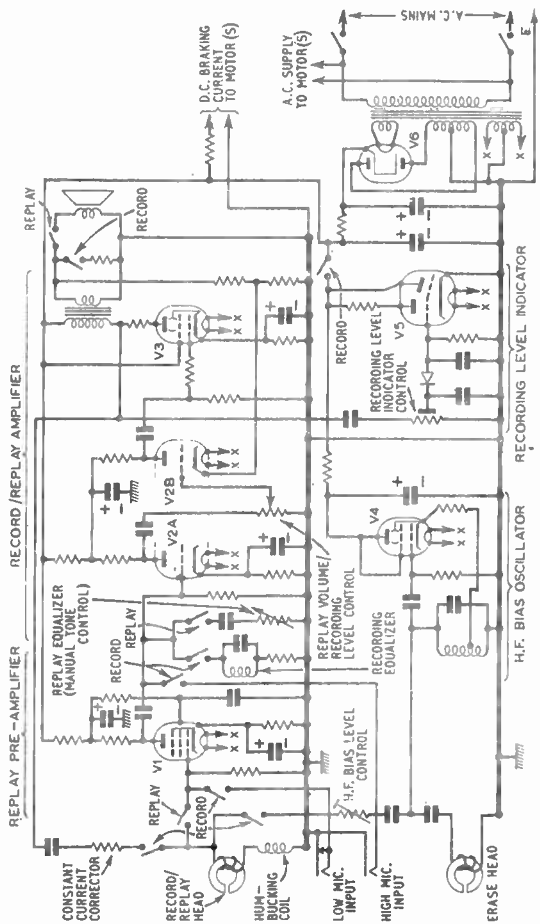


FIG. 20. REPRESENTATIVE CIRCUIT OF A DOMESTIC TAPE RECORDER.

to record a complex waveform, such as speech, containing a band of frequencies, and then to increase the bias level until the amplitude of the recording just begins to fall. A more accurate method is to feed a 1-kc/s signal into the recorder from an audio oscillator, and to adjust the bias for the maximum level of recorded signal. The amplitude of the recorded signal can be measured on playback by connecting a multi-range meter across the output-transformer primary. As can be seen

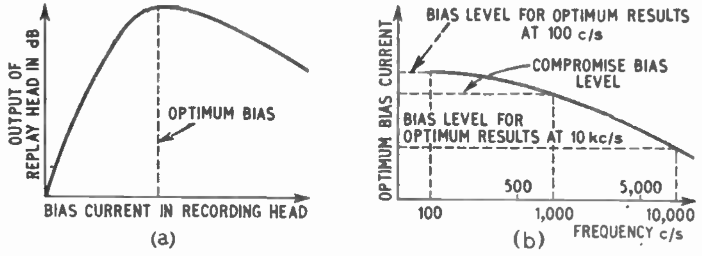


FIG. 21 (a).—SHOWS THE RELATIONSHIP BETWEEN H.F. BIAS LEVEL AND AMPLITUDE OF RECORDED SIGNAL; (b) SHOWS THE RELATIONSHIP BETWEEN THE OPTIMUM H.F. BIAS LEVEL AND THE FREQUENCY OF THE RECORDED SIGNAL.

from Fig. 21 (*b*), setting the bias for maximum output at 1 kc/s results in the lower frequencies being under-biased and the higher frequencies over-biased, but will represent the best compromise over the complete range.

Harmonic Distortion

As important as setting the H.F. bias to the correct level is to ensure that it is free from harmonics. The presence of second-harmonic distortion in the H.F. bias causes its waveform to be asymmetrical, or unbalanced. As an asymmetrical waveform contains a D.C. component, it will give rise to a D.C. magnetic field which will magnetize the tape and thereby increase the background noise in the recording. For this reason, the output valve in a two-stage bias generator is operated under Class A conditions. Another method of minimizing the effects of the harmonic content of the bias is to incorporate a filter, tuned to the second harmonic of the bias filter, in series with the H.F. bias feed to the head.

The bias waveform, which should be as near sine-wave shape as possible, may be inspected by connecting an oscilloscope across the bias output to the head. Any departure from a sine waveform normally indicates the presence of unwanted harmonics.

Background noise in a recording may also be caused by a magnetized recording head. The head may become magnetized if the H.F. bias is cut off abruptly. To prevent this happening when the bias H.T. is switched off, a large-value smoothing capacitor is connected across the H.T. line to the bias generator. This prevents the H.T. voltage falling rapidly when switching off, thus ensuring that the H.F. bias decays slowly.

Fault Finding

Failure of H.F. bias causes distortion, poor volume and an increase in background noise. In tracing the fault, any suspicion of the playback amplifier can be removed by checking it with a test tape. Provided that the fault does not lie in the coupling to the erase head, another symptom of H.F. bias failure will be the failure of the recorder to erase previous recordings from the tape. After making these preliminary tests, make A.C. voltage checks across the recording head and at the control grid or anode circuit of the bias oscillator. Absence of A.C. bias voltages at these points is usually due to an open-circuit oscillator coil, or a faulty oscillator valve or associated component.

Low H.F. bias level may cause only slight distortion in the recording. If the bias generator is also used to feed the erase head, however, previous recordings on the tape will not be completely erased. To confirm this condition, check the bias voltage across both the record and erase heads. Low H.F. bias is generally due to a low-emission oscillator valve, or a high-resistance bias or H.T. feed resistor.

Record/Playback Amplifier

The general design of the record/playback amplifier follows the lines used in good-quality record reproducers. Equalization is incorporated to compensate for the different record and playback characteristics: treble boost is inserted when recording, and bass boost when playing back. It is also common to vary the characteristic according to the speed.

The pre-amplifier stage is particularly sensitive to hum fields and to mechanical vibration. To reduce the effects of vibration, anti-microphonic valve-holders are generally used for this stage. Hum appearing in the output of the pre-amplifier valve may be due to heater/grid coupling, and it is common to guard against this by balancing the heater supply about earth potential, either by means of a centre-tapped heater winding or a potentiometer.

The replay head is also very sensitive to hum fields, and is usually fitted with a screen made from a high-permeability magnetic alloy. Another way of reducing the level of hum output from the head is to connect a hum-bucking coil in series with it. The coil is adjusted until the hum picked up by the head is exactly cancelled by that picked up by the coil. To adjust the coil, carefully orientate it for minimum hum, measuring this by means of an output meter or oscilloscope at the output of the amplifier, while the motor is running.

Microphony is generally due to a faulty valve in the first stage whose electrode system is no longer rigid. It may also be caused by the perishing or hardening of the rubber used in the anti-microphonic valve mounting; it is also not unusual to find the effect of a resilient mounting nullified by tight connecting leads.

Hum may be due to normal H.T. smoothing faults, unbalanced heater supply to the first stage, a hum-bucking coil out of adjustment or a heater/cathode leakage in the pre-amplifier valve.

In the case of distortion, first check that this is not due to faulty H.F. bias. Check the replay section of the amplifier by running a test tape through the recorder. Check the record section of the amplifier by making a recording on the machine and playing it back on a known good machine.

Recording Level Indicator

Tape recorders usually incorporate either a neon bulb, magic-eye indicator or meter to show the recording level. Magic-eye tuning indicators are used in most recorders and often a pre-set control is arranged so that the eye can be adjusted to just close on peaks. If the setting is in doubt, first find the correct recording level by trial and error. This can be done by recording a 1-kc/s signal from an audio oscillator and increasing the setting of the recording amplitude control in steps until distortion is detected on replay (this can best be done by connecting an oscilloscope across the loudspeaker and inspecting the waveform). Once the correct setting has been found, adjust the pre-set control of the record-level indicator, while feeding in the same 1-kc/s signal, so that the "eye" just closes.

The Tape Deck

A typical tape deck lay-out is shown in Fig. 23. The main components are the heads, tape guides, capstan, pinch roller and pressure pads. Note in particular that the heads consist of coils wound on high-permeability cores: on no account should the continuity of the windings

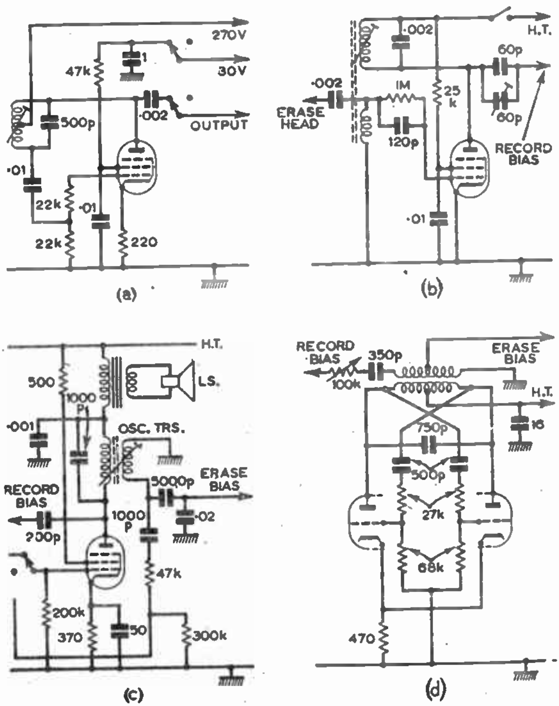


FIG. 22.—TYPICAL BIAS OSCILLATOR CIRCUITS.

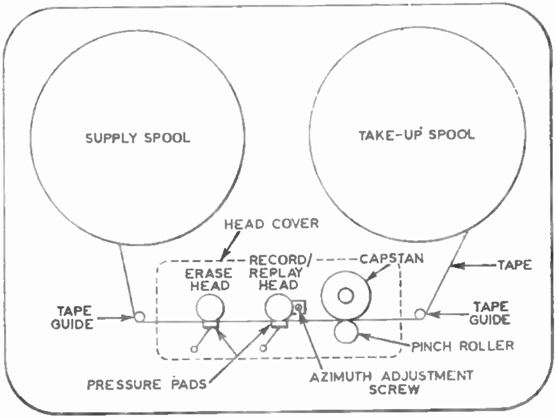


FIG. 23.—LAY-OUT OF TYPICAL TAPE DECK.

be checked with a D.C. continuity meter, as the magnetic field induced by a D.C. will leave the head magnetized and may reduce the permeability of the core material. The only safe way of checking the continuity of head windings is to measure the A.C. through the head, or to test the head on an A.C. bridge.

Magnetized Heads

The presence of residual magnetism in the record/playback head increases the level of background noise when recording: it will also introduce noise into an existing recording being played back.

When making a recording, the H.F. bias fed to the record/playback head provides an effective demagnetizing field to the head. However, the level of H.F. bias required to demagnetize fully the head is greater than the optimum level required for recording purposes. If, as may sometimes happen, the A.F. signal to the head exceeds the level which the H.F. bias can handle, the head will not be completely demagnetized. It is therefore advisible when servicing a recorder to demagnetize the head, particularly before carrying out any checks with a test tape or commercial recording.

To demagnetize the head, the core must be fully saturated with an A.C. field, and the field then reduced slowly to zero. Suitable defluxers designed for this purpose are available.

Head Wear

The head face is slowly worn away by the abrasive oxide-coated surface of the tape. As this occurs, a point is reached where the head gap begins to widen, causing a falling off at the high-frequency end of the response. The build up of a film of dirt on the face of the head can also cause a deterioration in top response, by preventing the tape from coming into intimate contact with the head.

Great care should be taken in choosing a cleaning fluid to clean the head, as some sealing compounds used in the head gap are attacked by tetrachloride or lighter fuel. In the absence of any manufacturer's recommendations, methylated spirits is perhaps the safest cleaning agent.

Head Alignment

Top response can also be adversely affected by the record/playback head being out of alignment. Note that the effect is self-compensating between recording and playing back, and so will only be noticeable when a known good recording is played. The output of a 10-kc/s signal recorded at 7½ in./sec. drops by about 0·6 dB if the head is moved out of alignment by 2 minutes. The alignment is correct when the head gap is exactly at right angles to the traverse of the tape. Screw adjustments are usually provided so that the head can be pivoted. Carry out the adjustment while running the high-frequency (10 kc/s) section of a test tape through the recorder. Connect an output meter across the loudspeaker, and tilt the head for a maximum reading. Several peaks will be found as the head is adjusted: the correct setting is that which gives the largest peak reading.

If a separate recording head is fitted, this can be aligned by making a series of recordings (using a 10-kc/s output from an audio oscillator) while adjusting the head-alignment screw in quarter-turn steps. After each recording, play back the tape and note the meter reading. Set the head in the position giving the maximum reading.

Before adjusting a head which is badly out of alignment, check with the owner of the recorder whether he has made a library of recordings with the machine as it is: if he has, he may prefer to leave the heads out of alignment.

Fault Finding

Poor top response may be caused by the heads being dirty, worn or out of alignment. Noisy recording may be due to a magnetized head.

Wow and Flutter

Wow is caused by a low-frequency variation in the tape speed, usually in the region of 1–3 c/s. It is generally associated with components which rotate at this speed, e.g., it may be caused by a bent capstan or pinch-roller spindle, or by irregularities in the surfaces of these components. Flats may develop on the pinch roller or on the driving surfaces between the capstan and the motor. A roller or wheel rubber which has developed a flat should be replaced.

Wow may also be caused by excessive pressure between the pinch roller and capstan, and the pressure of the pinch roller should be checked (use a spring balance). In the absence of any figures for the correct pressure, thread the tape between the capstan and the pinch roller in the normal way, switch the recorder to playback, take the free end of the tape, and pull it through between the capstan and pinch roller at a constant rate while adjusting the pinch-roller pressure until it is just sufficient to prevent any slipping of the tape on the capstan.

Another possible cause of wow is a distorted spool.

Flutter is due to a tape-speed variation from about 20 c/s upwards. It can often be associated with those parts of the tape deck which rotate at higher speeds than that of the capstan, e.g., by a bent spindle in the capstan-drive motor.

Another cause of flutter is insufficient tape tension or insufficient pressure on the pad which holds the tape against the face of the record/playback head. Both these faults allow the tape to flutter as it passes the head, although the actual flutter may be too small to be discernible to the eye.

Figures for the correct pressure on the record/playback and erase pressure pads may be stated by the manufacturers and can be checked with the aid of a spring balance. Whilst sufficient pressure must be applied to prevent the tape fluttering as it passes the heads, too much pressure will impose an excessive drag on the tape transport and may increase the tension to a point where the tape will snap. Too much pressure will also cause rapid wear of the head faces.

Other forms of speed irregularities may be caused by dirt or oil on motor-drive belts or pulleys. All belts, driving surfaces and friction clutches should be kept clean and free from oil or grease.

Pressure pads which have hardened should be replaced.

A test tape is particularly helpful in detecting wow and flutter, particularly if it is loaded in an endless-tape cassette.

Servicing Routine

The following routine adjustments should be made whenever a recorder is serviced:

(1) Thoroughly clean the deck, paying particular attention to the head faces and the surfaces of the capstan, pinch roller and tape guides.

(2) Demagnetize the record/playback head.

(3) Check the alignment of the record/playback head with the aid of a test tape.

(4) Check the level of the H.F. bias to the record/playback head, and inspect the bias waveform for linearity on an oscilloscope.

(5) Check the setting of the record-level indicator control.

(6) Check the hum level. If necessary, adjust the hum-bucking coil and heater-balancing potentiometer.

Test Equipment

The following test equipment is helpful for tape-recorder servicing: a multi-range meter; an audio oscillator; an oscilloscope; a spring balance (calibrated in grams and ounces) to provide a simple means of checking pressures; a test tape containing a range of fixed-frequency, constant-amplitude recordings (the tape can also be marked out in suitable lengths for checking the running speed of a recorder—18·75 ft. for a 1 minute run at 3¾ in./sec.); a defluxer; an endless tape cassette.

40. TELEVISION RECEIVER INSTALLATION AND SERVICING

40. TELEVISION RECEIVER INSTALLATION AND SERVICING

INSTALLATION

A television receiver, if possible, should be placed in a position where light does not fall directly on the screen. A low-wattage light should be provided, a darkened room may cause eye strain and is unnecessary with modern sets. Place the mains lead and aerial feeder where they will not be easily damaged, and keep the aerial feeder as short as possible.

The mains-supply socket must be rated at least at 2 amperes, and the receiver must be adjusted to suit the mains supply. The correct rating of the mains-supply fuses is also important. Note that A.C./D.C. receivers will operate from D.C. supplies only when the mains plug is inserted into the supply socket the right way round.

Adjusting Controls

The controls provided in a television receiver may be split into two main groups, those that are pre-set, including the time-base controls and mechanical picture adjustments, and the user controls, which may require slight adjustment before each programme to take into account variations in the electricity supply voltage and the effect of the internal temperature of the receiver.

In fringe areas it is advisable to adjust the pre-set controls for average signal-strength conditions, as in these areas a certain amount of fading is to be expected, and unless some such allowance is made they will require constant attention.

Time-base Controls

The controls usually found are the line and frame hold, the line and frame linearity (sometimes called line and frame form), the width (sometimes called line-amplitude) and height (sometimes called frame-amplitude) controls. In many receivers two frame-linearity controls are provided, one affecting the top of the picture only.

The correct procedure is first to adjust the frame- and line-hold controls to give a steady raster—a picture that tends to break into vertical strips or to move up or down indicates that an adjustment of one of the hold controls is necessary. Next adjust the width and height controls so that the picture just fills the screen. Then adjust the linearity controls to compensate for any irregularities, such as cramping, in the picture. Where two frame-linearity controls are provided, the one affecting the top of the picture only should be adjusted first, the overall control being adjusted last. It is sometimes desirable, in order to achieve good linearity, to adjust the height and width controls so as to overscan the tube face when making adjustments to the linearity controls.

Vision Interference and Sensitivity Controls

Two other pre-set controls usually found are the vision-interference limiter and the sensitivity control.

The vision-interference control may take the form of a pre-set potentiometer or a plug and socket adjustable in a number of fixed steps. Most forms of limiter, when set to give maximum clipping, will cause some flattening of the highlights of the picture. Vision-interference limiters should therefore be adjusted to the minimum clipping position consistent with satisfactory reception.

The sensitivity control is fitted to vary the gain of the receiver so that it may be adjusted to suit the signal strength of the area in which it is installed. If the signal strength is too great; sound-on-vision and vision-on-sound interference may be caused, and the sensitivity control will in such conditions need to be turned to the minimum usable position.

Miscellaneous Pre-set Controls

Other pre-set controls commonly fitted are for picture quality, line anti-striation and line drive adjustment. The picture-quality control usually takes the form of a compression trimmer or other arrangement for increasing or decreasing the capacitance in the cathode circuit of the video output valve. This control should be adjusted to give a sharp picture, free from smear, a compromise in some locations having to be sought between Bands I and III transmissions.

Line anti-striation or balancing trimmers are fitted on the deflection coils to balance the stray capacitances in the coils. The adjustment of this is normally carried out in the factory, and should only need readjustment if new deflection coils are fitted. The adjustment is for minimum waviness in the line scanning.

Line drive controls usually consist of a trimmer controlling the drive to the line output valve, and should need no adjustment unless valves in the line time-base are changed. The usual method is to adjust the control so that the E.H.T. voltage is a given value, as specified by the manufacturers. If out of adjustment, a faint white line will appear near the centre of the screen. Damage to the line output valve or transformer can be caused if this adjustment is not correctly carried out.

Mechanical Picture Adjustments

Around the neck of the picture tube are situated the deflection coils and also usually several magnets for such purposes as centring and focusing the picture, and for reducing ion burn.

The deflection coils are housed close up against the bulb of the picture tube, and may be rotated, after loosening the clamping screws, so as to level or square the picture. Correction magnets may also be found fitted against the bulb of the tube to correct for barrel and pin cushion distortion, etc., and to remove corner shadowing. Note that a certain amount of pin-cushion distortion may be desirable on 90° and 110° picture tubes to improve the overall focus. Sideways adjustment affects the distortion, up-and-down movement reduces shadowing. These are factory adjustments which should not normally require alteration. Behind the deflection coils are situated magnets controlling the focusing (if magnetic focusing is employed) and centring of the picture. These are adjustable by means of suitable levers or knobs. At the rear of the picture-tube neck is frequently fitted an ion-trap magnet.

Adjusting Ion-trap Magnets

Ion-trap magnets are normally secured to the neck of the tube by means of a clamp, and to facilitate fitting there is usually an arrow stamped on the magnet, and a line along the neck of the tube. The magnet is normally fitted above the neck with the arrow pointing in the direction of the screen, but alternatively may be fitted with the magnet underneath and the arrow pointing away from the screen. To fit and adjust the ion-trap magnet, the following procedure should be followed, preferably when a stationary test pattern is available.

With all power switched off, and reservoir capacitors discharged if necessary, the magnet is pushed over the base of the tube with the arrow pointing towards the screen, and placed immediately over the line marked on the tube neck. The cathode-ray-tube socket is then replaced, the receiver reconnected to the mains supply, ensuring that the chassis is not above earth potential, and the brightness control set to a position where the raster is just visible. To achieve this, it may be necessary to adjust the position of the magnet slightly.

Then with the arrow over the line, the magnet is moved towards the screen until the focused raster is at its brightest. The brightness control is then re-adjusted until the peak-white portions of the image are at a correct level, and, if necessary, the position of the magnet adjusted slightly to obtain maximum brilliance.

Where the picture cannot be centred by adjusting the position of the focus field, the ion-trap magnet may have to be rotated slightly around the neck; this operation, however, should not lead to any decrease in brilliance.

When the picture fulfils the above requirements, lock the magnet in position by tightening the thumbscrew, ensuring that the magnet does not change position while this is being done.

Should it not be possible to obtain a position of maximum brilliance, it may be necessary to substitute another magnet. The magnet should never be adjusted to remove a shadow if this involves reducing the brightness of the picture; this should be done by adjusting the focus coil, deflection coils and/or correction magnets.

Always handle an ion-trap magnet with care: it should not be subjected to strong magnetic fields or mechanical shocks. It should not be allowed to come into contact with metallic objects.

Tuning Signals

The B.B.C. and I.T.A. both radiate tuning signals for about five minutes before each transmission, to enable viewers to adjust their television receivers correctly in readiness for the start of the programmes. In addition, there is the Test Card "C", shown in Fig. 1. A test card is transmitted by both B.B.C. and I.T.A. during the mornings to help service engineers.

Test Card " C "

Test card " C " has been designed to give an immediate indication of the performance of the whole transmitting and receiving chain. As the performance of the transmitting equipment is maintained in accordance with the agreed standards during the normal periods of radiation for test purposes, this Test Card " C " can serve as a check on propagation and the performance of the receiving apparatus.

The card, which bears the identification letter " C ", incorporates a number of patterns, each designed to assess one particular characteristic of the system, thus:

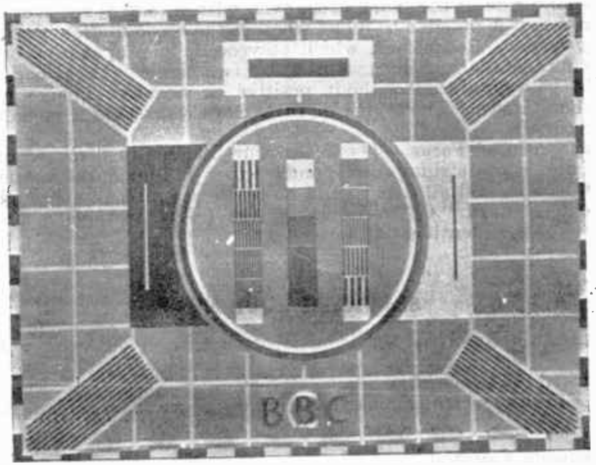


FIG. 1.—B.B.C. TEST CARD " C ".

Aspect Ratio.—Concentric black and white circles surrounding the five frequency gratings will appear truly circular when the width and height of the picture are adjusted to the standard aspect ratio of 4: 3.

Resolution and Band-width.—Within the circles there are two groups of frequency gratings, each consisting of five gratings having black and white stripes corresponding to fundamental frequencies of 1·0, 1·5, 2·0, 2·5 and 3·0 Mc/s. In the left-hand group the 1·0-Mc/s grating is at the top, the frequency increasing towards the bottom, and in the right-hand group the order is reversed. The response of the whole system is required to be uniform to 2·7 Mc/s, so that the 2·5-Mc/s grating should be clearly reproduced, but the 3-Mc/s gratings may be blurred. The picture must just fill the viewing aperture during the test, with the black-and-white border visible.

Contrast.—A five-step contrast wedge appears in the centre of the test card. The top square is white, corresponding to 100 per cent modulation, and the lowest square is black, corresponding to 30 per cent modulation. The three intermediate squares should be reproduced as pale, middle and dark grey.

Scanning Linearity.—The background of the test card is a middle grey, bearing a graticule of white lines. The areas enclosed between the lines should be reproduced in all parts of the picture as equal squares.

Synchronization Separation.—The border consists of alternate black and white rectangles which facilitate recognizing interference between the picture signals and the synchronization.

Low-frequency Response.—A black rectangle within a white rectangle is provided, and in a perfect system it would be reproduced as a rectangle of uniform blackness on a clean white background. At present imperfections in the transmitting system result in a slight streaking at the right-hand side of the black area, even with a perfect receiver, but by experience it is possible to judge whether the reproduction is abnormal.

Reflections.—Reflections, which may occur in propagation or in the receiving installation, are indicated by two single vertical bars, which should be reproduced without positive or negative images at their right-hand sides. The width of these bars represents a pulse of 0·25 microseconds.

Uniformity of Focus.—There are four diagonally disposed areas of black and white stripes corresponding to a fundamental frequency of about 1 Mc/s, and all four should be resolved uniformly throughout.

The Test Card "C" radiated by the I.T.A. stations is the same except for the identification letters.

SERVICING EQUIPMENT

Whilst radio-receiver servicing is often carried out with a limited amount of test equipment, test gear of a fairly comprehensive nature is required for television work.

A useful list, which is not exhaustive but serves as a general guide, is as follows :

Absolutely Essential

1. Signal generator.
2. Universal meter, 20,000 ohms per volt.
3. Pattern generator, giving a true B.B.C./I.T.A. pattern.

Very Desirable

4. Oscilloscope.
5. Wobbulator.
6. Insulation Tester.
7. Component Test Bridge.

Well Worth While

8. Valve Tester.
9. E.H.T. Voltmeter.
10. Signal-strength Meter.
11. Valve Voltohmmeter.
12. Crystal Calibrator.

When investing in test gear, it must always be borne in mind that the criterion is that the equipment should earn its keep. On the other hand, equipment unintelligently used will never pay. To be satisfactory, the gear must be reliable and always give the same reading in similar circumstances. It must be reasonably easy to handle, so that it does not mislead. Nevertheless, unless test gear is handled and used wisely it can be a time waster. Hours can be spent on the niceties of alignment, which, although they may appear to have considerable influence on the measured responses, hardly alter picture quality one iota.

Signal Generators

At one time these could be divided into two classes, more aptly named test oscillators and standard signal generators. The former consists of a tunable oscillator, which can be modulated, together with means of attenuating the radio-frequency output. The second class has all these features, but also has a means of setting the output level to a known and repeatable value on all frequencies. The test-oscillator output varies, perhaps widely, with frequency, rendering comparative tests sometimes quite misleading. Years ago the difference in performance was very great, but clever design has produced most efficient test oscillators with remarkably constant output which enables quite good relative checks to be made.

It would be preferable if service stations used metered signal generators exclusively, but the modern cheaper variety is so good, in its best examples, that cost often overrides. It is, however, suggested that all service stations should have at least one standard metered generator to serve as a check against deterioration of performance.

One essential feature of a signal source is good frequency stability and resetting accuracy. Unfortunately the requirements of television sets working on single side-band are very high, in fact higher than can be reasonably expected from the most expensive laboratory generator. No tunable oscillator of normal type can be expected to hold over a long period to as good as 0·1 per cent, including resetting accuracy, and few manufacturers guarantee better than 1 per cent. The variation of stability of the television set will absorb most of the tolerance available, therefore the source stability should be much better so that the error from it is negligible. This means that it should be certainly within 20–30 kc/s of nominal, which is round 0·03–0·04 per cent.

Clearly if we demand such stability from our signal generators we shall either delude ourselves or be disappointed.

The simplest way of overcoming this is to beat the carrier of the signal generator with the station, so synchronizing them, and noting how much the generator deviates and how it drifts during working periods. A little experience soon shows how frequently this must be done, and how closely the required channel can be set up without reference to the station. This is not an ideal method, and sometimes it is not possible to use it. The ideal method is a crystal calibrated wavemeter or oscillator.

Universal Meters

The ubiquitous Avometer here comes to mind, although there are other good instruments available. For television work it is often essential to have a high-resistance meter of 20,000 ohms/volt, such as the Avo Model 8. However, some manufacturers' service data still calls for a 1,000-ohms/volt meter such as the Avo Model 7, so that it is normally necessary to have both types at hand.

In choosing an instrument it is very important to be sure that good overload protection is incorporated. The extra cost is saved in every shop sooner or later.

Pattern Generators

A television pattern generator should be used not only when the actual transmissions are not on the air, but at all times. The real picture content is continually varying, and sometimes is of variable quality. It is also subject to interference. A constant pattern is far more satisfactory for servicing, and is not misleading. The statement that of course the final criterion is what the actual transmission looks like, is not as all embracing as might appear.

This assumes that the pattern generator and the actual transmission have equal effect on the receiver as to synchronizing, etc. Now the picture waveform is extremely complex, and this complexity is necessary. A television pattern generator must therefore give a fully sychronized interlaced pattern having all the characteristics of the real transmission. A receiver adjust l on such a pattern generator will then immediately give a current picture on the station, provided that the correct radio-frequency level is used. This applies to all phases of servicing adjustments—linearity, aspect ratio, interlace, hum level, sound on vision, etc. All can be checked with a good pattern generator with certainty that all will be well when the station comes on.

But if the pattern generator does not give exactly the correct waveform; if it does not incorporate interlace, or have line and frame frequencies derived from a single oscillator, or half-line pulses, or the front and back porches, or any such feature, many of the most valuable features may, in practice, be lost. The reason for this drastic statement is that such an instrument gives misleading results, which must be avoided at all costs. On such a pattern generator any adjustment made cannot be guaranteed correct for the actual transmission—it may look right on the pattern generator and be wrong on transmission,

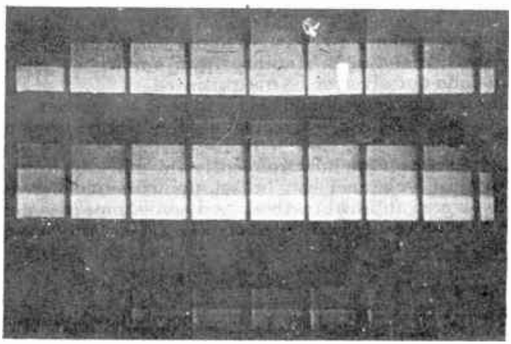


FIG. 2.—TELEQUIPMENT WG/42 PATTERN.
This photograph shows non-linearity and ringing. The number of vertical lines is adjustable from about 5 to 20.

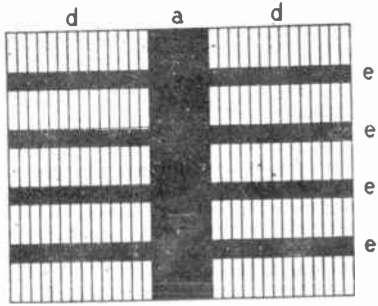


FIG. 3.—MURPHY TPG.11 PATTERN.

(a) 20,250-c/s master oscillator central vertical bar and line blanking pulse.
(d) 300-kc/s vertical grey bars.
(e) 250-c/s horizontal black bars and frame frequency blanking pulse.

or vice versa. Some receivers will in fact only function properly when fed with an interlaced pattern.

It may be mentioned at this juncture that the optimum pattern generator for use on sets with flywheel synchronization is not the optimum for sets without this facility. In order to cope with both cases a compromise has to be adopted, which leaves so little to be desired that most would be unaware of the deficiency.

Oscilloscopes

There is much difference of opinion over oscilloscopes, some finding them of inestimable value, and others not. Probably the reason for this lies partly in the handling, and in knowing how to interpret the results. If the 'scope amplifiers have a reasonably high gain, and the input of the 'scope is connected to a high-impedance circuit, the resulting trace may be modified or completely upset by spurious pick-up, mains hum, etc. The attachment of the 'scope input may also disturb the operation of the circuit under test. Only experience can show how to overcome these difficulties in different cases. In the writer's opinion the greatest value of a 'scope is in conjunction with a wobbulator.

Some points to look for in choosing a 'scope are: good sensitivity variable over a wide range, wide frequency range and a good linear time-base. For many purposes a good high-frequency response is

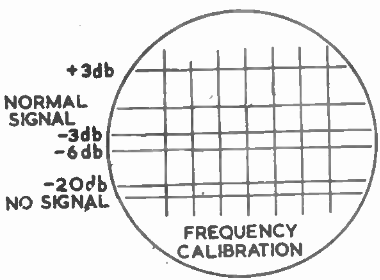


FIG. 4.—TYPICAL GRATICULE LAYOUT.

unnecessary, but it is invaluable when checking a pattern generator, or when looking for spurious high-frequency oscillations on low-frequency circuits.

Wobbulators

A wobbulator, or frequency modulated oscillator, together with an oscilloscope, is undoubtedly far the best apparatus for lining up television sets to the correct band-width accurately and quickly. As usual, however, there are snags. The apparatus must be right, it must be connected appropriately and must be used correctly. Matters have not been helped in that makers' instruction books have not emphasized the pitfalls sufficiently.

In the first place, optimum alignment should be made at the working sensitivity of the receiver, or at a lower input signal. As most reasonably priced wobbulators are not fitted with accurate attenuators, this sometimes leads to trouble. Secondly, the oscilloscope must be connected in such a way as not to affect the operation of the receiver. Practically, the output is normally taken from grid or cathode of the cathode-ray tube. If a clip is fitted to a resistor of 50,000 ohms upwards, which is connected in series with the conductor of a screened cable, with a further clip from the screening for earthing, this can be connected to the 'scope and will not usually upset the receiver at all. If it does, slight readjustment of leads or resistance should remove the trouble. Next factor of importance is the sweep frequency. This must be low; 16⅔ c/s (⅓ mains frequency) is good, 25 c/s should be the maximum. This is on the limit of a small oscilloscope (and some larger ones), which distorts the waveform if the frequency is too low for it. It is also good to connect a capacitor of about 0·01 μF across the input to the 'scope amplifier.

The method of operation of a wobbulator differs somewhat with different sets. Usually if the set is not far from alignment, injection into the aerial or at intermediate frequency, and alignment in order working from the circuits nearest the second-detector through finally to the aerial, is the drill. Sometimes it is necessary to inject into each

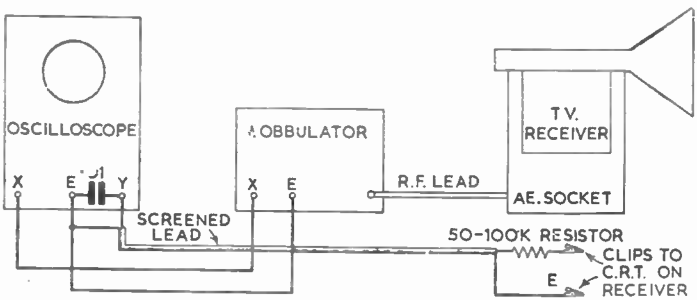


Fig. 5.—Wobbulator Connections.

N.B.—Series earth capacitors to be fitted where necessary.

grid in turn, dealing only with the immediately subsequent circuits in turn, arriving finally at the aerial end. Each type of receiver has its own idiosyncrasies, and experience is the only way to find the optimum method in any particular case. It is a good thing to make a graticule showing the calibration of the instrument in frequency in the horizontal (X) direction and calibrated in db in the vertical (Y) direction, say at 3 db up, 3 db, 6 db and 20 db down on the normal deflection (3 db down is 70 per cent, 6 db down is 50 per cent, 20 db down is 20 per cent of normal, but 3 db up is 140 per cent). Here is a wonderful opportunity for time-wasting trying to get a perfect curve. It must be remembered that getting the carrier at the 3-db or 6-db point is important on a single-side-band set, but that dips or rises of 2 or 3 db over the band will hardly affect the picture. As long as the carrier is right and the bandwidth adequate, small irregularities are of no consequence.

Another important thing is the frequency setting of the wobbulator. If a signal from the generator is injected into the wobbulator (if provided with a suitable terminal) or through a T pad with the wobbulator, a kink will be seen on the trace at the frequency corresponding to the generator frequency. The 0·01-μF capacitor across the scope will prevent the high-frequency beats from widening the trace, and enable an accurate adjustment to be made. Care must be taken that the signal-generator signal is not too strong, or it may modify the trace from the receiver, and so mislead. A better method is to feed from the T pad into a crystal rectifier with a load of say 10,000 ohms in series. The 'scope (with 0·01-μF condenser) is then connected across this 10,000-ohm resistor. The marker can then be used to check the linearity of sweep and calibrate a graticule.

A wobbulator correctly used often shortens alignment time to between one-tenth and one-fifth of the time taken by ordinary methods, including dampers, etc. The technique of using it is, however, of vital importance.

Insulation Testers

A large proportion of faulty components in television receivers have faulty insulation, and the writer believes that much inferior performance can be attributed to leaky condensers in particular. In the writer's laboratory all such components are tested on 500 volts D.C., with quite startling results. For such tests an instrument giving a maximum reading of 200 MΩ is the minimum requirement, but high accuracy is not needed. The Wee Megger Tester needs no introduction, and particularly when extreme portability is required, such types with hand-driven

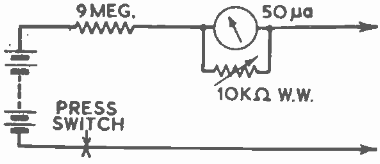


FIG. 6.—SIMPLE INSULATION TESTER.

Calibration of meter:

Current (μA)	50	45	40·9	32·1	28·1	23·6	18·8	15·5	7·6	4·1	2·1	0
Resistance (MΩ)	0	1	2	5	7	10	15	20	50	100	200	∞

D.C. generators have no rivals. A very convenient mains-operated device is the Taylor Model 130A, which derives the 500-volt supply from the mains. It is also possible to make a very simple instrument using eight 67½-volt deaf-aid-type batteries, a few resistors and a 50-μA meter. The battery drain never exceeds about 50 μA, and the life is extremely long; some have been in use for seven years before running down.

Component Test Bridges

Many faults arise due to components, mainly capacitors and resistors, changing in value. These are best checked on a bridge. The requirements are that the full probable range of values of capacitance and resistance can be measured reasonably accurately, say to within 5 per cent at worst. Even this is a fairly stringent requirement. It is sometimes an advantage to have the facility to measure inductance as well, but for a number of reasons this facility is not as useful as it would appear.

Valve Testers

A valve tester can be a great help, but it must be borne in mind that it is no reflection on the instrument that it will undoubtedly fail to reveal some valve faults. If it were not for this, a valve tester would rate higher in the priority list. The writer personally prefers a valve characteristic meter, such as the Avo, which enables values to be measured on a meter which is perhaps an understandable bias!

E.H.T. Voltmeters

The measurement of E.H.T. voltages is a very difficult problem. Very little current is available for a meter, and leakage due to humidity is a particular snag. Two methods, which can be satisfactory, are to use an electrostatic meter, or a very low current (20–25 μA) meter with high series resistance. Both methods are unfortunately costly. Some results can be obtained by connecting, say, three 5-kVA electrostatic meters in series, and summing the readings, but great circumspection is required.

Signal-strength Meters

Such a device as a signal-strength meter is valuable in two main connections. Firstly, in assessing strength of signal at new locations, and secondly, in testing existing aerial installations. It is important to make such measurements not on picture content, but on peak white or synchronizing levels.

Valve Volt-ohmmeters

Such instruments do approximately all that a Universal Test Meter will do, and much besides. Unfortunately they are fairly delicate and a shade tricky to handle. If such difficulties are appreciated, and care used in getting results, they are most useful adjuncts to any test shop.

The uses are legion, but good engineering skill or supervision is needed. The lack of these and a failure to appreciate the limitations of technique have given this type of instrument a bad name not altogether deserved.

Conclusion

Instruments are not the only items of importance in a repair shop. The obvious ones of good bench space and so on need no stressing. But a point overlooked too often is that television sets are lethal. Every test position must be equipped in accordance with the Factory Acts and regulations.

An essential is a properly screened double-wound transformer of 1 : 1 ratio. The screen should be well earthed. An earth bar may be fitted at the back of the bench, but it must only be connected to earth through a series capacitor of high test voltage and not more than 0·02 μF capacitance. Where earth terminals of instruments would be connected effectively direct to the mains via a live chassis, isolating condensers of similar type are essential.

The whole load of the bench or test position, for one engineer only, should come from one transformer secondary, connected to a multiplicity of sockets of all likely types and sizes. Two engineers should never share a transformer—they could get two chassis not very far apart connected to opposite sides of the A.C. supply. This is a very real problem, which has to be solved in the light of the particular circumstances in the shop.

ALIGNMENT

A circuit diagram of the R.F. and I.F. stages of a representative modern television receiver, fitted with a turret tuner, is shown in Fig. 7 (*a*). The operation of the circuit is, briefly, as follows: IFT V3 is an overcoupled transformer providing a simple double humped curve as shown in Fig. 7 (*b*). Almost the entire overall curve shape is determined by the network between V3 anode and V4 grid. This consists of IFT V1 (L11), L20 and IFT V2 (L12). L11 and L12 are a bandpass pair, bottom coupled by L20, which is bridged by C24, C25 and R20. This is known as a " bridge T " coupling, where L20, C24 and 25 are resonated at the sound frequency to produce a high degree of rejection by reducing the coupling between L11 and L12. R20 is a balancing resistor which is chosen during design for optimum rejection with the minimum effect upon the vision response curve. The overall curve from V3 grid is shown in Fig. 7 (*c*) and the sharpness of rejection is apparent. It is also seen that the overall shape is rounded. This shape is used to overcome picture distortions such as rings and overshoots which arise with flat-topped curves.

The connection between L11 and the " bridge T " filter passes through the primary of IFT S1. There is a considerable amount of sound-frequency energy available at the input of the filter, due to its resonance, and in this way it may be passed to the sound amplifier V11. The output of V11 is coupled to the sound detector via IFT S2, and the overall sound curve from V3 grid is shown in Fig. 7 (*d*).

The circuit coupling the mixer in the tuner is a broad one, since it has to transfer both vision and sound. Also, the tuner being a separate

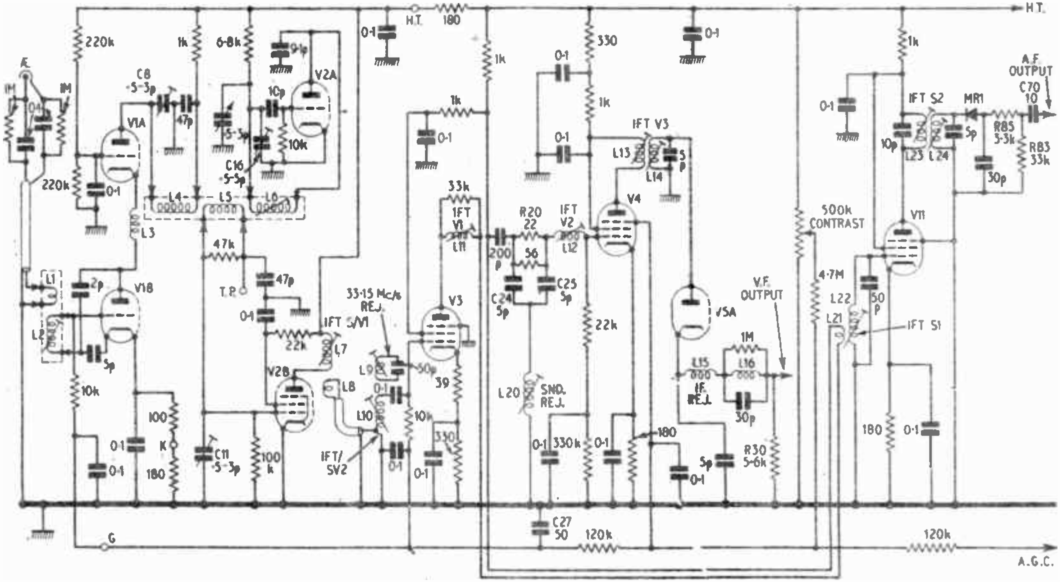
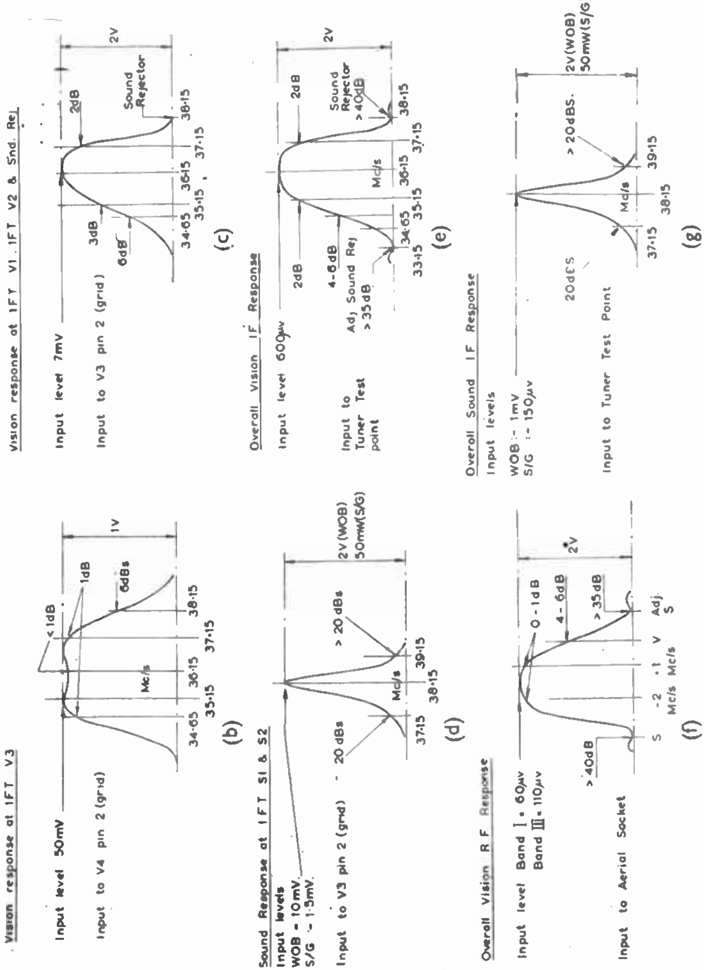


FIG. 7 (.).—CIRCUIT DIAGRAM OF THE TURRET TUNER UNIT AND I.F. STAGES OF AN R.G.D. RECEIVER.



unit, the circuits IFT SV1 (L7) and IFT SV2 (L10) are coupled via a cable link. L9 is a trap circuit coupled inductively to L10 to produce a rejection at the I.F. produced by sound frequency interference from the next channel. The type of response given by this network is broad and flat except for the rejection due to L9. The overall response at the grid of V2 (tuner test point) is given in Fig. 7 (*e*).

Further information on the formation of response curves in I.F. stages is given in Section 15.

The Turret Tuner

In the tuner (see Fig. 7 (*a*)), L4, L5 form a bandpass at signal frequency between V1A anode and mixer (V2B) grid. The response is again broad to transfer vision and sound without affecting the overall I.F. curve.

L6 is the oscillator coil appropriate to each channel, the oscillator voltage being fed into L5 by mutual coupling between the coils. L6 has a screw core adjustment.

L1, L2 is the aerial circuit, a coupled transformer, forming with the response of L4, L5 an overall flat curve for the tuner. The overall curve for the set is shown in Fig. 7 (*f*).

It is essential for the aerial circuit to be tuned correctly to the centre of the band for maximum gain and lowest noise on the picture (especially where signals are weak). L1, L6 are the only adjustments normally available in a turret tuner. The coils are built on to removable plastic " biscuits ", and the inner coils L4, L5 are set up and sealed in manufacture.

All other adjustments in the tuner are for manufacturing set up only, and should *not* be used unless a complete realignment is being attempted.

The Switch Tuner

A switch-type tuner can equally well be used, and the circuit of one is shown in Fig. 8. Here, instead of having the aerial and oscillator circuits of each channel to tune when set to a given channel, all the oscillator circuits are available through the front.

The aerial, R.F. anode and mixer-grid circuits have screw adjustments only for Band III as a whole and Band I as a whole.

Because all the circuits are in series, alignment of any circuit must be done by working down from Channel 13, one channel at a time. If, however, only one channel is in use in each band, it is permissible to use the adjustments appropriate to the whole bands only, *i.e.*,

Band	*Aerial*	*R.F. anode*	*Mixer grid*	*Oscillator*
Band III . .	L9	L11	L27	L37
Band I . . .	L7	L13	L22	L36

Normally only the oscillator circuit should be considered and the adjustment for the appropriate channel tried first.

The aerial circuit can well be trimmed after valve changing. The makers' instructions alone should be followed when attempting complete realignment.

Intermediate Frequencies

The first superheterodyne receivers used intermediate frequencies near 10·5 Mc/s for vision and 14·0 Mc/s for sound. Interference problems proved serious, and a movement was made to 16·0 Mc/s vision

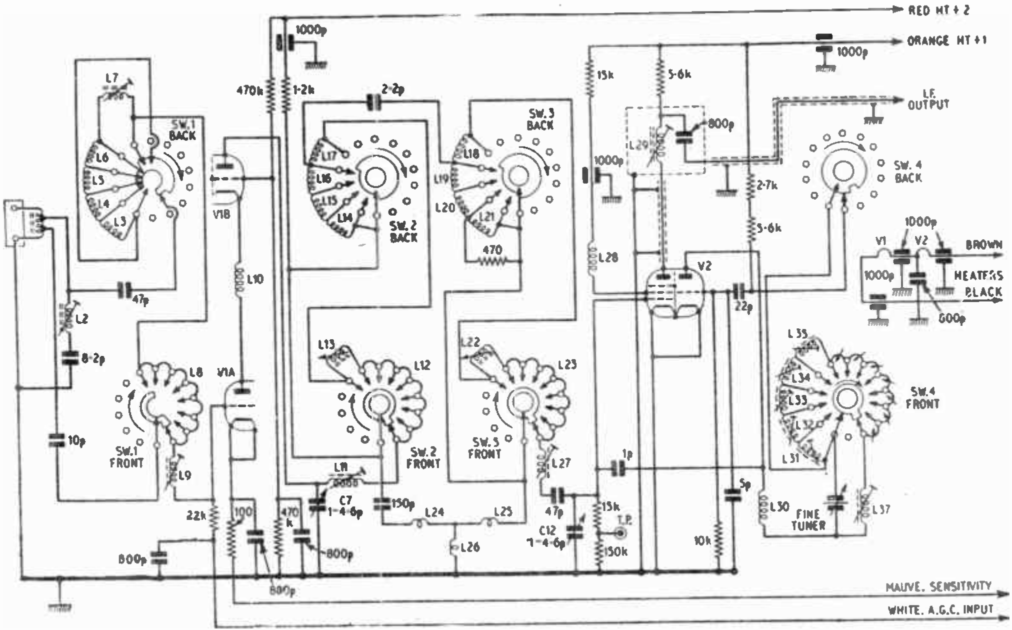


FIG. 8.—BAND I/III TUNER UNIT OF THE INCREMENTAL SWITCH TYPE.

and 19·5 Mc/s for sound. Difficulties also arose with these frequencies for sets tuning to Band III in that the "image" was too close for good rejection and oscillator radiation was often high. Finally, B.R.E.M.A. set a standard of 34·65 Mc/s for vision and 38·15 Mc/s for sound.

It is essential that I.F. amplifiers are aligned accurately to the frequencies specified, otherwise interference problems and difficulty in aligning the tuner may result.

A.G.C.

Automatic Gain Control circuits are usually found in both sound and vision channels. The sound A.G.C. is often simple and consists of a connection via a smoothing circuit between the detector output and I.F. valve(s) grid returns. The vision A.G.C. may be more complicated and the R.F. stage may have the A.G.C. voltage delayed. Makers' instructions should always be followed when aligning sets with A.G.C. When in doubt, apply −2 volts approximately to vision A.G.C., short-circuit sound A.G.C. and R.F. stage A.G.C. only when delayed.

Equipment Required for Signal Generator Alignment

The following list is based on the equipment required when aligning the circuits of Fig. 7 (*a*).

(1) An accurately calibrated signal generator giving C.W. and modulated output with an output impedance of 75 ohms, and having a range of 30–40 Mc/s for I.F. alignment, 40–75 Mc/s for Band I and 170–200 Mc/s for Band III R.F. alignment. The co-axial output lead should be terminated with a 82-ohm resistor (for I.F. alignment only) and the connection leads from the terminated co-axial lead *must* be kept as short as possible.

(2) A vision output meter. This may be a 20,000-ohms/volt meter switched to the 10 volt D.C. range (meter resistance not less than 50,000 ohms), e.g., an Avo Model 8 (not an Avo Model 7) in series with a 5·6k resistor on the hot side. Connect across R30. Alternatively, a 1,000-ohms/volt meter switched to the 1-mA D.C. range (meter resistance not more than 500 ohms), e.g., Avo Model 7 (not Avo Model 8), may be used. Connect one lead to chassis, and the other in series with the earthy end of R30, by-passing the leads with a 1,000-pF condenser.

(3) For use as an I.F. transformer shunt, a ¼-watt, 1k resistor in series with a 1,000-pF miniature ceramic condenser, with short leads, are required.

(4) A sound-output meter. This may be a 3-ohm sound-output meter or an A.C. meter switched to the 1-volt or 50-volt A.C. range.

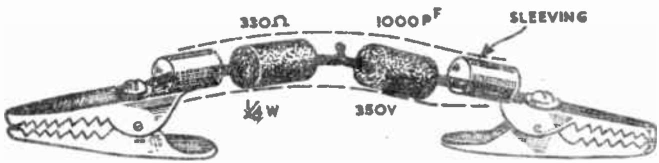


Fig. 9.—Damping Device.

A 3-ohm sound-output meter should be connected across the sound-output transformer secondary in place of the loudspeaker. If an A.C. meter is used switched to the 1-volt A.C. range, it should be connected across the sound-output transformer secondary: if the A.C. meter is switched to the 50-volt A.C. range, it should be connected across the shunt-output transformer primary. In either case the loudspeaker should be left connected or a 3-ohm load connected in its place.

Procedure for Signal Generator I.F. Alignment

The following procedure relates to the circuits shown in Fig. 7 (*a*).

It is good practice, whenever alignment of any stage is required, to carry out complete re-alignment of the receiver.

Connect the signal generator output to the tuner test point for all adjustments, and switch the tuner to a Band III channel for which coils are fitted (but not a channel which the set receives). Turn the contrast control to minimum and the volume control to maximum.

During vision I.F. alignment, adjust the input level of the unmodulated I.F. input so as to maintain 2 volts D.C. across, or 400 μA through, R30.

For sound I.F. alignment, use a 30 per cent modulated signal, and adjust the input level to maintain an output of 50 mW or 0·5 V r.m.s. across the sound output transformer secondary, or 25 V r.m.s. across the sound-output transformer primary.

Carry out the adjustments in the order given in Table 1

TABLE 1.—I.F. ALIGNMENT OF BAND I/III RECEIVER

Step	*Inject (Mc/s)*	*Mod./C.W.*	*Shunt*	*Adjust*	*Response*
1	38·15	C.W.	—	L20	Min. vision
2	38·15	Mod.	—	L24	Max. sound
3	38·15	Mod.	—	L23	Max. sound
4	38·15	Mod.	—	L22	Max. sound
5	Unscrew L10 core flush with base of former.				
6	Set L9 core down $\frac{9}{16}$ in. from top of former (this is approx. working position).				
7	35·75	C.W.	V4 pin 7 (Anode)	L14	Max. vision
8	35·75	C.W.	V5A pin 7 (Anode)	L13	Max. vision
9	36·75	C.W.	V3 pin 7 (Anode)	L12	Max. vision
10	36·75	C.W.	V4 pin 2 (Grid)	L11	Max. vision
11	Repeat operations No. 7, 8, 9, 10.				
12	35	C.W.	—	L10	Max. vision
13	36·75	C.W.	V3 pin 2 (Grid)	L7	Max. vision
14	33·15	C.W.	—	L9	Min. vision

The correct tuning position for all cores except those specified in the following paragraph is the peak nearest the adjustment end.

The correct tuning position for the cores of IFT S1 (L22), IFT V2 (L12) and IFT V1 (L11) is the peak nearest the top (above chassis).

Note that L9 will not tune unless L10 has been correctly adjusted, and that the first operation must always be the sound rejector (L20).

The signal-generator frequency setting should not be disturbed when carrying out steps 1, 2, 3 and 4 in Table 1, and the shunt should always be connected between the nearest point on the chassis and the point specified in Table 1, using the shortest possible leads.

When the I.F. alignment procedure has been completed, the overall sound and vision response curves should be checked against those shown in Fig. 7.

Sweep Generator I.F. Alignment

The following sweep-generator alignment procedure is for the circuits shown in Fig. 7 (*a*). The tuner should be switched to an unused Band III channel for which coils are fitted, and the contrast and volume controls turned to minimum. Connect a 1·5-V bias battery (negative lead) to the A.G.C. line at C27. The co-axial output lead from the sweep generator should be terminated with an 82-ohm resistor (for I.F. alignment only) and the connection leads from the terminated co-axial lead must be kept as short as possible. For vision alignment, connect the sweep generator "Y input" to the junction of L15/L16. For sound alignment, connect the sweep generator "Y input" to the junction of R85, R83 and C70. No I.F. transformer shunts are used.

The stages should be aligned in the following order: IFT S1 and S2 (to response curve shown in Fig. 7 (*d*)); IFT V3 (to response curve shown in Fig. 7 (*b*)); IFT V1, V2 and sound rejector (to response curve shown in Fig. 7 (*c*)). Then check that the overall vision and sound I.F. response curves are as shown in Figs. 7 (*e*) and 7 (*g*) respectively. Connect the wobbulator as indicated in the figures, adjusting its output to maintain the cathode-ray tube trace amplitude quoted on each curve.

Turret Tuner Adjustments

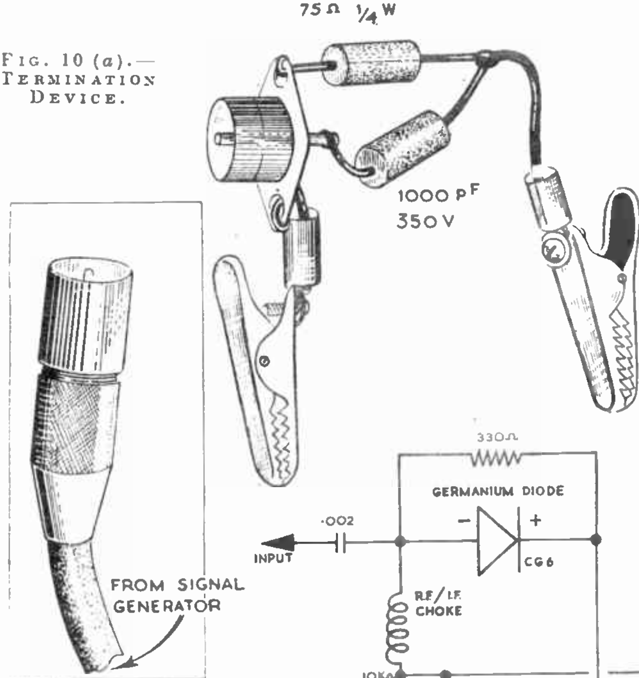
As has previously been pointed out, the only adjustments for the service engineer are the oscillator cores and aerial input transformer. The usual adjustment in each case is for maximum sound, with signal input at appropriate channel frequency, to the aerial socket, and fine tuner at mid-travel. Further alignment requires special test equipment not generally available to the service man.

Miscellaneous Information

A signal-generator termination device is shown in Fig. 10 (*a*). It is intended for use with a signal generator having an output impedance of the order of 75 ohms, and should be connected in circuit when injecting a signal directly to the grids of the valves. It is not required when injecting signals at the aerial socket.

The damping device shown in Fig. 9 is for damping the primary of I.F. transformers whilst the secondary is being tuned, and vice versa. The crocodile clips must make good electrical connection.

Fig. 10 (a).—Termination Device.



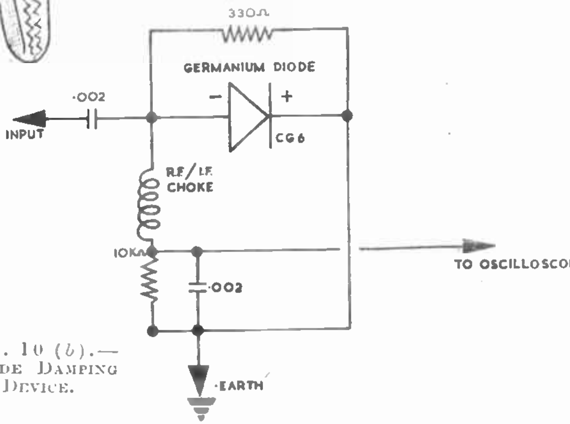


Fig. 10 (b).—Anode Damping Device.

For convenience, both these units may be made up ready for use.

A sweep generator and oscilloscope may be used to examine the response of tuned circuits individually in case of need. To do this, connect a detector made up compactly to the circuit of Fig. 10 (b) to the anode of the valve following the circuit under test, and connect the wobbulator to the preceding grid.

SAFETY PRECAUTIONS

In normal circumstances the risk of shock is no more pronounced with television receivers having mains-connected chassis than with broadcast receivers using similar power arrangements. However, it should be noted that, in the case of a television receiver, certain routine adjustments, such as centring the picture, permanent-magnet focusing and the like, can be carried out only with the protective back removed from the cabinet. While it is possible to check that the chassis is not above earth potential by means of a neon bulb or A.C. voltmeter, reversing the mains plug in its socket where this is found necessary, a safer method in the busy servicing department is always to feed the receiver from an isolating transformer having a low-leakage inductance.

The service engineer should ensure that during servicing no alterations are made to a receiver that might invalidate the manufacturer's safety precautions, and should bring to the notice of the owner any deficiencies in receivers which do not fully comply with modern practice.

E.H.T. Voltages

Where receivers incorporate E.H.T. mains-transformer windings, extreme care should be exercised when servicing, as a shock from such a power source may be lethal. Whilst other forms of E.H.T., such as line-flyback, radio-frequency or pulse oscillators, are less dangerous, serious burns and shocks are still possible, and due respect should be paid to all points at E.H.T.

General Precautions

(1) Never attempt to measure the voltage at the anode of a line-output valve directly, as the E.H.T. pulses will damage the meter.

(2) Before removing the picture tube make sure that the E.H.T. capacitor is discharged: remember that an Aquadag coating may hold its charge for long periods.

(3) Removal of a scanning-coil connection plug whilst the receiver is operating may cause picture-tube screen burn.

Projection E.H.T. Dangers

The 25-kV. electron beam of projection models produces soft X-rays, which are normally shielded from the operator by the optical box. Should it be necessary to operate the cathode-ray tube outside of the optical box, it is recommended that a lead-glass shield be used. The equivalent lead thickness of the shield should not be less than 0·5 mm.

Handling Cathode-ray Tubes

It is important that care should always be taken when handling cathode-ray tubes, in order to avoid the risk of implosion, the accidental dropping of the tube, or damage to the face of the tube.

Information on the simple precautions which should be taken when removing or replacing picture tubes is given in the Section on Cathode-ray Tubes.

B.S.I. Safety Recommendations

In order to protect the public from the dangers of mains-connected chassis, a number of recommendations have been drawn up by the British Standards Institution (B.S. No. 415). The most important sections of the B.S.I. are in regard to the measures to be adopted to prevent the user from having access to "live" parts of the apparatus. Whether or not any particular part is "accessible" is to be determined by consideration of a "standard hinged finger" consisting of a metal probe about the size and shape of the little finger of a human hand.

Points which are listed as normally requiring protection in receivers using A.C./D.C. technique include :

(a) *The Chassis.* The ventilation holes in the back-plate should be small enough to prevent access, and the back-plate itself should be removable only by means of a tool such as a screwdriver.

(b) *Control Spindles.* These should be either of insulating material or isolated from the chassis. Live spindles are considered to be acceptable only if the fixing holes for grub screws are subsequently filled with insulating material.

(c) *Fixing Screws.* All screws securing such parts as the chassis, loudspeaker, etc., to be isolated from the chassis. It is good practice to earth other large metal parts, including metal ornaments on the cabinet.

(d) *Chassis Outlets.* All outlets from the chassis, such as aerial and earth terminals, to be isolated from the chassis.

It is also recommended that apparatus should be tested for insulation by the application of a test voltage between the live parts and the safety earth provided.

FAULT FINDING

In attempting to diagnose faults in a television receiver much can be learnt from observing the operation of the cathode-ray tube. Obvious symptoms will be either a completely blank screen or a condition of uncontrollable full brilliance. Other conditions of the picture or raster on the screen will indicate faults occurring in the various sections of the complete receiver. It is in the recognition of these indications that the service engineer can diagnose and cure a fault in the minimum of time.

Before commencing the diagnosis of faults, it is as well to make a few preliminary checks. Ensure that the mains supply is present at the on/off switch and that the switch is operative. Fuses should be inspected, and if found open-circuited the receiver should be examined for obvious signs of H.T. short-circuits and burn-outs. A preliminary check with the multi-range meter can be made for breakdown of major components, e.g., smoothing and reservoir capacitors.

To summarize, a careful check should be made to ensure that upon switching on the receiver no further damage will be caused.

Faults in Radio-frequency Stage

Faults other than that of valve failure in the radio-frequency stage are not frequently met. On turning the contrast on full, the remainder of the set will produce fluctuation noise, which will be apparent both from the loudspeaker and the tube raster. The time-bases will be running freely, and the lower-pitch whistle from the line-output transformer will indicate the lack of synchronizing pulses from the carrier. Quick verification of the faulty stage can be obtained by connecting the aerial to the input grid of the frequency changer.

In areas of weak signal strength the aerial should first of all be checked for a possible short-circuit or open-circuit in the co-axial connection to the grid tuning coil. If the fault is at the dipole connection, and the centre conductor has become disconnected, some indication of signal can be obtained by making a temporary contact between the braiding and the centre connector of the co-axial input socket on the receiver. If the aerial is in order, the grid circuit of the first stage should be examined for a possible dry joint or open-circuit. If the correct D.C. potentials are on the anode and screen of the valve, the cathode return circuit should be checked for an open-circuit.

Component Deterioration

Quarter-watt resistors, when of more than 1,000 ohms in value, and included in the anode circuit of a high-frequency stage for the purpose of de-coupling, can cause trouble by overheating and increasing in value. As a rule it is advisable to replace any such components by equivalents of half-watt rating. During the course of examination of the vision strip it is sometimes very helpful to tap all components lightly with an insulated probe. A microphonic valve or a capacitor with an intermittent fault can often be located by this method.

Ordinary radio sets are usually quite tolerant of radio-frequency decoupling capacitors that, during the course of years, have developed leaks and are in effect resistors.. In the case of television receivers, however, the standard of circuit efficiency is far more exacting, particularly when the most has to be made of a weak signal. When a receiver of the T.R.F. type has been in operation for a number of years, and routine checking of valve emission and circuit voltages fails to reveal the cause of low modulation on the cathode-ray tube, it is worth while testing for " leaky " components in the signal-frequency stages of the circuit.

R.F. and I.F. Instability and Loss of Signal

Cases of R.F. or I.F. instability, usually apparent through uncontrollable peak white modulation on the tube, can be traced quickly by connecting the grids of the amplifier stages systematically to the chassis, starting at the point of the lowest signal. As a rule, the T.R.F. type of receiver is more inclined to develop this kind of trouble than the superhet. A faulty valve-holder quite often proves to be the cause of R.F. or I.F. instability. If all valve connections prove to be in order, then the possibility of feedback through the wiring of the set should be examined. The position of the aerial feeder relative to the final amplifier stage should be considered, as feedback is most likely to occur between points of high and low signal amplitudes.

Apart from incorrect alignment, loss of vision signal can sometimes be due to the incorrect adjustment of rejector circuits. This condition is usually indicated by vision-on-sound interference in the loudspeaker. The cure is best effected by re-aligning the receiver throughout. In the case of the superhet, lack of signal in both vision and sound sections can be due to failure of oscillation in the frequency changer. This is commonly due to either loss of emission or a low voltage on the anode of the oscillator valve. Check also for open-circuit by-pass capacitors in the I.F. amplifier stages. Loss of sensitivity may be caused by faults in A.G.C. circuits: note that gated A.G.C. systems depend for their operation on a pulse derived from the time-bases, so that a time-base fault can in such cases affect the sensitivity of the receiver.

Both loss of signal and instability may be caused by faulty decoupling capacitors in an R.F. or I.F. stage. Decoupling capacitors should alway be replaced with a direct equivalent and soldered into the *same position* as the one being replaced. Note that much of the wiring in I.F. and R.F. stages is critically placed.

Faults in Video-amplifier Stage

If the television receiver is considered in two sections, from the aerial to the detector, and from that point to the cathode-ray tube, it is in the second section that the majority of faults are found to occur. The video detector itself can cause a fault which may at first sight be wrongly attributed to a weak signal from the vision strip, or alternatively to a loss of drive from the video amplifier. If the detector is of the thermionic type, failure in emission will result in a weak picture, loss of synchronization in the line and frame time-bases—in short, a number of puzzling "red herrings" confront the fault tracer. It is well to remember that crystal diode detectors are also liable to breakdown.

The video-amplifier handles the rectified signal, and should be capable of useful gain over a frequency range extending to 3 Mc/s. The quality of the television picture depends on the efficiency of the video stage, and on the correct degree of frequency compensation which is applied to this circuit. In some video stages the value of the anode load is kept as low as possible. If quarter- or half-watt resistors are used in this part of the circuit, and there is evidence of over-heating, it is advisable to replace the resistors by 1- or 2-watt equivalents. Valve efficiency is rather critical in the video stage, and any falling off in emission will often cause an unsatisfactory picture, with poor synchronization if the signal strength is low.

Faults in Cathode-ray-tube Circuit

For reasons of economy and efficient circuit design, cathode-ray tubes are cathode driven, and are usually directly coupled to the anode of the video amplifier. This connection is sometimes made through a resistor of about 100,000 ohms, by-passed at video frequencies by a suitable capacitor of small size. If the capacitor is open-circuit, the picture will be very smeary and lacking in contrast.

In the case of vision-interference-limiter faults, the cause will almost certainly be the diode, which is biased to conduct on signals above peak white modulation level.

It is sometimes found that a brightness-control potentiometer with a worn track can produce a picture fault which may be confused with the symptoms of failure in the tube-heater-to-cathode insulation. The picture breaks up into strips of varying brightness, running horizontally, which resemble the blurred areas which occur during intermittent heater-to-cathode short-circuits.

Faults in Synchronizing Separator

When faults occur both in the line- and frame-deflection circuits, it is logical to trace the cause to a common origin, and the most likely component to suspect is the synchronizing separator, which, owing to loss of emission or faulty resistors in either the grid or anode circuit, may be operating on the wrong part of the valve characteristic. If anode current changes are caused by picture content instead of being limited to the synchronizing pulses, then the time-bases will run so irregularly that it may prove impossible to lock the picture into a steady position. The cause of trouble in this case will often be traced to the grid coupling capacitor to the synchronizing separator. This component should have a very low leakage current, and should show a resistance of about 20 MΩ on test.

Interlacing

Faulty interlacing of the picture-line structure is often due to unwanted coupling between the line and frame oscillators. Although it is very difficult to remove all trace of the line-synchronizing component from the frame pulse, considerable success has been achieved in producing well-interlaced pictures in most modern receivers.

Imperfect interlacing produces a serious loss of picture definition in the vertical direction, and may, in extreme cases, produce a loss in horizontal definition through the over-lapping of adjacent line information. Apart from the obvious "liny" picture structure, faulty interlacing can be recognized by a tendency for the start of the half-line trace to wander at the top of the raster. This indicates that the frame time-base is not being triggered at the correct intervals by the frame-synchronizing pulse. This irregularity is due to the variation in the intervals between successive fields caused by the uneven operation of the frame time-base. For correct interlacing, the shape and duration of the synchronizing pulses must not vary. It is through the addition of small voltages at line frequency to the envelope of the frame pulse produced by the integrator circuit that faulty interlacing arises. As very small intervals are concerned in producing these waveform irregularities, the underlying cause is to be found in minor circuit variations due to faulty insulation in capacitors. Induced voltages at line frequency in the frame-time-base circuits can arise from the inadequate screening of the line-output transformer or line-deflection coils

The cure for some cases of faulty interlacing can only be discovered after considerable time has been spent, using the oscilloscope, in the systematic examination of the waveform at various points in the frame-scan circuit. In some modern receivers a valve-operated buffer stage separates the frame-synchronizing pulses from the mixed waveforms in the synchronizing separator. An oscilloscope applied to the anode and cathode of the "pulse shaper" should reveal clean-cut negative-going waveforms.

The presence of pulses at line frequency can be revealed in the frame circuit by rendering the frame oscillator inoperative either by removing the valve or disconnecting the H.T. supply. Under these conditions the use of the oscilloscope at various points in the frame time-base will reveal the extent of the coupling effects from the line oscillator. During this experiment both the aerial and E.H.T. supply should be disconnected.

Faults in E.H.T. Circuit

As the majority of modern receivers obtain their E.H.T. suppli from the rectification of high voltages generated in an inductance forming part of the load on the line-output pentode, failures in this part of the set will usually be caused by the strain imposed by the high induced voltages. The line-output transformer with the extra windings to supply the high-voltage rectifier should be a high-grade component throughout.

Any partial breakdown in the insulation of the line-output transformer will be indicated by a characteristic fold-over on the side of the picture. If E.H.T. is also supplied from the winding there may be insufficient voltage to produce a picture. The faulty component may overheat until a complete winding breakdown occurs. The presence of a high potential on the cathode-ray-tube anode may be quickly verified by obtaining a brush discharge on to the blade of a well-insulated screwdriver. After a time the service engineer will become quite expert in assessing the E.H.T. voltages by this method. It is wise not to apply this test to mains-generated E.H.T. systems, which should be treated with extreme caution at all times.

Apart from breakdown in the line-output transformer, the E.H.T. supply may fail through a variety of additional causes, and faults in this part of the circuit are very common in some receivers. The presence of air inside the high-voltage rectifier may be recognized by a purple glow diffused throughout the envelope. If, however, the valve looks in order but still fails to develop E.H.T., the cause may be due to a short-circuit in the line-output deflection coils, especially if these components have a metal shroud. Extensive picture interference is commonly caused by corona discharge at high potential points. Careful listening or examination in subdued light will often reveal the source of the discharge. Seepage of E.H.T. across insulated surfaces due to dust and dampness can also cause this type of picture interference. Some manufacturers enclose the entire E.H.T. system in a screened compartment, and shorting across the E.H.T. from the cathode of the diode to the metal case can occasionally occur. A slight movement of the diode away from the nearest earth potential point will speedily cure this fault. If lack of voltage pulses in the line-output transformer is indicated by the absence of the characteristic high-pitched whistle, then the lack of oscillation can be due to a failure in the line feedback circuit, or, possibly, to a faulty oscillator valve or blocking oscillator transformer. The use of the oscilloscope is again essential for the rapid location of the fault. It should not be overlooked that quite high voltages can be found throughout the feedback circuit, and small mica condensers will sometimes fail or develop an intermittent fault, producing picture interference.

Cathode-ray-tube Faults

The modern cathode-ray tube is a precision instrument, and should last at least two years regular use without serious deterioration. In the past the most common cause of tube replacement was the presence of a relatively insensitive area of tube fluorescence due to heavy ion bombardment from traces of gas in the tube. In many picture tubes this defect is overcome by means of an " ion trap ", the presence of which is usually recognizable by the deliberate off-centre alignment of the cathode-gun assembly. A small permanent magnet clamped externally near the tube base deflects the lighter electron beam into the accelerating electrodes, leaving the ions behind in the "trap". It is important to ensure that the small correcting magnet is adjusted for maximum picture brightness before a set is sold or returned to the customer after servicing. Dull pictures and shortened tube life will result from incorrect adjustment of the ion-trap magnet.

Probably the most common cause of tube failure is due to a breakdown in the heater-to-cathode insulation. This defect causes blurred definition and uncontrollable tube illumination with complete loss of picture modulation. Sometimes the tube will function normally after the set has been switched rapidly off and on, but the trouble returns more and more frequently. There are temporary measures for this defect, such as running the tube heater from a floating winding on a separate heater transformer. Some improvement can usually be effected by this measure, but unless the transformer is a specially constructed component with a low inter-winding capacity, most of the video detail at high frequencies will be lost. Grid-to-cathode shorts rarely occur, and are usually accompanied by a "click" from the tube, the screen of which flares brightly as the bias disappears. Very occasionally a tube may show a faulty heater connection which may occur in the base connection. If this condition is suspected, the application of solder and flux may effect a cure. It occasionally happens that the tube E.H.T. connecting cap becomes disconnected and the wire broken off in the glass seal. A satisfactory contact may still be made by means of a quick-drying, plastic metal compound available from shops dealing with model making.

Time-base Faults

Time-base troubles account for a high proportion of the faults in which the resulting picture appears in a distorted or unsteady form, while faults in the line time-base/E.H.T. circuits are also a common cause of a complete absence of raster. Owing to the divergences in the circuits, similar faults in different time-bases may produce very different symptoms, but the following notes are generally applicable.

Line Time-base Faults

If the line oscillator is not functioning no E.H.T. will be produced, and there will be a complete absence of raster. If it is functioning, but at an incorrect speed, the picture will either not lock and remain a complete jumble (indicating an absence of synchronizing pulses) when the line hold control is varied or will lock with several side-by-side images repeated across the screen, indicating that the speed is too fast.

That the oscillator is functioning may be indicated by the presence of the 10-kc/s whistle, but, since this whistle comes mainly from the line output transformer, its absence is not a final test of non-oscillation. With a blocking oscillator a useful check is to measure the voltage with a high-resistance meter between the grid and chassis: oscillation is indicated by the presence of a negative voltage.

Where the speed of a blocking oscillator is too fast, this may be due to failing emission of the valve; high-resistance windings or leaky insulation in the blocking transformer; or a faulty capacitor between the "earthy" end of the grid winding and chassis.

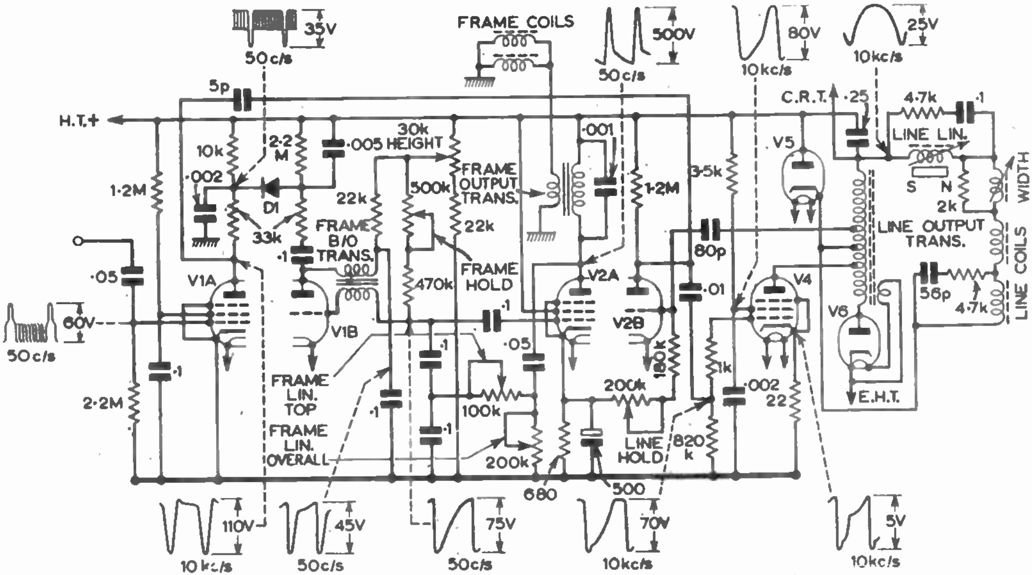


Fig. 11.—The Time-bases of a Typical Modern Receiver Showing the Waveforms to be expected at Various Points.

V1A is the sync. separator and D1 a frame interlace diode. V1B is a frame blocking oscillator and V2A the frame output valve. V2B forms, with V4, the line multi-vibrator (feedback to V2B via the 80 pF capacitor), V4 being the line output valve. V5 is an efficiency diode, and V6 an E.H.T. rectifier.

Where the speed of a blocking oscillator is too slow, the most likely cause is a resistor increasing in value, usually in the network which feeds the positive bias voltage to the "earthy" side of the grid winding of the blocking transformer.

With multi-vibrator circuits, too fast speed may be caused by low emission of the valves; low capacitance or leakage in the grid capacitors or change in the networks, if any, supplying the bias to the valve. Too slow a speed is often due to a resistor having gone "high"; suspects should include the anode and screen feed resistors and the grid resistor networks.

Line Output Stage

If the oscillator is functioning but there is a complete absence of raster the fault is almost certainly in the line output stage. First check the three valves: the line output, the efficiency diode and the E.H.T. rectifier. Failure may also be due to a short-circuit or "damping", across the output windings of the line output transformer (e.g., shorted turns in the transformer or in the scan coils, or leakage between line and frame scan coils). For some receivers a quick check for damping across the secondary is to unplug the scan coils, but care must be taken to avoid the risk of burning the screen. Also check any screened leads connecting the scanning coils for breakdown of insulation.

Cramping

Cramping on the left-hand side of the picture with a dark border showing along the mask, may be due to lack of emission in the efficiency diode, or in a high-resistance joint in the wiring to this valve.

Cramping on the right-hand side of the picture is most likely due to failing emission in the line output valve, but may be caused by insufficient power being applied to the heater (for example, due to wrong setting of the mains adjustment panel) or due to the screen feed or cathode bias resistor, if any, having increased in value.

Cramping at the centre of the picture (Test Card "C" showing an "egg-shaped" central circle) may be due to a fault in the cathode circuit (e.g., faulty cathode decoupling capacitor) producing negative feedback, o to an incorrect waveform being applied to the output valve, either due to a fault in the oscillator stage or in the correction network in the grid circuit of the output stage.

Frame Time-bases

The absence of frame scan output will result in a thin bright horizontal line across the screen, and when investigating such a fault take care to keep the "brightness" control as low as possible to avoid burning the screen.

A quick check to locate whether the fault is in the oscillator or output stage may be made by touching, as in sound receiver practice, the grid of first the output and then the oscillator valve (take care that the chassis is not at "live" potential). If the valve is worki g the thin line will open up a little due to hum pick-up. If there is no effect on the thin line it indicates that the valve is not working or that the grid circuit is short-circuited to chassis. Faults likely to produce incorrect speeds, etc., in the frame time-base follow roughly similar lines to those described for the line time-base circuits, although the frequencies involved are much lower (50 frames per second).

If the oscillator is running too fast the picture will appear to be moving downwards; if too slow the direction will appear to be upwards.

Insufficient Height.—Check output and then oscillator valves for low emission, anode feed resistor on the oscillator valve and grid resistor on output valve.

Cramping at Top.—Check the components in the negative feedback correction circuit, if any, between the anode and grid of the output valve. Check the grid coupling capacitors for leakage.

Cramping at Bottom.—This may be due to faulty emission of the frame output valve or loss of capacitance in the cathode by-pass capacitor. Faults in the negative feedback correction components may also give rise to this effect.

Synchronization Faults

Where the fault is due solely to poor or absent synchronization pulses it should be possible to obtain a picture momentarily by manual adjustment of the hold controls.

Where it is possible to lock the line but not the frame oscillator, the fault is most likely to be in the integrating circuit or in the interlace filter which may be used to eliminate line pulses from the frame synchronizing pulses.

When the frame but not the line oscillator can be locked, the fault may be in the differentiating circuit or in any fly-wheel synchronizing circuits which may be fitted.

Poor hold on both oscillators may be due to the synchronizing separator valve failing, incorrect screen voltage due to feed resistor increasing in value or leakage in the by-pass capacitor.

If the line hold only is weak examine the line synchronizing feed capacitor. Where line tearing is pronounced, it may be due to the coupling components in the grid circuit.

Turret Repairs

The repair of faulty turret tuners often requires most delicate workmanship, and less skilled work is likely to impair rather than to improve results. It is for this reason that many manufacturers recommend that turrets should always be returned to them for repair or adjustment. For example, the full alignment of a tuner is best carried out on the costly and complex equipment available to the manufacturer. There are, however, some repairs that a service engineer can tackle, provided that he has the necessary experience and deftness.

Poor or intermittent results on one channel are often due to poor contacts. In such cases it is important that no attempt be made to increase contact pressure by adjustment of the spring contacts, as these are usually carefully set in the factory. Careful and sparing application of turret lubricants is more often successful, and sometimes adjustment to the switch-locating device will change the point of contact slightly. The channel-selector mechanism on turrets should be kept lubricated, and after cleaning it is advisable to apply a smear of petroleum jelly to each stator spring contact.

A common cause of short-circuits is the wearing off of the insulating paint on the ends of the small ceramic capacitors, leaving an exposed wire which may rub against another component. Faulty components may be the result of overheating during soldering, either in manufacture or subsequently.

Where any attempt is made to replace a component, it is essential that the tuner should be handled with great care, and that only an exact replacement, physically as well as electrically, should be fitted in the precise position of the original component, with connecting wires of the same length and path.

Coils are often adjusted by spreading their ends, so that great care should be taken not to accidentally disturb them

Miscellaneous Picture Faults

While many picture faults provide by their symptoms an obvious clue as to the source of the trouble—linearity faults, such as cramping, insufficient height or width, for example, immediately suggest a time-base fault—there are a number of symptoms for which the location of the trouble will prove puzzling but which, once recognized, provide no less valuable a clue as to the likely fault.

" Ringing "

This common picture fault, which takes the form of a black line immediately to the right of a white object, or a black line following a white object, is almost invariably caused by a distorted vision-response curve, providing excessive high-frequency amplification. Response characteristics of this type tend to produce a series of damped oscillations in the tuned circuits, and these produce the effect already described. Where the fault is found when installing a new receiver, it often denotes mistuning of the local oscillator, while on older receivers it may either be due to local oscillator drift or to misalignment generally. However, where the fault develops suddenly and is accompanied by a marked reduction in the band-width of the receiver (as shown on Test Card C) and an increase in sensitivity, the damping resistors across the tuned circuits should be checked. Faulty components in the video compensating network of the video output stage, and the decoupling circuits, are other frequent causes of " ringing ", a useful pointer in this case being that the receiver sensitivity may not be unduly affected. It should be noted that in some receivers, particularly those intended for " fringe " reception, a certain degree of " ringing " may be introduced intentionally to sharpen the images; but since it is difficult to prevent such instability from later becoming more pronounced, designers must be wary of carrying this process too far. An effect somewhat akin to " ringing " may also be caused by " ghost " images (see Section 21) produced by the arrival of signals at the receiver along multiple paths.

" Line Ringing "

Vertical bars on the left-hand side of the picture may be caused by " ringing ", i.e., damped oscillation, of the scanning coils or line-output transformer. With improvements in the efficiency of these components and the use of efficiency and boost diodes, this fault has become less frequent; but in older models it is not uncommon for the high-wattage damping resistor (which is required to dissipate considerable energy) connected across the scanning coils to become open-circuited. On some modern receivers employing wide-angle tubes, a somewhat similar symptom may be caused by spurious signals from the line-output valve beating with the local oscillator to produce an interfering signal; the remedy is to provide a small high-voltage decoupling capacitance to the anode of the line-output valve.

Sound-on-vision

Alternate light and dark horizontal bars across the picture, occurring during the louder sound passages, and sometimes when severe destroying line hold, are the symptoms of this fault. The intermittent nature of the trouble will distinguish it from the effect of hum in the receiver. The fault may be due to misalignment, general inefficiency of the sound-rejector circuits or simply to slight oscillator drift which may be insufficient to produce other harmful effects. Another common cause—though one that is not likely to arise after the receiver has been working successfully—is overloading of the frequency changer valve, and this may be cured by fitting an attenuator pad in the aerial feeder. Other possible causes are valve microphony and feedback in the H.T. circuits.

This fault should not be confused with microphony of the valves employed in the vision chassis. Here, the vibrations set up by the loud-speaker cause a similar effect on the screen to that of sound-on-vision. If reduction of the volume control fails to eliminate the effect, the fault must be sound-on-vision.

"Pulling"

"Pulling on Whites" is the name given to a fault in which the picture is momentarily displaced horizontally whenever a white image moves across the right-hand side of the picture: it can most clearly be observed on Test Card C, when a castellated effect on vertical lines will be found to coincide with the changes from black to white in the right-hand border. It is caused by later triggering of the line-scan oscillator following a line ending in white light, due to the receiver not responding to the "front porch" preceding the line pulse. This may be due to poor high-frequency response of the vision receiver, and in this case will be accompanied by loss of the high definition patterns, or, where the patterns are unaffected, by loss of high frequencies in the circuits immediately preceding the synchronizing separator; a likely cause being high stray capacitance in the coupling between the video amplifier and the grid of the synchronizing separator.

"Pulling on Blacks", or as it is sometimes called "triggering on picture", is a somewhat similar condition, but with the picture displaced in the opposite direction. The cause is almost invariably incorrect clipping in the synchronizing separator, the clipping being above the 30 per cent black level, and thus allowing picture content to trigger off the time-bases. A faulty component in the synchronizing separator stage, such as an increase in value of the screen-feed resistor, is usually the reason for this condition.

Line pulling may also be caused by ghost images interfering with the synchronization. This will not occur, of course, in sets using flywheel synchronization.

"Plastic" Picture

Where the outlines of objects are clear but the picture has an overall grey appearance, the usual cause is poor low-frequency response in the vision receiver or video amplifier, and any of the components or adjustments affecting overall response may be at fault. These include misalignment or incorrect local oscillator setting, a decrease in value of a coupling or by-pass capacitor in the video amplifier.

"Flare" or "Streaking"

These terms are used to describe the condition when streaks or smudges appear to follow black-and-white images horizontally across the screen : the black horizontal bar at the top of Test Card C provides a most useful check. The cause is excessive low-frequency response, and here again may be caused by misalignment or defective compensation in the video-amplifier stage. A gradual increase in flaring usually suggests misalignment or oscillator drift, whereas any sudden increase points to a faulty component, such as a decoupling capacitor, or an increase in value of the anode-load resistor in the video amplifier.

Loss of Highlights

A condition may sometimes be encountered where the darker shades of the picture are reproduced normally, but the lighter tones tend to flatten out and become almost indistinguishable from one another. This is caused by the peak amplitudes of the vision signals being lost by clipping, and is commonly caused by over-advance of the vision-interference limiter, but may also be due to overdriving the cathode-ray tube or incorrect bias in the video-amplifier stage.

PICTURE-TUBE SALVAGE

The replacement of a picture tube represents a considerable item of expenditure to the average owner, and service engineers are often asked if they can avoid the need to purchase a new tube, even at some cost to picture quality.

Makeshift repairs are often inadvisable, as any temporary improvement may soon be reversed, with the result that a new tube has to be purchased after all, and money spent on extending the life of the original tube will have been wasted. There are, nevertheless, several methods of picture-tube salvage which have become fairly well established in practice, and often give reasonably satisfactory results for a worth-while period.

Low-emission Tubes

Where the picture has become "fuzzy" with little contrast or brilliance and with a tendency to turn negative when either of these controls is advanced, this is often a sign of low emission. Two methods of temporary rejuvenation of the cathode emission, provided that the tube is still " hard ", are fairly widely used, although in both systems it should be clearly recognized that there is the risk of the attempted " cure " causing the complete breakdown of the heater.

The first is to run the tube for a short period with the heater considerably over-run and with all other voltages removed (sometimes a potential of about 100 volts is put on the grid), by means of a tapped transformer. The other method, which is probably more widely used, is to install a permanent " boost " transformer or auto-transformer providing from 20 to 50 per cent additional heater voltage. The installation of a boost transformer has frequently extended the useful life of tubes by many months, particularly with the older type of low-voltage heater, although in other cases the life may be prolonged for only a short period; the extra life averages roughly about four to six months.

Electrode Short-circuits

Heater–cathode short-circuits, often of an intermittent nature, are not infrequent, and may result in flashing, uncontrollable brilliance, hum bars or absence of raster. Where a heater–cathode short-circuit has been traced, isolating transformers, which are specially made for this application and which are now readily available, provide a means of extending the useful life of the tube. Such transformers must have very low inter-winding capacitance, as otherwise there will be considerable loss of high-frequency video signals (which, with the heater–cathode short-circuit, will appear on the heater line) and hence deterioration of picture quality. The use of an isolating transformer, is, of course, practicable only on A.C. mains. With the aid of an isolating transformer, tubes with heater–cathode short-circuits can often be used quite successfully for relatively long periods; in fact, there is little evidence that the life of a defective tube fitted with an isolating transformer differs appreciably from that of a normal tube.

Similar fault symptoms may sometimes be traced to grid–cathode short-circuits. Where the tube has a tetrode gun, it is possible, though often at some cost to picture quality, to strap the grid to the cathode and then to rewire the tube with the first anode acting as control grid.

Intermittent inter-electrode short-circuits often occur only when the tube is at full operating temperature and can sometimes be eliminated by slightly lowering the heater voltage; a simple method with series-connected tubes is to wire a suitable resistor in parallel with the heater of the tube.

It is worth noting that premature tube failures and inter-electrode short-circuits are frequently caused by the over-running of the heater and consequent high cathode temperature. It is recommended, when replacing a faulty tube, to check the heater *voltage* in a parallel-fed receiver and to check the heater *current* in a series-fed receiver. Measurements should be made after checking that the mains-tapping is correctly adjusted. Currents should be within 5 per cent of the rated figure and voltages within 7 per cent.

Miscellaneous Picture-tube Faults

A picture-tube fault that may occasionally develop in fairly new cubes is a change in the grid cut-off characteristics; this may result in excessive brilliance even with the brilliance control set at minimum. This type of fault can sometimes be overcome by adjusting bias levels; sometimes by simply connecting a high-value resistor (e.g., 10M) between the first anode and chassis.

Scratched tube faces can often be repolished by the tube manufacturers provided that the scratches are not too deep.

Although not strictly speaking a picture-tube fault, it is worth noting that as picture tubes age any deficiency in E.H.T. voltage tends to produce results akin to that of a "soft" tube or failing emission, and before deciding that a tube has no further useful life it is advisable to check the E.H.T. supply.

Picture-tube Repair

A number of firms specialize in carrying out the repair of faulty picture tubes for the trade. The two main systems used are: (*a*) to open up the tube and fit a completely new gun assembly, or (*b*) to reactivate the cathode, followed by restoration of the vacuum by using an R.F. heater to heat the getter. The first system provides the tube with a completely new lease of life, apart from any deterioration which may have taken place in the flourescent screen. The second process does not give the equivalent of a new tube, but does extend the life of the tube by some months. The tubes which respond best to this treatment are those where the cathode has lost emission either through being run over long periods at too low a heater voltage or where the cathode surface has been affected by the tube being slightly soft. Little improvement can be made to cathodes which have lost their emission due to being over-run, or which have already had a very long life.

SERVICING PRINTED CIRCUITS

This subject is covered in Section 39, the notes given there being generally applicable in the case of television receivers also.

BAND III CONVERSION

Band III converters fall into two categories, those in which the incoming Band III signal is converted to the I.F. of the existing receiver (thereby replacing the "front-end" of the receiver), and those in which the incoming Band III signal is converted to the Band I frequency that the receiver is aligned to receive. The latter is the "universal" type, which may be used with any receiver.

For the conversion of T.R.F. receivers, a "universal" converter must be used. In the case of superheterodyne receivers, either type of converter may be used, though the "Band III to receiver I.F." type is often to be preferred, for reasons given later.

Universal converters may be designed so that the Band I signal is switched directly from the input to the output socket of the converter, undergoing no amplification. Alternatively, the converter may be designed so that both Band I and Band III signals pass through its circuitry, enabling arrangements to be included to provide equal signal outputs on both bands. This eliminates the necessity to readjust the receiver controls when switching from one band to another.

If the converter draws its power from the existing receiver, the following points should be noted: (*a*) The receiver power supply must be capable of supplying the extra load. (*b*) If the receiver has, as is usual, a mains-connected chassis, then the converter chassis will also be mains connected. This is dangerous unless the converter is specifically insulated for such usage. (*c*) The valves used in the converter must have heaters of voltage and current rating suitable for inclusion in the heater chain of the receiver.

Upper-sideband Receivers

Some early Channel 1 T.R.F. receivers were designed so that the full double-sideband transmission of the old Alexandra Palace transmitter was accepted by their input circuits. Subsequently, the sound channel was aligned to receive the sound signal, and the vision channel the upper sideband, thereby providing good sound rejection. This type of receiver will not accept lower-sideband transmissions without serious degradation of the picture, and is therefore not really amenable to conversion. It may be possible to realign the vision stages to accept more of the lower sideband, but sound rejection may then be found to be poor, necessitating the insertion of sound-rejector coils.

Patterning

The main disadvantage of a "universal" converter is that, whilst it is receiving the Band III programme from the converter, it may also pick-up a certain amount of the Band I transmission. This may either make viewing impossible, or, in less-severe cases, result in a background pattern being present on the screen.

A further disadvantage of the "universal" converter is that the converted Band III output of the converter, being at Band I frequency, may cause interference to neighbouring sets receiving Band I, particularly after amplification in the receiver. The signal may be radiated from the cable connecting the converter to the receiver, from the R.F. circuits of the receiver itself, or by inductive or capacitative coupling, within the converter, to the aerial circuits and thence to the dipoles. This latter type of radiation is overcome to some extent by incorporating an R.F. amplifier in the converter, as this acts as a buffer between the mixer-grid circuit and the aerial.

Aerial-to-I.F. Converters

Aerial-to-I.F. converters are similar in design to the tuner units used in Band I/III receivers. Conversion is generally carried out as follows: the R.F. amplifier and frequency-changer valves of the receiver are removed, the converter output plug is plugged into the frequency-changer valve-holder, and the power supply for the converter taken via a plug inserted into the R.F. amplifier valve-holder. In some cases a single connecting cable is provided, in which case this may be inserted into the frequency-changer valve-holder, or into the R.F. amplifier valve-holder. In the latter case, provided that the oscillator is rendered inoperative, the mixer will act as an additional I.F. amplification stage. The extra gain achieved in this way may be necessary in some cases.

Points to be borne in mind when selecting a converter for a particular receiver are: (*a*) type of valve heaters in use, i.e., whether series or parallel wired; (*b*) the I.F. of the receiver; (*c*) whether the receiver oscillator, before conversion, is higher or lower than the signal frequency; (*d*) in the case of tuners with clip-in coils, the channels it is desired to receive.

Conversion Problems

The position of the converter heaters in the heater chain of a universal-mains type of receiver may require consideration. Some receivers have their R.F. and mixer valves relatively " high up " the heater chain, that is to say away from the ground. This may give rise to modulation hum troubles on some conversions, and the cure is to rewire the converter heaters " down " the chain, nearer to ground.

Another fault, commonly encountered but mystifying on first acquaintance, is as follows: A picture is obtained, after conversion, but no sound can be obtained; tuning the converter oscillator control will bring in the sound but cause the picture to disappear. The trouble is that an incorrect type of converter has been obtained. The receiver oscillator has, for example, been working on the low side of signal frequency, whereas the oscillator in the converter is working on the high side of signal frequency. Or, conversely, the receiver oscillator may have been on the high side and the converter on the low.

41. TRANSMITTER INSTALLATION AND MAINTENANCE

41. TRANSMITTER INSTALLATION AND MAINTENANCE

INSTALLATION OF TRANSMITTERS

The construction of transmitters designed for fixed station operation falls broadly into three classes : (*a*) low-power receiver type chassis, (*b*) cabinet, cubicle or rack type, and (*c*) welded frame with enclosing panels.

Most of the recommendations given for installing receivers apply equally to low-power, naturally cooled transmitters of the first class. The primary requirements are a single-phase A.C. power supply, ample space for ventilation and access to controls, in a convenient position for leading in the aerial or aerial feeder, and a short, low-impedance earth connection.

The important features to be considered in planning the arrangement of medium- and high-power transmitting equipment are :

(1) Efficiency and ease of operation.

(2) Economy in floor space, wiring and installation materials.

(3) Access for servicing valves and components and for removing defective components.

(4) Adequate natural or artificial lighting of controls and meters.

(5) Ample doorway clearances for the entry of the largest piece of apparatus, if necessary in its travelling crate.

(6) Ventilation, either natural or assisted, for dissipating the heat generated by the equipment.

(7) Headroom for aerial leads-in, overhead radio-frequency connections and lifting tackle for transformers and rotary machines.

The installing engineer is usually furnished with an accommodation plan, installation diagram, cable schedule and wiring diagrams of each unit for guidance in carrying out the erection and wiring. With high-power installations, additional detail plans and diagrams are provided for power-supply equipment, switchboards, cooling plant, air trunking and cable ducts.

Installation time is reduced by preparing such items as plinths, machine foundations, cable and air ducts, aerial leads-in and the equipment earthing system before taking delivery of the equipment.

A shallow concrete or wood plinth, planed level and bolted to the floor, is a useful means of concealing interconnecting cables and protecting the transmitter panels from abrasion and dirt when the floor is being cleaned, and is also the most satisfactory way of taking up small irregularities in the floor surface, to ensure that cubicles stand level. Small differences in the base levels of adjacent cubicles can produce wide, diverging gaps between the sides, which make it difficult to bolt them together and make a well-finished job.

Taking Delivery

Before signing delivery receipts for equipment, the cases should be checked and inspected for visible signs of damage. If there is a shortage or apparent damage, the papers must be endorsed to this effect. As soon as possible, the carriers must be notified, and all invoices and correspondence filed until a settlement has been reached.

It will sometimes be found that components have been removed from an assembly and packed separately for safe transport. The components and the corresponding positions from which they were dismantled are invariably labelled with appropriate numbers or letters as a guide to reassembling when the equipment is installed. It is advantageous to erect frame-type cubicles with the doors and panels removed, to give easy access for bolting the frames to the floor, reassembling the components and connecting external cables. When reassembling, frequent reference should be made to the unit wiring diagrams to avoid wiring errors.

Earthing

The frames and cabinets of radio-frequency units must be earthed by as short and direct a route as possible, preferably with one or more copper strips laid in parallel and spaced apart. An earth connection approaching a quarter wave in length will have a high impedance at the transmitter, and appreciable radio-frequency voltages may be developed in the frame and screens. At the higher frequencies a capacity earth in the shape of copper sheets laid under and bonded to the units has a lower earth impedance than a direct connection, since the reactance to earth varies inversely as the frequency; but it is important to note that a metallic connection is equally necessary for anchoring D.C. and A.C. circuits and earthing the frames and screens of auxiliary equipment.

There is little to be gained by employing earth electrodes of large surface area. It is not so much the area in contact with the soil as the resistivity of the soil that determines the earth resistance. Two or more copper stakes, spaced well apart and driven into moist soil, are often more efficient than a single plate having two or three times the equivalent contact area.

Further information on earthing will be found in Section 21.

Installation Wiring

Cable schedules and estimates are prepared by taking each run on the installation diagram in turn, and either scaling the length on the accommodation plan or surveying the station building. Sizes and voltage gradings are determined from the duty of the cable, the cable-maker's list and the wiring regulations.

The systems of wiring best adapted for radio transmitting installations and their applications are summarized in Table 1.

Power supply and interconnecting cables can often be run in advance of erection. Plan the runs so as to avoid unnecessary cross-overs where cables enter branch ducts and cubicles, and lay all cables preferably between two selected units before proceeding to the next. Run the cables by the shortest route, making ample allowance for terminating and bending where they leave a duct or deviate inside cubicles, before cutting. Where the runs are long, it is advisable to bunch the cables and earth the sheaths at regular intervals along the run to avoid the possibility of radio-frequency pick-up. Radio-frequency and audio-frequency leads should be separated from A.C.-supply leads to prevent

TABLE 1.—WIRING SYSTEMS

System	*Description*	*Applications*
Box ducts	Enamelled sheet-metal ducts with screw-on or spring-clip covers. Cables run by removing covers and laying loose or in clips or grips.	Used extensively for concealing several small or medium-sized TRS or PVC cables. Screwed to walls, ceilings and recessed wall cavities, or laid in floor trenches.
Cable trays	Perforated steel trays without covers. Cables secured to bottom of tray by saddles.	Used for bunched wiring of most types in cabinets, racks, cubicles and overhead runs between cabinets and walls.
Screwed-steel conduit	Screwed-steel tubes, enamelled or galvanized, and equipped with draw-through boxes at bends.	Suitable for running a few small TRS or PVC cables with small bending radius. Cables drawn through angle boxes at ends of each straight section. Universal applications.
Flexible metal conduit	Tubes of spiralled metal strip, easily bent to any desired form. Cables drawn through.	For short terminations in confined spaces between wall or floor ducts and machine or transformer terminal boxes, particularly where the cables are subjected to vibration or equipment is resiliently mounted.
Saddles and cleats	Metal strip, bent and drilled to straddle cables run on walls. Clamping bars of insulating material.	Used for small lead-covered, PVC or TRS cables in basements and other situations where appearance is not important.
Porcelain and ceramic insulating supports	Porcelain, ceramic or pyrex insulators in conical and various other shapes.	Bare copper radio-frequency and high-voltage connections.
Cable hangers	Steel hangers, sometimes with brackets, shaped to support one or more cables, and drilled for wall or ceiling fixing.	Heavy lead-covered or armoured cables carried along walls or suspended from ceilings.
Earthenware conduit	Glazed stoneware circular sections with spigoted ends and radiused bends.	Suitable for underground and underfloor runs of lead-sheathed PVC or TRS cables.

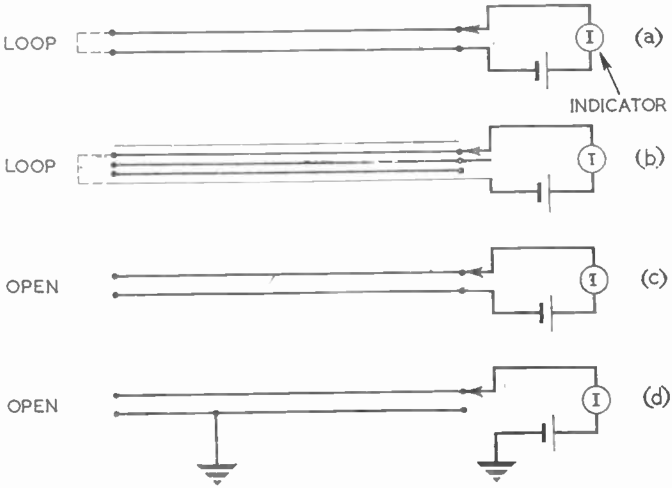


FIG. 1.—WIRING TESTS FOR CONTINUITY, SHORT-CIRCUIT, EARTH AND LOW INSULATION RESISTANCE.

possible interference, and single-phase A.C. go-and-return leads should be twinned to neutralize the external field.

Wiring Tests

On completion of the station wiring, each run should be checked against the installation diagram. Faults can be broadly separated into wiring and equipment faults, and fault-finding when making final tests will be simplified if the wiring has first been checked for continuity, short-circuits, earths and low insulation resistance. These tests are readily performed with almost any kind of current or voltage indicator and one or two dry cells or other source of D.C. supply. Fig. 1 (*a*)–(*d*) shows the connections for a current indicator such as a headphone receiver, a galvanometer or a milliammeter.

Continuity Tests

The continuity test (Fig. 1 (*a*)) consists of looping the far end of the conductor to the return conductor or sheath, and tapping the indicator and battery across the near ends. If a galvanometer or milliammeter is used, it is necessary to connect a resistance or lamp in series to limit the current to a safe value for the meter. Continuity is proved by a sharp click heard in the receiver or a reading on the meter. This test can also be used to identify conductors in a multi-core cable (Fig. 1 (*b*)). The far end of the conductor to be identified is connected to the sheath, while the indicator and battery are tapped across the sheath and each conductor in turn at the near end, until an indication is obtained.

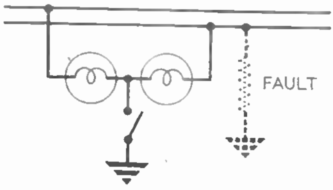


FIG. 2.—LIVE TEST FOR CONTINUITY, EARTHING OR LOW INSULATION RESISTANCE.

Short-circuit Tests

In the short-circuit test (Fig. 1 (*c*)), the far ends are open-circuited, and a click in the receiver or a reading on the meter reveals a short-circuit. To test for an earth on either conductor or for low insulation resistance, the far ends are left open and one terminal of the battery earthed (Fig. 1 (*d*)). A loud click in the receiver or a reading on the meter indicates an earth or low insulation resistance on the particular conductor under test. It should be noted that, when making insulation tests, a long cable may have appreciable capacitance, and a click heard on making the connection may be produced by the charging current. The true indication of faulty insulation is an equally pronounced click heard when the connection is broken.

Open-circuit Tests

A quick method of testing live conductors for open-circuit, low insulation resistance or an earth is shown in Fig. 2. Two metal-filament lamps in series, with their junction point connected to earth through a switch, are tapped across the conductors. With the switch open, the lamps will glow equally if conditions are normal, but if they fail to light, either the supply has failed or an open-circuit exists. When the switch is closed, no change in brightness should occur, but if the insulation is faulty, the lamp connected to the faulty line will glow less brightly. If there is a direct earth, the lamp will be extinguished.

Locating Faults

The position of an earth on a long radio-frequency co-axial cable can be located by measuring the resistance between conductor and sheath at each end, with the opposite end open-circuited, as shown in Fig. 3 (*a*). If R_x and R_y are the measured values at ends X and Y respectively, and l is the length of the cable, the distance of the fault from end X

$$l_x = l \frac{R_x}{R_x + R_y}$$

An earth on a twin-wire cable can be traced in a similar manner. In this case the resistance is measured between each conductor and sheath at one end, with both conductors looped at the far end, as in Fig. 3 (*b*). Let the two values be R_x and R_y. Then the distance of the fault from the near end

$$l_x = 2l \frac{R_x}{R_x + R_y}$$

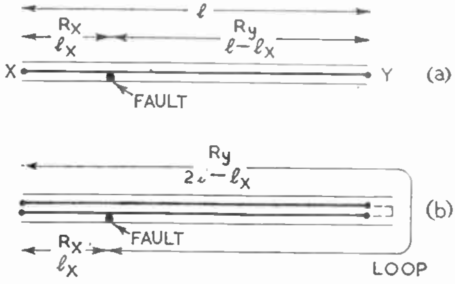


Fig. 3.—Location of Earth Fault in Radio-frequency Cables.

Before finally connecting the installation to the power supply, see that all distribution switches on the equipment and the main supply isolator or circuit-breaker are in the open position.

ALIGNING AND TESTING THE TRANSMITTER

Before aligning the transmitter circuits, the internal wiring must be inspected to ascertain that there are no loose or misplaced connections, particularly where components have been dismantled for transportation, and followed by a mechanical check of all components.

No hard-and-fast rules can be laid down for dealing with the many variations of transmitter design, and the description will here be confined to an outline of the basic principles of alignment. The order of procedure can be regarded as the reverse of that for receivers. A receiver is aligned by injecting an artificial signal obtained from a

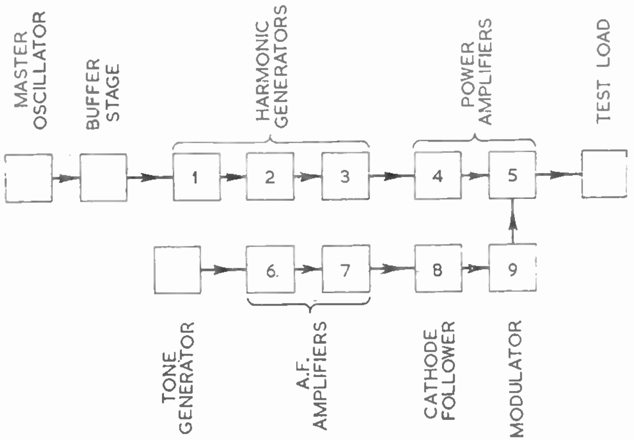


Fig. 4.—Alignment Sequence for an Amplitude-modulated Transmitter.

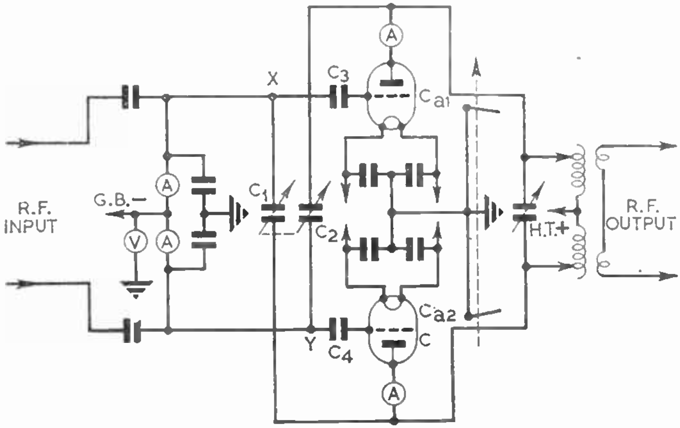


FIG. 5.—NEUTRALIZING CIRCUITS IN A PUSH-PULL RADIO-FREQUENCY AMPLIFIER.

calibrated signal generator, and adjusting for maximum reading on an output-level meter, working back, stage by stage, from the output-power amplifier to the aerial coupling. In a transmitter the master oscillator is the source of excitation and the standard of reference for the radio-frequency stages, and a tone generator provided with an attenuator pad for the audio-frequency stages. The procedure commences with the low-level stages and finishes at the output stage, the complete operation being carried out in two parts : firstly, alignment of the radio-frequency stages to produce an unmodulated carrier of the required amplitude and frequency in a test load, and secondly, adjustment of the audio-frequency stages to deliver an undistorted audio-frequency output at the correct level to fully modulate the carrier.

The frequency of high-frequency and medium-frequency transmitters is generally governed to a degree of accuracy of at least 1 part in 10^5 by a crystal oscillator, which serves as a standard of reference for tuning purposes. The correct amplifier settings have usually been recorded on the test bed, and can be repeated from the test figures or calibration charts supplied with the equipment. Each adjustment is checked by observing the readings of the anode-current and grid-current meters on the stage concerned. Having set the grid bias and anode voltages to the specified values, the alignment of any one stage entails three main adjustments with a steady input applied from the master oscillator :

(1) Tuning the input circuit and adjusting the input coupling for optimum excitation, as indicated by the grid current peaking to a maximum. Too loose a coupling is revealed by a deficiency of grid current; too tight a coupling is indicated by a double peak of grid current, and causes instability.

(2) Tuning the output circuit to resonance. In this condition

the conversion efficiency is greatest and the anode dissipation is least, as indicated by a fall of anode current.

(3) Neutralizing the anode-grid capacitance of the valve in the case of triodes, to eliminate unwanted feedback.

Neutralization of Amplifiers

Fig. 5 illustrates the operation of neutralizing a push–pull stage. The H.T. supply lead is disconnected and a miniature lamp bulb is tapped across a fraction of a turn of the anode tuning inductance. First the circuit is tuned to the signal input to obtain maximum glow in the lamp. The neutralizing capacitors C_1C_2 are then varied to reduce the glow until it is just visible. At the higher frequencies it is sometimes necessary to insert capacitive reactances, C_3C_4, in the grid circuit to neutralize the inductance of the leads. The circuit is again tuned for maximum glow, and the neutralizing capacitors readjusted, the process being repeated if necessary until the glow is extinguished. When this condition is fulfilled, the points X and Y are at equal potential, no radio-frequency energy is being transferred to the anode circuit, and conversely no feedback to the input circuit can occur under operating conditions.

Valve Voltages

The correct control grid bias and H.T. voltages for Class A, Class B, Class C or intermediate modes of amplification are determined from the valve characteristic curves and the operating conditions.

Class A Amplifiers. Lightly biased negatively to set the datum point at the centre of the I_a/E_g characteristic, as shown in Fig. 6. Positive and negative excursions of input voltage are confined to the linear part of the curve. Grid current is zero, both with and without excitation. There is a standing anode current which does not vary appreciably when excitation is applied.

Class B Amplifiers. Biased negatively to the lower bend of the I_a/E_g characteristic. Excursions of input voltage extend into the upper bend and beyond the lower bend. Grid current is zero and anode current small in the absence of excitation. Grid current appears and anode current increases with excitation.

Class C Amplifiers. Biased heavily negative beyond cut-off. Excursions of input voltage extend well beyond the upper and lower bends of the I_a/E_g characteristic. No grid or anode current flows in the absence of excitation. Heavy grid current and anode current flows during excitation.

Use of Grid-dip Oscillators

A grid-dip oscillator is a useful aid to tuning when the settings are not known. This is a miniaturized uncalibrated oscillator, equipped with a grid current microammeter and compact plug-in coils. A selected coil, adapted for insertion in confined spaces, is fitted to the end of a probe to establish a coupling with the circuit. The oscillator is brought near the L–C circuit of the amplifier and tuned to resonate. At resonance the amplifier absorbs maximum energy, which is observed by a fall of grid current in the oscillator. The oscillator frequency is then checked by offering it to a frequency meter, which need not have an accuracy

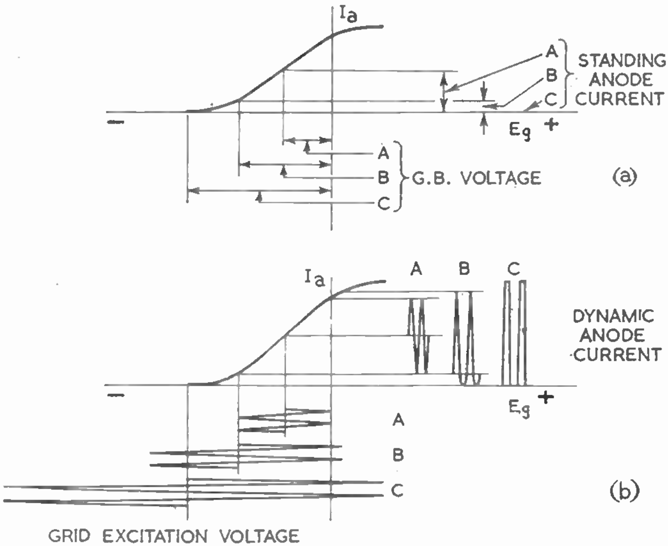


FIG. 6.—CLASSES A, B AND C AMPLIFICATION. GRID VOLTAGE AND ANODE CURRENT READINGS.

greater than about 1 part in 1,000. If the frequency is low, the tuning capacitance of the amplifier must be decreased; if it is high, the capacitance must be increased. The amplifier circuit is adjusted accordingly, and the process is repeated two or three times, if necessary, until the circuit frequency is approximately correct. Fine adjustments may then be made, and the circuit neutralized as previously described.

Typical Setting-up Procedure

The circuit arrangements of transmitters are many and varied, and although the general procedure is the same for all, it will necessarily differ in detail for each type. The routine is, perhaps, best illustrated for an amplitude-modulated broadcast transmitter.

(1) Check that power at the correct voltage and frequency is available at the main isolator, and that fuses of the correct rating are in position.

(2) With power switched on to the control circuit and the main supply switched off, see that all control switches, relays and interlocks operate satisfactorily.

(3) Start the air-cooling fan or water pump, and verify that the cooling system is functioning normally.

(4) Switch on the filament-heating supply and adjust the individual heater voltages to the specified values.

(5) Switch on the grid-bias rectifiers and check the bias voltages.

(6) Switch on the auxiliary H.T. and the main H.T. supply rectifiers, checking the voltages in turn.

(7) Connect the master oscillator and buffer stage to the first amplifier or harmonic generator and verify that grid current is indicated on the meter.

(8) Proceed to tune and align each stage, neutralizing where necessary to eliminate unwanted feedback, as previously described.

(9) With the test load connected to the final amplifier, check the power output by the reading on the load ammeter. It is assumed that the test load has previously been calibrated by a photometric or calorimetric method. See that there is sufficient excitation, and readjust each stage to obtain maximum output, taking care that the anode dissipation is not exceeded.

(10) With the master oscillator switched off, test through for stability by varying the tuning adjustment on each stage. In the absence of excitation the anode current should remain stationary, and the power in the load should be zero.

(11) Proceed to adjust the audio-frequency stages in the same way as the radio-frequency stages, except for the absence of tuning. Check by increasing the signal input that no grid current appears in the Class A stages, and that the rated anode dissipation limit on any one stage is not exceeded.

(12) Verify with an oscilloscope that the modulation wave shape at the output of the modulated amplifier is sinusoidal, and that the full modulation is achieved without limiting.

(13) Connect the feed-back circuits, and check that feedback is in the correct sense. Re-check with full power applied.

ACCEPTANCE TESTS

Final acceptance tests are more stringent and comprehensive. The most important is an extended test run for a specified period of time on full load with modulation applied, to ensure that no overheating or flash-over occurs in the valves, components and dielectric materials.

The frequency stability must conform to Atlantic City Standards (see Section 44, Table 2). Permissible frequency tolerances vary from $\pm 0{\cdot}01$ per cent for high-frequency telephony to ± 20 c/s for broadcasting, according to the class of transmission. To satisfy these requirements the frequency constancy is closely checked against a time-controlled frequency standard, and any abnormal drift from the assigned frequency is observed on a protracted run.

Noise-level Tests

Simple noise-level tests are made by sampling the modulated radio-frequency output with an oscilloscope or noise-level meter and an attenuator pad calibrated in decibels. If a noise-level meter is used, it is first necessary to demodulate the sample signal. The output level is first observed and noted with modulation applied. Modulation is then cut off, and the residual noise frequency brought up to the modulation level by cutting out attenuation. The amount by which the attenuation has been reduced is then a direct measure of the noise level in decibels, referred to modulation level. Noise level should be at least 60 db (unweighted) below 100 per cent modulation for broadcasting or 40 db for commercial telephony.

Frequency-response and Distortion Tests

The overall frequency-response characteristic and harmonic distortion are important factors in judging the performance of broadcast and commercial telephone transmitters. A response curve is obtained from a plot of output modulation voltage against representative frequencies within the audio-frequency band. Typical requirements are that the response should not deviate from the ideally flat characteristic by more than ± 2 db over a band of 30–10,000 c/s for broadcasting or ± 1 db over a band of 200–5,000 c/s for commercial telephony. The signal source used for the test is a tone-frequency generator covering the specified frequency band, and adjusted to apply a constant input voltage to the first audio-frequency stage.

Harmonic distortion may be determined with a harmonic analyser, which measures the effective voltage of the individual harmonics. Alternatively, the percentage total harmonic content may be obtained with a distortion factor meter, which measures the r.m.s. value of the harmonic voltages. For satisfactory performance the total harmonic content should not exceed 4 per cent for broadcasting or 6 per cent for commercial telephony.

Although the harmonic content may be negligibly small, unwanted inter-modulation products can still seriously degrade the audio-frequency fidelity. A knowledge of the inter-modulation products is often necessary for a broadcast transmitter. This test is made by applying a two-tone input to the modulation chain and measuring the unwanted combinations produced in the output on a harmonic analyser in the same way as harmonic distortion.

TRANSMITTER MAINTENANCE

It is convenient to consider maintenance under two headings: (*a*) running maintenance, and (*b*) periodical systematic maintenance.

Running maintenance consists of keeping a continuous watch on the transmitters while in service, making occasional adjustments and keeping an operational record. In this record entries are made of the times of switching on and off, changing frequency and varying the power output; readings of anode and grid currents and voltages, filament voltages and power output; and notes of any abnormalities and changes in meter readings since they were last recorded.

Regular systematic maintenance helps to ensure an efficient service and forestall possible breakdowns. It is often difficult to organize a regular programme where transmitters are in operation for long periods of the day and night, but as a general policy a weekly or fortnightly routine of inspection and cleaning should be arranged, and a record kept of overhauls and replacements.

The keeping of valve-record sheets is a great help in analysing the causes of valve failures and short operational life, and is often a condition of the maker's guarantee for large power valves. A record sheet is started for each of the larger valves as soon as it has been removed from its crate and subjected to an initial test. Entries are made of the initial test figures, the date the valve is put into circuit, its position and operating conditions, the number of hours in service (entered weekly or monthly), and finally the date and cause of failure. Where guarantees apply, valves with a life below average must be returned to the maker, accompanied by completed record sheets, in order to obtain credit.

Routine Inspection

Before making a round of inspection and cleaning, switch off the isolator and see that smoothing capacitors are fully discharged, as a protection against shock. If the isolator is not within sight of the apparatus being serviced, or is not interlocked, affix a warning label as a safeguard against an unauthorized person switching on in error.

Remove deposits of dust and dirt from insulators, bushings, air dielectric capacitors, coils and relay contacts with a portable blower or a brush and cloth. Lubricate control mechanisms and timing devices with a little fine oil where necessary, and finish by cleaning and polishing panels, plated metal parts and controls with a soft cloth.

Inspect power-supply equipment, cooling plant and rotary machines for visible signs of mechanical defects and charred or defective insulation.

With power switched on, check the meter readings for any deviation from normal. Verify that conductors, transformers, reactors, motors and contactor coils do not overheat under normal load. Audible manifestations are a further aid in diagnosing faults. A change in the characteristic rhythm of a motor is often evidence of overloading or excessive friction in a bearing. A hissing or crackling noise may be caused by brush discharge or intermittent arcing at insulators or between high-voltage conductors.

From time to time sample the oil in oil-filled transformers, reactors and circuit-breakers by draining a little from the bottom of the tank, and see that it is reasonably free from sludge and moisture. If necessary, subject the sample to a crackle test, and if the moisture content is found to be excessive, empty the tank and refill with fresh clean oil.

Fire Hazards

Every precaution must be taken to reduce fire risks. Fire-extinguishing appliances must be in efficient working order. No inflammable material must be placed near transmitters, aerial leads or oil-filled transformers, and insulating oil should be stored in a separate chamber or out-building. Fires are caused by defective insulation, overheating, arcing set up between high-voltage conductors and earth, and radio-frequency leakage over combustible materials.

Valve Maintenance and Testing

Large power valves should be stored in suitably designed racks in a vertical position, at an even ambient temperature between the limits of 15° and 35° C., in positions shielded from draughts and strong sunlight. When removing them from the rack and transferring them to their mountings, they must be handled with care to prevent stress being put on the glasswork and glass-to-metal seals, and must not be subjected to jerky, irregular movements or rotated in their seatings without first relieving the weight. Large, heavy valves are often transported from store to mounting position in the transmitter on a specially designed carriage to reduce lifting and risk of damage to a minimum.

Air- and water-cooled valves are tested initially for filament continuity and vacuum before removing them from their crates. Since some types require a heating current of 100 or 200 amperes, which is inconveniently large for a full-scale test, a modified test, known as the Pirani test, is carried out at reduced voltage and about 10 per cent of the normal heating current. Readings are taken at 5-minute intervals of the voltage drop across the filament terminals. When the voltage has reached a steady value, it should be within ± 5 per cent of some average value specified by the maker. Too high a voltage indicates an open-circuit in one or more of the parallel wires forming the filament cage; too low a voltage indicates that the vacuum is impaired.

During normal opeation the rated filament voltage must be closely adhered to, as a continuous over-voltage of 5 per cent can halve the life of a valve. In the interest of obtaining the maximum life, filament voltmeters should be checked occasionally with a meter capable of measuring voltages to an accuracy of ± 1 per cent. Transmitters designed for unattended operation, where the necessary manual adjustments cannot be made, should be equipped with automatic voltage regulators or flux-regulating filament transformers.

The flow of air or water must be applied to cooled anode valves before the filaments are switched on, maintained at the specified rate of flow during operation, and not removed until the filaments have been switched off and allowed time to cool down.

Hot-cathode Mercury-vapour Rectifiers

Hot-cathode mercury-vapour rectifiers contain free metallic mercury when cold, and must be stored vertically and handled so as not to disturb the distribution of mercury inside the envelope. Special precautions are necessary when putting them into commission. After storage or transport, or if they have been out of service for several hours, the cathode must be heated at the operating temperature for at least 15 minutes before applying the high-tension voltage, in order to volatilize any mercury that may have become attached to the cathode or anode. In operation the heater voltage must be maintained strictly to the rated value. If it is allowed to fall by more than 5 per cent of the rated voltage, back bombardment will deactivate the cathode and seriously reduce the life of the valve. Wherever voltage fluctuations exceed ± 5 per cent, it is advisable to install an automatic voltage regulator to stabilize the supply voltage well within these limits.

Frequency Monitoring

Although the frequency of crystal-controlled transmitters is initially preset to a high degree of precision, slow drifts occur with seasonal changes of ambient air temperature and ageing of materials, and short-period variations may result from voltage variations, keying, maladjustments and circuit defects. It is important that any deviation from the authorized frequency should be confined to the limits defined by the International Radiocommunications Regulations (Atlantic City, 1947) by checking the radiated frequency from time to time against a standard of reference.

Three general methods of checking frequency are available: (*a*) a secondary standard, (*b*) a frequency monitor, and (*c*) checking with a receiving station equipped with a secondary standard.

A common type of secondary standard is a calibrated heterodyne frequency meter adjusted to beat with the transmitted carrier and produce a directly measurable difference frequency. These instruments can be maintained to an accuracy of up to ± 1 part in 10^4 when checked regularly against standard frequency transmissions.

Frequency monitors are commonly employed with broadcasting and high-frequency communication transmitters, where the frequency is subject to close tolerances. They consist of a stabilized crystal oscillator designed to oscillate at a fixed frequency differing by, say, 1 kc/s from the carrier frequency. The difference frequency is monitored directly on a frequency meter, which provides a continuous indication of any divergence from the pre-set value.

The method of obtaining checks from a receiving station is useful for low-power fixed and mobile transmitters, where relatively wide frequency tolerances are permissible. Secondary standards at checking stations are compared regularly with standard frequency transmissions. These transmissions are broadcast daily at specified hours and frequencies by certain stations in different parts of the world, and are controlled by primary standards based on astronomical data and stabilized to a precision of ± 1 part in 10^7.

FAULT FINDING

Transmitters are monitored chiefly by meter readings. The ability to translate readings is best acquired from experience of aligning and testing, and the preceding notes will serve as a general guide. Circuit faults form a subject involving so many variants of circuit design and modes of operation, that a detailed survey is not possible in this section. The majority of faults, when narrowed down to one or two components, can be verified with a multi-range test meter.

Broadly speaking, there are two ways of checking components :

(*a*) by measurements of voltage, voltage drop or current with the circuit alive ;

(*b*) by testing for continuity, short-circuit and low insulation resistance, or checking the values with a test meter or A.C. bridge.

The first is suitable for checking small resistors, capacitors, reactors and transformers, if the voltages are not dangerously high. Insulated test prods must be used for this purpose, and it is advisable to stand on an insulating mat. The second method can be applied to most types of components, but in these tests the component must be isolated from the circuit to eliminate possible errors introduced by alternative current paths.

The following summary of the faults commonly encountered, and the tests that can be applied, will be found useful :

Fixed and Variable Resistors

Check the voltage drop with the resistor in circuit. Alternatively, test for continuity or measure the resistance with the resistor disconnected. Check for intermittent contact in a variable resistor by making a continuity test, or measure the resistance between the slider and either end of the winding while adjusting the resistance.

When testing composition resistors, it should be noted that they are less stable with time than wire-wound types, and have a negative

TABLE 2.—SYMPTOMS OF FAULTS IN RESISTORS

Symptom	*Fault*
Overheating	Inadequate ventilation. Overloading.
No current flow	Element burnt out. Open-circuit at end connector, resulting from electrolytic action.
Intermittent current flow (variable resistors)	Faulty contact, caused by excessive wear, insufficient contact pressure, dirt or corrosion.

temperature coefficient, which causes the resistance to fall somewhat when a voltage is applied.

Capacitors

Test for leakage by tapping a milliammeter in series with a D.C. source not exceeding the voltage rating across the capacitor. An initial

TABLE 3.—SYMPTOMS OF FAULTS IN CAPACITORS

Symptom	*Fault*
Overheating	Dielectric leakage.
Overheating (electrolytic capacitors)	Disintegration of dielectric film. Drying out of dielectric, accompanied by excessive leakage current.
Overheating (waxed paper capacitors)	Punctured dielectric, caused by too high a voltage or voltage surges when load is disconnected or suddenly reduced.

current surge, falling gradually to zero on making contact, shows that the capacitor takes and retains a charge. A steady current reading indicates abnormally low leakage resistance. Absence of a reading indicates an open-circuited condition.

When a voltage is applied to an electrolytic capacitor after a period of disuse, the initial leakage current is high, but falls to a normal value of 1–2 mA/μF after a minute or two, as the anode film reforms. Electrolytic capacitors have a definite storage life, and when this is exceeded the leakage current cannot be reduced by reforming.

Power Rectifiers

Test for capacitor failure as described under *Capacitors*, and for short-circuited turns or open-circuit in transformer as described under *Transformers*.

Transformers and Reactors

Thermometers are usually fitted to oil-filled transformers and reactors. The temperature of the oil should never be allowed to exceed

TABLE 4.—SYMPTOMS OF FAULTS IN POWER RECTIFIERS

Symptom	*Fault*
No D.C. voltage	Failure of A.C. supply or primary fuses. Operation of overload release, caused by flash-over in valve, overloading or breakdown of dielectric in smoothing capacitor. Open-circuited transformer winding.
D.C. voltage low.	A.C. supply voltage low. Loss of emission, failure of heater or de-activated cathode in rectifier valve. Short-circuited turns in transformer secondary winding, resulting from breakdown of insulation. Leaky smoothing capacitor.
Absence of blue glow (mercury vapour rectifiers)	Cathode de-activated. Open-circuited anode connection.

TABLE 5.—SYMPTOMS OF FAULTS IN TRANSFORMERS AND REACTORS

Symptom	*Fault*
Overheating	Overloading. Short-circuited turns in winding, due to failure of insulation and deterioration of insulating oil.
No secondary voltage	Open-circuited winding. No primary voltage.
Secondary voltage low	Overloaded secondary winding. Short-circuited turns in winding. Primary voltage low.

75° C. on normal load. If there is an abnormal rise of temperature, first check the oil level by the gauge, and if necessary refill the tank to the normal level mark. The presence of moisture in the oil is a common cause of insulation failures, short-circuited turns and overheating. Short-circuited turns reduce the resistance of the winding, but if the trouble is not revealed by resistance measurements, remove the transformer from its tank and inspect the windings for signs of charred and perished insulation.

An open-circuit is most likely to be caused by the winding burning out as a result of heavy overloading.

Rotary Machines

Overheated bearings on a normal load run will quickly make the bearing run dry and seize up. Check the oil level in the reservoir or grease in the grease-cups, and see that oil rings revolve freely.

Test the insulation resistance between windings and frame. If it is less than 1 MΩ, test each coil separately. If caused by excessive

TABLE 6.—SYMPTOMS OF FAULTS IN ROTARY MACHINES

Symptom	*Fault*
Overheated bearings	Excessive friction, due to defective balls or rollers or deficiency of lubricant.
Overheated winding and slow running (motors)	Mechanical overloading. Short-circuited turns, due to breakdown of insulation. Single phasing (in three-phase induction motor) due to one of the three supply fuses failing.
Overheated windings (generators)	Electrical overloading. Short-circuited turns, due to breakdown of insulation.
Poor commutation and overheated commutator	Excessive brush pressure and friction. Insufficient pressure or worn brushes, accompanied by sparking.
Low insulation resistance of windings	Excessive dampness when machine is installed in a damp position or has been out of service for some time.
Failure to start (motors)	Failure of supply or fuses. Faulty starter. Burnt-out turns in winding. Faulty connector, caused by excessive vibration.
Absence of output voltage (generators)	Burnt-out turns in winding. Faulty connector, caused by excessive vibration.

dampness, dry out by placing a lamp or low-wattage heater near the windings. If the condition persists, inspect the windings carefully for defective insulation.

The temperature rise of windings in open-type machines must not be allowed to exceed 72° F. during normal running. It is usually impossible to measure the temperature rise directly with a thermometer, but it is easily determined by measuring the increase of resistance of the windings. The temperature rise in ° C. corresponding to a percentage resistance increase r is approximately equal to $2{\cdot}5r$ for copper.

Short-circuited turns can be confirmed by measuring the resistance of the winding or by exciting it at reduced voltage for a short period, and observing which coil overheats or shows signs of smoking. An open-circuited winding is easily detected by making a continuity test with a test meter or a test lamp and battery.

If a commutator overheats, check that the brush pressure is correct and wear is not excessive. If necessary readjust or renew the brushes. Clean worn or grooved commutators with fine emery cloth, or in severe cases with an abrasive block.

Contactors and Relays

Before checking and adjusting contactors and relays, switch off the power supply. If it is necessary to make adjustments with the circuits energized, use tools with insulated handles to avoid risk of shock.

See that the armature pulls in and drops out freely when operated by hand and when the coil is energized. Adjust multi-pole contacts to close and open simultaneously, and equalize the spring tension if necessary. If noise and chatter is excessive, check the adjustments of the main

TABLE 7.—SYMPTOMS OF FAULTS IN CONTACTORS AND RELAYS

Symptom	*Fault*
Overheating of control coil	Excessive voltage. Excessive friction or tension in main or buffer springs.
Excessive noise and chatter	Incorrect settings. Maladjustment of springs. Dust or dirt on pole faces of A.C. contactors.
Armature fails to drop out	Excessive friction. Deficient tension in main springs. Misalignment.

and buffer springs and contact gaps. Inspect securing screws and grub screws for tightness. Verify that mercury switch tubes are securely clamped and accurately balanced.

Contacts should be cleaned with a fine-cut file or emery cloth only when deep pitting has occurred, and finished with a metal burnishing strip. A certain amount of oxidation will not seriously affect operation. High-resistance contacts may overheat. The contact resistance must be low compared with the circuit impedance, and may be checked by tapping a millivoltmeter across the contacts with the normal current flowing and dividing the reading by the current. Delayed-action contacts should be timed periodically with a stop-watch. Pivots, cam and clutch mechanisms should be inspected and lubricated occasionally with small quantities of clock oil.

TABLE 8.—AVERAGE FAILURES AND RECOMMENDED QUANTITIES OF SPARE COMPONENTS

Type of Component	*Average Percentage of Total Failures*	*Minimum Percentage Number of Spares of Each Type*
Valves, receiver type	25	500–800
Valves, glass power type	12	200–300
Valves, cooled-anode type	6	200
Indicator lamps, medium-voltage, incandescent	15	500
Indicator lamps, low-voltage, incandescent	12	250
Resistors, low-voltage, fixed	3	200
Resistors, small variable	4	200
Capacitors, electrolytic	5	200
Capacitors, small mica	2	100
Small dry rectifiers	3	100
Porcelain insulators	1	100
Small relays	2	100
Indicating meters	2	50–100
Miscellaneous	8	—

Replacements

An adequate stock of valves, components and consumable materials must be maintained for replacements and emergencies. The stock list should be checked and amended regularly and deficiencies made good.

The quantity of consumable spares, such as valves, is based on the number of each type in use and the average operational life. Spare components whose life is limited only by the possibility of accidental damage and mechanical or electrical breakdown, such as insulators, are determined from the quantity of each type in use and the yearly average percentage breakages. By far the most frequent failures, and the heaviest item of operational costs, are valves. Small receiver types have an average life expectancy of 8,000–10,000 hours; glass power valves 10,000–12,000 hours; cooled-anode transmitting valves 12,000–15,000 hours. The figures on which life guarantees are based are naturally much lower.

The probability of failure of components depends on the loading, care in handling and regular servicing. The percentage of the total failures for each type of component varies, of course, with the proportions of each type in a given installation. Typical average figures for a high-power installation, and the suggested quantity of spares for one year's working, on the basis of a 12-hour working day, are given in Table 8.

There should be a minimum of one spare of each type of component, with a few exceptions. If, for example, the equipment includes two milliammeters, one scaled 0–100 and the other 0–200 mA, a single common spare scaled 0–200 may serve temporarily in either position. The same consideration can be given to resistors of equal resistance but different wattage or small fixed capacitors of equal capacitance but different voltage rating.

MOBILE INSTALLATIONS

Mobile installations may be divided into two general groups: The first includes light equipment which can be accommodated under the dashboard or in the boot of a car. The second comprises bulky and heavy equipment, which must be installed in heavy-duty vehicles.

Widespread use is made of light, mobile V.H.F. transmitter–receivers rated up to 15 watts carrier power to maintain two-way communication between moving vehicles or with a fixed headquarters station, notably for police, ambulance, fire-fighting and taxi services.

In the heavy-duty class the installations of chief interest are: medium-power transmitters for long-range communication, used by the Services, Civil Defence and expeditions; outside broadcast, television and "Roving Eye" units, equipped for relaying outside events; test transmitters and field-strength measuring equipment for site prospecting; radar and demonstration units. Before reviewing the principles governing the different types of installation it will be useful to consider briefly the equipment of this type of vehicle.

Constructional Features

In accordance with orthodox practice, the body is built on a framing of ash, welded mild steel or aluminium-alloy sections, enclosed by sheet steel or aluminium panels, and secured to the chassis by U-bolts, reinforced by hardwood blocks. The lining is plywood, hardboard or

Formica, the space between the lining and outer panels being filled with insulating material.

Racks, benches and cupboards are laid out so as to give easy access to the equipment for operation and servicing, and avoid obstructing doors or interfering with natural lighting from windows. Spares, testing instruments and tools are stowed in lockers, cupboards, shelves or drawers, and must be firmly secured to prevent them jolting or breaking loose. Robust catches are fitted to doors, cupboards and drawers to prevent them opening when the vehicle is in motion. Stirrup or folding steps give access to the main door. A fixed or detachable ladder, stowed under the chassis or inside, is provided when accessories or a cat-walk are to be accommodated on the roof.

Vehicles are ventilated by louvres or grilles, but where the heat dissipated by the radio equipment exceeds about 1 kW, natural ventilation is supplemented by exhaust fans, equipped with weather shields to prevent the ingress of rain and snow. Hinged or detachable side panels are often fitted to power-plant vehicles to give the maximum possible ventilation.

Installation

As a first step in planning a mobile installation, it is important to note that :

(1) The rated carrying load of the vehicle must on no account be exceeded.

(2) Equipment must be arranged to distribute the load uniformly on the chassis, the centre of gravity falling as nearly as practicable midway between the front and back axles and the sides. Any out-of-balance load should preferably be biased to the off-side to counteract the effect of road camber.

(3) The overall dimensions of the body, including demountable equipment stowed or attached outside, must not exceed those permitted by the statutory regulations.

Fixed equipment liable to resonate mechanically or suffer from the effects of vibration is installed on anti-vibration mounts, rigidly fixed to the chassis, floor or bench. Light floor-mounted units may be fastened to the wood floor-boards, but heavy items must be bolted through to the chassis. Where the fixing centres do not correspond with the structural members, steel angles or channels are bolted across the chassis in suitable positions to provide the required fixing centres. Items to be fixed to walls or ceiling should be screwed to battens fitted between the ribs of the body structure, and not superficially to the lining panels.

Economy in space is important. "Forward-drive" vehicles give maximum accommodation space in the body. Cubicles and racks designed for front access only occupy least floor space, as they can be ranged against the sides of the vehicle with a central gangway between, whereas units with both front and back access require a minimum clear space 2 ft. wide back and front. Further space may be saved by the use of folding chairs and tables. When folded away, these items must be made captive by means of clamps or straps.

Essential accessories include CO_2-type fire extinguishers, placed in accessible positions for an emergency, and a panel-mounted or cabin clock for keeping operational records.

The more comprehensive type of installation, which includes power-generating plant, masts and aerials for external erection, is often accommodated in a train of vehicles equipped with a telephonic inter-communication system. Masts, aerials, R.F. feeders, power-supply and signalling cables for interconnection must be capable of being rapidly dismantled for stowage when the vehicle is mobile, and quickly reassembled for operation. To assist assembly the component parts should have distinctive identification and location marks.

Electrical Equipment of Vehicles

Except for very low-power, battery-operated equipment, power is derived from A.C. supply mains, when available, with or without a stand-by engine-alternator, or solely from an engine-alternator. If the total load does not exceed 5 kVA, one phase or phase and neutral of a three-phase supply system may be used, but for higher loading it is advisable to take power at three-phase to avoid excessive unbalance of the phase loading.

Wiring and fuse protection should conform to the I.E.E. Regulations for the Electrical Equipment of Buildings, in as far as they apply. The mains isolator and fuses must be placed in an accessible position. and a kWh meter may be fitted to meter the mains supply. With dual supply a changeover switch or separate plug-and-socket connections are necessary for changing to mains or generator. Distribution facilities usually include a main switch-fuse for the radio-equipment, with branch ways if necessary, and an independent switch and distribution fuses for lighting, heating and ventilation. The use of miniature circuit-breakers in place of fuses has much to commend it. It is sometimes an advantage to feed lights, heaters, fans and bench sockets from a ring circuit.

Wiring can be run behind panels, under floor-boards or secured by cleats to wood battens screwed to the framework, the cleats being sufficiently closely spaced to prevent fracture under stress of vibration. Sheaths, metal conduits and casing must be bonded to the vehicle earth system, continuity being maintained through joints by bonding strips.

The radio equipment is most conveniently illuminated by semi-recessed incandescent lamps with diffusers or strip lights in the ceiling or well panels, and benches by adjustable bracket lamps. Sockets for tools and inspection lamps may be fitted to work-benches and in other convenient positions. It is usual to supplement the A.C. lighting by D.C. battery lighting on a separate distribution system, for use when the vehicle is mobile. Shielded tubular heaters are commonly installed. but heating from the engine is sometimes employed when in transit.

To facilitate rapid connection and disconnection of external power supply, signalling and radio-frequency cables, waterproof plugs and sockets, fitted with captive screw caps are placed in an accessible position under the body or behind a hinged cover in the side panels. These cables are wound either on removable drums stowed in a rear compartment or on drums designed to spin on a fixed frame.

The earth system normally consists of a copper continuity strip connected to a substantial earth bolt fitted to the nearside of the chassis. All metal parts of the frame are bonded together, and metal cabinets. racks and containers are connected to the continuity strip. An earth stake and flexible lead are provided for connecting the earth terminal to ground when the vehicle is parked. An independent earth system or counterpoise may be required when an external mast and aerial system have to be erected.

Mobile High-frequency and Medium-frequency Transmitters

This class of installation is used for long-distance telegraphic and/or telephonic communication. Low-power transmitters rated up to 500 watts radio-frequency output, with associated power-supply plant are commonly installed in a light-duty vehicle on a 1½- or 2-ton chassis. Power is supplied either by a rotary transformer energized by the vehicle battery or a petrol engine generator, with optional facilities for a supply from external A.C. mains. Small engine-generators may be fitted in a separate compartment or furnished with carrying handles for off-loading by means of a portable ramp when the vehicle is stationary. The transmitter, sometimes combined with a receiver, may be bench or floor mounted; or the receiver, together with signalling key, automatic sender, recording apparatus and/or a telephone hand-set may be mounted separately on a bench.

Medium-power transmitters delivering up to 10 kW radio-frequency, with a power-distribution unit and monitoring apparatus will usually fully occupy a 5–10-ton vehicle. A comprehensive two-way high-speed telegraphic communication system might require a convoy of four vehicles, comprising a transmitting station, a receiving station with central-telegraph-office facilities, a power-generating unit and an auxiliary unit, conveying masts, aerials, cable drums and other accessories.

There are many possible variants, which it is impossible to review here. Apart from the need for a compact layout and special attention to the load distribution on the chassis, the specifications follow, in general, well-established practice for fixed stations.

Mobile V.H.F. Stations

V.H.F. transmitter–receivers are housed in compact screening cases on shock absorbers, designed for installation in the boot of a car. For convenience of operation, a remote-control unit, loudspeaker and hand microphone are fitted separately under the dashboard or near the rear seats, and connected to the main equipment through a multi-core cable. A quarter-wave whip aerial on a flexible mount, which is less liable to damage from overhead obstructions when the car is mobile, is fitted to the roof or in front of the driving-position. A flexible co-axial cable connects the aerial to the transmitter–receiver, and a change-over relay, controlled by a pressel switch in the hand-set transfers the aerial and H.T. supply connections from the receiver to the transmitter when speaking.

A selective calling device is sometimes incorporated, by means of which a pre-selected tone is made to modulate the fixed transmitter and permit any one of a number of mobile units to be called without disturbing the others. The received tone frequency is fed through a sharply tuned filter, which rejects tones of other than the assigned frequency. It is then rectified to produce a biasing voltage on the grid of a control valve, which actuates a sensitive relay.

Normally, the audio-frequency stage of the receiver is inoperative until a call is received, but on receipt of a call of the correct frequency, the relay contacts switch on the audio-frequency stage and simul-

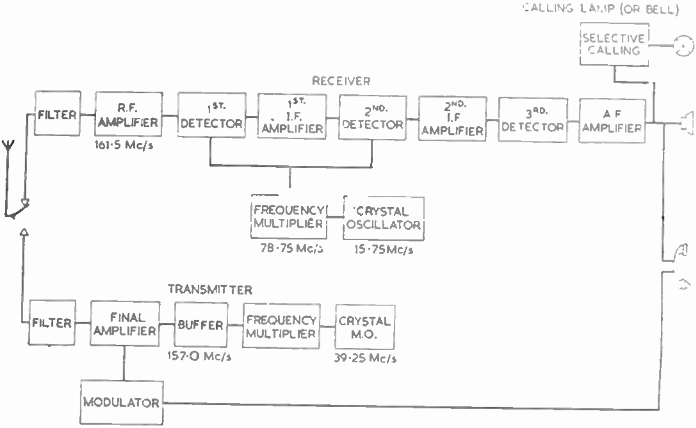


FIG. 7.—TYPICAL MOBILE V.H.F., A.M., TRANSMITTER/RECEIVER.

taneously ring a bell. Once the call has been made and the carrier is received, the receiver remains operative until the carrier is switched off, when the receiver again becomes quiescent. Fig. 7 shows the circuit arrangement of a typical V.H.F. transmitter–receiver in block schematic form.

Equipments rated up to 5 watts are designed for operation from a standard 12-volt car battery and a rotary transformer or vibrator converter. Transmitters of higher rating generally require a larger-capacity battery and a special heavy-duty charging dynamo.

Mobile Radar Units

Mobile radar installations follow the practice previously described for the heavier types of radio equipment. Like television units, these vehicles are generally air-conditioned and the windows are equipped with shutters or blinds to improve the viewing of radar displays in semi-darkness.

W. E. P.

42. MEASUREMENT OF NOISE FACTOR

42. MEASUREMENT OF NOISE FACTOR

The advent of the thermionic valve made it possible, in principle, to increase the sensitivity of radio receivers without limit merely by cascading the required number of amplifying stages. It was soon found, however, that an upper limit to the useful sensitivity was set by spontaneous fluctuations of voltage or current in the valves and circuits of the receiver. In the early days of radio these fluctuations were nearly always observed by listening to the output of the receiver, and hence became known as "noise". The random movement of electrons under the influence of heat, for example, sets up oscillations in the circuits of an amplifier; these in turn interfere with the wanted signal, so that unless this is sufficiently strong compared with the "noise" its information content will be reduced or obliterated, regardless of the amount of amplification employed.

Noise is generated not only in the receiver but also in the aerial system. If the receiver noise can be made negligible compared with that picked up by or generated in the aerial, there is little point in further improvement of the receiver. Aerial noise therefore provides the appropriate yardstick for assessing the "goodness" of the receiver, and this state of affairs has been formalized by the introduction of the noise-factor concept.

The sensitivity of a receiver is normally measured with the aid of a "dummy aerial". This should simulate as closely as possible the impedance Z of the aerial from which the receiver is designed to work, but for the purpose of noise-factor measurement it is usually sufficient to represent the aerial by a resistance R. For complete generality, however, Z should be expressed in the form $R \pm jX$, where both R and X are functions of frequency.

Across the terminals of any resistance R at temperature $T°$ (absolute) there exists a noise voltage given, for the open-circuit condition, by

$$\overline{V} = \sqrt{(4kTBR)} \qquad . \quad . \quad . \quad . \quad (1)$$

where $\overline{V}$ is the r.m.s. noise voltage, k is Bolzmann's Constant, equal to $1{\cdot}37 \times 10^{-23}$ joules/° C., and B is the energy-band-width (in cycles) over which the noise is integrated by the device used for measuring it. This formula, due to Nyquist, is often written more conveniently in the form

$$P_{av.} = \overline{V}^2{}_2 4R = kTB \qquad . \quad . \quad . \quad . \quad (2)$$

where $P_{av.}$ is the "available noise power", i.e., the power which the resistance, regarded as a generator of noise, would deliver into a matched load R.

The aerial noise level, as given by these expressions, clearly sets a limit to the useful sensitivity of the receiver but, owing to additional noise generated in the receiver itself, this limit is never quite reached in practice.

Receiver noise, like aerial noise, is proportional to band-width, and may therefore be taken into account by assuming an appropriate increase ΔT in the effective " temperature " of the aerial. The noise factor N is then the power ratio $(T + \Delta T)/T$, and represents the number of times by which the required input signal power for any given signal-to-noise ratio is increased as a result of noise generated in the receiver; it therefore provides for any specified value of T a realistic measure of the " badness " of a receiver, assuming, of course, that the gain is sufficient to ensure that noise is the limiting factor.

The noise-factor concept may be applied either to part or the whole of a receiver, and it is common to define the noise-factor of any linear network as the signal-to-noise power ratio at the input divided by that at the output. The equivalent, though less familiar, form of definition used above has been chosen, as it helps to emphasize the principles of the preferred methods of noise-factor measurement, in which the " temperature " of the source is increased by a known amount and N is computed from the effect this has on the noise output of the receiver; thus if the receiver is linear and the noise-output power is doubled, it follows that the known added noise power must be equal to $T + \Delta T$: hence knowing T it is a simple matter to evaluate $N = (T + \Delta T)/T$. The added temperature may be either an increase of actual temperature or, for example, an apparent increase obtained by passing the anode current of a saturated diode through R, and it will be noticed that noise-factor measurements made in this way are independent of band-width.

Under the normal conditions of measurement in the laboratory T is usually in the region of 290–300°, and it is therefore usual, in defining noise-factor, to specify a temperature in this region, although the " equivalent noise temperature " of the actual aerial may be very different. Difficulties may arise when high accuracy is required and the laboratory temperature differs from the specified temperature, since, although a change of T could be easily allowed for, ΔT may itself be some unknown function of T; for most purposes, however, these effects are not of practical importance.

Many different devices may be used for indicating the noise output from the receiver. It is often stated that the output must be measured " before the second detector ", but any device may be used which gives a reading proportional (or related in some known way) to the noise power averaged over the required band-width, usually that of the circuits preceding the second detector, the noise-factor being then known as the " full-band noise-factor ". In some cases, however, the " single-frequency noise factor " may be of greater interest; this is the noise factor observed when the indicator is preceded by a narrow-band filter tuned to any desired frequency within the overall pass-band, but as a rule it does not differ very greatly from the full-band noise factor unless the overall band-width is very large. In this case difficulties are liable to arise not only in the measurement of noise factor but also in its definition and interpretation; this is because (for example) the aerial can no longer be represented by a single-valued resistance, and noise contributions from various sources are not only amplified by different amounts but integrated over different band-widths. Complications also arise in the case of non-linear (e.g., crystal-video and super-regenerative) receivers. The following sections have been simplified by assuming conventional linear receivers with band-widths up to and including those normally employed in radar and television, and caution should be exercised in applying the methods and formulæ to other cases.

Hot Noise Sources

The most obvious method of measuring noise factor is to raise the temperature T' of the dummy aerial by a known amount and observe the number of times (m) by which this increases the receiver-noise power level. We then have the relation

$$(T' + \Delta T) = m(T + \Delta T)$$

whence we find that

$$N = (T + \Delta T)/T = (T' - T)/(m - 1)T \quad . \quad . \quad (3)$$

For this purpose the dummy aerial may consist of a lamp fed from an A.C. or D.C. source through radio-frequency chokes.[1] Alternatively, it may be a suitably designed resistance heated in an oven, and lengths of waveguide loaded with tapered polyiron wedges have been used in this way at wavelengths of 1–3 cm.[2]

As compared with diode noise sources, hot sources are inconvenient to use and produce much less noise power; moreover, the effective temperature of a hot wire is difficult to measure accurately, since corrections have to be made for end-cooling effects, and the arrangement is inflexible, since R usually varies with the temperature, which must therefore be fixed at the appropriate value. For these reasons hot sources are now rarely used, but they have some merit as laboratory standards at frequencies too high for accurate measurements to be made with noise-diodes, particularly when employed in conjunction with the technique described later under " Measurements of Very Small Noise Increments ", which enables quite small incremental temperatures to be used.

The Noise Diode

Any thermionic device may in principle be used as a source of noise, but there is usually a space-charge present which exerts a considerable " smoothing " action on the noise. In the case of a diode operated at a sufficiently high anode voltage, however, all the available electrons are attracted to the anode, there is no space charge, and the noise current flowing in the anode circuit is given accurately by

$$I_n^2 = 2eIB \quad . \quad . \quad . \quad (4)$$

where I is the D.C. current flowing and e is the charge on an electron. If this current flows through a resistance R the noise power available at the terminals of the resistance, due to the current, will be $eIBR/2$. If I is adjusted (by varying the filament current and hence the supply of electrons) so that the total noise power is doubled, we have

$$eIBR/2 = k(T + \Delta T)B$$

Hence

$$N = (T + \Delta T)/T = IR.e/2kT \quad . \quad . \quad . \quad (5)$$

By making $T = 290°$, which is typical of normal room temperatures, we obtain for $e/2kT$ the convenient value of 20. In other words, we may write, to an accuracy sufficient for most purposes,

$$N = 20IR \quad . \quad . \quad . \quad . \quad (6)$$

where I is measured in amperes and R in ohms.

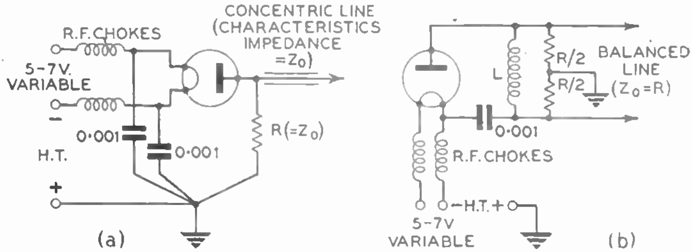


FIG. 1.—TYPICAL NOISE-DIODE CIRCUITS.

(*a*) Unbalanced line. (*b*) Balanced line.

Diodes with oxide-coated cathodes normally operate under conditions in which the flow of anode current is limited by the space-charge, and they tend to be unsatisfactory under temperature-limited conditions. It is advisable to use diodes with pure tungsten filaments, although thoriated filaments have been used successfully.

The circuit arrangements of typical noise sources [3] using the CV172 diode are shown in Fig. 1, and Fig. 2 illustrates the general form of the characteristics. The rated maximum anode current of the CV172 is 30 mA, although the writer has frequently used currents up to 80 or 100 mA for brief periods. An anode voltage of 100 is adequate to ensure saturation at currents up to 100 mA. Higher voltages increase the dissipation, and therefore the risk of valve failure at high currents.

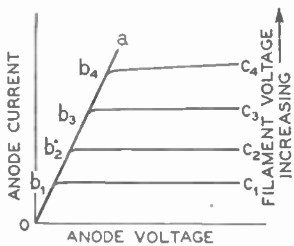
In using noise diodes, errors can arise in the following ways:

(*a*) The anode impedance (dI/dV) is in shunt with the resistance R, and this may reduce the noise output appreciably with large values of R and I. The effect may be estimated from the slope (dV/dI) of the diode characteristic at the working point; thus in a typical case (CV172) at $I = 30$ mA, $dI/dV = 20{,}000$ ohms and the error, if uncorrected, will be about 5 per cent (i.e., 0·2 db) if $R = 1{,}000$ ohms.

(*b*) *Transit-time Effects.*—These set an upper limit to the frequency for useful operation, and as this limit is approached it becomes necessary to apply correction factors; these have been calculated for the case of cylindrical diodes intended for use at micro-waves.[4]

FIG. 2.—VARIATION OF DIODE CHARACTERISTICS WITH FILAMENT VOLTAGE.

O_a represents the space-change limited region as used for rectification and mixing. b_1c_1, b_2c_2, etc., represents the temperature-limited region as used for noise-sources.



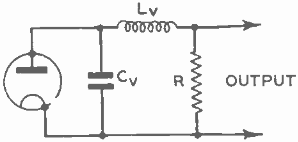


FIG. 3.—NOISE-DIODE EQUIVALENT CIRCUIT ILLUSTRATING EFFECT OF LEAD INDUCTANCE AND INTER-ELECTRODE CAPACITANCE.

In most cases, however, an earlier limit is set by the capacitance and inductance effects discussed below. A simple way of checking that transit-time effects are negligible is to raise the anode voltage, which should (in the absence of such effects) have a negligible effect on the noise output.

(c) *Lead-inductance and Inter-electrode Capacitance.*—These combine to introduce errors at high frequencies, in accordance with the equivalent circuit, Fig. 3; thus at the resonant frequency of L_v and C_v the noise-output is multiplied by Q^2 where $Q = \omega L_v/R$. In the case of the CV172, provided reasonable care is taken with the wiring, the error should be negligible up to at least 100 Mc/s. In one particular case at 210 Mc/s, with $R = 40$ ohms, the noise factor as measured with the CV172 was optimistic by about 2 db, but the error became negligible after doing away with the valve-holder, a screening can which had been placed over the valve, and the metal baseplate of the valve. The upper frequency limit for the CV172, subject to these precautions, is in the region of 300 Mc/s for less than 1 db error.

(d) The value of R is generally measured with D.C. and assumed to be the same at radio frequency. This is usually justifiable with small carbon or composition resistors having the values normally used in noise-factor measurements, but errors may be appreciable if values greater than a few hundred ohms are used at several hundred megacycles.

Specially constructed noise diodes [4] have been used at frequencies of 3,000 Mc/s and above, but, being difficult to make, fell into disfavour with the advent of fluorescent-tube noise sources, which have proved simpler and more reliable. These have the limitation that they require to be mounted in waveguides, and are therefore inconvenient to use at frequencies lower than 3,000 Mc/s.

The range from 300–3,000 Mc/s has received relatively little attention as regards the development of noise sources.

It is possible to derive a known amount of high-frequency noise from a lower-frequency diode source by means of a mixer, applying the reciprocity principle as described later under "Generation of Radio-frequency Noise Power from an Intermediate-frequency Source."

Fluorescent-tube Noise Sources

An ordinary commercial fluorescent tube may be mounted across a waveguide in such a way that energy fed into the guide is almost entirely absorbed in the discharge. This arrangement can be thought of as a "resistance", more or less matched to the guide, and at a temperature in the region of the electron temperature of the gas. Such a device can be used as a "hot" noise source, and possesses important advantages over the "solid" hot sources described earlier; thus very little heat is generated, and yet the effective temperature is relatively

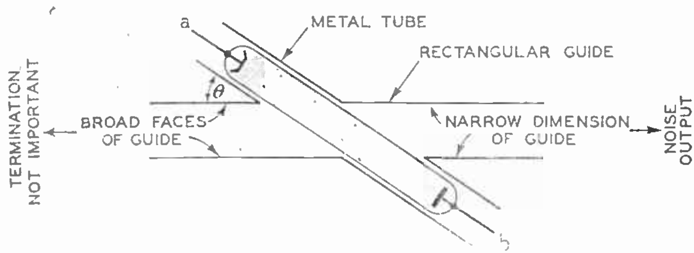


FIG. 4.—SIMPLIFIED DRAWING OF FLUORESCENT-TUBE NOISE SOURCE.

Angle θ should be as small as possible. Tube is mounted centrally, i.e., in a plane parallel to, and half-way between the narrow faces of the guide. The metal tubes act as chokes, preventing the escape of radio-frequency energy. Main supply is connected to a, b.

high, nearly independent of current and capable of being rapidly switched.

It is usual to mount the tube in the E-plane, as shown in Fig. 4, the angle θ being made as small as possible. In this way an attenuation of 20 db or more can easily be obtained through the discharge, with standing-wave ratios better than 0·9. H-plane mounts have been used, but are not quite so good, and have the disadvantage of producing more reflection when the tube is cold. Measurements are made by turning on the supply to the tube and observing the effect this has on the receiver-noise output, and since the noise level of a "good" receiver is appreciably dependent on radio-frequency matching, any difference of match between the "hot" and "cold" conditions will be a possible source of error.

Errors may also be caused by low-frequency oscillations such as are frequently present in tubes of this type. These oscillations do not affect the temperature directly, but may modulate the value of impedance presented to the guide, and this in turn can affect the mixer in various ways. Such oscillations can be suppressed by using gas mixtures, as in some commercial lamps containing argon and mercury, and their micro-wave effects can be reduced by any measure which increases the plasma density, e.g., increasing the current or gas pressure, or using an inert gas of greater atomic weight. Since these measures increase the attenuation through the device they also help to prevent errors due to reflections from beyond the device, but with the E-plane mount, and except with abnormally low pressures and currents, errors from all these causes are likely to be fairly small.

In addition to commercial fluorescent lamps, argon-filled tubes are often used. The noise level is usually taken in both cases as 15·5 db above room-temperature noise over the entire micro-wave range, this figure being about 1 db below the calculated and measured electron temperatures. Noise factor may therefore be computed from Equation (3) putting $T' = 10{,}500°$. Variations between lamps of the same type are mostly fairly small fractions of a decibel, and the variations of any given lamp during its life are even less, but there is still some lack of general agreement about the absolute level, and the reader would

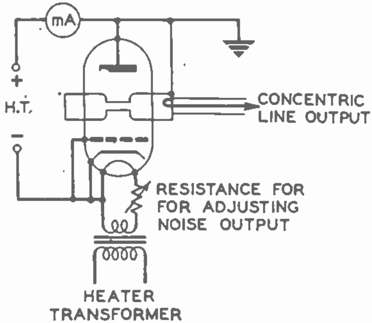


FIG. 5.—TYPICAL CIRCUIT OF NOISE KLYSTRON.

perhaps be well-advised to assume a possible error of at least $\pm 0{\cdot}5$ db in the figure quoted.

Other Noise Sources

Both klystrons and crystals have often been found useful as noise sources for comparative noise-factor measurements, although unsuitable as absolute standards. Klystrons may be connected as shown in Fig. 5 and the anode current adjusted to a convenient value by variation of either anode voltage or heater current, the noise-output power being proportional to current for small currents up to, say, 2 mA. The noise level obtainable at 2 mA may be of the order of 300 kTB, but, since the cavity is in general fairly selective, this device is unsuitable for measurements on very wide-band receivers.

A crystal for use as a noise source may be mounted in any convenient holder, e.g., a spare mixer assembly coupled to the receiver under test through a suitable matching transformer, and energized by a D.C. current of 2–4 mA in the reverse direction. Noise temperatures of the order of 3,000° are readily obtainable; considerably higher temperatures may be obtained with partially "burnt-out" crystals, but these tend to be rather unstable.

Measurement Techniques Using Noise Sources

The concept of noise factor is restricted to linear systems, and it is often stated that the output noise must therefore be "measured before the detector". This is a useful convention for purposes of definition, but does not exclude the possibility of using the detector, either on its own or in conjunction with subsequent circuitry, as the measuring device. There are divergences of opinion as to the best form of indicating device, and some workers replace the second detector of the receiver by thermocouples or bolometers; apart from their inconvenience, such devices tend to be sluggish in operation, and this can lead to errors due to changes of gain during the measurement. The majority of receivers employ diode detectors, which normally provide a linear relation between input-noise voltage and rectified current for voltages of the order of 0·5 V. up to some value determined by overloading of the preceding intermediate-frequency stages. It is usually a simple matter to connect a milliammeter or microammeter (depending on the load resistance) in series with the load, and an increase of $\sqrt{2}$ times in the meter reading then corresponds to a doubling of the noise power. Doubling the meter reading implies a four-fold increase of noise power, so that in this case the *added* noise power is three times as great, and the noise-diode current or temperature increment must be divided by three before applying the appropriate formula. It is a simple matter to check whether the linear-law holds, since, if it does, measurements made at these two levels will agree. In the writer's experience departures from linearity have always been due to one of the following causes:

(*a*) Excessive diode standing-current I_0. It is advisable to use a working level not less than about 4 I_0.

(*b*) Attempts to correct for the standing current by subtracting it. Correction is unnecessary if condition (*a*) is satisfied.

(*c*) Non-linearity of preceding circuits.

(*d*) Insufficient input signal.

(*e*) Meter errors.

If the above test indicates non-linearity, it may be worth-while to repeat it at a higher or lower gain setting.

Square-law detectors present no difficulty, the rectified current being proportional to the total power input. A silicon crystal operating at low level (rectified current $<5\ \mu$A) is a convenient form of square-law detector, and may be made up as a probe-type voltmeter. Such a voltmeter may be applied directly to an intermediate-frequency circuit, the effect of this on the circuit being equivalent to a load of 10,000 ohms or so, which, though it appreciably restricts the general scope of the instrument, does not interfere with its use for noise-factor measurement; care must be taken, however, not to damage the crystal by excessive voltages or switching surges. Germanium diodes may also be used; these are more robust, and in general load the circuit less heavily, but may have a significant effect on tuning.

It is not essential for the detector to be either linear or square-law, as the following ingenious procedure enables any law to be used; the receiver gain is adjusted to give a convenient meter reading M_1 due to $(T + \Delta T)$ alone. On raising T to T' (or adjusting the noise-diode current to a value I_1) we obtain a reading M_2. The gain is now turned down so that the reading is again M_1, and T' is then increased to a value T'' (or noise diode current to I_2) such that the reading is again M_2. We now have the obvious relation:

$$(T' + \Delta T)/(T + \Delta T) = (T'' + \Delta T)/(T' + \Delta T)$$

Knowing T, T' and T'' we may therefore evaluate ΔT, and hence N from Equation (3).

This procedure requires relatively large and easily adjustable noise powers, conditions which tend to restrict it to the case of diode-noise sources, for which we obtain the relation

$$N = \frac{20 I_1^2 R}{I_2 - 2I_1} \qquad \qquad (7)$$

in place of equation (6).

Similar procedures may be used for measuring the equivalent noise temperatures of T_a of an actual aerial; we first observe the output noise with the aerial connected, and then replace the aerial by the equivalent resistance R in conjunction with a noise diode, which is adjusted to a current I, such that the noise level is the same as before. We then have the relation

$$T_a - T = 20\ IR.T$$

e.,

$$T_a = (20\ IR + 1)T \qquad \qquad (8)$$

Note that in this case it is not necessary to know the detector law. If T_a is too large to measure in this way, it may be reduced to any desired extent by inserting attenuation between the aerial and the receiver, subject to the appropriate correction.

Measurement of Noise Factor using C.W. Signal Generators

If a known signal power $P_s = m.k(T + \Delta T)B$ is made available from the dummy aerial by means of a C.W. signal generator, we have the relation

$$N = (T + \Delta T)/T = P_s/m.kTB \quad . \quad . \quad (9)$$

The presence of P_s increases the output power of the receiver by $(m + 1)$ times, so that by observing m we may evaluate N, provided the energy band-width B is known.

To determine B it is necessary to plot the power input P to the second detector as a function of frequency (f) for a constant available power from the signal generator. If P_0 is the on-tune value of P we have the relation

$$B = \frac{1}{P_0}\int_{-\infty}^{\infty} P\,df$$

Disadvantages of this method, in addition to the need for evaluation of the integral, are: (*a*) the difficulty of measuring P_s accurately; (*b*) effects of signal-generator leakage, which, even though smaller than the direct power, can cause relatively large errors by combining in or out of phase with it; (*c*) the detector, unless square law, must be replaced by a power-measuring device, owing to the difference in waveform between the noise and the test signal, which makes it inaccurate to use a comparison of (for example) mean values as a measure of the relative powers.

Generation of Radio-frequency Noise Power from an Intermediate-frequency Source

This may be done [7] by means of a mixer as shown in Fig. 6.

The first diode works into a load R in the usual way, and the second diode has as its load the intermediate-frequency output impedance R_x

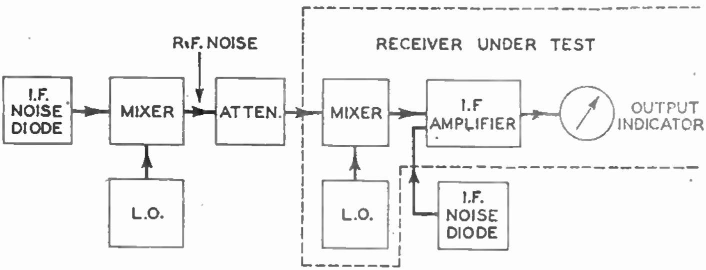


FIG. 6.—USE OF MIXER FOR OBTAINING MICRO-WAVE NOISE FROM AN INTERMEDIATE-FREQUENCY SOURCE.

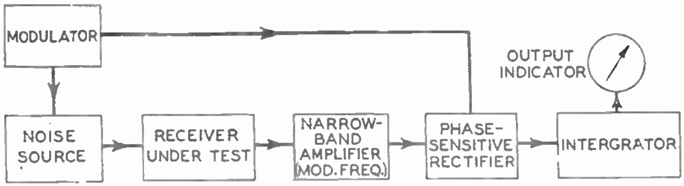


FIG. 7.—SKETCH SHOWING APPLICATION OF DICKE METHOD TO THE MEASUREMENT OF RECEIVER SENSITIVITY.

of the mixer in the receiver. If I_1, I_2 are the first and second diode currents for identical deflections of the output meter, and if the mixers are carefully matched, we have

$$20 I_1 R_s L_a L^2 = 20 I_2 R_x$$

where L is the conversion loss of each mixer. L_a and R_x being measurable, we can evaluate L and hence, if I_1 doubles the noise output (i.e., $m = 2$),

$$N = 20 I_1 R / L L_a \quad . \quad . \quad . \quad (10)$$

This assumes the reciprocity principle, i.e., that the conversion gain of a mixer from radio frequency to intermediate frequency is the same as from intermediate frequency to radio frequency, which has been reasonably well established for silicon mixer crystals. The two local oscillators are tuned to opposite sides of the signal frequency. The input circuit of the first mixer should be tuned to the intermediate frequency, or alternatively a rejector circuit used to eliminate noise at three times the intermediate frequency, as this results in an image-frequency component at radio frequency. L_a should be at least 12 db, and it will usually be necessary to work with fairly small values of m, e.g., in the region of 1·4–2, unless an amplifier is used after the noise source.

R_x may be measured in various ways. One method is to substitute known resistances for the crystal; variation of resistance usually affects the receiver output voltage when an intermediate-frequency signal is loosely coupled (e.g., through a small capacitance) into the input circuit, and a calibration curve may be drawn accordingly.

Measurements of Very Small Noise Increments

If noise is integrated over a wide band-width B, for a relatively long period t, the result is a steady D.C. voltage V with a small residual r.m.s. fluctuation of the order of $V\sqrt{tB}$. Practical values of t and B are 10 seconds and 10 Mc/s respectively, so that the residual fluctuation is only of the order of 1 part in 10,000, and an incremental noise voltage of this order of magnitude should be perceptible, i.e., in terms of Equation (3) the lowest possible value of m becomes 1·0002. In practical cases a residual fluctuation of this order would be completely "swamped" by the effect of variations of receiver gain, flicker noise, etc. This difficulty may be overcome by a method due to R. H. Dicke,[8] which in effect employs a much smaller value of t but achieves an equivalent result by averaging a large number of observations (Fig. 7).

The incremental noise is switched on and off at, say, 30 c/s, and the resulting modulation of the receiver output is detected by applying it in the same phase to each of a pair of diodes having a common load resistance. A relatively large 30-c/s voltage synchronized with the switch is applied to the same two diodes but in opposite phases, so that there is no residual load current unless the receiver output contains a voltage which is phase-correlated with the switching waveform, in which case it increases the rectified current of one diode and reduces that of the other. This detector is followed by an integrator having a time constant of several seconds. With this method, using an intermediate-frequency band-width of 16 Mc/s and a meter time-constant of 2·5 seconds. Dicke has been able to detect micro-wave source temperature changes of 0·5° C., so that an incremental temperature of only a few degrees should in principle suffice for reasonably accurate noise-factor measurements. Errors are liable to be introduced, however, owing to the difficulty of maintaining identical matching conditions for both positions of the switch, and hence the temperature change should be as large as possible.

This technique has hitherto been employed mainly for measurement of aerial temperatures.

Direct-reading Noise-factor Meters [9]

Instruments have been described for giving a direct reading of the relative noise factors of, for example, mixer crystals.

The noise source is modulated and an automatic-gain-control system holds the noise output of the receiver at the peak of the modulation cycle constant at some predetermined level. The worse the receiver, the less the effect of the modulation on the total noise-output level, and therefore the less the difference between the peak level (which is constant) and the trough. An A.C. voltmeter reading the modulation-frequency component of the receiver output may therefore be calibrated in terms of noise factor.

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L. A. M.

43. PROJECTION TELEVISION SYSTEMS

43. PROJECTION TELEVISION SYSTEMS

The difficulties in manufacturing large cathode-ray tubes at an economic price led to the development of projection systems in which a small-diameter cathode-ray tube gives a brilliant picture, which is then passed through a suitable optical system on to a large screen. In practice three main types may be distinguished:

(1) THE BACK PROJECTION SYSTEM for domestic receivers, in which the optical unit is situated behind the screen, the latter being of ground glass or some other suitable translucent material. This method was used in the Philips and a number of other manufacturers' large-screen projection sets.

(2) THE FRONT PROJECTION SYSTEM for domestic receivers. In this case the screen is separate from the receiver unit, and an optical system is used to project the enlarged picture on a screen of high reflective power suspended on a wall directly facing the lens of the receiver. This was the system employed in the Decca and Philips large-screen front-projection receivers.

(3) LARGE-SCREEN PROJECTION suitable for cinemas and theatres. This employs the ordinary Schmidt optical system, whereas systems 1 and 2 employ the Schmidt "folded optical" system using plane reflecting mirrors.

The manufacture of projection receivers is now confined largely to special markets, such as schools and clubs, very few sets now being made for the domestic market.

The main objective of most projection receivers is to produce a large picture which is at least as acceptable as a smaller direct-viewing one. This may not be the case with larger systems intended for cinemas or other public showing, where the necessity for satisfying a large number of patrons simultaneously may to some extent override the ideal of technical perfection. The following remarks are therefore more applicable to viewing with the smaller projection receivers.

The quality of the final picture is first affected by the small cathode-ray tubes used, in so far as difficulties may be encountered in providing the required resolution and contrast at the high brightness which is necessary at the tube face. The optical system then has its effect. In place of a lens it is now more usual to employ a version of the Schmidt camera originally designed for astronomical use. The small cathode-ray-tube face is placed at the centre of a spherical mirror, and the distortions which necessarily occur in the reflected beam of light are compensated by a corrector plate of special shape placed in its path. In order to make the system compact, the light path is frequently folded by means of a plane mirror set at 45° to the axis of the beam, the cathode-ray-tube face projecting through a hole in the centre. This mirror also obscures the rather bulky scanning yoke from the light beam, as it would otherwise reduce the aperture considerably. An effective f/No. of about 0·7 may be obtained with this system, and it is therefore an efficient way of imaging the small cathode-ray-tube picture on to a larger screen. Each system is designed for a definite optical-path length, and hence the distance from the screen is fixed. Many sets are constructed with a self-contained screen, which is therefore fixed in its proper position, while other sets having separate screens are provided with means for making the necessary adjustment in distance. Despite the complication and possible sources of degradation introduced by these optical arrangements, it is stated that they are more than adequate for the reproduction of a 625-line picture, so that they should easily cope with the B.B.C.'s 405-line transmissions.

Another influence on picture quality is the projection screen itself. Reflecting screens for front projection vary from simple white flats to screens with special curvature and surface which reflect most of the available light in and around a preferred direction, giving the maximum brightness over a small field of view. Most self-contained receivers use back projection with translucent plastic screens, and for the purpose of providing a uniformly bright picture over as wide an angle as possible the screen surfaces are frequently ruled with straight or circular lines so as to form a multiple lens of the Fresnel type. It has already been mentioned that spot wobbling may be a great asset in projection sets where the cathode-ray-tube screen loading is high, and it is somewhat surprising that it is not more generally used, especially when it is remembered that the visual processes demand better quality from large pictures than from small ones.

The questions of ambient illumination and contrast need reconsideration with projection sets. On one hand, it is generally true to say that the maximum brightness attainable with good quality is lower than with a direct-viewing set. The contrast also is usually lower due to the combined effects of the small cathode-ray tube, the optical system and the special screen. On the other hand, as there is no specular reflection from the screens, the room lights and windows do not appear as images on the picture. With a translucent screen too, some of the stray light which reaches it passes through and is mostly lost from view, so that it can be said that projection sets are perhaps less sensitive to the effects of ambient illumination and stray lights than direct-viewing receivers. Whilst it is therefore apparent that direct viewing and projection television pictures may not look precisely equal in all respects, it is possible for the physical and psychological factors affecting the quality to be so interchanged that the two appear equally attractive.

The advantages of a projection receiver over its directly viewed counterpart may be summarized as follows :

(1) The replacement cost of the tube is small.
(2) The screen is perfectly flat.
(3) There is no danger from implosion of the tube.
(4) Larger pictures are possible.
(5) The projected picture has good blacks, which gives good contrast without excessive luminance.
(6) There is almost complete absence of deflection-defocusing, because of the small angle of deflection.

Its disadvantages are :

(1) The cost of the optical equipment.
(2) More skilled maintenance and setting up are required.

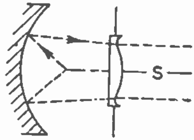


FIG. 1.—BASIC PRINCIPLE OF THE SCHMIDT OPTICAL SYSTEM.

The spherical mirror is the main focusing element.

(3) More difficult to repair because of the optical equipment, and the tube-protection circuit, which cuts off the spot if the time-base fails. This makes diagnosis more difficult.

For very large-screen pictures, a directly viewed tube is, of course, not technically possible.

THE SCHMIDT OPTICAL SYSTEMS

It will be appreciated from the above that the two important items in any system of projection television are the special type of cathode-ray tube employed and the optical projection system.

(1) THE DIRECT SYSTEM. The Schmidt optical system is shown in its simplest form in Fig. 1, in which figure it will be seen that light from an object in the focal plane of the mirror is reflected back through a correcting plate or lens, the purpose of which is to eliminate spherical aberration.

Fig. 2 shows how this principle can be applied in a television projection unit. From this it will be seen that the high-intensity tube is arranged co-axially with the spherical mirror, the position of the tube along the axis being adjusted during practical tests. The light emitted from the fluorescent screen of the cathode-ray tube is reflected by this mirror, and passes through the correcting plate or lens on to the screen of the television receiver. The hole at the centre of the mirror serves a double purpose. First, if the mirror had no aperture, most of the light falling on the centre portion would be reflected back on to the face of the projection tube, with a consequent blurring of the image. Secondly, the aperture allows for the air cooling of the tube face, should this be found desirable with high-intensity tubes.

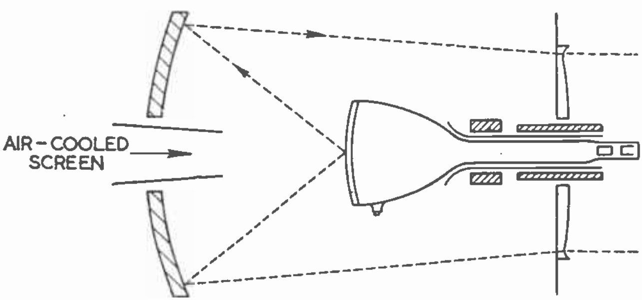


FIG. 2.—SCHMIDT OPTICAL SYSTEM APPLIED TO PROJECTION TELEVISION.

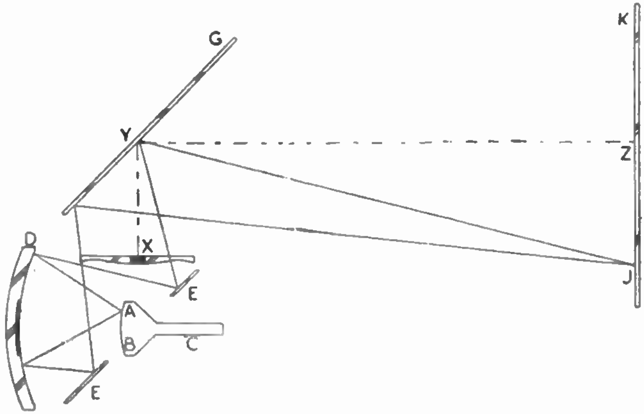


Fig. 3.—The Schmidt "Folded Optical" System.

(2) The "Folded" System. In the "folded" version of the Schmidt optical system, which is shown in Fig. 3, the cathode-ray tube C projects through a central aperture in a mirror E, which is inclined at an angle of 45° to the axis of the spherical reflector. The inclined mirror E catches the light from the spherical reflector and directs it through the correcting lens X on to a second plane mirror G, from which it is again reflected on to the projection screen.

CATHODE-RAY TUBES FOR PROJECTION SYSTEMS

Tubes for Large-screen Projection

The projection tubes used for large-screen television equipment have an anode voltage of 50–80 kV. The average beam current is 1–2 mA with a peak value of 15 mA, the cut-off and drive voltages being respectively minus 500 and plus 400 volts. A high-frequency vertical oscillatory motion is imparted to the beam to produce "spot-wobbling", as it has been found that this increases the brightness of the picture and eliminates field line structure. The diameter of the face of the tube is 9 in., and the outer surface of the tube face is cooled by a current of air blown through the centre aperture of the spherical mirror. A magnetic focusing coil is used; the deflection system also employs magnetic coils. An eight-pole magnetic field is used for correcting pin-cushion distortion of the picture.

Tubes for Smaller Projection Receivers

Those receivers employing the folded optical system use a smaller picture tube generally having a screen diameter of 2½ in. The following notes apply to the Mullard type MW6·2, which is designed to operate at an anode voltage of 25 kV. At a beam current of 100 μA, the spot size is 170 μ (0·0068 in.). The tube is quite conservatively rated and can, if necessary, pass higher peak currents. The spot size is then somewhat increased, with the consequent reduction in picture definition, but, as this current corresponds to peak white, slight loss of definition can be tolerated. The design of the electrode assembly is, of course, governed mainly by the consideration of beam size. The requirements are that the beam must be of such dimensions as to produce a spot of the required diameter, yet not so concentrated that mutual repulsion of the electrons produces blurring of the spot. Further, the beam at its widest diameter must be small enough to keep the electrons well clear of the tube walls and the anode when fully deflected. Finally, the I_a/V_g curve must be sufficiently steep to allow the tube to be driven by a normal video-output valve.

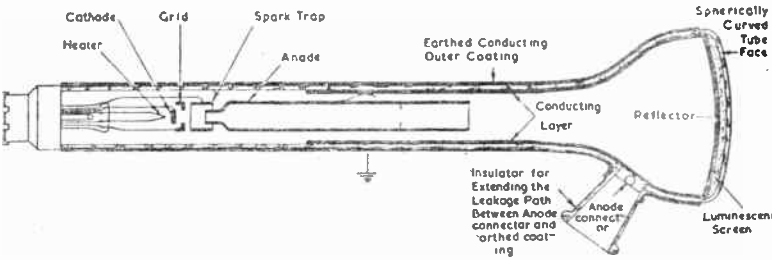


FIG. 4.—SECTIONAL DRAWING OF THE MULLARD MW6-2 CATHODE-RAY TUBE FOR PROJECTION TELEVISION.

A sectional drawing of the tube is reproduced in Fig. 4. A spark trap, consisting of a ring-shaped electrode, is situated between the grid and the anode, and is connected to one of the base pins, which should, in turn, be connected to chassis. Any discharge which might occur, due, for example, to the release of a small amount of gas as the result of unintentional overload, will take place between the anode and the spark trap, thus avoiding damage to the cathode.

The external surface of the neck and cone is coated with a graphite preparation, and must be earthed. This coating, with the glass envelope and the internal metallizing of the tube, forms a capacitance of approximately 450 pF, which, with a 1M resistor in the 25 kV lead, serves as the final smoothing for the E.H.T. supply.

The most conspicuous external feature of the tube is the glass shield surrounding the anode terminal. This shield obviates risk of flash-over or leakage between the E.H.T. connection and the deflection coils or the graphite coating.

The luminescent screen is backed by a metallic coating. The metal used is aluminium, which has the necessary high optical reflectivity and also a low atomic weight to permit easy penetration by the electron beam. The thickness of the layer is in the order of 0·4 μ. The metal is evaporated within the tube by conventional methods, and is deposited as a thin film on the rear surface of the screen and on the flare of the tube, to which has previously been applied a thin layer of nitro-cellulose to provide an even surface.

The metal-backed screen has many advantages. In the first place, the metal coating reflects outwards a large proportion of the emitted light, which would otherwise be directed to the rear of the tube. The increase in the output of forward-going light due to this feature is, under correct operating conditions, as much as 75–80 per cent. The metal backing also eliminates the internal reflections and screen irradiation which would otherwise result from the light directed towards the rear of the tube. Picture contrast is therefore considerably enhanced. Finally, the metallic film serves as an effective ion trap, as the negative ions, which would otherwise ultimately be responsible for ruining the screen, are not able to pass through the film.

The tube face is made from a special type of glass which is free from the discoloration so often occurring in tubes after a long period of operation at anode voltages above some 15 kV. This discoloration, which ranges in colour between purple, brown and black according to the type of glass, appears over the area on which the raster is formed, and is mainly due to the generation of soft X-rays resulting from the bombardment of the screen by the high-velocity electron beam.

The glass face of the tube is formed, by pressing, to the required convex radius to suit the restricted depth of focus of the Schmidt optical system, and this radius is controlled to a very high degree of accuracy.

Tube Data

Heater. This tube is suitable for series operation.

V_h	6·3 volts
I_h	0·3 amperes

When the heater of the MW6-2 is used in a series heater chain a current-limiting device may be necessary in the circuit to ensure that the surge voltage across the tube heater does not exceed the rated limiting value of 9·5 volts r.m.s. when the supply is switched on.

Mounting Position. The tube may be mounted in any position except screen downward within a cone with a top angle of 100°.

Capacitances.

$c_{g-\text{all}}$	6·3 pF
$c_{k-\text{all}}$	6·3 pF
c_{a-M}	450 pF

M indicates external conductive coating.

Screen Properties.

Colour	White
Colour temperature	6,500° K.
Maximum useful screen diameter	57·5 mm.

Magnetic Focusing. The number of ampere-turns required for focusing is approximately 920 at $V_a = 25$ kV and with an air gap of 11–13 mm. in the magnetic circuit at the end nearer the screen.

The inner diameter of the focus coil should be 27·5 mm., and provision should be made for adjusting the axis of the coil over a range of 2·5–3° in both the horizontal and vertical directions to allow for picture centring.

Double Magnetic Deflection.

Deflection angle 38°

Deflection sensitivity $\dfrac{0{\cdot}3P.L.a.}{\sqrt{V_a}}$ cm./gauss

where P = length from centre of deflection coils to screen in cm.;
V_a = anode potential in volts;
L = length of the electron path through the field of the deflection coils in cm.;
a = correction factor, which depends upon the shape of the coils and the field strength, and is normally about 0·5.

Typical Operating Conditions.

V_a	25 kV
$I_{a(av)}$	150 μA
$v_{drive(pk)}$ excluding synchronizing pulses	65 volts
* V_{g1} for beam cut-off	− 40 to − 90 volts

Spark trap must be at earth potential.

In order to keep the dissipation of the phosphor and the tube face within safe limits the raster area should not be less than 14 sq. cm.

Permanent-magnet Focusing. In special circumstances it may seem advantageous to employ permanent-magnet focusing in projection receivers. Owing, however, to the difficulties then experienced in ensuring adequate protection to the tube on switching off and also in the event of time-base failure, no standard equipment with the feature has been introduced.

Deflection-coil Data.

Line-deflection coil :	
Inductance	3·24 mH
Resistance	4·4 ohms
Peak-to-peak deflection current for full scan	825 mA
Frame deflection coil :	
Inductance	4·0 mH
Resistance	12·2 ohm ± 10 per cent
Peak-to-peak deflection current for full scan	500 mA

Tube Protection and Compensation for Pin-cushion Distortion

It is essential that means be provided for the instantaneous removal of the beam current in the event of failure of either one or both of the time-bases. Unless such a safety device is incorporated, a failure of this type will result in the immediate destruction of the cathode-ray-tube screen.

An optical system incorporating a spherical mirror introduces a certain degree of pin-cushion distortion, which, in the Mullard projection television system, is compensated in the design of the deflection coils.

X-ray Protection

It has been recommended that in circumstances under which the whole body may be exposed, over an indefinite period, to X- or gamma-radiation of quantum energy less than 3 MeV the maximum permissible

* The direct value of grid bias must not be allowed to become positive with respect to the cathode.

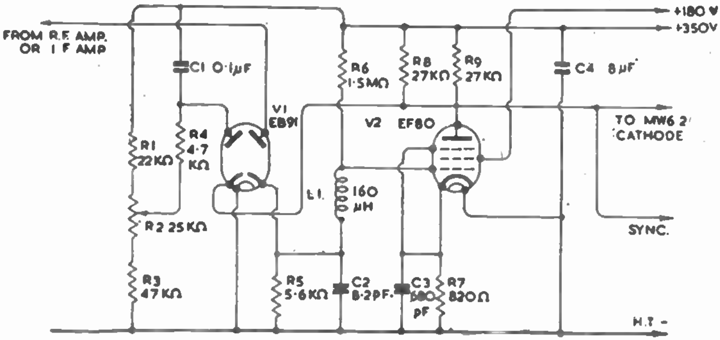


FIG. 5.—VIDEO OUTPUT STAGE USING EF80.

dose received by the surface of the body should be 0·5 röntgen in any one week. Only when the tube is outside the projection unit is a dosage rate of this order exceeded. For the purpose of normal adjustment, centring, etc., when it is necessary to view the tube outside an optical unit, the tube should be operated at the minimum brightness consistent with examination of the picture. No protection against X-rays is then necessary.

With the tube in the projection unit the highest rate of X-ray radiation is only about one quarter of the maximum permissible, while with the projection unit fitted in the plywood case the highest radiation rate is about 1/40 of the maximum permissible and at distances of 1 ft. or more from the cabinet it is never greater than 1/100 of the maximum permissible and in most directions is less than 1/1000 of this value.

Should it be desired, however, to operate the tube at full brightness outside an optical unit, the use of a lead-glass shield is recommended. The location and dimensions of the shield should be such as to interrupt the X-rays between the tube and the observer. The equivalent lead thickness of the shield must not be less than 0·5 mm. This precaution is desirable in the region within 40-in. of the tube screen.

CIRCUITRY

When it is intended to incorporate a projection unit in a television receiver, there are several points in the circuit and general design which should be borne in mind.

Video Drive

By using cathode compensation, it is possible to use an EF80 R.F. pentode in projection video stages quite satisfactorily. A suitable circuit is shown in Fig. 5. This circuit has a high-value cathode bias resistor (820 ohms) and a high value of anode load (13,500 ohms). With this load resistance a current swing of only 6·3 mA is required for 85 volts output, and this arrangement can be used because of the high feed-back factor provided by the cathode resistor.

The ideal cathode resistance for feed-back purposes is a little too

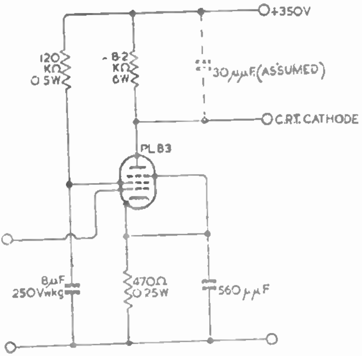


FIG. 6.—VIDEO OUTPUT STAGE USING PL83

high for bias purposes, and a correcting positive bias (0·9 volt) is applied to the grid through resistor R6. The demodulator, also shown in Fig. 5, is biased by this small potential, and this clips the tips of the synchronizing pulses.

The potential drop across the anode load under quiescent conditions is approximately 55 volts, and the maximum swing from the H.T. line may therefore be in excess of 155 volts. For this reason the anode must be fed from an H.T. line of at least 275–300 volts, and preferably 350 volts.

The video stage shown in Fig. 5 has a gain of approximately 11·5, so that the last R.F. stage and demodulator must be capable of delivering at least 10 volts peak-to-peak to the video stage.

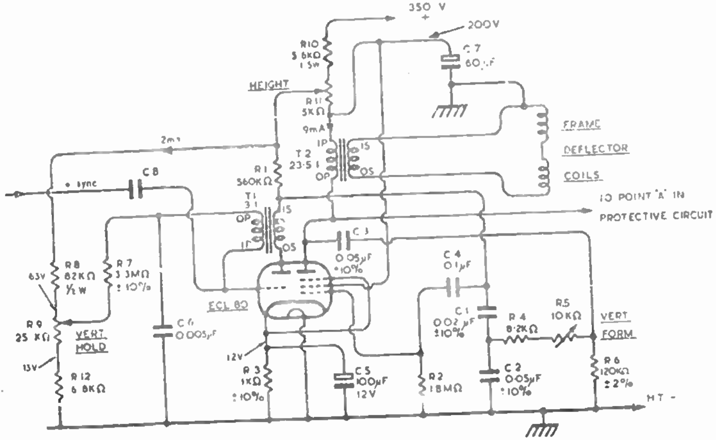


FIG. 7.—FRAME TIME-BASE GENERATOR CIRCUIT.

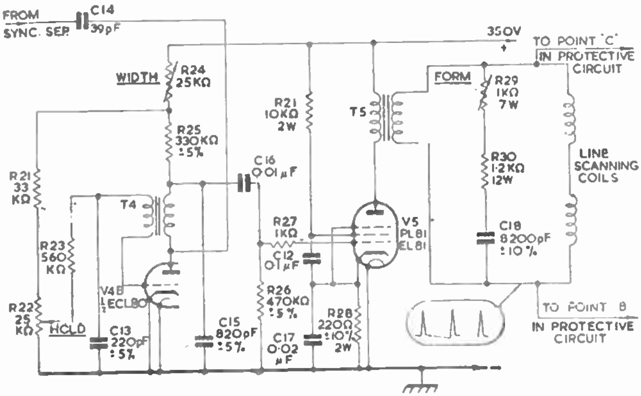


FIG. 8.—LINE TIME-BASE GENERATOR CIRCUIT.

Video Drive Using PL83

Where it is required to take full advantage of the rating of the MW6-2 projection picture tube and the improvement in spot-size at zero bias, a video stage capable of a greater output voltage than that obtainable with EF80 may be desirable. A suitable circuit using PL83 is shown in Fig. 6.

The cut-off limits for the MW6-2 tube are −40 and −90 volts. In order to cope with the highest cut-off value, the required peak-to-peak signal, assuming 30 per cent synchronizing ratio, will be 129 volts. To allow for all tolerances, a nominal output of 150 volts has been assumed in the circuit calculation.

Frame Time-base Generator

The frame-deflection coils require approximately 500 mA peak-to-peak for full deflection, and have a resistance of 12·2 ohms.

A suitable circuit for frame scanning, using an ECL80 triode-pentode, a silicon-iron-cored blocking oscillator transformer and a 25·5 : 1 output transformer, is shown in Fig. 7.

The circuit has ample reserve of scan, and draws a total current of 13 mA only from the 350-volt H.R.T. line.

Indications of the currents and potentials appearing in the circuit are also given in Fig. 7.

Line Time-base Generator

The line-deflection coils require 825 mA peak-to-peak to scan fully, and have an inductance of 3·24 mH.

A projection receiver employing the Mullard 25 kV E.H.T. unit normally has a 350-volt H.T. line available for the line time-base, and there is therefore little advantage in employing an energy-recovery system.

A suitable circuit employing the triode section of an ECL80 as blocking oscillator and a PL81 as output valve is shown in Fig. 8.

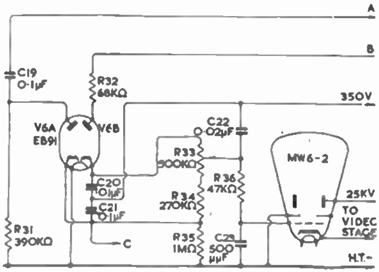


FIG. 9.—PROTECTIVE CIRCUIT FOR USE IN CONJUNCTION WITH A VIDEO OUTPUT STAGE USING AN EF80.

Protective Circuits

If either of the time-base units should become inoperative, the high-velocity electron would destroy the screen surface along a line which would be horizontal or vertical according to which time-base was out of action. Means must therefore be provided to cut off the beam in the event of the failure of either or both the time-base units. A circuit possessing all the necessary characteristics would be very complex, and it is therefore necessary to make some compromise. The circuit shown in Fig. 9 has the advantage of simplicity and also has the essential characteristics.

The voltage appearing at the anode of the frame-output valve (Point A in Fig. 9) is taken via a capacitor–resistor coupling to one diode anode (V6A) of the double diode V6, type EB91. A potential of approximately 90 volts positive to the chassis is obtained across resistor R35.

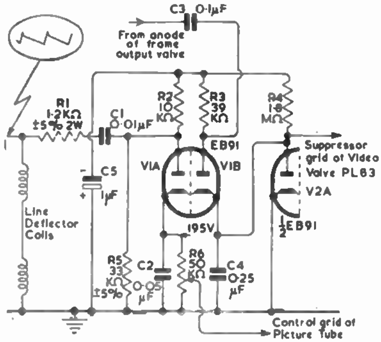
The end of the line-deflection coils at which the flyback pulse is positive-going (Point B in Fig. 9) is connected via a limiting resistor R32 to the second anode (V6B) of V6, and a steady potential of at least 150 volts is produced across resistors R33 and R34. R33 is the brightness control, and its slider is connected to the picture-tube grid via resistor R36. This sets the grid of the picture tube over a range of potentials suitable for a brightness control when the cathode of the picture tube is taken to the anode of the video valve as shown in Fig. 5.

If the frame circuit fails, the potential at all points on the chain R33 and R34 falls by 90 volts—sufficient to cut off the picture tube. Similarly, if the line circuit fails, the total voltage across R33, R34 and R35 falls by 150 volts, and the proportional change at the grid of the picture tube is again sufficient to cut off the tube.

This circuit is suitable for use in the video stage employing an EF80. It does not, however, provide adequate protection with a video stage employing a PL83. This is due to the lower mean anode potential of the PL83, and also to the greater video-drive voltage applied to the cathode-ray tube. For this reason the circuit shown in Fig. 10 has been devised.

In this circuit the end of the line-deflector coils at which the flyback pulse is positive-going is connected to the anode of the diode V1A via a limiting resistor R1 and coupling components C1 and R5. The positive, rectified output from the diode cathode is utilized by the

FIG. 10.—PROTECTIVE CIRCUIT FOR USE IN CONJUNCTION WITH A VIDEO OUTPUT STAGE USING A PL83.



brightness control R6, which supplies the grid potential of the picture tube. Thus, a failure of the line time-base biases the picture tube beyond cut off. The coupling components C1 and R5 are included to ensure that a waveform with a mean component negative to earth appears at the anode of the diode V1A. This mean potential is applied, via the smoothing components R2, R4 and C4, to the suppressor grid of the video valve. In the absence of any other applied potential, the anode current of the video valve is then cut off and the picture tube, having its cathode connected directly to the anode of the video valve, receives no signal and assumes the positive potential corresponding to that of the H.T. supply. The beam is therefore cut off.

H.T. Power Supply Unit

The requirements of the E.H.T. unit in large measure determine the design of the H.T. power-supply unit in a projection television receiver.

The E.H.T. unit must be supplied at 350 volts from a source having a resistance of between 250 and 550 ohms. It will draw from 25 to 55 mA, according to the picture content.

E.H.T. Unit

Production of a 25-kV supply direct from the normal 50 c/s mains is uneconomic by reason of the large physical dimensions and weight of such a unit and its high cost—particularly the cost of insulation. In addition, the unit would be inherently capable of a much greater output than the few hundred microamperes required for the cathode-ray tube, so that operation at supply frequencies would involve considerable danger to life.

The pulsed oscillator system has therefore been widely adopted, and a compact unit is achieved, for instance in the Mullard unit, by the use of a ferroxcube transformer core and miniature rectifier tubes. Fundamentally the unit comprises a blocking oscillator, a driver valve capable of withstanding high peak anode voltages and a cascade voltage tripler rectifier circuit.

In the circuit, Fig. 11, the first valve, an EBC33, is a double diode triode,

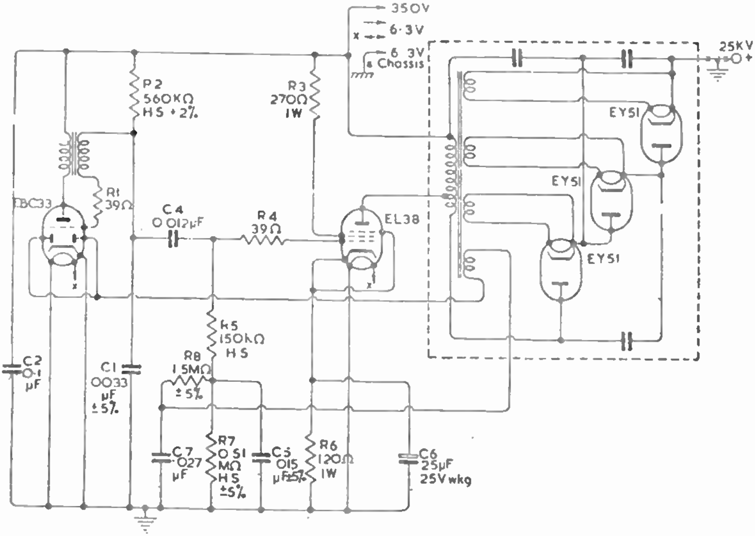


FIG. 11.—TYPICAL E.H.T. UNIT FOR PROJECTION TUBES.

the triode portion of which is connected as a blocking oscillator. The voltage on the capacitor of the R.C. network rises exponentially until the short-circuit pulse from the EBC33 discharges it rapidly, and the cycle repeats. The resulting voltage, which has an approximately saw-tooth waveform, is applied to the grid of the driver valve, an **EL38.**

This valve is biased well below cut-off, so that anode current flows during only a part of the saw-tooth input cycle. On cessation of anode current the energy stored in the inductive component of the anode circuit causes the circuit to "ring" at a frequency determined by the inductance of the anode circuit tuned by all the stray capacitance. In the equipment described this frequency is of the order of 25 kc/s.

The peak value of the oscillatory voltage can be derived from the energy equation.

$$\tfrac{1}{2}L_a i_0^2 = \tfrac{1}{2}C_p V_0^2 \qquad . \quad . \quad . \quad . \quad (1)$$

where L_a = inductance in anode circuit;
i_0 = instantaneous value of anode current at the moment of interruption;
C_p = total stray capacitance, including that of the valve;
V_0 = peak value of output voltage.

From the above equation :

$$V_0 = i_0 \sqrt{\frac{L_a}{C_p}} \qquad . \quad . \quad . \quad . \quad . \quad (2)$$

and substituting practical values ($i_0 = 100$ mA; $L_a = 0{\cdot}5$ H and $C_p = 50$ pF) the peak voltage is:

$$V_0 = 0{\cdot}1\sqrt{\frac{0{\cdot}5}{50 \times 10^{-12}}} = 10{,}000 \text{ volts}$$

The interruption frequency is given by the formula:

$$L_a \frac{\delta i_0}{\delta t} = \frac{L_a i_0}{aT} = V_a \quad \ldots \ldots \quad (3)$$

Whence

$$f_i = \frac{aV_a}{L_a i_0} \quad \ldots \ldots \quad (4)$$

where $f_i = \dfrac{1}{T}$ = interruption frequency;

a = anode current pulse ratio;

V_a = permissible anode voltage drop.

Substituting practical values ($a = 0{\cdot}2$; $L_a = 0{\cdot}5H$; $i_0 = 100$ mA and $V_a = 250$ volts), the interruption frequency is:

$$f_i = \frac{250 \times 0{\cdot}2}{0{\cdot}5 \times 0{\cdot}1} = 1{,}000 \text{ c/s}$$

The peaks of the damped oscillatory wave trains produced by the driver stage are available for rectification and smoothing to form the E.H.T. supply. The maximum voltage, as calculated from formula (2) above, is, however, somewhat limited, as it is impossible to reduce C_p to much below 50 pF, while i_0 is restricted by the valve used to about 120 mA, and there are practical limitations to the physical dimensions of the inductor L_a, particularly where a very compact power unit is desired.

The circuit is therefore designed to generate a voltage in the order of 8–9 kV, which is applied to a three-stage voltage tripler rectifying circuit of conventional design using three EY51 half-wave rectifiers and producing a rectified output at 25 kV.

It is essential that the E.H.T. power unit for a projection television system has good voltage regulation over the operating range of beam current. Normally, good regulation demands a supply unit of low internal resistance, but it can be shown that, in a unit of the type described, the internal resistance is inversely proportional to the input power, and since the output power required is small, low internal resistance could be achieved only at very considerable sacrifice of overall efficiency.

The problem of voltage regulation has, however, been solved in a simple manner by automatic regulation of the biasing voltage applied to the grid of the driver valve.

An additional winding is provided on the "ringing" transformer, and the voltage induced in this winding is rectified by the diode section of the EBC33 and is then fed via a filter network as grid bias to the driver valve.

As a result of this feature, the regulation of the E.H.T. unit is very good for outputs up to about 250 μA, after which it falls off rapidly.

Should, however, the control circuit fail for any reason, the E.H.T. voltage will be free to rise to over 30 kV, resulting in contraction of the picture and possible flashover.

ADJUSTMENT OF DOMESTIC PROJECTION UNITS

Projection Television Focusing

In a direct-viewing television receiver focusing is a matter of adjusting the current through the focusing coil or, in the case of permanent-magnet focusing, adjusting the permanent magnet on the neck of the tube. In projection television, however, in addition to accurate focusing of the picture on the face of the cathode-ray tube, it is necessary to adjust the position of the face of the tube with respect to the spherical mirror to ensure that the picture is accurately focused on the viewing screen. Focusing a projection television receiver should preferably be carried out while a test pattern is being transmitted. It can, however, be done when no signal is available, the image then consisting only of the scanning lines.

If the picture is hopelessly out of focus the normal electrical focus control must first be adjusted, but if it is found that a strip of picture is already in correct focus, only the mechanical adjustment described below need be made.

Note that these instructions, and Fig. 13, refer to a picture which emerges from the corrector lens with the scanning lines parallel to the axis of the tube. In some receivers the lines may be at some other angle. This will result in the strip of picture which is in focus being

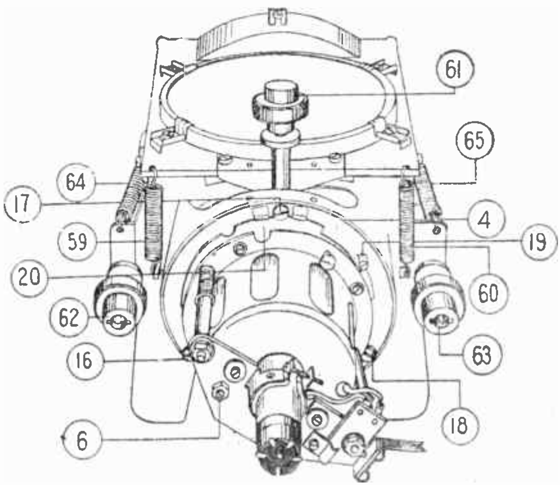
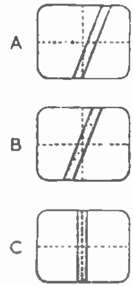


Fig. 12.—Adjustments for Mechanical Focusing.

FIG. 13.—SUCCESSIVE STAGES IN MECHANICAL FOCUSING

The shaded area represents the strip of picture in focus.



at a different angle from that shown in Fig. 13, but does not affect the method of adjustment.

(*a*) Remove the red locking plate from the front of the optical unit by withdrawing four screws, see Fig. 12.

(*b*) Slacken the knurled lock-nuts of screws 61, 62 and 63.

(*c*) Turn screw 61 to one extremity of its travel. This results in a narrow strip of picture only being in focus (see "A", Fig. 13) if the raster is correctly focused on the face of the cathode-ray tube. If a strip of picture is already in correct focus on the viewing screen, operation (*c*) may not be necessary.

(*d*) Turn screws 62 and 63 *simultaneously in the same direction* until the strip of picture which is in focus passes through the exact centre of the picture (see "B").

(*e*) Turn screws 62 and 63 in *opposite directions* until the strip of picture which is in focus is both central and vertical (see "C").

(*f*) Adjust screw 61 until the strip of picture which is in focus widens and ultimately covers the whole picture area, and adjust for best results.

(*g*) Tighten the knurled lock-nuts of screws 61, 62 and 63.

(*h*) *Do not* replace the red locking plate, as this may throw the focus out of adjustment. The locking plate with its four screws should, however, be preserved for use if the receiver has to be sent away for service.

Blocking Oscillator Frequency

The power delivered to the filaments of the three EY51 rectifiers in the E.H.T. can is governed by the repetition frequency of the blocking oscillator. In order to avoid over-running or under-running the filaments, therefore, the blocking oscillator frequency must be kept between the stated limits.

Adjustment of this frequency should be necessary only when a frequency-determining element such as the blocking oscillator transformer T1 is replaced.

Small variations from the specified value can be corrected by replacing C1 or R2 with components of different value. If, however, the frequency departs considerably from the correct value, the cause will most probably be a defective blocking oscillator transformer.

When replacing this transformer it should be noted that the winding with the lower resistance is connected in the grid circuit of V1 (EBC33).

The frequency of the blocking oscillator can be measured with a cathode-ray oscilloscope in conjunction with an accurately calibrated A.F. generator.

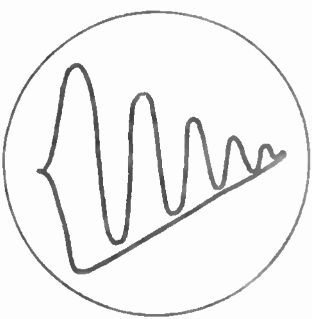


FIG. 14.—OSCILLOSCOPE TRACE WITH CORRECT ADJUSTMENT OF AUDIO-FREQUENCY GENERATOR.

The Y deflection terminal of the oscilloscope should be loosely coupled to the E.H.T. unit; usually a wire connected to the Y deflection terminal and brought close to the E.H.T. unit will pick up sufficient energy for a reasonable vertical deflection.

The horizontal deflection can be obtained from the A.F. generator. The frequency of the A.F. generator should be adjusted until a steady Lissajou's figure similar to that represented in Fig. 14 is obtained. This should occur at a frequency of between 930 and 1100 c/s. If the A.F. oscillator frequency is outside these limits, C1 or R2 must be replaced with components of suitable value.

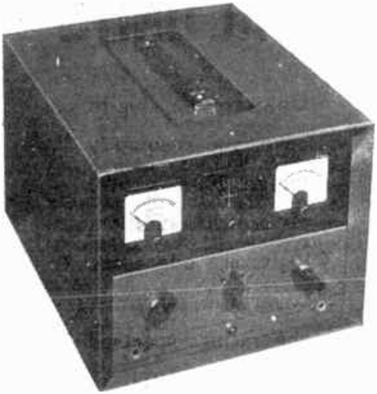
THEATRE TELEVISION SYSTEMS

There are two basically different ways of providing the light which is spread over the big television screen. One is to take the intense light pattern from the fluorescent screen of a projection cathode-ray tube and spread it thinly over the greater area of the cinema screen by static optical devices, lenses and mirrors, and in particular by the economical and powerful devices of Schmidt optics. The other is to use a "light-valve" which modifies or modulates the light from a powerful arc projector of conventional type (save again for the high intensity necessitated by the low efficiency of most light-valves) on its way from arc to screen. The light-valves in turn may be roughly subdivided into two different classes. The first uses as a permanent and reproducible light-valve the familiar cinema film; this gives us the "Intermediate Film" process of theatre television. The other class contains light-valves in which the modulation is as transient as that from the original scanner; it comprises devices such as the Kerr cell, depending on modulation by polarization, the "Eidophor", depending on modulation by the continuously varying diffraction pattern from a continuously deformed liquid surface, and—with close affinities with the photographic image of the intermediate film—the Skiatron or dark-trace cathode-ray tube, in which a quasi-permanent picture is produced by electron bombardment of a potassium halide screen, and after use effectively as a single-frame transparency, is wiped out by infra-red irradiation to clear the screen for the next frame image.

Acknowledgment

The sections on "The Schmidt Optical Systems", "Cathode-ray Tubes for Projection Systems", and "Circuitry", have been compiled largely from information supplied by Messrs. Mullard Ltd., on their projection television system.

E. M.



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Input Impedance	1 Megohm
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Effective limiter range	± 10 dB
Meter scaling—" Peak wow "	0 to ± 1% (centre zero
Meter scaling—" Wow and Flutter "	0 to 1% R.M.S. and 0 to 2%
Crossover frequency	20 c.p.s.
Flutter meter response	−3 dB at crossover −3 dB at 200 c.p.s.
" Wow " meter response ...	1 dB at 5 c.p.s.
C.R.O. output frequency response	level down to zero frequency −3 dB at 200 c.p.s.
3000 c.p.s. oscillator output level	5V R.M.S. into ·5 Megohm 100 mV. R.M.S. into 500 ohms
Accuracy	Meter presentations ± 2% f.s.d.
Power consumption	35 watts
Operation..................	45 to 60 c.p.s.

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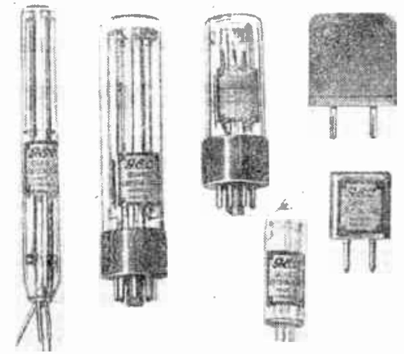

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OC 328	B	100 Kc/s 800 Kc/s 10 Mc/s	500 Kc/s 9999·9 Kc/s 20 Mc/s	2 15 10
OC 287 O.C.327	C AND D	200 Kc/s 800 Kc/s 10 Mc/s	500 Kc/s 9999·9 Kc/s 20 Mc/s	2 10 5
OC.193	E	120 Kc/s 800 Kc/s 10 Mc/s	500 Kc/s 9999·9 Kc/s 20 Mc/s	0·4 10 5
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QUARTZ CRYSTALS

Quartz crystals are widely used in modern telecommunications practice as frequency stabilizing elements and as resonant elements in filter circuits. They are relatively simple and robust in construction and extremely stable in their properties. Quartz, itself, is chemically inert, has extremely low electrical and mechanical losses, and because of its piezo-electric properties is readily applied in high-frequency electrical circuits as the equivalent of a stable, high Q, tuned circuit, having a stability and Q factor much higher than could be achieved by conventional means.

Quartz, SiO_2, the most widely occurring crystalline form of silica, crystallizes in the trigonal (hexagonal) class, and well-formed crystals have a striking symmetry, the form of the crystal being a hexagonal prism terminated by hexagonal pyramids. Perfectly shaped crystals would have the form shown in Fig. 1, but natural crystals rarely develop all these faces.

Quartz occurs in the two forms known as left-hand and right-hand. The two kinds have identical properties and physical constants, except that each is the mirror image of the other. They are distinguished in the natural form by the characteristic small facets marked x and s in Fig. 1, where it will be seen that the faces m, x, s, z progress in the order of a right-hand screw about the principle, Z, axis for right-hand quartz and vice-versa for the left-hand form.

Quartz rotates the plane of polarization of a beam of polarized light traversing the crystal in the direction of the Z axis, and the sense of this rotation is opposite in the two forms.

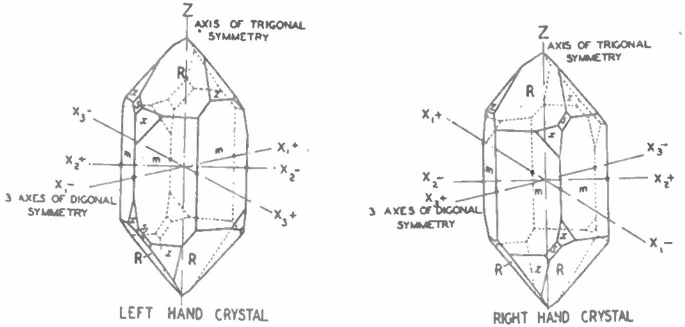


FIG. 1.—THE FORM OF QUARTZ CRYSTALS.

Quartz has four axes of crystalline symmetry, the Z (or principal) axis around which it exhibits trigonal (threefold) symmetry, and three X axes, equally spaced at 120° to each other in a plane perpendicular to the Z axis. Each X axis is an axis of diad or twofold symmetry, and all three are, of course, identical and indistinguishable from each other.

As in all crystalline materials, the mechanical, electrical and optical properties vary with direction, the variations being in accordance with the symmetry. There are therefore small but significant changes in the elastic constants (Young's modulus and shear modulus) and in the thermal-expansion coefficient, with change in the direction of measurement. These variations have important practical significance, in that they permit crystal plates to be cut having low- or zero-temperature coefficient of frequency.

Piezo-electric Properties

Quartz is of value for telecommunications purposes, mainly on account of its piezo-electric properties, that is, its ability to develop electrical charges on its faces when subjected to mechanical stress, and conversely to change its dimensions slightly when electrically stressed.

Consider a small rectangular block of quartz cut with its major faces perpendicular to an X axis, and its other faces perpendicular to, and parallel to, the Z-axis respectively. For convenience the third axis to complete a set of rectangular co-ordinates will be designated Y. Each pair of faces can then be identified by the symbol corresponding to the axis normal to the faces.

If pressure is applied between the x faces, charges are developed on those faces and the charge developed is proportional to the applied force and independent of the dimensions of the plate. Tensional forces also result in charges, the sign of the charge being reversed for such forces.

Quartz also develops charges on its x faces when compressional or tensional forces are applied to the y faces, and the charge density for a given stress is the same in magnitude as for the same stress on the x faces, but the polarity is of opposite sign.

Tensional or compressional forces in the direction of the Z-axis do not result in the appearance of charges on any faces. It is usual to express these results in terms of charge density, P, and stress S; hence we obtain the following equation :

$$P_x = d_{11}S_x - d_{11}S_y \qquad (1)$$

where P_x is the charge density on the x faces, S_x is the stress on the x faces, S_y the stress on the y faces and d_{11} is the constant known as the piezo-electric constant. In the m.k.s. system, d_{11} has a value of about $2{\cdot}15 \times 10^{-12}$ coulombs per m.k.s. unit of force.

If the quartz is deformed in shear about the X-axis, i.e., in the yz plane, a charge is developed on the x faces. In terms of charge density and shear stress this is given quantitatively by :

$$P_x = d_{14}S_{yz} \qquad (2)$$

where S_{yz} is the magnitude of the shear stress and d_{14} is the piezo-electric shear constant. In the m.k.s. system, d_{14} has a value of $-0{\cdot}5 \times 10^{-12}$ coulombs per unit of force. Shear forces about the Y

and Z axes do not produce charges on the x faces; they do, however, result in charges on the y faces. For shear about the Y axis, i.e., in the zx plane, the relationship is:

$$P_y = -\,d_{14}S_{zx} \qquad . \quad . \quad . \quad (3)$$

The constant in this equation (3) is the same numerically as in the equation (2). In the case of shear force about the Z axis, i.e., in the xy plane, the equation is:

$$P_y = -2d_{11}S_{xy} \quad . \quad . \quad . \quad . \quad (4)$$

where d_{11} has the numerical value quoted earlier (see equation (1)). The plate cannot be stressed in any manner which results in charges on the Z faces.

Thus the full piezo-electric equations for the direct effect can be written in terms of the applied stresses as:

$$P_x = d_{11}S_x - d_{11}S_y + d_{14}S_{yz} \quad . \quad . \quad . \quad (5)$$

$$P_y = \qquad -d_{14}S_{zx} - d_{11}S_{xy} \quad . \quad . \quad (6)$$

A similar set of equations can be derived written in terms of the strains s, set up by the applied forces, viz.,

$$P_x = e_{11}s_x - e_{11}s_y + e_{14}s_{yz} \quad . \quad . \quad . \quad (7)$$

$$P_y = \qquad -\,e_{14}s_{zx} - 2e_{11}s_{yx} \quad . \quad . \quad (8)$$

e_{11} and e_{14} are known as the piezo-electric moduli and in the m.k.s. system have values:

$$e_{11} = +0{\cdot}16 \text{ coulombs/m.}^2$$
$$e_{14} = +0{\cdot}045 \text{ coulombs/m.}^2$$

When a voltage is applied to the opposite faces of an unstressed plate, there is a corresponding change in dimensions. This phenomenon is known as the converse piezo-electric effect, and the equations, corresponding to equations (7) and (8), and relating the resultant strain, s, to the applied electric stress are:

$$\left.\begin{aligned} s_x &= d_{11}E_x \\ s_y &= -d_{11}E_x \\ s_{yz} &= d_{14}E_x \\ s_{zx} &= -d_{14}E_y \\ s_{xy} &= -2d_{11}E_y \end{aligned}\right\} \qquad (9)$$

where Ex, Ey are the electric stresses in volts/metre along the X and Y directions respectively. The piezo-electric constants, d_{11} and d_{14} have the numerical values quoted earlier.

If the plate is simultaneously under mechanical and electrical stress, the equations become much more complicated. For a full treatment of the piezo-electric effect in quartz see references given at the end of this article.

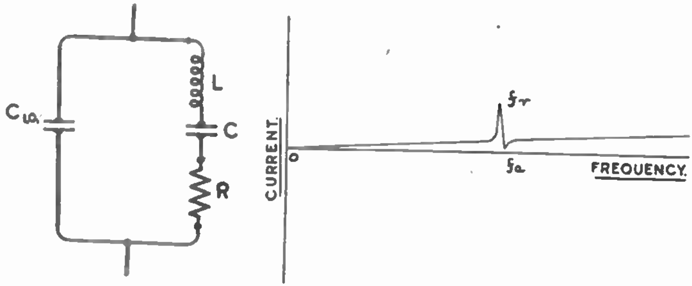


FIG. 2.—THE EQUIVALENT CIRCUIT OF A QUARTZ CRYSTAL.

Resonant Frequencies

We have, so far, dealt only with the static properties of the plate. Since there is hardly any time lag between the application of the voltage and the change in dimensions, it follows that the dimensions of a quartz plate will vary cyclicallv in step with an applied alternating voltage. Suppose that a low-frequency voltage is applied to a plate and that the frequency is slowly increased. If we were to measure the current flowing into the plate and then plot current against frequency, a curve such as Fig. 2 would result. A small current would flow initially and would increase proportionally to the frequency in a similar manner to the current flowing into a small capacitance. As we approach closely to the frequency f_r the current rises sharply, reaching a maximum at f_r, and then falls sharply to a minimum at a slightly higher frequency, f_a. With further increase in frequency, it then increases slowly, resuming its linear rate of increase with frequency.

The frequency corresponding to maximum current is the resonant frequency (f_r), and that at the minimum current is known as the anti-resonant frequency (f_a).

The behaviour of the plate at its resonant frequency is directly comparable to that of a series-tuned electrical circuit. The plate resonates mechanically, and the mechanical resonance is reflected in its electrical behaviour. At the resonant frequency f_r the amplitude of the mechanical oscillations of the plate is some thousands of times greater than the amplitude at a slightly lower frequency on the linear part of the curve, and consequently the magnitude of the current flowing into the plate is similarly increased. The behaviour is identical to that of the equivalent series tuned circuit L–C–R, provided we assign suitable values to these elements. A significant difference between the behaviour of a quart plate and a conventional tuned circuit lies, however, in the sharpness and extent of the peak. Both these are dependent on the Q value, which for a quartz plate usually lies between some 10,000 and 1,000,000, compared with about 300 for a very high-grade inductance and capacitor.

The sharp dip at the anti-resonant frequency, f_a, is the result of the series arm exhibiting inductive reactance at frequencies above the resonant frequency. At the frequency f_a, the value of this inductive reactance is equal in magnitude to the capacitive reactance of the direct

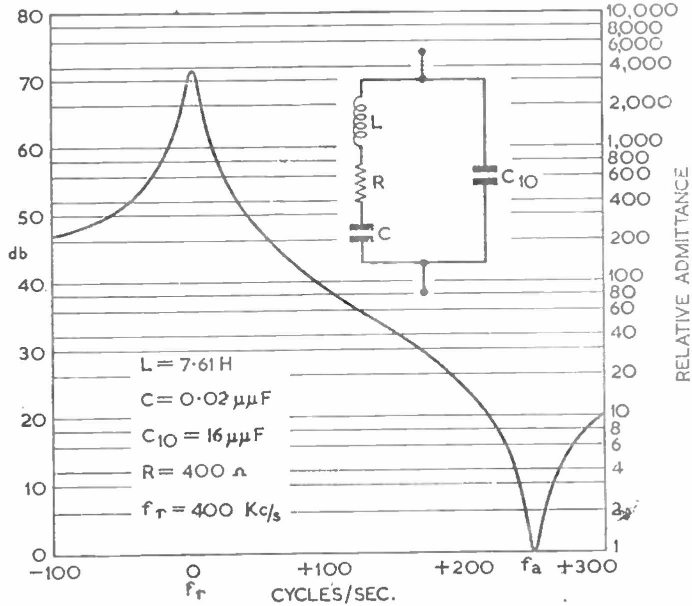


Fig. 3.—Resonance Curve for a Quartz Crystal.

capacitance between the electrodes represented in Fig. 2 by the capacitor C_{10}. Thus at this frequency the crystal with its electrodes behaves as a parallel resonant circuit which is similarly characterized by a very high Q value. At frequencies higher than f_a the plate behaves again substantially like a small capacitor.

Fig. 3 is an accurate plot of a typical resonance in a quartz plate vibrating in a longitudinal thickness mode. A logarithmic vertical scale is necessary to show the wide range of admittance encountered. For this specimen with a Q value of about 50,000 there is a ratio of some 4,000 : 1 between minimum and maximum, and the change occurs over a frequency range of only some 250 c/s in 400,000, i.e., band-width only 0·06 per cent of mean frequency.

Modes of Vibration

The frequency at which a plate or bar resonates depends on the mode of vibration employed and the plate dimensions. For most purposes fundamental modes are employed, but for special requirements—and especially for the highest frequencies—overtone modes have to be used, otherwise the plate would be too thin or too small to permit manufacture.

We have, so far, considered the plate as exhibiting only a single

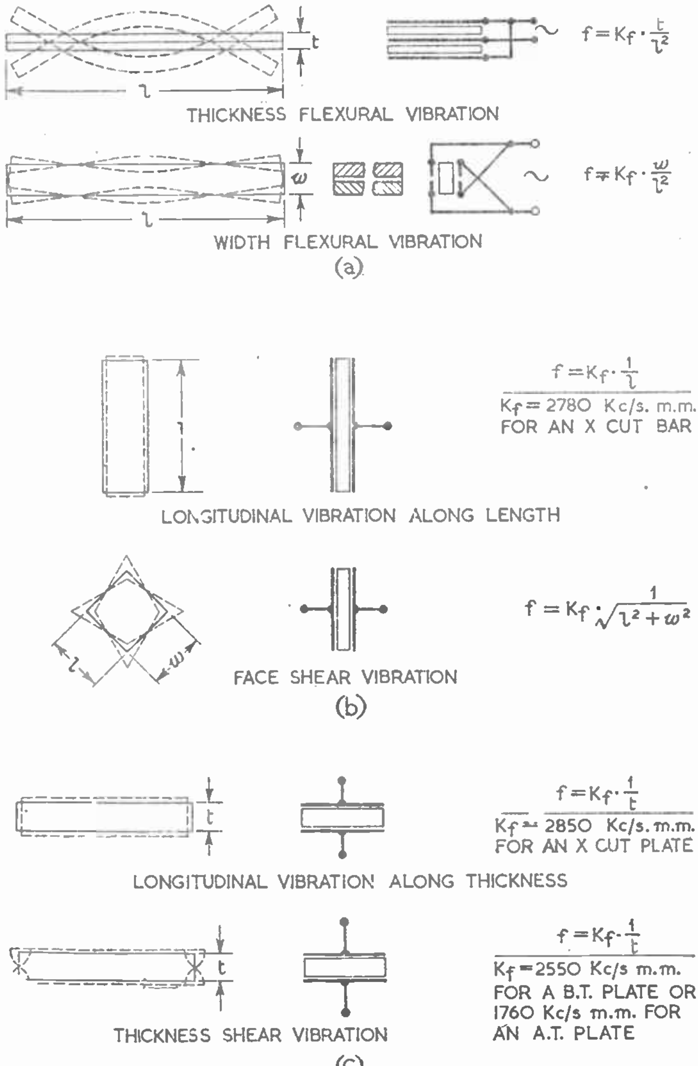


Fig. 4—Modes of Vibration of Quartz Crystals.

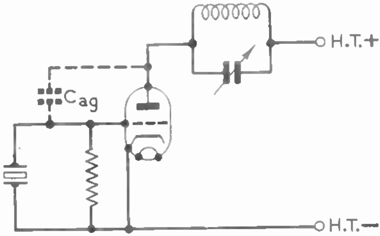


FIG. 5.—MILLER–PIERCE CIRCUIT.

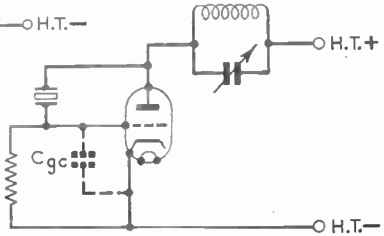


FIG. 6.—PIERCE CIRCUIT.

resonance. In practice, any plate or bar has an infinite variety of modes in which it can resonate, and its behaviour can be extremely complex. Fortunately many of the modes can be inhibited by the choice of the electrode system and the manner of mounting, or by suitable choice of plate dimensions. In making and designing crystals the aim is to produce an element which has a single predominant response over a more or less wide-frequency band according to the application. The wider the band required to be free from spurious responses, the more difficult it is to achieve a satisfactory design.

The principal types of quartz resonator and the manner in which the plate or bar vibrates are indicated in Fig. 4. With each diagram is the equation showing the principal resonance as a function of dimensions. The shear and longitudinal modes can be easily excited simply by placing electrodes on the major faces. The flexurals require different electrode systems, as indicated in the figure, such that one half of the plate is caused to contract whilst the other expands.

CRYSTAL OSCILLATOR CIRCUITS

We have already seen that a quartz plate has two frequencies of zero reactance associated with each mode of vibration, viz., the series resonant frequency and the parallel or anti-resonant frequency, and oscillator circuits can be broadly divided into two classes according to which resonance is used. The anti-resonant circuits have been the most widely employed hitherto, the most popular arrangements being as shown in Fig. 5 and Fig. 6. There is a close similarity between these circuits. In both cases the operation of the circuit depends on the stray inter-electrode capacitance represented by the capacitor shown dotted.

Replacing the crystal in Fig. 5 by its equivalent circuit, L–C–R, C_{10}, and neglecting D.C. power supplies, the circuit is as shown in Fig. 7 where C_{ag} is the grid–anode capacitance of the valve, and the anode–cathode and grid–cathode capacitances are included in C_1 and C_{10} respectively. The grid circuit is shown broken, so that we may consider the circuit as an amplifier, having a signal input of e_g and an output of E_a at the anode. The voltage E_g, derived from the potential divider formed by C_{ag} and the crystal, is available at the crystal terminals. Obviously, if e_g were exactly equal to E_g the two terminals could be joined without effect on the circuit; thus we can have stable operation as an oscillator when, and only when, e_g is identically equal to E_g. This identity can be resolved into two conditions, equality of amplitude and equality of phase. The equality of amplitude is normally taken care of by the amplitude-limiting effect in the valve itself. For zero overall phase change from e_g to E_g, any phase difference between e_g and E_a must be equal to, and in opposite sense to, that between E_a and E_g. To illustrate the operation of the circuit we have therefore calculated these phase changes for typical circuit values and plotted the results in Fig. 8.

The curve marked α gives the phase change of E_g relative to E_a, whilst the group of curves marked β give the relative phase of E_a to e_g. Each of these β curves corresponds to a particular value of C_1, the figure at the right of each curve indicating how much the circuit is off tune at the crystal frequency. In general, both these phase differences vary with frequency. The frequency of oscillation, the frequency at which these two phase changes are equal and opposite, is defined by the points of intersection of the α with the β curves.

Over the small frequency range for which the curves have been prepared, the rate of change of phase of E_a relative to e_g is negligible; that is, each β curve shows an almost constant phase change except for the small kinks near the right-hand intersection which are due to the shunting effect of the crystal and C_{ag} in series, on the tuned anode circuit. The most striking feature of the α curve is the high rate of change of phase in the region of f_r and f_a, a direct consequence of the

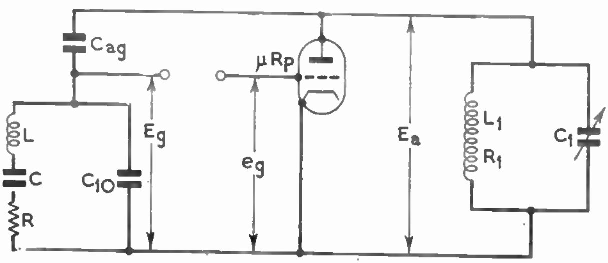


FIG. 7.—THE EQUIVALENT CIRCUIT OF THE MILLER–PIERCE OSCILLATOR.

$$\frac{E_g}{E_a} = A\angle\alpha \quad \frac{e_g}{E_a} = B\angle\beta$$

Conditions for oscillation:

(a) *Amplitude* . . . $\frac{A}{B} \leqslant 1$

(b) *Frequency* . . . $\alpha = \beta$

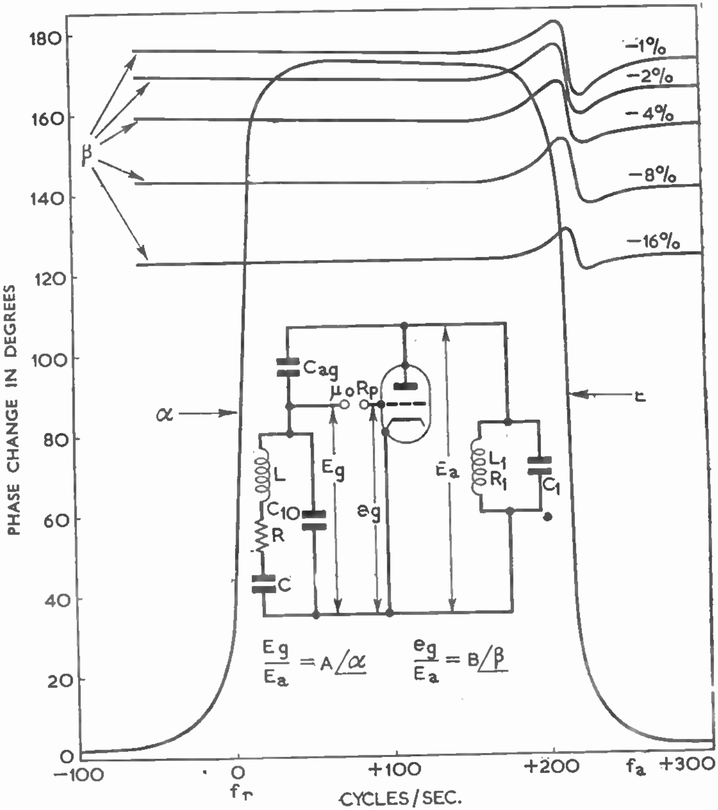


FIG. 8.—PHASE RELATIONS IN THE MILLER–PIERCE CIRCUIT.

high Q value of the quartz resonator. Consideration of the amplitude condition excludes all the intersections with the left-hand half of the α curve, leaving the right-hand intersections as the points defining the frequencies of oscillation corresponding to each setting of C_1.

A very small change in frequency, therefore, suffices to correct for any change in the phase angle β which results from a change in the valve characteristics due either to ageing or to changes in operating potentials, or any similar cause. Even changes in the tuning capacitance have only a minor effect on frequency, especially when the anode circuit is well off tune. This is illustrated again in Fig. 9 showing measured results for a typical 3-Mc/s crystal in a Miller–Pierce circuit.

In a circuit of this kind it can be readily shown that the frequency

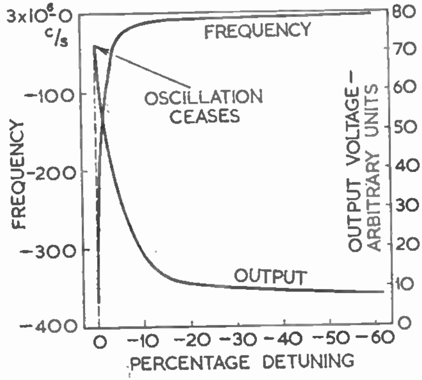


FIG. 9.—FREQUENCY CHANGE WITH TUNING IN THE MILLER-PIERCE CIRCUIT.

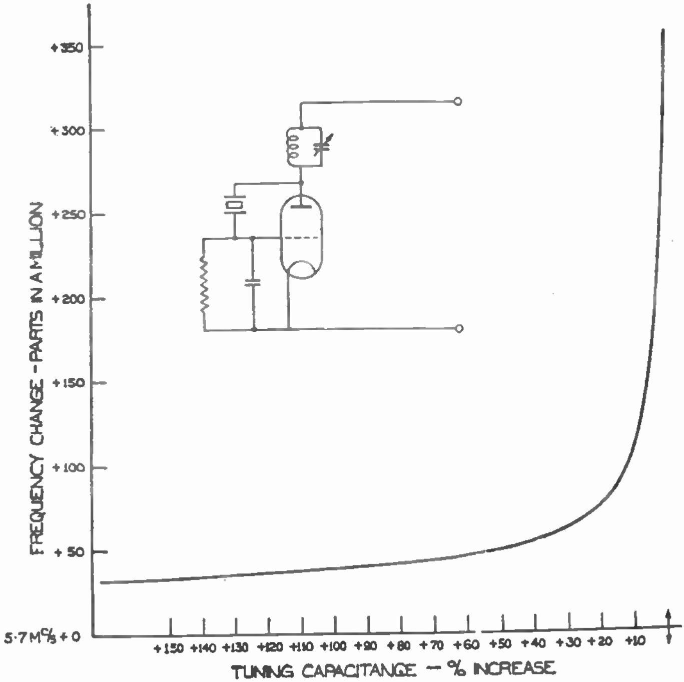


FIG. 10.—FREQUENCY CHANGE WITH TUNING IN THE PIERCE CIRCUIT.

corresponding to 90° phase change, point E in Fig. 8, is related to the resonant frequency f_r of the crystal by the equation :

$$f'_a = f_r\left[1 + \frac{1}{2}\frac{C}{C_{10} + C_{ag}}\right] \quad . \quad . \quad . \quad (10)$$

f'_a, and hence the oscillation frequency, is therefore directly dependent on C_{10}, which includes the grid–cathode capacitance of the valve, and C_{ag}, the grid–anode capacitance. The stability of these capacitances is therefore of prime importance. Where frequency stability must be improved it will be desirable to supplement one or both of these stray capacitances by adding additional stable fixed capacitances in parallel. A similar analysis can be made for the Pierce circuit, resulting in the same general conclusions. The frequency f'_a is again given by equation (10), but the change of frequency with tuning is in the opposite sense, the frequency increasing slightly as the anode circuit is brought into resonance at the crystal frequency (see Fig. 10).

Colpitts-derived Oscillator

A circuit which is also very much used is derived from the Colpitts oscillator circuit, and is shown in Fig. 11. Essentially, the crystal is acting as an inductance, having the combination of C_{ag} and C_{10} in parallel with its terminals and also shunted by the capacitances C_{ac} and C_{gc}, which are effectively in series across its terminals. The effective capacitance across the crystal is $\left(C_{ag} + C_{10} + \frac{C_{ac} + C_{gc}}{C_{ag}C_{gc}}\right) = C_e$ say, and the frequency of oscillation is given by :

$$f'_a = f_r\left(1 + \frac{1}{2}\frac{C}{C_e}\right) \quad . \quad . \quad . \quad (11)$$

This circuit is especially useful for the higher frequencies, and where several crystals have to be switched to give alternative frequencies, since no adjustment or tuning is required.

Frequency-multiplication

In practice, a designer is seldom content to use a separate valve for the crystal-oscillator stage; he usually prefers to frequency multiply in the same stage, or use the valve as a mixer or modulator as well. With tetrodes or pentodes this can usually be done without serious effect on the frequency stability. Many circuit variations are possible, and typical arrangements are illustrated in Fig. 12.

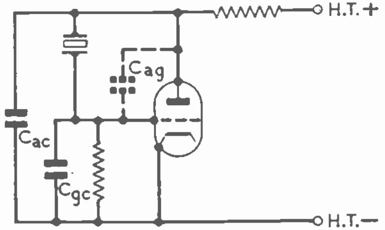
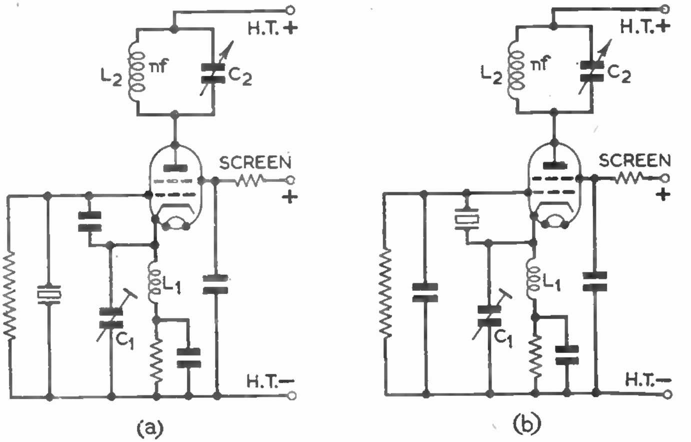


FIG. 11.—CRYSTAL COLPITTS CIRCUIT.



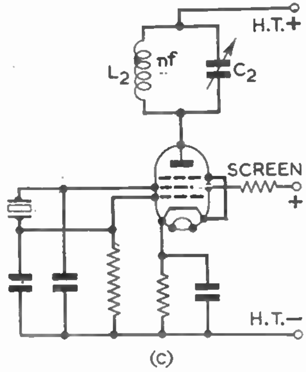


FIG. 12.—VARIATIONS OF THE PIERCE, MILLER-PIERCE AND COLPITTS OSCILLATOR CIRCUITS PERMITTING FREQUENCY MULTIPLICATION IN THE SAME STAGE.

Series-mode Oscillators

Whilst crystals used at parallel resonance give results which are adequate for most purposes, it has been shown that the frequency of oscillation is dependent directly on the circuit capacitances. These capacitances may not be very stable in themselves, and therefore such circuits do not permit the highest degree of frequency stability which the crystals themselves can provide.

When the crystal is operated at its series resonance, it exhibits a relatively low impedance, and the effect of the capacitance between its terminals is negligible, except at the very highest frequencies.

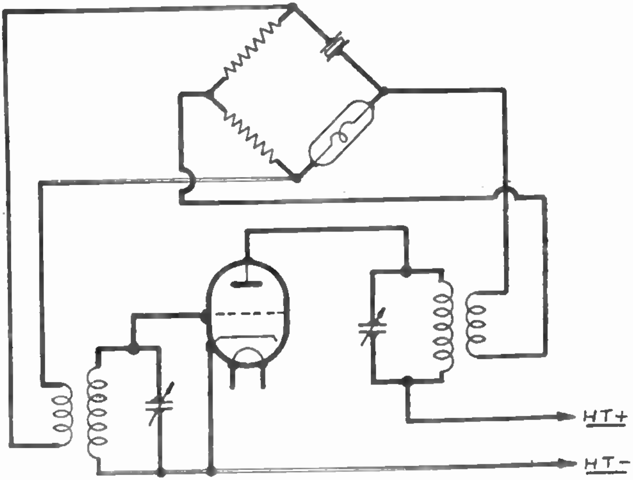


FIG. 13.—MEACHAM BRIDGE CRYSTAL OSCILLATOR.

With suitable circuit arrangements it can therefore provide a stability of frequency at least one order better than when operated in the parallel mode.

The most widely known circuit for series mode operation is the Meacham bridge circuit (Fig. 13). In this circuit, which is generally employed in modern frequency standards, the crystal forms one arm of a resistance bridge which constitutes the feedback network between the output and input of a linear amplifier. Another arm of the bridge is in the form of a load-dependent resistor such as a tungsten lamp or a thermistor.

The bridge is designed so that an increase in the voltage applied to the bridge results in a reduction in the input signal available at the other diagonal of the bridge, hence the amplitude is stabilized independently

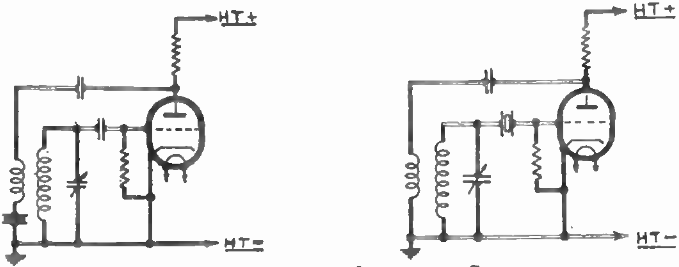


FIG. 14.—SERIES-MODE OSCILLATOR CIRCUITS.

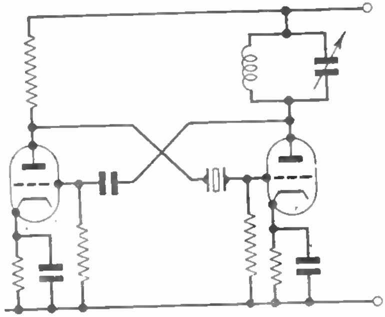


FIG. 15.—TWO VALVE SERIES-MODE CRYSTAL OSCILLATOR CIRCUIT.

of the valve amplifier, which can then operate strictly in the linear mode. Provided that the phase change in the amplifier is constant, the crystal alone determines the frequency, since there are no other reactive elements in the bridge. Using this circuit, long-term stabilities better than one part in a hundred millions have been obtained.

The arrangement shown in Fig. 14 (*a*) is also suitable for series mode operation, especially at the lower frequencies. The grid circuit is tuned to the crystal frequency, and feedback through the coupling coil occurs only at the crystal series resonance; Fig. 14 (*b*) shows an alternative form of circuit which has also been successfully applied for series mode operation.

Two other circuits, Fig. 15 and Fig. 16, deserve mention. The circuit of Fig. 15 has been frequently used in the past, especially with small high-impedance crystals where a two-valve oscillator was necessary to provide sufficient gain. The circuit of Fig. 16, frequently known as the Butler circuit, uses the crystal as a coupling element between the valve cathodes. In this arrangement the first valve is a grounded-grid amplifier and the second valve is essentially a cathode-follower stage matching the anode circuit to the low input impedance of the amplifier. This arrangement is specially suitable for low-

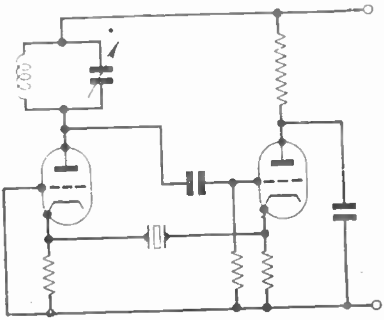


FIG. 16.—BUTLER OSCILLATOR CIRCUIT.

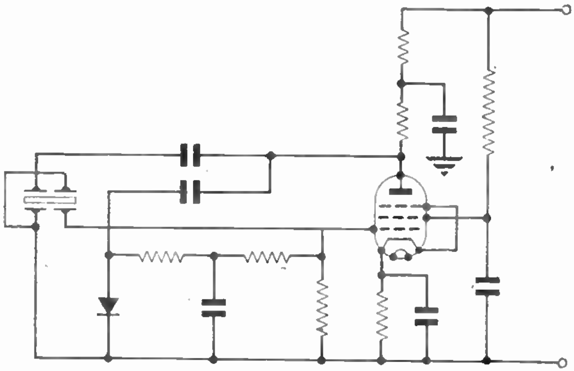


FIG. 17.—FORM OF OSCILLATOR USED FOR FLEXURAL CRYSTALS.

impedance crystals, and is useful for driving high-frequency and overtone crystals in the series mode.

The flexural types of crystal normally have three or four electrodes. They are most frequently used in circuits of the type shown in Fig. 17, where they behave like a tuned transformer in coupling the anode circuit back to the grid. These crystals are usually extremely active, and unless some form of amplitude limiting is applied it is easy to fracture the plates by over driving, hence the automatic-gain-control circuit designed to provide additional negative bias to the valve grid.

Overtone Oscillators

The limit to fundamental mode operation at high frequencies is set by the practical difficulties in manufacturing and handling very thin plates. In the present state of the art this limit is at about 20 Mc/s. For the higher frequencies, which are of increasing importance in modern practice, overtone modes have to be utilized. In theory, all thickness shear mode plates have resonances at the odd multiples of the fundamental frequency.

The effective activity decreases rapidly, however, as the order of the overtone mode increases, so that for practical purposes only the third, fifth and possibly the seventh overtone modes can be utilized. By these means, oscillation frequencies up to 100 Mc/s or higher can be directly stabilized by means of a quartz plate.

Whilst third overtone plates having frequencies up to about 35 Mc/s have been used for many years, there are many problems still to be solved before the higher-frequency plates come into general use. Circuit design for these high frequencies is also currently under investigation. The Butler circuit, Fig. 16, has been successfully used at frequencies up to at least 50 Mc/s, but, since it is capable of oscillation at frequencies other than the crystal frequency, as a result of the coupling through the residual capacitance of the crystal, its use is not recommended except under laboratory conditions, where spurious oscillation is easily checked. At frequencies of the order of 50 Mc/s, for example,

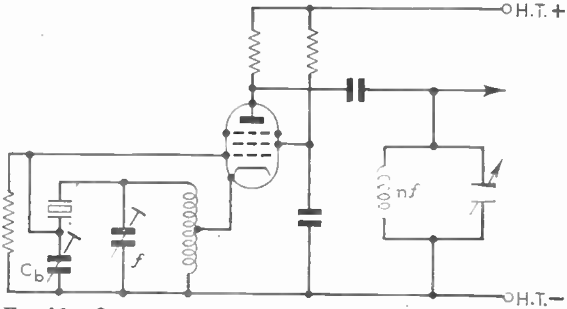


FIG. 18.—OSCILLATOR FOR USE AT OVERTONE FREQUENCIES.

the inter-electrode capacitance, together with wiring and other stray capacitances, can easily amount to 10 pF, giving a reactance of the order of some 300 ohms. An overtone crystal, which might have an equivalent series resistance of some 30 or 40 ohms at series resonance, cannot therefore show the same variations in terminal impedance as the fundamental type does at the lower frequencies.

The effects of the inter-electrode capacitance can be overcome, to a degree, by the use of circuits in which the crystal is used as one arm of a reactance bridge. The reactance elements are arranged to be in balance at the operating frequency, and the bridge therefore transmits a signal only at the series resonance of the crystal. A circuit of this type is illustrated in Fig. 18. In some cases the balance of the reactive elements is achieved by utilizing natural stray capacitances arising in the circuit, such as valve inter-electrode capacitances and wiring strays.

The Pierce circuit of Fig. 6 can also be utilized for overtone operation up to about 50 Mc/s, but may require careful adjustment. When the tuned anode circuit is tuned to a frequency higher than the crystal frequency, oscillation is not possible; for oscillation to be possible the anode circuit must be capacitive at the operating frequency. Hence, by tuning to a frequency a little below the required overtone mode, we have the right conditions for inhibiting oscillation at the fundamental or lower-order overtones and facilitating oscillation at the desired mode. With crystals of good activity oscillation can frequently be obtained up to the seventh order, and up to about 60 Mc/s. This arrangement is, however, operating the crystal at its parallel overtone mode, and there are fundamental limitations to this mode of operation at the higher-order overtones.

Crystal Activity Measurement

Apart from the obvious importance of the frequency of a quartz crystal, the equipment designer, or user, is primarily interested in its activity, for this determines the level of oscillation in his circuit, and indeed, may restrict the choice of circuit in some cases. The word "activity" has been used for a long time in a qualitative sense to describe the relative ease with which a crystal could be made to oscillate. It has, however, acquired a more precise meaning, and is now generally

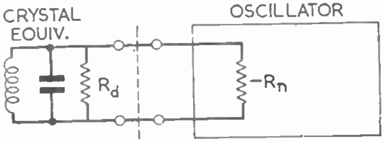


FIG. 19.—DIAGRAMMATIC REPRESENTATION OF A PARALLEL MODE CRYSTAL OSCILLATOR.

understood to mean the measured value of the factor which would determine the amplitude of oscillation if no other circuit losses were present. For a crystal used at series mode this is simply the value of the resistance R in the equivalent circuit. This is generally written

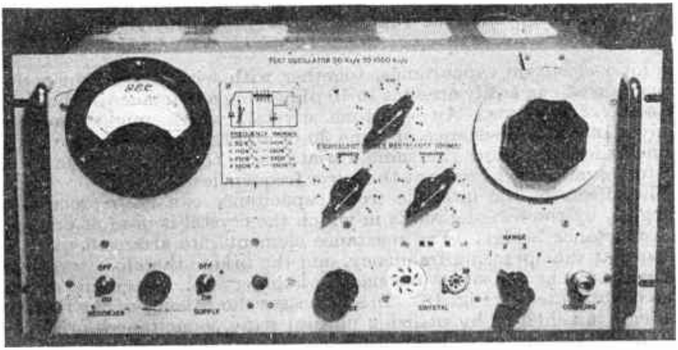


FIG. 20.—QUARTZ CRYSTAL ACTIVITY TEST SET FOR DIRECT MEASUREMENT OF THE EQUIVALENT SERIES RESISTANCE.

(Courtesy of the General Electr.c Company Ltd. of England)

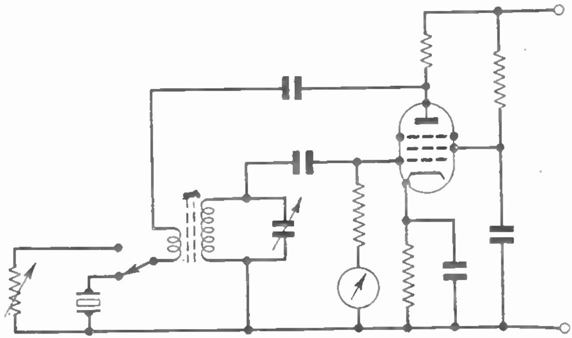


FIG. 21.—SIMPLIFIED CIRCUIT OF AN EQUIVALENT SERIES RESISTANCE MEASUREMENT SET.

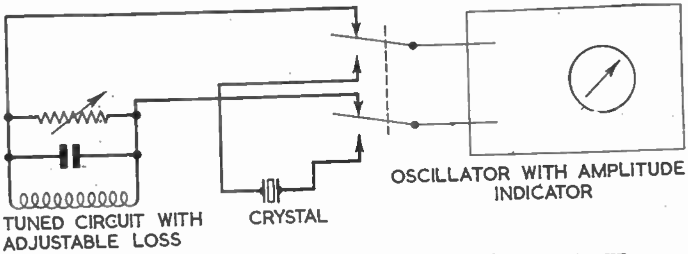


FIG. 22.—DIAGRAMMATICAL ARRANGEMENT FOR MEASURING THE EQUIVALENT PARALLEL RESISTANCE OF A CRYSTAL.

as e.s.r., an abbreviation for equivalent series resistance, and is measured in ohms. For parallel-mode operation, the crystal, together with any shunt reactances, including its inter-electrode capacitance and any other stray capacitances, can be shown to be equivalent at the operating frequency, to an ideal parallel-tuned circuit with a paralleled resistor, R_d. Under these conditions the maintaining oscillator is equivalent to a further negative resistor, $-R_n$ (see Fig. 19). When $R_d = R_n$ in magnitude there is no resultant loss, permitting oscillations to occur in the tuned circuit at constant amplitude. The value of R_d associated with the tuned circuit in this arrangement is therefore a measure of the value of $-R_n$ required to be provided by the maintaining oscillator. In other words, R_d is a measure of the crystal activity. A close analogy can be shown with a conventional tuned circuit comprising a coil and capacitor, and in fact the value of R_d, which would be the dynamic resistance of the tuned circuit, is given by :

$$R_d = \frac{1}{\omega^2 C_t R} \qquad . \quad . \quad . \quad . \quad (12)$$

where ω = operating frequency in radians/second;

C_t = total effective capacitance in parallel with the crystal;

R = crystal series resistance.

It will be seen that it is dependent on both the equivalent series resistance, R (e.s.r.) of the plate and on the value of the total capacitance in parallel with the crystal terminals. This measure of the activity is usually known as the equivalent parallel resistance (e.p.r.) of the plate. It is also referred to sometimes as " Performance Index ".

Values of e.s.r. can be measured by conventional means, such as radio-frequency bridges, provided that the generator supplying the bridge has adequate frequency stability and can be adjusted to precisely the resonant frequency of the plate. It is, however, more usual to drive the plate at its series mode in a specially designed oscillator and then determine its e.s.r. by substituting a resistance and determining the value of the resistor which gives the same amplitude of oscillation as when the plate is in circuit. Typical equipment is shown in Fig. 20 and a simplified circuit in Fig. 21.

FIG. 23.—QUARTZ CRYSTAL ACTIVITY TEST SET FOR DIRECT MEASUREMENT OF THE EQUIVALENT PARALLEL RESISTANCE.

(Courtesy of the General Electric Company Ltd. of England)

Test Set 193

For the direct measurement of e.p.r. an instrument, known as "the 193 Test Set" from its Services designation, is widely used. This set permits direct comparison between the crystal and a calibrated tuned circuit of variable loss. The e.p.r. can be measured easily, rapidly and with good accuracy by this means.

A diagrammatic representation of the circuit is given in Fig. 22, and the instrument is shown in Fig. 23. As will be seen from equation (12) the e.p.r. value can be readily converted into the e.s.r. value and vice versa, if the appropriate value of C_t is also known.

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[2] HEISING, "Quartz Crystals for Electrical Circuits", Van Nostrand.
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[4] BIGGS and WELLS, "The Measurement of the Activity of Quartz Oscillator Crystals", *Journal I.E.E.*, Part III, 1946.

E. A. F.

REGION 1 FREQUENCY ALLOCATIONS

The frequency allocation tables given on the following pages are based on the recommendations of the Atlantic City Conference, 1947, of the International Telecommunications Union, as applicable to Region 1, which roughly comprises Europe, Africa, the U.S.S.R. and Turkey A number of minor reservations made by the U.S.S.R. and certain other countries have been omitted from this table, but modifications relating to the United Kingdom are shown under "Remarks".

These frequency allocations are being revised at the Geneva Conference, 1959.

The following abbreviations have been used in the table :

Areas

W.W. . . . . . World-wide (i.e., applicable to Regions 1, 2 and 3)

1 . . . . . . Region 1

Services

Aero. Fx.	Aeronautical Fixed Service
Aero. Mob.*	Aeronautical Mobile Service
Aero. R. Nav.	Aeronautical Radio-navigation Service
Amat.	Amateur Service
B'cst.	Broadcasting Service
Fx.	Fixed Service
Land Mob.	Land Mobile Service
Mob.	Mobile Service
M. Mob.	Maritime Mobile Service
M. R. Nav.	Maritime Radio-navigation Service
R. loc.	Radio-location Service
R. Nav.	Radio-navigation Service
S. Freq.	Standard Frequency Service
Spec.	Special Service

* Aero. Mob. (R). Frequencies reserved for any communications between aircraft and those aeronautical stations primarily concerned with the safety and regularity of flight along national or international civil air routes.

* Aero. Mob. (OR). Frequencies reserved for any communications between aircraft and those aeronautical stations other than those primarily concerned with flight along national or international civil air routes.

In connection with the table of frequency allocations, the following international definitions should be noted :

Radio-location : Determination of a position or of a direction by means of the constant-velocity or rectilinear-propagation properties of Hertzian waves.

Radio-navigation : Radio-location intended solely for the determination of position or direction or for obstruction warning, in navigation.

Radar : Radio-location system where transmission and reception are carried out at the same location, and which utilizes the reflecting or retransmitting properties of objects in order to determine their positions.

Primary Radar : Radar using reflection only.

Secondary Radar : Radar using automatic retransmission on the same or on a different radio frequency.

Radio Direction-finding : radio-location in which only the direction of a station is determined by means of its emissions.

Fixed Service : A service of radio-communication between specified fixed points.

Broadcasting Service : (*a*) A radio-communication service of transmissions to be received directly by the general public. (*b*) This service may include transmissions of sounds or transmissions by television, facsimile or other means.

Region 2 is the Western Hemisphere, comprising the Americas, Greenland and all countries under the control of the Federal Communications Commission.

Region 3 is Australasia, Oceania and Asia, except those territories included in Regions 1 and 2.

TABLE 1.—FREQUENCY ALLOCATIONS APPLICABLE TO REGION 1

Frequency Band (kc/s)	*Area*	*Services*	*Remarks*
10–14	W.W.	R. Nav.	
14–70	W.W.	(*a*) Fx.	
		(*b*) M. Mob.	Coastal A1 stations only
70–90	1	(*a*) Fx.	
		(*b*) M. Mob.	Coastal A1 stations only
		(*c*) R. Nav.	Exclusive use of 70–72 kc/s and 84–86 kc/s
90–110	W.W.	(*a*) Fx.	
		(*b*) M. Mob.	Coastal A1 stations only
		(*c*) R. Nav.	Eventually to have exclusive rights
110–130	1	(*a*) Fx.	
		(*b*) M. Mob.	
		(*c*) R. Nav.	Exclusive use of 112–115 kc/s and 126–129 kc/s
130–150	1	M. Mob.	Ship stations only: calling frequency 143 kc/s
150–160	1	(*a*) B'cst.	
		(*b*) M. Mob.	On non-interference basis
160–255	1	B'cst.	
255–285	1	(*a*) Aero. R. Nav.	U.K. to share portions with M. Mob.
		(*b*) B'cst.	
		(*e*) M. Mob.	On non-interference basis
285–315	1	M.R. Nav.	
315–325	1	Aero. R. Nav.	U.S.S.R. to use for M. Mob.
325–405	W.W.	(*a*) Aero. Mob.	325–365 and 395–405 kc/s for B'cst. in certain European areas
		(*b*) Aero. R. Nav.	Priority over Aero. Mob. 333 kc/s calling frequency
405–415	1	(*a*) Aero. R. Nav.	Has priority on 410 kc/s
		(*b*) M.R. Nav.	Restricted types
		(*c*) Mob except Aero. Mob.	
415–490	W.W.	M. Mob.	Telegraphy only
490–510	W.W.	Mob.	Distress and calling only. 500 kc/s international distress frequency
510–525	1	M. Mob.	Telegraphy only
525–535	1	B'cst.	
535–1,605	W.W.	B'cst.	
1,605–2,000	1	(*a*) Fx.	In certain countries 1,800–2,000 kc/s may be used by Amateurs
		(*b*) Mob. except Aero. Mob.	
2,000–2,045	1	(*a*) Fx.	

TABLE 1.—*contd.*

Frequency Band (kc/s)	*Area*	*Services*	*Remarks*
2,000–2,045		(*b*) Mob. except Aero. Mob. (R)	
2,045–2,065	1	Met. Aids.	
2,065–2,300	1	(*a*) Fx. (*b*) Mob. except Aero. Mob. (R)	2,182 kc/s International distress and calling frequency
2,300–2,498	1	(*a*) B'cst. (*b*) Fx. (*c*) Mob. except Aero. Mob. (R)	Restrictions on use
2,498–2,502	1	S. Freq.	Standard frequency 2,500 kc/s
2,502–2,625	1	(*a*) Fx. (*b*) Mob. except Aero. Mob. (R)	
2,625–2,650	1	(*a*) M. Mob. (*b*) M. R. Nav.	By special arrangement
2,650–2,850	1	(*a*) Fx. (*b*) Mob. except Aero. Mob. (R)	
2,850–3,025	W.W.	Aero. Mob. (R)	
3,025–3,155	W.W.	Aero. Mob. (OR)	
3,155–3,200	W.W.	(*a*) Fx. (*b*) Mob. except Aero. Mob. (R)	
3,200–3,400	W.W.	(*a*) B'cst. (*b*) Fx. (*c*) Mob. except Aero. Mob.	Restrictions on use 3,200–3,230 kc/s Mob. except Aero. Mob. (R)
3,400–3,500	W.W.	Aero. Mob. (R)	
3,500–3,800	1	(*a*) Amat. (*b*) Fx. (*c*) Mob. except Aero. Mob.	
3,800–3,900	1	(*a*) Aero. Mob. (OR) (*b*) Fx. (*c*) Land Mob.	
3,900–3,950	1	Aero. Mob. (OR)	
3,950–4,000	1	(*a*) B'cst. (*b*) Fx.	
4,000–4,063	W.W.	Fx.	
4,063–4,438	W.W.	M. Mob.	

TABLE 1.—*contd.*

Frequency Band (kc/s)	*Area*	*Services*	*Remarks*
4,438–4,650	1	Fx.	
4,650–4,700	W.W.	Aero. Mob. (R)	
4,700–4,750	W.W.	Aero. Mob. (OR)	
4,750–4,850	1	(*a*) Aero Mob. (OR)	
		(*b*) B'cst.	Restrictions on use
		(*c*) Fx.	
		(*d*) Land Mob.	
4,850–4,995	W.W.	(*a*) B'cst.	Restrictions on use
		(*b*) Fx.	
		(*c*) Land Mob.	
4,995–5,005	W.W.	S. Freq.	Standard frequency 5,000 kc/s
5,005–5,060	W.W.	(*a*) B'cst.	Restrictions on use
		(*b*) Fx.	
5,060–5,250	W.W.	Fx.	
5,250–5,430	1	(*a*) Fx.	
		(*b*) Land Mob.	
5,430–5,480	1	(*a*) Aero. Mob. (OR)	
		(*b*) Fx.	
		(*c*) Land Mob.	
5,480–5,680	W.W.	Aero. Mob. (R)	
5,680–5,730	W.W.	Aero. Mob. (OR)	
5,730–5,950	W.W.	Fx.	
5,950–6,200	W.W.	B'cst.	
6,200–6,525	W.W.	M. Mob.	May exceptionally be used for Fx. (50 watts limit)
6,525–6,685	W.W.	Aero. Mob. (R)	
6,685–6,765	W.W.	Aero. Mob. (OR)	
6,765–7,000	W.W.	Fx.	
7,000–7,100	W.W.	Amat.	
7,100–7,150	1	(*a*) Amat.	
		(*b*) B'cst.	
7,150–7,300	1	B'cst.	
7,300–8,195	W.W.	Fx.	
8,195–8,815	W.W.	M. Mob.	
8,815–8,965	W.W.	Aero. Mob. (R)	
8,965–9,040	W.W.	Aero. Mob. (OR)	
9,040–9,500	W.W.	Fx.	
9,500–9,775	W.W.	B'cst.	
9,775–9,995	W.W.	Fx.	
9,995–10,005	W.W.	S. Freq.	Standard frequency 10,000 kc/s
10,005–10,100	W.W.	Aero. Mob. (R)	
10,100–11,175	W.W.	Fx.	
11,175–11,275	W.W.	Aero. Mob. (OR)	
11,275–11,400	W.W.	Aero. Mob. (R)	
11,400–11,700	W.W.	Fx.	
11,700–11,975	W.W.	B'cst.	

TABLE 1.—*contd.*

Frequency Band (kc/s)	*Area*	*Services*	*Remarks*
11,975–12,330	W.W.	Fx.	
12,330–13,200	W.W.	M. Mob.	
13,200–13,260	W.W.	Aero. Mob. (OR)	
13,260–13,360	W.W.	Aero. Mob. (R)	
13,360–14,000	W.W.	Fx.	13,560 kc/s ± 0·05% Industrial, scientific and medical
14,000–14,350	W.W.	Amat.	
14,350–14,990	W.W.	Fx.	
14,990–15,010	W.W.	S. Freq.	Standard frequency 15,000 kc/s
15,010–15,100	W.W.	Aero. Mob. (OR)	
15,100–15,450	W.W.	B'cst.	
15,450–16,460	W.W.	Fx.	
16,460–17,360	W.W.	M. Mob.	
17,360–17,700	W.W.	Fx.	
17,700–17,900	W.W.	B'cst.	
17,900–17,970	W.W.	Aero. Mob. (R)	
17,970–18,030	W.W.	Aero. Mob. (OR)	
18,030–19,990	W.W.	Fx.	
19,990–20,010	W.W.	S. Freq.	Standard frequency 20,000 kc/s
20,010–21,000	W.W.	Fx.	
21,000–21,450	W.W.	Amat.	
21,450–21,750	W.W.	B'cst.	
21,750–21,850	W.W.	Fx.	
21,850–22,000	W.W.	(*a*) Aero. Fx. (*b*) Aero. Mob. (R)	
22,000–22,720	W.W.	M. Mob.	
22,720–23,200	W.W.	Fx.	
23,200–23,350	W.W.	(*a*) Aero. Fx. (*b*) Aero. Mob. (OR)	
23,350–24,990	W.W.	(*a*) Fx. (*b*) Land Mob.	Inter-ship M. Mob. permitted 23,350–24,000 kc/s
24,990–25,010	W.W.	S. Freq.	Standard frequency 25,000 kc/s
25,010–25,600	W.W.	(*a*) Fx. (*b*) Mob. except Aero. Mob	
25,600–26,100	W.W.	B'cst.	
26,100–27,500	W.W.	(*a*) Fx. (*b*) Mob. except Aero. Mob.	27,120 kc/s ± 0·06% industrial, scientific and medical
27,500–28,000	1	Met. Aids	
28,000–29,700	W.W.	Amat.	

TABLE 1.—*contd.*

Frequency Band (Mc/s)	*Area*	*Services*	*Remarks*
29·7–31·7	1	Aero. R. Nav.	
31·7–41	1	(*a*) Fx.	40·68 Mc/s ± 0·05% industrial, scientific and medical
		(*b*) Mob.	Aero. R. Nav. may be permitted in Region 1
41–68	1	B'cst.	In U.K. 66·5–68 Mc/s may be used for Land Mob. and Fx.
68–70	1	Aero. R. Nav.	
70–72·8	1	(*a*) Fx.	In France and U.S.S.R. for Amat. service
		(*b*) Mob. except Aero. Mob.	
72·8–75·2	1	Aero. R. Nav.	75 Mc/s designated for Aero. marker beacons with guard-band ±0·2 Mc/s
75·2–78	1	(*a*) Fx. (*b*) Mob. except Aero. Mob.	
78–80	1	Aero. R. Nav.	
80–83	1	(*a*) Fx. (*b*) Land Mob.	
83–85	1	Aero. R. Nav.	
85–87·5	1	(*a*) Fx. (*b*) Mob. except Aero. Mob.	In U.K. 85–90 Mc/s shared with M. R. Nav.
87·5–88	1	B'cst.	In U.K. 85–90 Mc/s shared with M. R. Nav.
88–100	W.W.	B'cst.	In U.K. 85–90 Mc/s shared with M. R. Nav. In U.K. and France 94·5–95 Mc/s shared with Met. Aids In U.K. 95–100 Mc/s shared with Fx. and Land Mob.
100–108	1	Mob. except Aero. Mob. (R)	
108–118	W.W.	Aero. R. Nav.	
118–132	W.W	Aero. Mob. (R)	121·5 Mc/s Aero. Emergency Frequency
132–144	1	Aero. Mob. (OR)	
144–146	W.W.	Amat.	
146–156	1	Aero. Mob. (OR)	Met. Aids may use 151–154 Mc/s
156–174	1	(*a*) Fx.	156·80 Mc/s safety frequency for M. Mob.
		(*b*) Mob. except Aero. Mob.	In France 162–174 Mc/s for B'cst.

TABLE 1.—*contd.*

Frequency Band (Mc/s)	*Area*	*Services*	*Remarks*
174–216	1	B'cst.	In U.K. 174–200 Mc/s for Fx., 200–216 Mc/s for Aero. R. Nav., 200–235 Mc/s for distance measuring equipment
216–235	1	Aero. R. Nav.	
235–328·6	W.W.	(*a*) Fx. (*b*) Mob.	
328·6–335·4	W.W.	Aero. R. Nav.	
335·4–420	W.W.	(*a*) Fx. (*b*) Mob.	400–420 Mc/s may be used for Met. Aids
420–450	W.W.	(*a*) Aero. R. Nav. (*b*) Amat.	Has priority
450–460	1	(*a*) Aero. R. Nav. (*b*) Amat.	
460–470	W.W.	(*a*) Fx. (*b*) Mob.	
470–585	W.W.	B'cst.	
585–610	1	R. Nav.	In France and Italy 585–685 Mc/s for Fx. and B'cst.
610–940	W.W.	B'cst.	In France and Italy 585–685 Mc/s for Fx. and B'cst.
940–960	1	B'cst.	
960–1,215	W.W.	Aero. R. Nav.	
1,215–1,300	W.W.	Amat.	
1,300–1,600	1	(*a*) Fx. (*b*) Mob.	In U.K. 1,300–1,365 Mc/s restricted to surveillance radar
1,600–1,700	1	Aero. R. Nav.	
1,700–2,300	W.W.	(*a*) Fx. (*b*) Mob.	1,700–1,750 Mc/s may be used for Met. Aids
2,300–2,450	W.W.	Amat.	In U.K. 2,450 Mc/s ± 50 Mc/s may be used for industrial, scientific and medical purposes
2,450–2,700	W.W.	(*a*) Fx. (*b*) Mob.	
2,700–2,900	W.W.	Aero. R. Na	May be used for Met. Aids
2,900–3,300	W.W.	R. Nav.	3,246–3,266 Mc/s for Racons Shipborne radar 3,000–3,246 Mc/s
3,300–3,900	1	(*a*) Fx. (*b*) Mob. (*c*) R. Nav.	

TABLE I.—*contd.*

Frequency Band (Mc/s)	*Area*	*Services*	*Remarks*
3,900–4,200	W.W.	(*a*) Fx. (*b*) Mob.	
4,200–4,400	W.W.	Aero. R. Nav.	
4,400–5,000	W.W.	(*a*) Fx. (*b*) Mob.	
5,000–5,250	W.W.	Aero. R. Nav.	
5,250–5,650	W.W.	R. Nav.	5,440–5,460 Mc/s for Racons 5,460–5,650 Mc/s for ship-borne radar
5,650–5,850	W.W.	Amat.	5,850 Mc/s ± 75 Mc/s for industrial, scientific and medical
5,850–5,925	1	(*a*) Fx. (*b*) Mob.	
5,925–8,500	W.W.	(*a*) Fx. (*b*) Mob.	
8,500–9,800	W.W.	R. Nav.	9,300–9,320 Mc/s for Racons 9,320–9,500 Mc/s for ship-borne radar
9,800–10,000	W.W.	(*a*) Fx. (*b*) R. Nav.	
10,000–10,500	W.W.	Amat.	
Above 10,500	Not allocated		

STANDARD FREQUENCY TRANSMISSIONS

The frequencies of 2·5, 5, 10, 15, 20 and 25 Mc/s have been internationally allocated to the transmission of standard radio and audio frequencies, and time signals. Continuous services are in operation by the National Bureau of Standards from station WWV, Washington, D.C., and from WWVH, Puunene, Hawaii; and by the National Physical Laboratory from the Post Office Station MSF at Rugby. In addition, special experimental transmissions between the hours of 14.29 and 15.30 G.M.T. daily are made from MSF on 60 kc/s.

MSF

The transmission schedule of MSF is shown diagrammatically in Fig. 24 : carrier frequencies 2·5, 5, 10, 15, 20 Mc/s (three frequencies only are broadcast simultaneously, initially these are 2·5, 5 and 10 Mc/s); power 0·5 kW; modulation frequencies 1,000 c/s tone, 1 c/s pulses, the fifty-ninth pulse in each minute being omitted. The carrier and modulation frequencies are all derived from the same 100-kc/s standard and are maintained within ±2 parts in 10^8 of their nominal values. The frequency of the received signal may vary throughout the day,

Table 2.—Transmitter Frequency Tolerances
(Atlantic City Regulations)

Frequency Bands and Categories of Stations	*Tolerances* (%)
Band 10–535 *kc/s*	
(1) Fixed Stations	
From 10 to 50 kc/s	0·1
From 50 kc/s to end of band	0·02
(2) Land Stations	
(*a*) Coast stations—	
Power above 200 W	0·02
Power below 200 W	0·05
(*b*) Aeronautical stations	0·02
(3) Mobile Stations:	
Ship stations	0·1
Aircraft stations	0·05
Emergency, lifeboat stations, etc.	0·5
(4) Radio-navigation Stations	0·02
(5) Broadcasting Stations	20 c/s
Band 535–1,605 *kc/s*	
Broadcasting Stations	20 c/s
Band 1,605–4,000 *kc/s*	
(1) Fixed Stations:	
Power above 200 W	0·005
Power below 200 W	0·01
(2) Land Stations:	
(*a*) Coast stations—	
Power above 200 W	0·005
Power below 200 W	0·01
(*b*) Aeronautical stations—	
Power above 200 W	0·005
Power below 200 W	0·01
(*c* Base stations—	
Power above 200 W	0·005
Power below 200 W	0·01
(3) Mobile Stations:	
Ship stations	0·02
Aircraft stations	0·02
Land mobile stations	0·02
(4) Radio-navigation Stations:	
Power above 200 W	0·005
Power below 200 W	0·01
(5) Broadcasting Stations	0·005
Band 4,000–30,000 *kc/s*	
(1) Fixed Stations:	
Power above 500 W	0·003
Power below 500 W	0·01
(2) Land Stations:	
(*a*) Coast stations	0·005
(*b*) Aeronautical stations—	
Power above 500 W	0·005
Power below 500 W	0·01
(*c*) Base Stations—	
Power above 500 W	0·005
Power below 500 W	0·01
(3) Mobile Stations:	
Ship stations	0·02
Aircraft stations	0·02
Land mobile stations	0·02
Lifeboats, etc.	0·02
(4) Broadcasting Stations	0·003
Band 30–100 *Mc/s*	
(1) Fixed Stations	0·02
(2) Land Stations	0·02
(3) Mobile Stations	0·02
(4) Radio-navigation Stations	0·02
(5) Broadcasting Stations	0·003
Band 100–500 *Mc/s*	
(1) Fixed Stations	0·01
(2) Land Stations	0·01
(3) Mobile Stations	0·01
(4) Radio-navigation Stations	0·02
(5) Broadcasting Stations	0·003
Band 500–10,500 *Mc/s*	0·75

however, if there are ionospheric reflections in the transmission path. This error, which is due to the movement of the reflecting layers, seldom exceeds ± 2 parts in 10^7, and for a large part of the day is not more than a few parts in 10^8. The transmitted frequencies do not, in general, vary from day to day by more than ± 2 parts in 10^9.

The seconds pulses are derived from the standard by division, and consist of five cycles of 1,000 c/s tone. The precision of the pulses is $\pm$ 1 microsecond and the time interval between two pulses is therefore accurate to ± 2 parts in $10^9 \pm 2$ microseconds. For example, if the frequency is 1×10^{-8} high, then the time interval between corresponding pulses on consecutive days is 1×10^{-8} (approximately 1 millisecond) less than 1 day. The time error is integrated, and in general no attempt is made to alter the phase of the pulses so as to make them coincident with uniform time. If, however, they are in error by more than 50 milliseconds, an adjustment of 50 or 100 milliseconds is made. Such

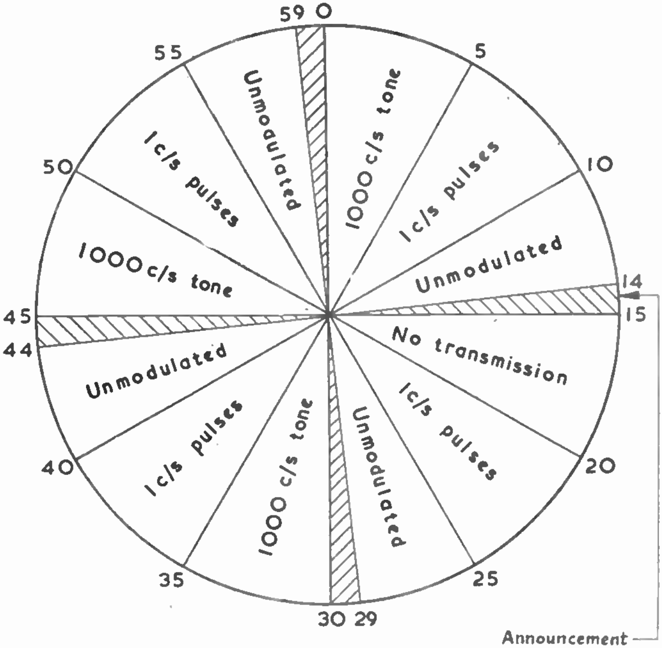


FIG. 24.—TRANSMISSION SCHEDULE OF MSF.
(National Physical Laboratory)

adjustments are made on the first day of the month, and the extent of the adjustment is announced.

WWV

The broadcasts from WWV are continuous on 2·5, 5, 10, 15, 20 and 25 Mc/s, the modulation being 440 or 600 c/s tone and 1 c/s pulses. The transmission schedule consists of four-minute tone periods (alternatively 440 and 600 c/s) followed by 1-minute announcement periods. The transmitted frequencies are correct to within ± 2 parts in 10^8. During the announcement periods at 20 minutes past and 10 minutes before the hour, information on propagation conditions over the North Atlantic route is transmitted in Morse code: a letter indicating the conditions at the time of the broadcast, followed by a number denoting the conditions expected during the following 12-hour period. This code is as follows: W—ionospheric disturbance in progress or expected; U—unstable conditions, but communication possible with high power; N—no warning; 1—impossible; 2—very poor; 3—poor; 4—fair to poor; 5—fair; 6—fair to good; 7—good; 8—very good; 9—excellent.

IONOSPHERIC FORECASTING

A radio wave propagated in empty space travels in a straight line, and in this context it may be considered that a wave propagated in un-ionized air—such as exists in the lower regions of the atmosphere—does likewise. It is evident, therefore, that under these conditions high-frequency long-distance radio transmission could not be accomplished, for the wave would not bend its path so as to reach a receiver on the other side of the spherically shaped earth, but would go straight off into space. That it does, in fact, bend around the earth is due to the existence, in the upper atmosphere, of a region where the air is ionized by the action of the sun: the region called the "Ionosphere". Because of the action of the solar radiation the ionospheric air contains free electrons, and thus acts as an electrical conductor, having the property of reflecting or refracting radio waves and of so bending their paths that they are returned to the earth's surface at distant points.

Formation and Structure of the Ionosphere

The upper atmosphere may be likened to an enormous chemical laboratory, in which the various gases—the principal of which are nitrogen and oxygen in both molecular and atomic forms—are subjected to the action of ultra-violet, X-ray and particle radiations which arrive there from the sun, and which give rise to the processes of chemical reaction, photo-dissociation and photo-ionization. Absorption of the solar radiations occurs during these processes, and they do not, therefore penetrate to the lower atmosphere. The different gases in their various forms are not, therefore, uniformly distributed throughout the upper atmosphere, but concentrations of a particular gas occur at different levels, as determined by the effects of the solar radiations. Consequently the photo-ionization process results in "layers" of ionization, each having its maximum electron concentration at a certain well-defined altitude. These are the D layer at an altitude of about 70 km., the E layer at about 110 km., and F_1 at about 200 km., and the F_2 at an altitude varying with time of day and season from about 280 to 480 km. The level of ionization in the different layers varies widely with time of day and season, latitude and period in the sun's cycle of activity, and is a function of the rate of ion production and the recombination rate, which latter is highest in the lowest layers, where the gas density is highest. The free-electron density is therefore highest in the F_2 layer, and lower in each other layer in inverse order of altitude, an approximate estimate being 1 million electrons per cubic centimetre in the F_2, 100,000 in the E and 10,000 in the D. The F_1 is observable only during the day, and the D also disappears shortly after sunset. The E layer ionization falls to a very low value during the night, but some "residual" ionization persists throughout the dark hours. The F_2 remains in being day and night. The three upper layers all possess the property of reflecting or refracting radio waves of different frequencies, but the D layer, so far as the short-waves go, acts as an absorbing layer and not as a reflector.

Summing up the ionospheric structure, we have, during the day, the F_2, F_1 and E as reflecting and the D as an absorbing region, and at night the F_2 with the E as a weak reflector only.

Measurement of Ionospheric Characteristics

In order to make the best use of the ionosphere as a transmission medium it is necessary to have available a forecast map detailing the ionospheric conditions likely to prevail all over the world. The basis upon which such maps are prepared is the regular hourly measurements now being made at some 100 ionospheric observations spread over the earth's surface.

A pulse of radio energy sent vertically upwards from the ground will travel at a constant velocity through the un-ionized air, but when it enters an ionospheric layer its velocity is reduced, to a degree depending on the density of the free electrons and inversely on the square of the frequency. In other words, the ionosphere has a refractive index which becomes progressively less than 1 the denser the free electrons become. At the level where the relationship between the wave frequency f and the electron density N reaches a certain critical value the refractive index becomes zero, and the pulse can no longer be propagated but is reflected downwards. If N is the *maximum* density of electrons in the layer all frequencies less than that necessary to produce the critical relationship will be reflected back to earth: all those greater than this will pass through the layer. The value of f when N is the maximum density of electrons is called the "critical" frequency, and is the highest frequency to be returned at vertical incidence.

Measurements at the ionospheric observatories are made by an equipment which automatically switches on at hourly intervals and sends pulses vertically upwards, starting on a low frequency and sweeping rapidly up in frequency. The pulse "echoes" from the ionosphere are received and displayed on an oscilloscope calibrated in terms of frequency and effective height of reflection, and the display is photographed, the film being moved in suitable steps corresponding to the steps of the frequency sweep. In this way a curve of effective height against frequency (an h'–f curve) is traced out, from which the significant ionospheric quantities may be read off. Fig. 25 shows a typical h'–f curve, in which $f^0{}_E$ and $f^0{}_{F_1}$ are the critical frequencies of the E and F_1 layers, and $f^0{}_{F_2}$ and $f^x{}_{F_2}$ are respectively the critical frequencies for the ordinary and extraordinary waves for the F_2 layer. The wave in this latter case is split into two components by the action of the earth's magnetic field, but $f^0{}_{F_2}$ is the more important quantity.

Normal Variations in Electron Density

Regular world-wide measurements of the ionosphere have disclosed the nature of the normal variations in electron density, as shown by the large variations in the critical frequencies of the layers. These are, first, variations following those in the sun's zenithal angle, which result in large diurnal, seasonal and geographical variations, secondly, those consequent upon the eleven-year cycle of activity of the sun itself, which result in a long-period variation in the critical frequency, and thirdly, those produced by upper atmosphere variations and by geomagnetic effects which produce anomalous variations in the F_2 layer only. The graphs of Fig. 26 will help to clarify the matter. All the layers except the F_2 behave similarly and in a regular manner, the geographical

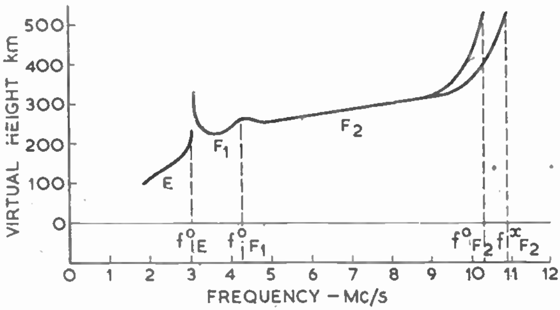
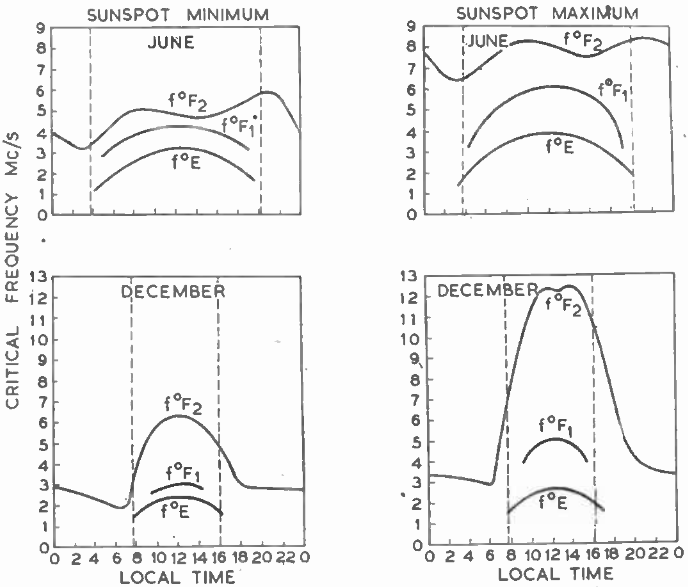


FIG. 25.—AN h'–f CURVE FOR A WINTER DAY SHOWING HEIGHTS AND CRITICAL FREQUENCIES FOR E, F_1 AND F_2 LAYERS.

FIG. 26 (*below*).—MONTHLY MEAN CRITICAL FREQUENCIES MEASURED AT SLOUGH FOR SUMMER AND WINTER—SUNSPOT MINIMUM AND SUNSPOT MAXIMUM PERIODS.



distribution being such that their critical frequencies are highest at the sub-solar point, i.e., where the sun is overhead at noon, and decrease more or less symmetrically towards the north and south poles, and east and west towards the dark zone. At a given point (see Fig. 26) they are either non-existent or have an unmeasurably low critical frequency at night, one that increases from sunrise to reach a peak at noon and a low value again at sunset, and one that is higher in summer than in winter. The geographical variation in the F_2 layer is such as to produce peaks in the day hemisphere a few degrees north and south of the magnetic equator, with the critical frequencies decreasing towards the north and south poles, and being low in the night hemisphere. In mid-latitudes (see Fig. 26) there is a general decrease during the night and an increase around sunrise. But the layer behaves anomalously in two respects: (1) its daytime critical frequency is lower in summer than it is in winter; (2) in summer the peak critical frequency occurs, not near noon as in winter, but around sunset. As to the effect of the eleven-year solar cycle it is the same for all the layers, resulting in a gradual increase in critical frequency from the time when the sun is least active (sunspot minimum) to peak values, when its ionizing radiations are most intense. The order of increase is, however, much greater for the F_2 than for the other layers, and is greater for that layer at noon in winter (2·0–2·5 times) than at any other time.

All these variations must be taken account of in preparing the forecast maps of ionospheric conditions.

Radio Transmission via the Ionosphere

In radio transmission over a distance the wave is, of course, incident on the ionosphere not vertically, as in measurement work, but obliquely. It has already been mentioned that when a wave enters a region where the air is ionized its velocity is altered, and the wave velocity is, in fact, increased * to a degree depending on the frequency and the density of the free electrons. It can easily be seen that, with electron density increasing from the lower boundary of a layer upwards, a wave incident on the layer obliquely will always have the upper part of its wave front in a region of higher electron density than its lower part. Therefore the upper part will travel faster than the lower and the wave path can no longer be a straight line, but will curve round so as to describe a trajectory which is finally in a downward direction again. Thus the wave is refracted and returns to the earth's surface, where it is reflected upwards again, and so travels onwards in a series of hops, as in Fig. 27.

In applying the vertical-incidence measurement data to transmission over a distance we make use of an obliquity law, according to which there is, for every frequency f which is reflected at a given height at vertical incidence, a frequency f' which is reflected at the same height at oblique incidence, and $f' = f \, . \sec\triangle$, where $\triangle$ is the angle the wave makes with the normal to the layer (see Fig. 27). Thus when f is the critical frequency of the layer $f \, . \sec\triangle$ is the highest frequency returned at any distance, and is called the Maximum Usable Frequency (M.U.F.) for that distance.

* Though the velocity of the pulse as a whole is decreased, i.e., the group velocity is decreased, the wave velocity is increased.

TABLE 3.—M.U.F. FACTORS

Distance (km.)	*E layer*	*F_1 layer*	*F_2 layer*							
			Sunspot maximum				*Sunspot minimum*			
			June		*December*		*June*		*December*	
			Noon	*Mid-night*	*Noon*	*Mid-night*	*Noon*	*Mid-night*	*Noon*	*Mid-night*
500	2·00	1·35	1·08	1·08	1·15	1·08	1·20	1·20	1·32	1·16
1,000	3·36	1·90	1·24	1·24	1·45	1·24	1·55	1·55	1·75	1·53
1,500	4·22	2·58	1·55	1·55	1·85	1·55	2·06	2·06	2·40	2·00
2,000	4·77	3·05	1·80	1·80	2·25	1·80	2·45	2·45	2·90	2·33
2,500	5·20	3·40	2·15	2·15	2·62	2·15	2·75	2·75	3·25	2·70
3,000	—	3·65	2·40	2·40	2·90	2·40	3·10	3·10	3·60	3·00
4,000	—	—	2·70	2·70	3·18	2·70	3·38	3·38	3·84	3·28

The secant law is strictly true only for a flat earth, and is modified by the ionospheric curvature and also by the effect of the earth's magnetic field on the free electrons. It is usual to take these effects into account and to express the result as a M.U.F. Factor, by which the vertical incidence critical frequency must be multiplied in order to obtain the M.U.F. for any layer and distance. In the case of the E and F_1 layers there is little variability in the Factors from time to time, but the F_2 Factor varies considerably with time of day, season and over the sunspot cycle, owing to variations in its height and semi-thickness. Table 3 gives M.U.F. Factors for various layers and distances, appropriate to these latitudes.

It is usual, in ionospheric work, to use the monthly mean values of the measured data, and since in the F_2 layer there is considerable day-to-day variation, the monthly mean M.U.F. is the value which will not be exceeded on 50 per cent of the days. By deducting 15 per cent from this a quantity called the Optimum Working Frequency (O.W.F.) is found, which is considered to be applicable to all the undisturbed days of the month.

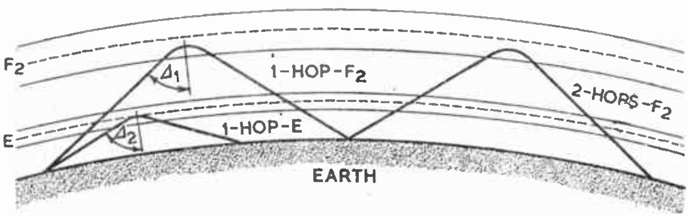


FIG. 27.—PROPAGATION BY HOPS BETWEEN IONOSPHERIC LAYERS AND EARTH.

Ionospheric Contour Maps

There are several different methods in use today for applying the world-wide ionospheric data to practical short-wave problems, but we shall confine ourselves here to the one used in the United Kingdom. This makes use of the ionospheric contour maps published by the Department of Scientific and Industrial Research (D.S.I.R.).

Suppose that the monthly mean measured F_2 critical frequency data from the 100 observatories are assembled, and that the value for each station appropriate to every second hour Universal Time is entered upon twelve Mercator maps in its correct latitude and longitude (one map for each second hour), then by joining up the values of equal critical frequency twelve critical frequency contour maps are produced, appropriate to every two hours from 0000 to 2200 Universal Time, showing the world-wide distribution of critical frequency. Similarly, by first multiplying the critical frequency values by the 4,000-km. M.U.F. Factors appropriate to each station, the maps can be made to show the 4000-km. M.U.F. contours, instead of those of critical frequency.

But what we need for practical use are contour maps showing, not conditions as they were in some past month, but forecast maps showing conditions as they will be during any month up to six months ahead. It will be remembered that the principal cause of variation in the world-wide hourly distribution of critical frequency as between a given month

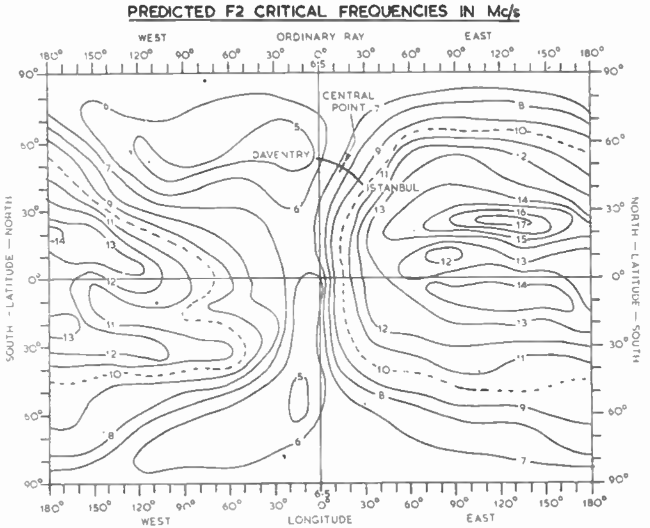


FIG. 28.—USING A PREDICTED FREQUENCY CONTOUR MAP (0600 U.T. MARCH 1959) FOR FINDING THE M.U.F. FOR A SINGLE-HOP CIRCUIT.

(Crown Copyright Reserved.)

and the same month of a future year are the effects of the eleven-year solar cycle. These variations have different values at different times of day, seasons and geographical positions, but it can be seen that if the measured values are corrected for the effects of these variations, then it is possible to produce a set of critical frequency or 4,000-km. M.U.F. contour maps for a month in the future. This description gives only the bare principles of the production of such maps, but they are the basic material with which the short-wave engineer has to work in the planning of his frequency schedules.

Figs. 28 and 29 are examples of such maps. They are published by permission of the Director of Radio Research, Slough of the Department of Scientific and Industrial Research. Maps are prepared on a similar basis showing the contours for the E and F_1 layers.

In order to use the maps to give the M.U.F. and O.W.F. for a one-hop circuit it is only necessary to draw in the great-circle path from transmitter to receiver as in Fig. 28 and to locate the position of the centre of this path. This is the ionospheric point where refraction occurs, and by reading off the value for this point from each of the twelve F_2 maps in turn, a circuit curve of M.U.F. and O.W.F. may be constructed for the 24-hour day of the month in question. (It is not, of course, necessary to draw the great-circle path on all the maps; but merely to locate the centre point.) The values may be either read off from the cri ical-frequency map and multiplied by the M.U.F. Factor appropriate to the

PREDICTED F2–4000Km–MUF IN Mc/s

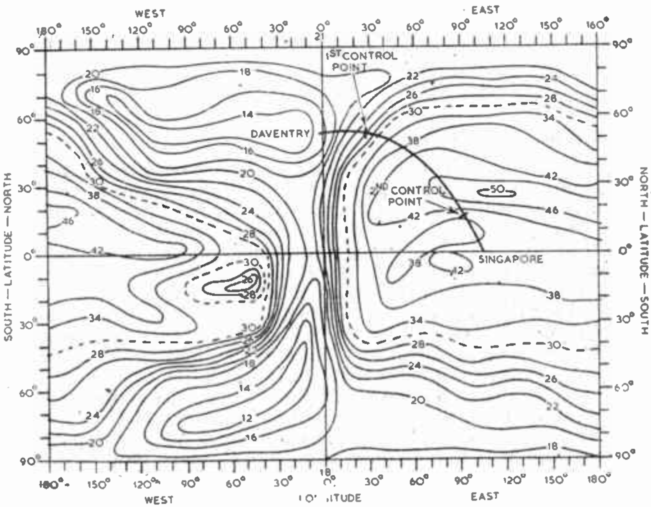


FIG. 29.—USING A PREDICTED 4,000-km. M.U.F. MAP FOR FINDING THE M.U.F. FOR A MULTIPLE-HOP CIRCUIT.

(Crown Copyright Reserved.)

distance, or from the 4,000-km. M.U.F. map and multiplied by a Distance Factor which is less than unity, or obtained by a combination of these methods.

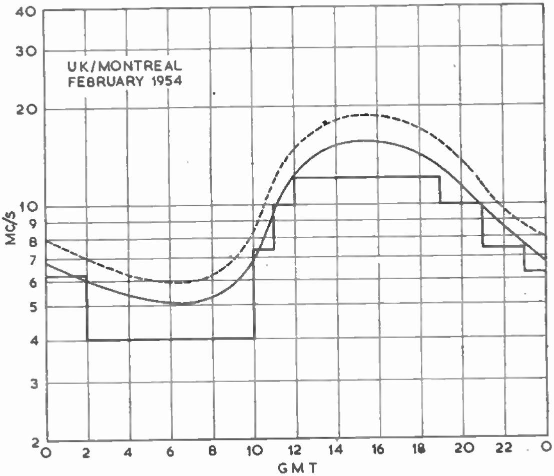
If the length of the transmission path is greater than 4,000 km. it cannot be covered in one hop, and up to about six hops between ionosphere and earth may be necessary to cover the longest paths. Where multiple-hop transmission occurs it cannot be treated as a simple extension of the one-hop case because of complexities introduced by the ionosphere, such as diffusion of the rays, scattering of energy, focusing, etc. Resort is therefore made to an empirical method for solving the problem, and this is at present widely used as giving the best results. No matter what the distance between transmitter and receiver, the great-circle path between them is drawn, as in Fig. 29, and a point 2,000 km. from either end is located. These are called the " control points " and the M.U.F. is read off for *both* points from each of the twelve 4,000-km. M.U.F. maps in turn, but for each map reading the *higher* of the readings is struck out (again it is unnecessary to draw the great-circle path on all the maps, but merely to locate the two control points). It will be noted that the 2,000-km. control points locate the ionospheric positions where a radio wave, travelling by the most oblique possible path so as to traverse 4,000 km. in a single hop, would undergo refraction either after leaving the transmitter or before reaching the receiver. The 4,000-km. M.U.F. maps are therefore used for all multiple-hop transmission paths, and the geographical variations along the path are thus taken account of by using the M.U.F. indicated for the control point where it is lowest. If refraction occurs at this frequency at this point, it will occur at that frequency (or even higher ones) at all other points along the path. A circuit curve of M.U.F. and O.W.F. is then constructed for the multiple-hop circuit, using the *lower* of the two readings for each hour.

Effects of E and F_1 Layers

In the case of multiple-hop transmission it is generally assumed that transmission is by way of the F_2 layer at all times (except when there is high-density Sporadic E) and the procedure described above suffices to find the M.U.F. and O.W.F. For one-hop circuits, however, it is possible that, over a certain range of distances and for a limited period of time around noon, the E or F_1 layer may control the M.U.F. This is because, though the electron density in these layers is much less than in the F_2, because they are lower down in the atmosphere, the angle of incidence on the layer in covering a short distance is greater than on the F_2 for the same distance. Thus the F_1 and E M.U.F. Factors for a given distance are greater than those for the F_2 (see Table 1), and this may result in their having the higher M.U.F. This matter is dealt with by reading, from the E or F_1 contour map, the M.U.F. for transmission by those layers for a few hours around noon, comparing the results with the M.U.F.s obtained from the F_2 maps, and striking out whichever M.U.F. is the *lower*. One is then left with the controlling M.U.F. for the circuit, be it by way of the E, F_1 or F_2. No allowance for day-to-day variation is necessary for E and F_1 transmission, i.e., the O.W.F. is assumed to coincide with the M.U.F.

Fig. 30 is an example of a monthly circuit curve as obtained by the

Fig. 30.—Circuit Curve Indicating Predicted M.U.F. and O.W.F. for a Month near a Sunspot Minimum.

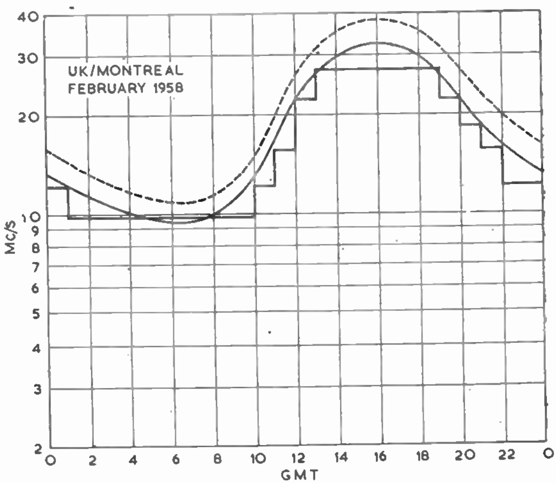


methods described. Fig. 31 is a curve for the same circuit for the same month, but for a year when sunspot activity was much higher.

The Lowest Useful Frequency

Circuit curves such as those of Figs. 30 and 31 contain all that is necessary to indicate the best frequency on which to operate a long-distance

Fig. 31.—Circuit Curve Indicating Predicted M.U.F. and O.W.F. for a Month near a Sunspot Maximum.



circuit, if the principle is observed of always working on the frequency nearest to the O.W.F. It is, however, useful to have one further piece of information, and that is an indication of the lowest frequency on which it is possible to work. For as frequency is reduced below the O.W.F. the absorption to which the wave is subject increases, until eventually a frequency is reached on which the field strength falls to a value which is inadequate for the service. The lowest frequency on which satisfactory field strength can be obtained is the "lowest useful frequency" (L.U.F.). The problem of finding this quantity is complicated, as it involves radiated power and directivity of aerials, world-wide variation of *D* layer absorption, world distribution of atmospheric radio noise, etc. But, as has been said, if the working frequency is chosen so as to be as near below the O.W.F. as possible, then the absorption on that frequency will be as low as it is possible to achieve for that particular circuit. It will be obvious, also, observing that the absorbing *D* layer disappears at night, that for a darkness path the L.U.F. will fall to an unimportant low value.

Ionospheric Disturbances

The above is a brief description of the main features of forecasting the *normal* ionospheric conditions which govern radio transmission. From time to time, however, the ionosphere is subject to violent disturbances, which temporarily produce abnormal conditions and so disrupt communications. For many years attempts have been made to forecast the onset of these ionospheric disturbances, but this has proved to be a difficult business. There are two distinct kinds of disturbances—one very short lived—but the principal kind, or ionospheric "storm", is almost certainly caused by the arrival in the ionosphere of streams of corpuscles which have been emitted from the sun about 30 hours earlier: the same corpuscles which give rise to displays of the auroræ and to disturbances in the earth's magnetic field. The forecasting problem is to identify some phenomenon which would positively indicate that the corpuscles were leaving the sun. Sunspots, solar flares and many other phenomena have been studied and, though the corpuscles are known to be associated with these, it has not yet been found possible to obtain accurate forecasts on this basis. In fact, the most reliable forecast material at present is the known strong tendency of the storms, during the decreasing phase of the sunspot cycle only, to repeat themselves at intervals of twenty-seven days. This corresponds to the approximate rotation period of the sun upon its axis, so that if a persistent corpuscular stream is emerging from a certain solar region so as to reach the earth (and such persistent streams occur only during the decreasing phase of the cycle), then in twenty-seven days time that region will again be in the same position relative to the earth, and the stream will start another ionospheric disturbance. It should be added that the observation of noise bursts from the sun on a number of different frequencies appears, at the present time, to offer promising hope of identifying the position of the corpuscles as they travel outwards from the sun through the solar corona.

T. W. B.

45. INDUSTRIAL TELEVISION

45. INDUSTRIAL TELEVISION

For an industrial television system the basic requirements are reliability and operational simplicity, combined with ruggedness and initial low cost. The equipment must be designed so that it can be operated by personnel whose only knowledge of television may lie in owning a domestic television receiver. The equipment may have to be operated twenty-four hours a day, every day, and may be situated in dusty, hot or cold, wet or humid surroundings.

SCANNING SYSTEMS

When the scanning standards for a public television service are chosen, a great many points have to be weighed up before a final decision is reached. These points are largely concerned with economic factors, the quality of the resulting picture being chosen on a value-for-money basis.

When television is used for industrial purposes—in this Section the term " Industrial " is taken to refer to all non-entertainment applications of television—the scanning standards can be chosen so that they best fulfil the needs of a particular situation. Where, for example, it is required to transmit still pictures over a low band-width line, a slow scanning rate would be adopted and the received information viewed on a cathode-ray tube having a long-persistence screen.

However, the majority of applications require the use of scanning rates similar to, or identical with, broadcasting standards, i.e., a scanning raster of some 400–800 lines per picture. Out of the television standards employed by the various world authorities, the only universal feature is the adoption of a 2 : 1 scanning interlace, this feature being worthwhile because of the twofold saving in band-width obtained for a given picture flicker rate. Some industrial television applications, however, benefit by the use of a sequential scanning system, especially those involving photographic recording of the televised image, as happens, for example, in the new electronic method of film making,[1] or when a television system is used in astronomy.

Where the televised scene emits very feeble quantities of light, as can happen in astronomy, it may be necessary to employ to the fullest advantage the storage action [2] of the camera tube, the storage action being the accumulation of the picture charge before its evaluation by the scanning beam. Since with normal repetitive periodic scanning and a suitable type of camera tube the maximum storage time is the duration between successive scannings, in order to derive fullest benefit of the storage action it is necessary to employ a non-repetitive mode of scanning.

This could mean scanning the camera tube with a single frame of, say, 400 lines in $\frac{1}{25}$ second at a selected time, recording photographically the picture briefly displayed on the monitor cathode-ray tube.

Apart from these highly specialized applications of television, the majority of industrial television requirements—telemetering and remote viewing in general—can be fulfilled by equipment operating on, or close to, normal scanning standards. Indeed, in many cases a simple "random" type of scanning is adequate or advisable because of its extreme simplicity. The term "random" is used because the raster produced has neither the appearance of a definite interlaced raster nor a true sequentially scanned raster, the timing of such a system being dependent on two master oscillators, one running at the line-scanning frequency and the other at the frame-scanning frequency. This mode of operation produces a raster whose lines will twinkle in and out of interlace, as there is no definite relationship between the line- and frame-scanning frequencies.

SYNCHRONIZING PULSE GENERATION

The generation of the synchronizing signal can follow conventional broadcast standards and produce a waveform similar to that used by the B.B.C. (Section 10, Fig. 3) or that recommended by the C.C.I.R. In many cases the production of a composite waveform is not necessary, since the video signal and the two synchronizing signals can be fed to the appropriate circuits separately. Where a composite waveform is required, the serration of the frame-synchronizing pulse can often be dispensed with, as the signal is generally fed over a closed circuit and there is no noise in the pulses to cause timing errors.

A typical composite waveform for an industrial television equipment, operating on the 405- or 625-line, 50-field systems, is shown in Fig. 1. As will be apparent, the only essential difference between this and the B.B.C. type waveform is that the frame-synchronizing signal consists of a single pulse of approximately the same duration as a scanning line.

The 2 : 1 interlace relationships of the line and frame pulses can be achieved by the general scheme outlined in Fig. 2, where if a 405-line, 50-field system is considered, the master oscillator must operate at a frequency 405×50 c/s = 20,250 c/s. From the master oscillator

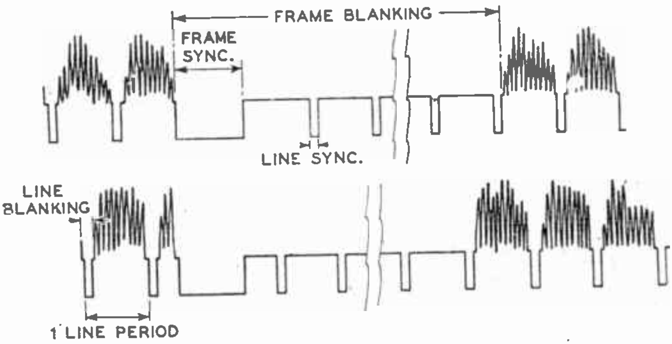


Fig. 1.—Two Successive Fields of a Composite Waveform for an Industrial Television Equipment.

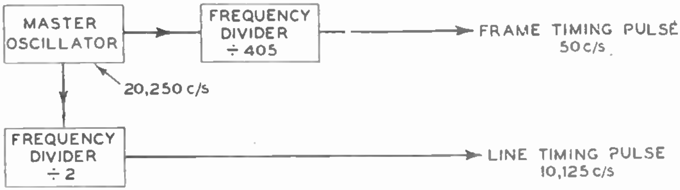


FIG. 2.—OUTLINE OF FREQUENCY DIVISION SYSTEM.

operating at 20,250 c/s, a divider circuit dividing by 405 will produce the frame-timing pulse, while a circuit dividing the master oscillator frequency by two will produce the line-timing pulse.

The master oscillator could be a multi-vibrator, a blocking oscillator or a stable sine-wave oscillator feeding a pulse-forming circuit. The divider circuit could employ stages of binary counters with feedback applied, as is the normal television station technique, the feedback being necessary to enable the odd number ratio of 405 to be counted. Since 405 has the factors 27 and 15, two blocks of binary counters dividing in these ratios will produce the frame-timing pulse.

While the binary counters have excellent characteristics, it is possible to obtain larger counts per stage with a good degree of stability by using step-counter circuits, blocking oscillators or flip-flops.

A selection of these circuits is shown in Fig. 3, and with these stable operation is possible with counts of up to seven per stage. Considering

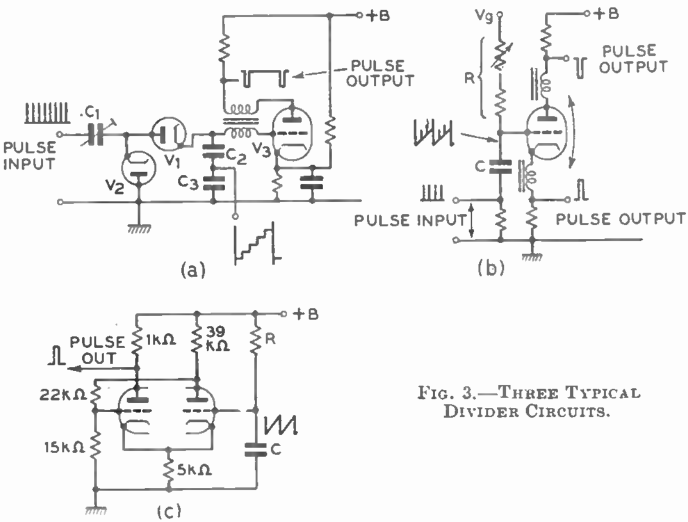


FIG. 3.—THREE TYPICAL DIVIDER CIRCUITS.

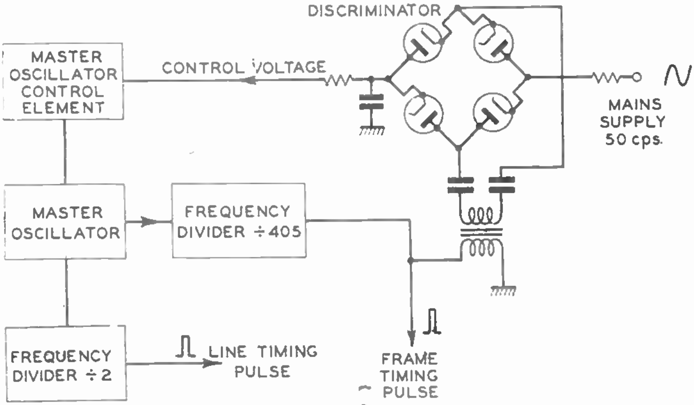


FIG. 4.—BLOCK DIAGRAM OF FREQUENCY DIVIDER WITH MAINS LOCK CIRCUIT.

405-line operation, the fact that 3 × 3 × 3 × 3 × 5 is equal to 405 means that the frame pulse can be derived from the master oscillator in five stages. Fig. 3 (*a*) shows a step counter, so called from the wave-shape obtained at the test point at the junction of C_2 and C_3. C_3 is about ten times C_2, and C_2 is about ten times C_1. When a series of positive pulses is fed into the counter through C_1, the capacitor C_2 receives a charge increment for each input pulse, the increment depending on the ratio C_1/C_2, while the diode V_2 prevents the capacitor C_1 from accumulating any charge. By proportioning correctly the input pulse voltage and the ratio C_1/C_2, it can be arranged that the blocking oscillator V_3 remains biased off until, say, five input pulses have been counted. When this occurs, the positive charge on capacitor C_2 has just exceeded the cut-off point of the blocking oscillator, which conducts, discharging C_2 and causing a single pulse in the output circuit.

By making C_1 variable, it is possible therefore to set the " count " of the stage to the desired value.

Although a blocking oscillator is shown and discussed as the discharging element of the step counter, a multi-vibrator circuit would perform a similar task.

It is quite possible to use blocking-oscillator or multi-vibrator circuits as frequency dividers directly. Fig. 3 (*b*) shows a blocking oscillator suitable for frequency division where the oscillator frequency is controlled by the time-constant R-C and the voltage V_g. Positive pulses of higher frequency are inserted at the synchronizing terminal, and appear at the valve grid added to the grid waveform; the frequency of the oscillator can be adjusted so that the oscillator fires on the required synchronizing pulse, say, the second, third or fifth, producing one pulse in the output.

Fig. 3 (*c*) shows a multi-vibrator circuit which has been used for frequency division.[3] As in the case of the blocking oscillator, the frequency is determined primarily by the values of R and C. By adding positive pulses to the sawtooth voltage developed across C, the circuit can be operated as a frequency divider.

So that the effects of residual "mains hum" in the equipment shall not be troublesome, it is common practice to lock the derived field frequency to the supply-mains frequency, wherever possible. In order to preserve the division ratio it is necessary to employ a discriminator circuit to compare the generated field frequency with the supply mains frequency, any frequency difference producing a voltage used to correct the frequency of the master oscillator. Such a scheme is shown in Fig. 4.

The line and frame (field) pulses obtained from the divider circuits could be used as synchronizing pulses directly and, by applying them to widening circuits, blanking signals can be generated.

Although the frequency dividers were described in connection with interlaced scanning systems, by merely altering the division ratio from an odd to an even number they apply to sequential scanning systems.

For the random type of scanning referred to under "Scanning Systems", blocking oscillators of the type shown in Fig. 3 (*b*) are admirable. Good results can be had by employing one oscillator locked to the supply frequency to generate the frame pulses, and another running free at line speed to produce the line pulses, the number of lines in the raster being adjustable by varying the line-oscillator frequency control.

THE FLYING-SPOT SCANNER

The simplest television system uses the "Flying Spot" technique to generate the picture signal.

The basic principle of a cathode-ray-tube flying-spot scanner is illustrated in the three sketches comprising Fig. 5. The essential components are the flat-faced scanning tube with the conventional means of focusing and deflecting the scanning beam, and the photocell usually combined with a secondary emission multiplier in order to obtain optimum signal-to-noise ratio. The transparency or picture to be televised is placed between the tube and the photomultiplier. The first two examples in Fig. 5 refer to the transmission of slides (transparent pictures), the simplest scheme being indicated in (*a*), where the slide is placed in contact with the front plate of the scanning tube and a photocell with a small aperture is positioned some distance away. While the luminous spot traces out the raster on the screen, the light received by the photocell is modulated by the local transparency of the slide, and in this manner the picture content is translated into a sequence of electrical impulses. Whilst this arrangement can be very useful as a simple and inexpensive signal generator, an optical imaging process—as shown in Fig. 5 (*b*)—is required to produce pictures of high quality. Here the plane of the screen is imaged on to the slide, the transmitted light being collected by the condenser lens and passed to the photomultiplier. Fig. 5 (*c*) shows a means of televising opaque pictures. The scanning spot is again imaged on to the object, while an arrangement of photomultipliers picks up the diffusely reflected light, which will change from point to point, depending on the picture content.

It is necessary that the phosphor of the scanning cathode-ray tube

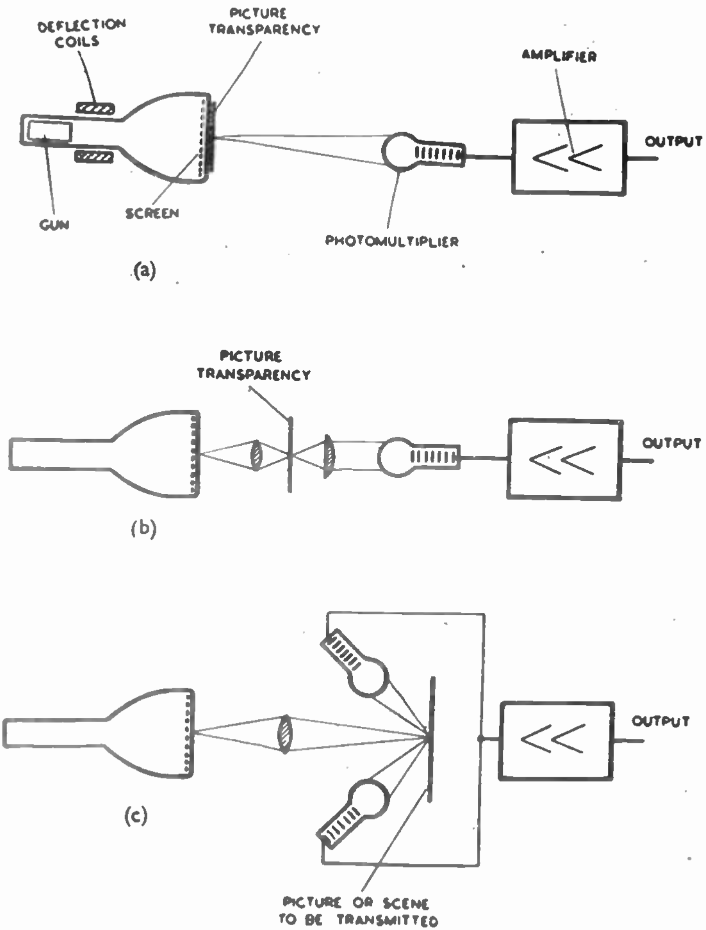
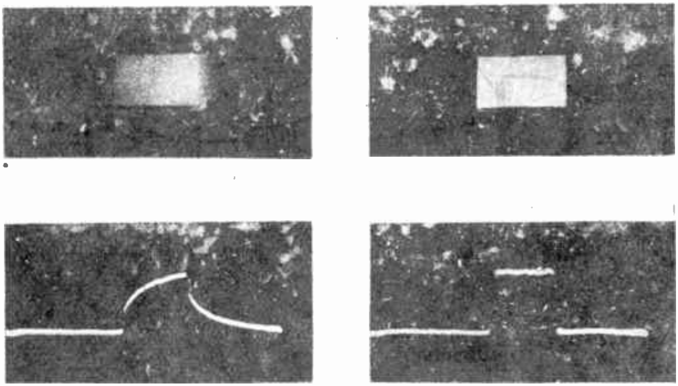


Fig. 5.—Three Examples of Flying-spot Scanning with Cathode-ray Tubes.

should have a short afterglow, otherwise effects will be produced in the picture similar to those caused by a poor high-frequency response in the amplifier.

Afterglow effects are illustrated in Fig. 6 (*a*), where photographs show the results obtained when a slide consisting of a white rectangle on a



(a) (b)

Fig. 6.—Television Pictures and Oscillograms Showing the Effects of Scanning-tube Phosphor Afterglow: (a) Uncorrected; (b) Corrected by Suitable Differentiating Circuits.

black field is scanned by a cathode-ray tube of the type used in a normal television receiver. The oscillogram exhibits the waveform along one scanning line crossing the white rectangle, and the picture shows the image reproduced on the receiver tube smeared by the afterglow effects. The smearing or integration of the picture waveform indicates that the circuits necessary for the correction of this distortion must take the form of differentiating circuits. These can be frequency dependent cathode feedback circuits [4] of the type shown in Fig. 7.

Generally, the screen-decay characteristic is not a single exponential, but consists of several components of different time constant; and the waveform is a superimposition of the individual decay components. Closer approximation to the screen material decay characteristic results if several compensating circuits of different time constant are used.

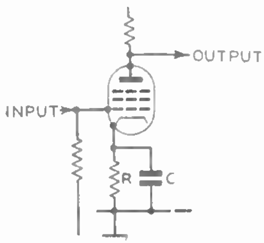


Fig. 7.—Example of a Circuit for Effecting Afterglow Correction for a Flying-spot Scanner.

Fig. 6 (*b*) shows that the slide can be faithfully reproduced when the correcting circuits are adjusted for optimum results.

Since the signal output from the flying-spot scanner is at all times proportional to the slide transparency, it follows that the signal presented to the viewing receiver will have unity gamma. Due to the "up-gammaing" action of the receiver cathode-ray tube, it follows that the overall gamma of the system would be greater than unity, resulting in the loss of details in the darker tones of the picture. For the transmission of half-tone pictures of optimum quality it follows that a gamma-correction stage in the video amplifier is necessary.

Applications

Flying-spot scanners of the type shown in Figs. 5 (*b*) and 5 (*c*) are particularly suitable for the transmission of documentary information; this can take many forms, such as diagrams of all kinds, microphotographic records, finger-prints, the checking of signatures in banks, etc.

A particular application could be the inspection of photographic negatives before having prints made, since it is a simple matter to reverse the polarity of the displayed picture.

THE PHOTOCONDUCTIVE TYPE CAMERA TUBE

Prior to the development of the photoconductive type of camera tube, i.e., the American "Vidicon"[6] and the British "Staticon", industrial television camera designers were restricted to using the highly sensitive Image Orthicon,[7] a tube from the Orthicon [8] or Image Iconoscope [9] family, or the very insensitive Farnsworth Image Dissector.[2] While the storage-type camera tubes mentioned are capable of excellent performances, they have a high initial cost and are large and bulky, especially when associated with their focus and deflection coils.

The photoconductive type camera tube shown in Fig. 8 provides an opportunity for the expansion of television in industry because of its comparatively low initial cost, operational simplicity and small dimensions, the tube being only 6½ in. long and 1 in. in diameter. The construction and mode of operation of the tube can be appreciated from Fig. 9.

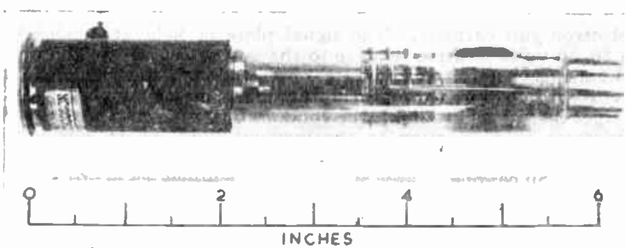


FIG. 8.—PHOTOGRAPH OF A PHOTOCONDUCTIVE TYPE OF CAMERA TUBE, THE STATICON.

(*Cathodeon Ltd.*)

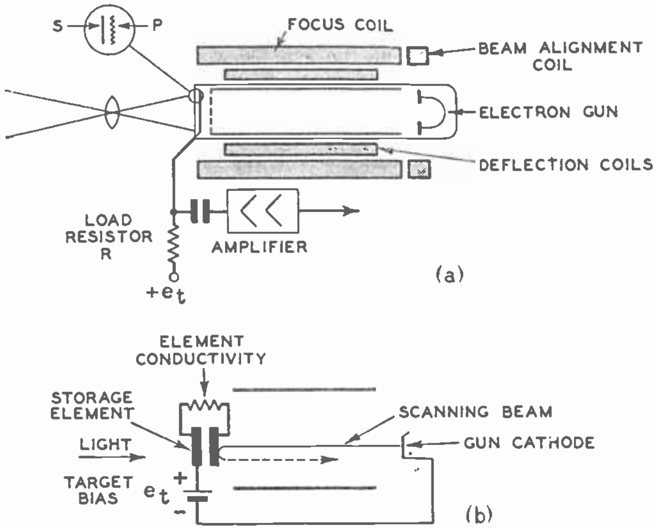


Fig. 9.—Functional Diagram of a Photoconductive Type of Camera Tube.

Principle

The storage target consists of a homogeneous layer of semi-conducting material (P) backed by a transparent signal plate (S). The transverse resistance of the semi-conductive layer is such that the storage target can be considered as a large number of discrete storage elements or capacitors, each with a common electrode—the signal plate. One of these storage elements is depicted by the sketch Fig. 9 (*b*). The scene to be televised is optically focused on the storage target. This is scanned orthogonally by a low-velocity beam of electrons emitted by the electron gun, and the target surface potential stabilized to that of the electron-gun cathode. The signal plate is held at a potential e_t some 20–30 volts positive relative to the gun cathode. In the absence of light the conductivity of the target material is so low that no appreciable current can flow through the target to the signal plate. Should the target be exposed to light, the conductivity of each elementary area will increase in proportion to the incident light level, causing these storage elements to behave as charged capacitors which have a leaky dielectric. Since one plate of these elementary capacitors has its potential fixed by the target bias e_t, the potential of the other plate will rise towards e_t by an amount depending on the light level falling on the element. During the following scanning cycle the scanning beam will deposit electrons on the discharged parts of the target, restoring the storage surface to cathode potential. This action will cause a current to flow in the signal-plate lead, developing a voltage across the load resistor R. This voltage constitutes the video signal. The polarity of the signal

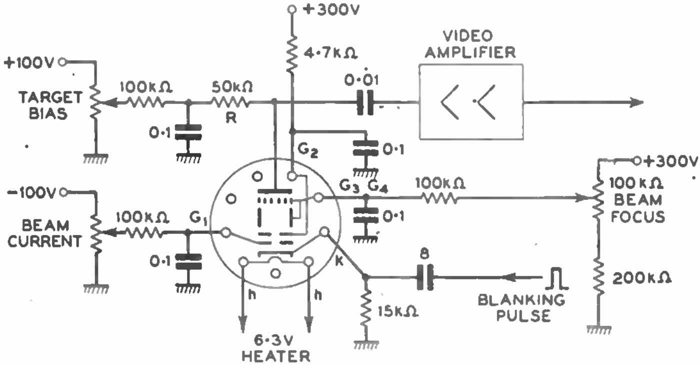


FIG. 10.—THE SUPPLIES NECESSARY TO OPERATE A VIDECON OR STATICON TYPE OF CAMERA TUBE.

is such that a negative swing corresponds to illuminated parts of the scene.

Fig. 10 indicates the supplies necessary to operate the Staticon or Vidicon. The electron beam is produced by the gun assembly, comprising the heated cathode k, control grid g, and the accelerator anode g_2. The beam is focused by the combined action of the magnetic field produced by the focus coil extending over the length of the tube, and the electric field from the cylinder anode and mesh g_3, g_4, the mesh being located across the end of g_3 and parallel to the surface of the target. The function of the mesh is to provide a uniform electric field over the target surface, which, in conjunction with the uniform magnetic field, ensures that the scanning beam approaches the target with normal incidence at all times, a condition essential for even signal generation over the storage surface.

Routine beam-focus adjustments are normally made by varying the voltage applied to the cylinder anode g_3 g_4, while keeping the magnetic field constant. The usual value of magnetic flux is 40 gauss measured in the centre of the coil.

The magnitude of the signal current is of the order of 0·1–0·2 μA for a peak white signal, and the sensitivity of the tube is such that satisfactory pictures can be obtained with wide aperture lenses, e.g., F1·9 with an incident scene illumination of the order of 50 foot-candles.

The target size is a format of 4 × 3 aspect ratio having a diagonal of 16 mm. allowing the standard range of 16-mm. ciné camera lenses to be used, while the colour response of the tube is similar to that of panchromatic film.

From the foregoing description of the manner of signal generation it will be obvious that it is necessary for the scanning-beam intensity to be sufficiently high to discharge fully the picture highlights. If this is not so, the highlight signal will appear "clipped" and devoid of details; also, should the scene content change, the highlights will not have been discharged by the scanning beam, and will therefore contribute a spurious signal.

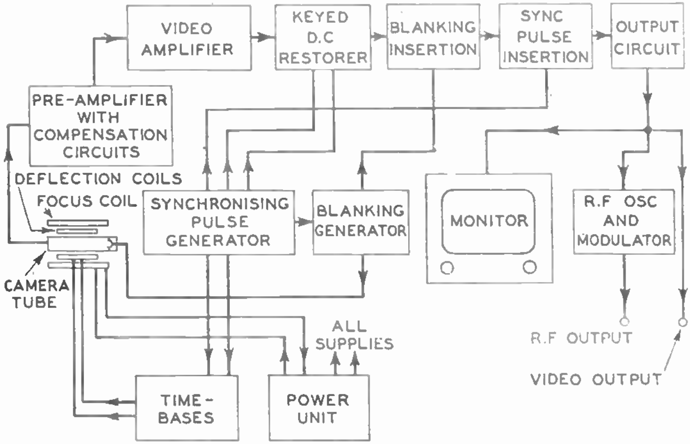


Fig. 11.—Block Schematic of a Television System Using a Photoconductive Type of Camera Tube.

TYPICAL PORTABLE EQUIPMENT

The small size of the photoconductive type pick-up tube enables portable industrial television equipment to be manufactured. Such equipment is available in several forms in the U.S.A.[3] and in Great Britain. A typical style of industrial equipment employs a complete television system in one package of about 16 × 6 × 4 in.; twenty-three valves are utilized to generate the picture, which may be on any of the 405-, 525- or 625-line systems, including a modulated radio-frequency outlet. This is shown in action in Fig. 19 (*a*).

The circuits necessary to control a Staticon or Vidicon type pick-up tube and "shape" the television signal are indicated in the block schematic Fig. 11.

Video Amplifier

Considering the design of the video amplifier and the band-width required to do full justice to the resolving power of the pick-up tube—as these tubes can resolve 800 vertical black and white bars across the picture—it follows that, for a 405-line system, it is necessary to aim at a video band-width of about 5 Mc/s. Since a typical value of signal output from the pick-up tube is 0·1 μA, the signal voltage developed across the load resistor (50 kΩ in Fig. 10) will be of the order of 5 mV. Due to the capacitance which inherently exists in parallel to this resistor, this value of signal voltage will apply to the low-frequency components only. Should this shunt capacitance be, say, 20 pF, at 5 Mc/s the input signal would be developed across about 1600 ohms, making the signal input to the amplifier about one-thirtieth of the low-frequency value. Suitable differentiating circuits must be incorporated in the amplifier to effect the necessary compensation of this integrated input signal. This can be done by cathode-feedback circuits of similar time constant to the input (integrating) circuit, or by feedback techniques as described in recent papers.[10, 11]

The signal-to-noise ratio of the video signal generated by a Vidicon or Staticon is determined by the noise threshold of the preamplifier, and in particular by the first amplifier valve. For best results, therefore, valves having a low figure of equivalent noise resistance should be chosen. This will enable good signal-to-noise ratios to be maintained while operating under low, light-level conditions.

Timing and Scanning Circuits

The synchronizing pulse generator can be built round any of the basic circuits discussed earlier, while the time-bases can be of conventional design.

Blanking Generator

The blanking generator must produce pulses at line and frame frequencies suitable for suppressing the scanning beam of the pick-up tube during the retrace times. The pulses may be applied to the cathode of the electron gun (see Fig. 10) and should be about 25 volts, positive-going.

D.C. Component

It will be apparent from the mode of signal generation that black in the picture will correspond to zero signal current: this, of course, happens also in the intervals when the scanning beam is suppressed; thus the signal from the tube contains a black-level reference period. By employing a " keyed clamp " circuit in the amplifier, operating only during the line-blanking periods, the D.C. component of the signal can be restored. Blanking and synchronizing pulses can then be mixed with the video signal to complete the composite waveform.

Signal Distribution

To enable domestic television receivers to be employed as picture monitors without modification, it is necessary to have available a signal output in the form of amplitude-modulated radio frequency. This can be provided by a simple two-valve unit consisting of an oscillator covering the range 41–68 Mc/s and a modulator stage capable of feeding a 75-ohm line with about 100 mV of signal. Several receivers can be simultaneously used on such a signal, using a simple resistive distribution network.

In practice, the oscillator frequency is adjusted to one of the B.B.C. television frequencies other than that of the local transmission.

Where long distances—of the order of some thousand feet—separate the camera and monitor, it is an advantage when distributing the signal by transmission line to do so at radio frequency. Should a system of video distribution be used, the higher-frequency components of the signal will be attenuated by the transmission line, necessitating frequency-correction at the received end. The radio-frequency signal is merely attenuated by the transmission line, and due to the large signal available from the camera, in many cases no further amplification is necessary.

Since many industrial television installations comprise several cameras and one viewing point, a further advantage can be gained by radio-frequency distribution. By assigning to each camera a different channel frequency and feeding the camera outputs through a simple combining circuit into a tunable television receiver, the output of any camera can be selected by merely operating the receiver-channel switch.

THE IMAGE CONVERTER

Principle

When an optical image is thrown on to the photo layer of an image converter a corresponding image is caused to appear on a fluorescent screen in the tube.

Photosensitive materials can respond to electromagnetic radiations having wavelengths between 2,000 Å in the ultra-violet [19] and 12,000 Å in the infra-red.[20] Since the response of the human eye is confined to a spectrum bounded by 4,000 and 6,500 Å, it is possible to extend the range of vision by using an image-converter tube with a suitable photo-surface. The relative response to equal amounts of electromagnetic energy of the human eye and two different photosensitive surfaces [12] is shown in Fig. 12. The photosensitive surfaces form photoemissive layers which can be incorporated in simple photocells or image converters. As can be seen, the antimony/cæsium layer has greater response to the shorter-wavelength radiations (violet and ultra-violet), whilst the cæsium/silver/silver-oxide layer has greater response in red and infra-red regions of the spectrum.

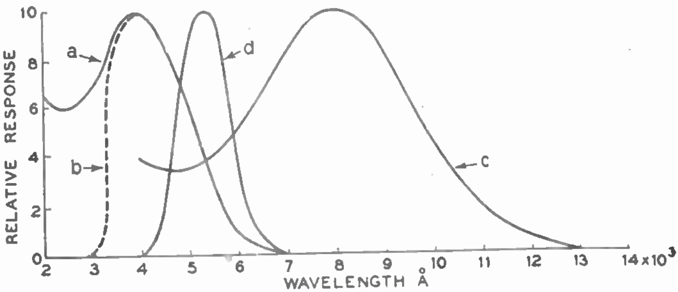


FIG. 12.—RELATIVE RESPONSE TO EQUAL AMOUNTS OF ELECTROMAGNETIC ENERGY OF THE HUMAN EYE AND VARIOUS PHOTOELECTRIC SURFACES.

(*a*) Caesium-antimony surface, with quartz windows and mosaic dielectric.
(*b*) Caesium-antimony surface, with normal construction, showing absorbent effect of glass window.
(*c*) Caesium/silver/silver oxide surface.
(*d*) Human eye.

(From Williams, Industrial and Professional Applications of Television Technique, Proc. I.E.E., Pt. 111A., Vol. 99, p. 651.)

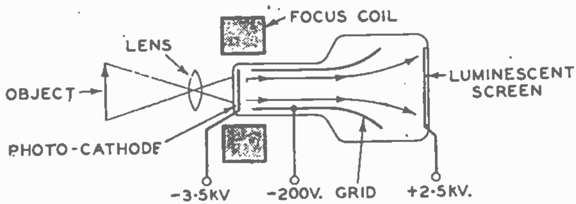


FIG. 13.—FUNCTIONAL DIAGRAM OF AN IMAGE CONVERTER.

Construction

A typical image converter tube is shown in outline in Fig. 13. The photocathode is formed on the inside surface of the end-plate of the glass bulb, the photocathode being semi-transparent to allow the electron emission to occur from the inner surface of the photo-electric material. The opposite end-plate carries a fluorescent screen which is bombarded by the electrons emitted from the photocathode, since the potential of the photocathode is several kilovolts negative relative to that of the fluorescent screen. The object to be viewed is optically focused on the plane of the photocathode, and the resulting electron emission focused on the fluorescent screen by the combined action of an electric field and a magnetic field set up by the focus coil surrounding the tube. In order to protect the phosphor from poisoning by the material forming the photolayer during processing and to render the tube opaque to light penetrating the photocathode, it is common practice to provide the luminescent screen with a backing layer of, say, aluminium.

In the sketch of the tube in Fig. 13 the electron paths from photosurface to luminescent screen are suggested, showing that there is an image magnification, the magnification depending on the geometry of the tube and the electron optical operating conditions.

Referring to Fig. 12, it is interesting to note the transmission loss of the shorter wavelengths through normal glass surfaces; for this reason image converters designed to operate in the ultra-violet region have the photosurface formed on a quartz plate (see curve (*a*)).

Applications

Much work was carried out during the 1939–45 war on the use of image converters to extend vision, one outcome of this being the infrared telescope for night observation.

The image converter has opened new fields in the observation of high-speed phenomena.[13,14] With a conventional camera shutter the upper limit of shutter speed is approximately $\frac{1}{1250}$ second. The image converter operated as a high-speed shutter enables extremely short photographic exposures to be made. These exposures can be of the order of 10^{-8} second. To operate the image converter as a high-speed shutter, the voltage applied to the grid electrode, shown in the diagram Fig. 13, can take the form of a positive pulse allowing the electrons from the photolayer to reach the fluorescent screen only while the grid is positive.

Pulsing the grid electrode is preferable to switching the voltage applied to the fluorescent screen, since a much lower pulse voltage will perform the switching action. The converter will therefore operate as a closed–open–closed shutter, the object to be photographed being reproduced as a fluorescent image for the duration of the switching pulse. As the pulse-repetition rate is controllable, it enables the image converter to be used as a photographic stroboscope for the recording of high-speed phenomena.

APPLICATIONS OF INDUSTRIAL TELEVISION

Microscopy

Recent investigations [15, 16] have shown that great benefits can result from the application of television techniques to the field of microscopy. Viewing objects through an optical microscope in the conventional manner is fatiguing and has the limitation that only one person at a time can view the specimen; this can be a serious drawback in educational establishments.

Viewing the specimen by a "Television Microscope", besides removing these drawbacks, has several important advantages over the conventional microscope. These can be summarized as follows:

(*a*) The resulting picture can be presented in any convenient position and may also be of any desired size.

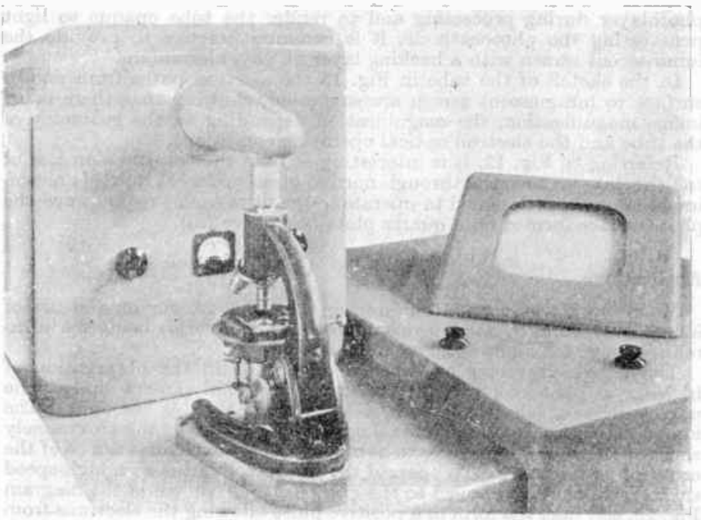


FIG. 14.—FLYING-SPOT MICROSCOPE.

(*Cinema Television Ltd.*)

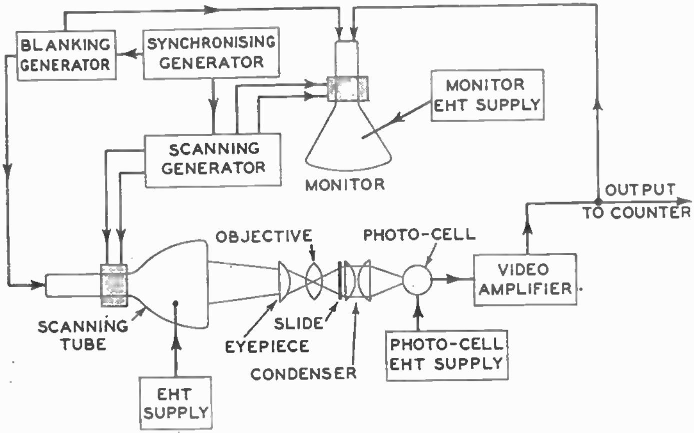


FIG. 15.—BLOCK DIAGRAM OF A FLYING-SPOT MICROSCOPE.

(*b*) Since the brightness of the viewed image is no longer determined by the specimen illumination, there is less chance of damaging the specimen by burning.

(*c*) The contrast of the viewed picture is no longer wholly dependent upon the specimen contrast.

(*d*) Should the specimen consist of a quantity of discrete particles, the number of these can be automatically ascertained with a high degree of accuracy in a very short time.

There are two ways in which microscopic objects can be televised: the first employs the flying-spot principle, being known, therefore, as the Flying-spot Microscope.[15] Such an instrument is illustrated by the photograph in Fig. 14 and the block diagram, Fig. 15.

As will be seen, the light spot on the scanning cathode-ray-tube screen is imaged by the optical system, so that the specimen mounted on the slide is scanned by a light spot of minute dimensions, the optical system consisting of a compound microscope.

The instrument operates on the normal British television standards, so that the specimen is scanned by a raster of 405 interlaced lines. Also included is a non-linear amplifier, to enable the overall "gamma" of the system to be adjusted, "gamma" being the term denoting the exponent of the contrast law of the system. This feature can be used to advantage when viewing objects of low contrast, and may render unnecessary the specimen-staining technique sometimes adopted to increase contrast artificially.

The other way in which microscopic objects can be televised is to "look" through the eyepiece of an optical microscope with a television camera. Such a scheme is shown in Fig. 16, where an industrial television equipment employing a photoconductive type pick-up tube is used. The resulting picture is displayed on the monitor alongside the microscope-camera set-up.

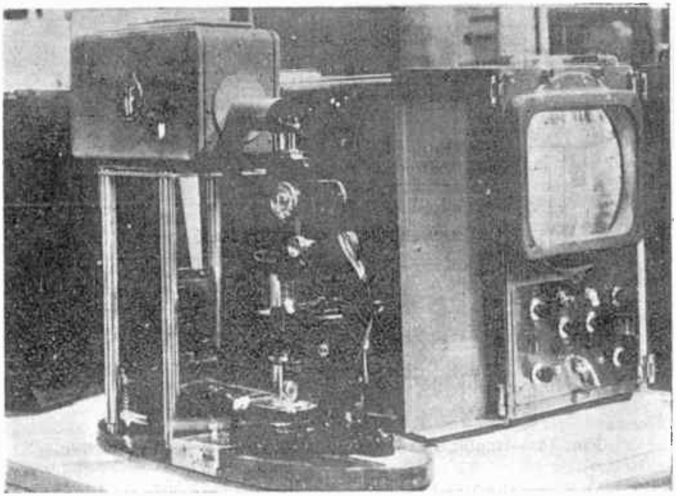


FIG. 16.—A TELEVISION MICROSCOPE.
(Pye Ltd.)

Increased resolving power results from irradiating the specimen with ultra-violet instead of visible light, as resolving power is a function of the irradiating wavelength; this can be taken advantage of in normal microscopy, but recourse must be made to photography in order to view the results. Since the phosphor used in the flying spot microscope is a wide-band emitter with radiations in the ultra-violet, and the Staticon or Vidicon [16] pick-up tube used in the television microscope can be rendered sensitive to these shorter wavelengths, it follows that television technique provides a continuously viewable microscope of increased resolving power.

Specimen irradiation by ultra-violet light has further benefits in the study of biological materials, since many of these materials, such as tissue, which may appear quite colourless to visible light, exhibit selective absorption in the ultra-violet spectrum. Pictures of good contrast can therefore be obtained by proper selection of the irradiating wavelength.

In extending this principle, a recent paper [16] has shown that by the adoption of colour-television technique it is possible to display "colour" pictures of the specimen, where each colour corresponds to the periodic exposure of the specimen to irradiation by a selected wavelength.

Perhaps the greatest contribution made by television to quantitative microscopy is the facility by which rapid and accurate counts can be made of the number of discrete particles forming a specimen. Circuits [15,17] have been developed which record the number of particles in a given field, regardless of the particle shape, size or position.

The photographs comprising Fig. 17 were taken from the monitor

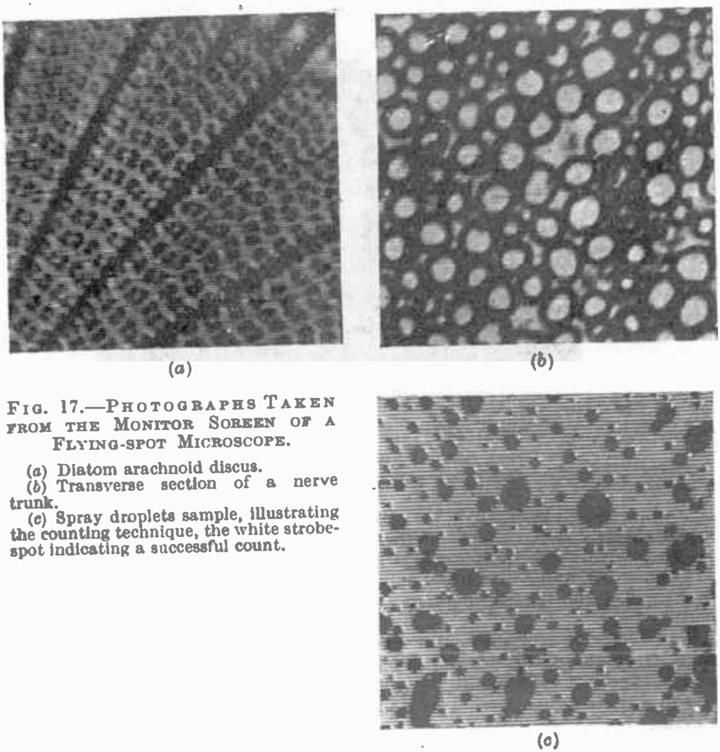


FIG. 17.—PHOTOGRAPHS TAKEN FROM THE MONITOR SCREEN OF A FLYING-SPOT MICROSCOPE.

(a) Diatom arachnoid discus.
(b) Transverse section of a nerve trunk.
(c) Spray droplets sample, illustrating the counting technique, the white strobe-spot indicating a successful count.

screen of the Flying-spot microscope, and illustrate the fine results obtained from this instrument; Fig. 17 (c) is worthy of special note, since it illustrates clearly the value of television and electronic techniques in particle counting.

Underwater Television

Underwater television, brought into prominence by the location of the sunken submarine *Affray* in 1951, and later the salvaging operations on the Comet aircraft, has proved to be an important new aspect of television.

With accurate visual information available on the picture monitor in the salvage ship, it is possible for expert direction to be given to the diver, who may be operating under severe mental stress, and who may have to handle complex equipment. In cases where careful exploration of the sea floor is required, the underwater camera functions as a means of rapid identification of objects located by Asdic or echo-sounding apparatus.

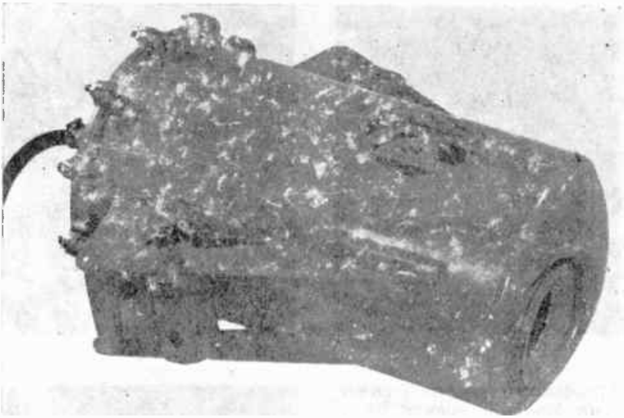


FIG. 18.—DIVER-HELD UNDERWATER TELEVISION CAMERA.
(*Pye Ltd.*)

Equipment to operate under water has generally been designed round the highly sensitive image orthicon camera tube, which can generate pictures with scene illuminations lower than 1 foot-candle.

Deep-sea television cameras have been constructed to operate at depths of 1,200 ft., the camera casing carrying the necessary lighting equipment.

Fig. 18 shows an underwater television camera intended for operation by a diver at depths down to 350 ft. Optical focus, lens-iris setting, and lens selection is remotely controlled from the surface, the supplies to the camera being carried by a multi-core cable heavily sheathed with P.V.C.

Other underwater television applications include marine biological research work and inspection of docks works and the hulls of ships.

Electronic Camera Method of Film Making

An interesting new application of television is its use in the manufacture of films [1]; the electronic camera method of film making being a complete breakaway from conventional film-making technique.

The essence of the system is the replacement of the film camera by a group of electronic cameras televising the scene to be " filmed " and the recording on film of the selected television picture. This places the film director in a position similar to that of the television producer, with the facilities of cutting and mixing from one camera shot to another while the scene is being enacted, and being able to see and control at all times the final artistic visual effect, as against being restricted to glimpses through a camera viewfinder before the commencement of shooting. Films made by this method can therefore be completed quickly and economically in sequences of any desired length instead of in brief, sometimes disjointed scenes which require subsequent editing.

Obviously the television system used for this mode of film making must have an extremely high resolving power and be above criticism in geometry and scanning linearities. A sequential scanning system is used of about 1,000–1,300 lines per frame and a video band-width of 15–20 Mc/s. The recording-film camera operates on an intermittent motion system, the film pull-down occurring during the television frame-suppression period.

Department Stores Aid

Compact, "easy-to-operate" industrial television equipments have proved to be a valuable sales aid in large department stores. Particular items of merchandise can be featured on television screens mounted in eye-catching positions; the television department can use the pictures as an alternative programme source for the receivers in showroom, making at least one television programme available during shop hours.

Furnace Control

Television has been successfully applied to the control of furnace burners in a power station in the U.S.A. At this power station the controls for the steam-producing plant and the generating sets are located in a central control room. Small television cameras looking into the combustion chambers of the oil or coal-dust-fired boilers enable the station engineers to continuously check the combustion conditions. In this type of boiler the oil or coal-dust is fed into the rectangular combustion chambers by four corner burners so arranged that a swirl of flame is produced. Since the position and symmetry of this flame swirl is critical for optimum boiler efficiency, continuous observation of the combustion chamber allows adjustments to the burners to be made as necessary, resulting in efficient boiler operation.

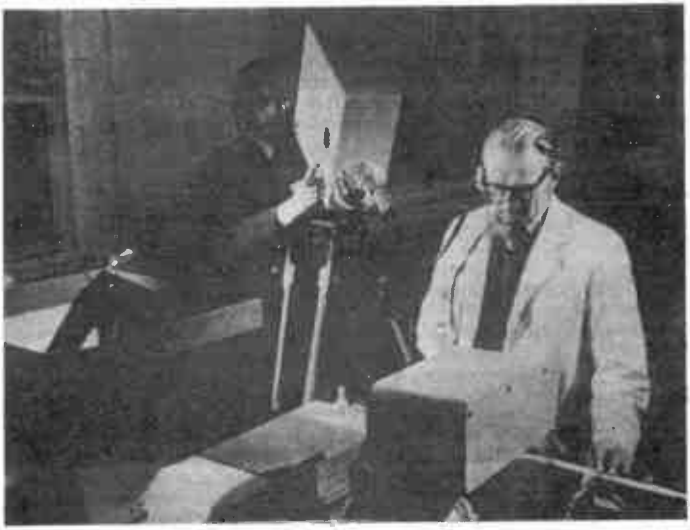
Security Aid

Television as a security aid is an aspect of industrial television as yet undeveloped, but if thought of as a means of instantaneously conveying visual information from police force to police force, there can be little doubt as to the future of television in this sphere.

Educational Aid

The potential contribution of television to education is immense. With it, the problem of teaching and demonstrating adequately, intricate mechanisms, surgical operations and the like, to large groups of students, is diminished considerably.

The training of surgeons demands attendance at many operations watching the master surgeon's technique, and it is difficult for one student, let alone several, to obtain an unobstructed view of the proceedings. It is here that a television camera, mounted, for example, in the lighting system [18] of the operating-table obtains a view of the operation equal to that of the surgeon. The pictures obtained (which could be accompanied by a commentary spoken by the surgeon) can



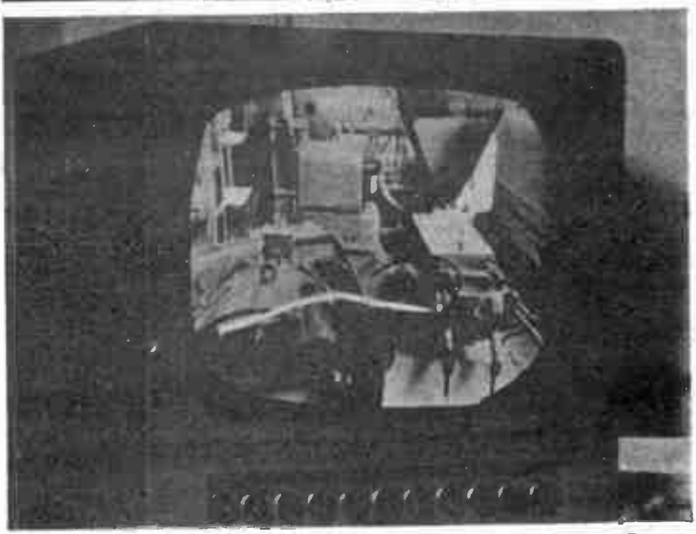


Fig. 19 (*a, top*).—Industrial Television Camera, Televising a Demonstrator; (*b, below*) Close-up of the Picture Displayed in the Lecture Theatre. (*Pye Ltd.*)

be relayed to lecture rooms capable of accommodating the student audience comfortably.

While monochrome television is useful in this application, colour television can play an even more important role, enabling the anatomical details to be more easily distinguished.

Fig. 19 (*a*) and (*b*) are photographs taken at the National College of Rubber Technology, London, where closed-circuit television is used to link the machine's laboratory (Fig. 20 (*a*)) in the basement of the building and the lecture theatre, which is at the top of the building. This enables the lectures to be illustrated by demonstrations of machine-handling technique.

The photographs show a demonstrator explaining the operation of a plastic extruding machine, whilst the students watch the demonstration on a large-screen monitor (Fig. 20 (*c*)) in the lecture theatre. A two-way sound system allows the lecturer to control the demonstration, for example, requesting particular items to be brought into close-up for specific discussion.

Scientific Research Aid

In the field of scientific research television can become an indispensable tool. To be able to view easily dangerous operations may mean that greater control can be exercised over the process, in turn leading to greater safety. Efficient remote observation can be a problem in the handling of radioactive materials where the technician has to manipulate these materials by a system of tongs from behind a protective barrier.

In many cases the use of optical viewers is not possible and television provides the only solution.

* * *

Only a few applications of television to industrial problems have been cited here, but they may serve to indicate the countless ways in which television will be applied to industry in future years.

H. A. McG.

Figs. 5 and 6 originally appeared in the Journal of Brit. Inst. of Radio Engineers, *see reference 2 below.*

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General

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46. UNITS AND SYMBOLS

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UNITS

Units are required for mechanical, physical, electrical, thermal and similar quantities to permit of recording or comparing their magnitudes. Each unit must be represented by some kind of natural or artificial standard, or must be capable of derivation from a combination of such standards.

Metric System

This is based on prototype standards, the *metre* and the *kilogram*, of length and mass respectively. The originals are preserved at the International Bureau of Weights and Measures, at Sèvres. Copies are available in certain National Laboratories. The metric system derives its units in decimal steps from the basic units, the *metre* and the *gram*. The names are formed by prefixes indicating the power of 10.

The third fundamental mechanical unit is the *second* of time, defined as $\frac{1}{86400}$ of the mean solar day.

Absolute Systems

Any system that includes length, mass and time in its fundamental units is called an absolute system, and the units are *absolute units*.

Centimetre–Gram–Second System

This is an absolute system using the units named, with proportionality factors in equations for physical relationships made, as far as possible, unity. It is the commonest physical system and is referred to by the letters c.g.s.

Metre–Kilogram–Second System

This is also an absolute system using the named units. The magnitudes of many of the derived units are more convenient from the engineering standpoint than for the c.g.s. system. It was adopted for use as a basis for the electrical units by the International Electrotechnical Commission in 1935. It is referred to briefly as the m.k.s.μ_0 system.

The British National Committee of the International Electrotechnical Commission has recently issued B.S. 1637 : 1950, entitled "Memorandum on the M.K.S. System of Electrical and Magnetic Units", which now forms a British Standard defining the system. This standard also sets out the facts which were taken into account by the I.E.C. in reaching their decisions relating to the adoption of this system. The rationalized m.k.s. system was finally adopted internationally in 1951.

Derived Mechanical Units

See Table 1. The c.g.s. unit of force is the *dyne*, which gives the mass of one gram the acceleration of one centimetre per second per second. The c.g.s. unit of work or energy is the *erg*, being the work done when a force of one dyne acts over a distance of one centimetre.

The corresponding m.k.s.μ_0 units are the *newton* (force) and *joule* (work) or energy. The definitions utilize the metre and the kilogram instead of the centimetre and the gram. Thus 1 newton $= 10^5$ dynes, and 1 joule $= 10^7$ ergs. The newton is more easily recognized in the form: 1 newton $=$ 1 joule per metre.

Electrical Units

There are several sets of units in use, the differences lying chiefly in the choice of constants relating electrical and magnetic quantities both to each other and to mechanical quantities.

Absolute units are obtained from the basic mechanical units, with one additional unit representing the magnetic or electric properties of space. The three absolute systems are:

(i) the *absolute electromagnetic c.g.s.* system, with the addition of unity for the magnetic space constant;
(ii) the *absolute electrostatic c.g.s.* system, with the addition of unity for the electric space constant;
(iii) the *absolute m.k.s.* system, with the addition of $\mu_0 = 4\pi/10^7$ for the magnetic space constant.

Both (i) and (ii) have been in common use in electrophysics for many years. The magnitudes of the e.m. system (i) are inconvenient, and a "practical" set of units having a metric relation has been preferred by engineers. The "practical" units have the familiar names of ampere, coulomb, volt, etc., and it has become customary to refer to their e.m. originals under the names abampere, abcoulomb, abvolt, etc. The magnitudes of the e.s. units (ii) are also inconvenient, but their relation to the "practical" units is not metric: it involves powers of c, the velocity of light (3×10^8 m. per sec.) as well as powers of 10. The names of the e.s. units are sometimes taken as those of the corresponding "practical" units with the prefix *stat*, as statampere, statcoulomb, statvolt, etc.

The m.k.s.μ_0 system has the advantage that the magnitudes of its electrical units are precisely those of the "practical" system. Unfortunately, engineers have been accustomed to mixing up the e.m. magnetic units with the practical units, and the m.k.s.μ_0 system requires a new set of magnetic unit quantities; however, these are related to the e.m. units by powers of 10. The unit of force, the *newton* $= 10^5$ dynes, is also unfamiliar, although its magnitude is far more convenient than that of the dyne. (Taking $g = 9{\cdot}81$ metres/second/second, 1 newton $= 10^5$ dynes $= 0{\cdot}102$ kg. force $= 0{\cdot}225$ lb. force.) The practical, m.k.s. and c.g.s. electric units have the relations listed in Table 2.

International Units

The definition of an absolute unit is one matter: its realization is another. The international Electrical Units (1893) were intended for actual realization in physical terms, so that the set-up of a specified arrangement of apparatus, employed under specified conditions, would permit of the measurement or comparison of electrical quantities.

INTERNATIONAL OHM: the resistance offered to an unvarying electric current by a column of mercury at the temperature of melting ice, 14·4521 g. in mass, of uniform cross-sectional area, and 106·300 cm. in length.

INTERNATIONAL AMPERE: the unvarying electric current which, when passed through a standard specified solution of nitrate of silver in water, deposits silver at the rate of 0·001118 g./second.

INTERNATIONAL VOLT: defined from the International Ohm and Ampere. It is that unvarying potential difference which produces a current of one International Ampere in one International Ohm.

The International Coulomb, Farad, Henry, Joule and Watt follow from the Ampere, Volt and Ohm. The measurements that have to be made to realize the International Units are absolute: i.e., for the Ohm, length and mass; for the Ampere, time and mass.

Improvements in the accuracy of measurements in absolute terms have shown that the International Units differ from absolute units. The following figures give the divergencies in parts per 100,000:

Ohm and Henry, 49 high; Ampere and Coulomb, 15 low; Volt, 34 high: Watt and Joule, 19 high; Farad, 49 low.

The International Units have been superseded, and electrical quantities are determined by absolute measurements made in terms of length, mass, time, and the assigned absolute permittivity of free space. Absolute current is measured by a form of current balance, the coils of which are wound on formers of permanent material so that their dimensions are measurable to great accuracy; the current is then found in terms of mutual inductance and length. Mutual inductance is measured in terms of the dimensions of a coil assembly; resistance by means of the rotation of a conducting disc in terms of mutual inductance.

For measurements that require to be precise but not absolute, the Weston Normal Standard cell is employed. Regularized methods of preparation and assembly are such that individual cells differ from each other by no more than about 20 μV from one standard e.m.f. of 1·01860 V absolute at 20° C. Together with standard resistors as shunts and multipliers, a full range of derived measurements (e.g. current, power, energy, etc.) can be made by potentiometer.

Physical Bases

The e.s. system of units was based on the observable mechanical force exerted between electric charges, and the variation of this force inversely as the square of the distance, so that the mechanical force f between charges q_1 and q_2 separated by distance d cm. was

$$f = q_1q_2/\kappa_0 d^2 \text{ dynes}$$

in free space. The value κ_0 was assumed to be unity, and the e.s. unit of charge, the *statcoulomb*, is that concentrated point charge which, at unit distance (1 cm.), exerts unit force (1 dyne) on a similar charge in free space.

The e.m. system was based on the highly artificial concept of the unit magnetic pole, which was defined on lines similar to those employed in the e.s. system, using the observable mechanical force f on magnetic point poles:

$$f = m_1 m_2/\mu_0 d^2 \text{ dynes.}$$

The value μ_0 was assumed to be unity. The two aspects of electromagnetic phenomena lead to units of different sizes for the same electrical quantity.

In the m.k.s.μ_0 system the electrostatic force is measured in newtons (= 10^5 dynes), the charges are in coulombs and the distances in metres: κ_0 is taken as $1/36\pi \times 10^9$, and μ_0 as $4\pi/10^7$ (since $\mu_0\kappa_0 = 1/c^2 = 1/9 \times 10^{16}$ approx., with c = velocity of electromagnetic propagation in free space = 3×10^8 metres/second approx.).

Magnetic phenomena can be described in terms of circuital currents, so that there is no need for a m.k.s.μ_0 unit pole. It is, however, necessary to define magnetic flux in order to deal with the magnetic field phenomena.

Definitions of Units

PRACTICAL UNITS: Units which have been adopted for practical use owing to the c.g.s. units being in many cases inconveniently large or small; each is a decimal multiple or sub-multiple of the corresponding c.g.s. unit. The practical units are the same as the units in the m.k.s.μ_0 system.

DYNE: The c.g.s. unit of force. It is that force which, acting on a mass of one gram, gives to it an acceleration of one centimetre per second per second.

NEWTON: The m.k.s.μ_0 unit of force. It is that force which, acting on the mass of one kilogram, gives to it an acceleration of one metre per second per second.

CALORIE: (cal). A unit of heat. It is the quantity of heat required to raise the temperature of one gram of water, at 15° C., by one degree Centigrade. A calorie is approximately equivalent to 4·18 joules.

KILO-CALORIE: (k.-cal.). A unit of heat equal to 1,000 calories.

BRITISH THERMAL UNIT: (B.Th.U.). A unit of heat. It is the quantity of heat required to raise the temperature of one pound of water from 60° to 61° F. It is approximately equivalent to 1,054 joules. A *mean British Thermal Unit* is $\frac{1}{180}$ part of the quantity of heat required to raise the temperature of one pound of water from 32° to 212° F. It is approximately equivalent to 1,055 joules.

AMPERE: (A). The practical unit of electric current. An ampere = 0·1 e.m. unit = 3×10^9 e.s. units. For the definition of the *International Ampere*, see page 4.

COULOMB: (C). The practical unit of quantity of electricity. One coulomb = 0·1 e.m. unit = 3×10^9 e.s. units.

OHM: (Ω). The practical unit of resistance. The true ohm = 10^9 e.m. units = $1/9 \times 10^{11}$ e.s. unit. For the definition of the *International Ampere*, see page 4.

VOLT: (V). The practical unit of e.m.f. and p.d. A volt = 10^8 e.m. units = 1/300 e.s. unit. For the definition of the *International Volt*, see page 4.

WATT: (W). The practical unit of power. It is the amount of energy expended per second by an unvarying current of one ampere under a voltage of one volt. A watt = 1 joule per second = 1 Newton-metre per second = 10^7 ergs per second.

Table 1.—Fundamental and Mechanical Units

Quantity	*Symbol*	*Defining Equation*	*British Units*	*Metric Units (c.g.s.)*	*Metric Units (m.k.s.)*	*Conversion*
Length	l	Fundamental	Foot (ft.) Inch (in.)	Centimetre (cm.)	Metre (m.)	1 ft. = 30·48 cm.; 1 in. = 2·54 cm.; 1 m. = 100 cm.
Mass	m	Fundamental	Pound (lb.)	Gram (g.)	Kilogram (kg.)	1 lb. = 453·6 g. = 0·4536 kg.; 1 kg. = 2·205 lb.
Time	t	Fundamental	Minute (min.) Second (sec.)	Second (sec.)	Second (sec.)	1 min. = 60 sec.
Area	a	$a = l_1 l_2$	Square foot (sq. ft.) Square inch (sq. in.)	Square centimetre (cm.²)	Square metre (m.²)	1 sq. ft. = 929·03 cm.² = 0·0929 m.² 1 sq. in. = 6·45 cm.²; 1 m.² = 10,000 cm.²
Volume	U	$U = l_1 l_2 l_3$	Cubic foot (cu. ft.) Cubic inch (cu. in.)	Cubic centimetre (cm.³)	Cubic metre (m.³)	1 cu. ft. = 28,317 cm.³ = 0·02832 m.³ 1 cu. in. = 16·39 cm.³
Angle	α	$\alpha = l_1/l_2$	Degree (°) Minute (′) Second (″)	Radian (rad.) Degree (°) Minute (′) Second (″)	Radian (rad.) Degree (°) Minute (′) Second (″)	1 rad. = 57° 17′ 45″ = 57·296° 1° = 60′ 1′ = 60″
Velocity (linear)	u	$u = \frac{dl}{dt}$	Feet per minute Feet per second	Centimetres per second (cm. per sec.)	Metres per second (m. per sec.)	1 ft. per min. = 0·508 cm. per sec. 1 ft. per sec. = 30·48 cm. per sec. = 0·3048 m. per sec.
Velocity (angular)	ω	$\omega = \frac{d\alpha}{dt}$	Revolutions per second	Radians per second (rad. per sec.)	Radians per second (rad. per sec.)	1 rev. per sec. = 2π rad. per sec.
Acceleration	a	$a = \frac{du}{dt}$	Feet per second per second	Centimetres per second per second (cm. per sec. per sec.)	Metres per second per second (m. per sec. per sec.)	1 ft. per sec. per sec. = 30·48 cm. per sec. per sec. = 0·3048 m. per sec. per sec.
Force	f	$f = ma$	Pound force	Dyne	Newton or Joule per metre	1 lb. force = 4·448 × 10⁵ dynes = 4·448 newtons 1 newton = 10⁵ dynes 1 lb.-ft. = 1·357 × 10⁷ dyne-cm. = 1·357 newton-m.
Torque	M	$M = \frac{W}{\alpha}$	Pound-foot	Dyne-centimetre	Newton-metre	
Energy	W	$W = fut$	Foot-pound	Erg or centimetre-dyne	Joule or Metre-newton	1 J = 10⁷ ergs; 1 ft.-lb. = 1·357 × 10⁷ ergs = 1·357 J.
Power	P	$P = fa$ $= W/t$	Foot-pound per minute	Erg per sec.	Joule per sec. or Watt (J. per sec., W.)	1 ft.-lb. per min. = 0·0226 W; 1 W. = 10⁷ ergs per sec.
Temperature	θ	—	Degree Fahrenheit (° F.)	Degree Centigrade (° C.)	Degree Centigrade (° C.)	1° F. = 5/9° C.; 0° C. = 32° F. on scale

$g = 980.241$ cm/sec^2 @ Cleveland
$g = 979.96$ cm/sec^2 @ San Francisco

TABLE 2.—RELATIONS BETWEEN PRACTICAL AND C.G.S. UNITS

Quantity	*Symbol*	*Practical and m.k.s.μ_0*	*C.g.s. Electromagnetic*	*C.g.s. Electrostatic*
Electromotive force and potential difference	E, V	VOLT: 1 V = 10^8 abvolts 1 V = 1/300 statvolt	1 abvolt = 10^{-8} V 1 abvolt = $1/3 . 10^{10}$ statvolt	1 statvolt = 300 V 1 statvolt = $3 . 10^{10}$ abvolts
Resistance	R	OHM: 1 Ω = 10^9 abohms 1 Ω = $1/9 . 10^{11}$ statohm	1 abohm = 10^{-9} Ω 1 abohm = $1/9 . 10^{20}$ statohm	1 statohm = $9 . 10^{11}$ Ω 1 statohm = $9 . 10^{20}$ abohms
Current	I	AMPERE: 1 A = 10^{-1} abampere 1 A = $3 . 10^9$ statamperes	1 abampere = 10 A 1 abampere = $3 . 10^{10}$ statamperes	1 statampere = $1/3 . 10^9$ A 1 statampere = $1/3 . 10^{10}$ abampere
Charge or quantity	Q	COULOMB: 1 C = 10^{-1} abcoulomb 1 C = $3 . 10^9$ statcoulombs	1 abcoulomb = 10 C 1 abcoulomb = $3 . 10^{10}$ statcoulomb	1 statcoulomb = $1/3 . 10^9$ C 1 statcoulomb = $1/3 . 10^{10}$ abcoulomb
Capacitance	C	FARAD: 1 F = 10^{-9} abfarad 1 F = $9 . 10^{11}$ statfarad	1 abfarad = 10^9 F 1 abfarad = $9 . 10^{20}$ statfarads	1 statfarad = $1/9 . 10^{11}$ F 1 statfarard = $1/9 . 10^{20}$ abfarad
Inductance	L	HENRY: 1 H = 10^9 abhenrys 1 H = $1/9 . 10^{11}$ stathenry	1 abhenry = 10^{-9} H 1 abhenry = $1/9 . 10^{20}$ stathenry	1 stathenry = $9 . 10^{11}$ H 1 stathenry = $9 . 10^{20}$ abhenrys
Energy	W	JOULE: 1 J = 10^7 ergs	1 erg = 10^{-7} J	1 erg = 10^{-7} J
Power	P	WATT: 1 W = 10^7 ergs per sec.	1 erg per sec. = 10^{-7} W	1 erg per sec. = 10^{-7} W
Magnetic flux	Φ	WEBER: 1 Wb = 10^8 maxwells	1 maxwell = 10^{-8} Wb	
Magnetic flux density	B	WEBER PER SQ. METER: 1 Wb./m.^{2} = 10^4 gauss	1 gauss = 10^{-4} Wb./m.	

Horse-power: (H.P.). A practical mechanical unit of power. The British horse-power is equal to 33,000 foot-pounds per minute or approximately 746 watts.

Erg: The c.g.s. unit of energy. It is the energy expended when a force of one dyne is exerted through a distance of one centimetre. An erg = 10^{-7} joule.

Joule: (J) The practical unit of energy. It is the energy expended when a force of one newton is exerted through a distance of one metre. A joule = 10^{7} ergs.

Watt-hour: (Wh.). A unit of energy. It is the energy expended in one hour when the power is one watt. A watt-hour = 3,600 joules. A *kilowatt-hour* (kWh) = 1,000 watt-hours = 3,415 British Thermal Units = $2{\cdot}6552 \times 10^{6}$ foot-pounds: it is called a "unit of electrical energy" for commercial purposes.

Henry: (H). The practical unit of self- or mutual inductance. A henry = 10^{8} e.m. linkages per ampere = 10^{9} e.m. units = 1 m.k.s.μ_0 linkage per ampere.

Farad: (F). The practical unit of capacitance. A farad = 10^{-9} e.m. unit = 9×10^{11} e.s. unit.

Φ — **Maxwell**: The c.g.s. e.m. unit of magnetic flux.

Weber: The m.k.s.μ_0 unit of magnetic flux. A weber = 10^{8} maxwells.

B — **Gauss**: The c.g.s. e.m. unit of magnetic flux density or induction. It is equal to one maxwell per square centimetre.

(B **Weber per Square Metre**: The m.k.s.μ_0 unit of magnetic induction. A weber per square metre = 10^{4} gauss.

Ampere-turn: A unit of magnetomotive force, expressed as the product of the number of turns of a coil and the current in amperes which flows through it.

F — **Gilbert**: The c.g.s. e.m. unit of magnetomotive force. A gilbert = $10/4\pi$ ampere-turn.

H **Oersted**: The c.g.s. e.m. unit of magnetizing or magnetic force.

SYMBOLS

Table 3.—Symbols for Practical Electrical and Radio Engineering Units

Ampere	A	Watt-hour	Wh
Volt	V	Volt-ampere	VA
Ohm	Ω	Volt-ampere reactive	VAr
Coulomb	C	Ampere-hour	Ah
Joule	J	Cycles per sec.	c/s
Watt	W	Kilocycles per sec.	kc/s
Farad	F	Megacycles per sec.	Mc/s
Henry	H		

Prefixes

Abbreviated prefixes have the following meanings:

milli-	m	= 1/1,000	($\times 10^{-3}$)
micro-	μ	= 1/1,000,000	($\times 10^{-6}$)
pico-	$\mu\mu$ or p	= 1/1,000,000,000,000	($\times 10^{-12}$)
kilo-	k	= 1,000	($\times 10^{3}$)
mega-	M	= 1,000,000	($\times 10^{6}$)

Examples

Milliampere	mA	Picofarad	$\mu\mu$F or pF
Microampere	μA	Microvolt	μV
Kilovolt-ampere	kVA	Millivolt	mV
Millifarad	mF	Kilovolt	kV
Microfarad	μF	Megavolt	MV

TABLE 4.—SYMBOLS FOR QUANTITIES USED IN ELECTRICAL AND RADIO ENGINEERING

Acceleration due to gravity	g
Admittance	Y
Aerial current	I_{ae}
Aerial resistance	R_{ae}
Aerial radiation resistance	R_r
Amplification factor (amplifier)	m
Amplification factor (valve)	μ
Angles	θ, ϕ, ψ
Angular velocity	ω
Anode A.C. resistance	r_a
Anode D.C. resistance	R_a
Anode current	I_a or i_a
Capacitance	C
Conductance	G
Coupling coefficient	k
Current	I
Decay coefficient	α
Effective height	h_e
Efficiency	η
Electric force	ξ
Electromotive force (e.m.f.)	E
Electrostatic flux	Ψ
Energy	W
Flux density (electrostatic)	D
Flux density (magnetic)	B
Form factor	K
Frequency	f
Fundamental frequency	f_1
Fundamental wavelength	λ_1
Grid A.C. resistance	r_g
Grid D.C. resistance	R_g
Grid current	I_g or i_g
Grid voltage	E_g or e_g
Impedance	Z
Intensity of magnetization	J
Magnetic field strength	H
Magnetic flux	Φ
Magnetic flux density	B
Magnetomotive force (m.m.f.)	F
Mutual inductance	M
Natural frequency	f_0
Natural wavelength	λ_0
Period or periodic time	T
Permeability	μ
Permittivity (dielectric constant)	ϵ or κ
Phase difference	ϕ
Potential difference, electric (p.d.)	V
Power	P
Power factor	$\cos\phi$
Quantity of electricity	Q
Radiation efficiency	η_r
Reactance	X
Reluctance	S
Resistance	R
Resistivity	ρ
Revolutions per unit of time	n
Self-inductance	L
Susceptance	B or b
Temperature	t or θ
Temperature (absolute)	T
Wavelength	λ
Wave velocity	v

TABLE 5.—ABBREVIATIONS FOR TERMS USED IN ELECTRICAL AND RADIO ENGINEERING

General	
Alternating current	A.C.
Direct current	D.C.
Audio frequency	A.F.
Video frequency (voice frequency)	V.F.
Radio frequency	R.F.
Low frequency	L.F.
Medium frequency	M.F.
High frequency	H.F.
Very high frequency	V.H.F.
Low voltage	L.V.
High voltage	H.V.
Extra high voltage	E.H.V.
Low tension	L.T.
High tension	H.T.
Extra high tension	E.H.T.
Root mean square	R.M.S.
Continuous wave	C.W.
Interrupted continuous wave	I.C.W.
Modulated continuous wave	M.C.W.
Double sideband	D.S.B.
Independent sideband	I.S.B.
Single sideband	S.S.B.
Amplitude modulation	A.M.
Frequency modulation	F.M.
Full wave	F.W.
Half wave	H.W.
Carrier telegraphy	C.T.
Voice frequency telegraphy	V.F.T.
Frequency shift keying	F.S.K.
Voice operated carrier suppressor	V.O.C.S.
Transformers	
Air, natural cooled	A.N.
Air blast cooled	A.B.
Oil-immersed, natural cooled	O.N.
Oil-immersed, air blast cooled	O.B.
Oil-immersed, water cooled	O.W.
Oil-immersed, forced air circulation and water cooled	O.F.W
Oil-immersed, forced air circulation and air blast cooled	).F.B.
Cables	
India rubber	I.R.
India rubber, pure	P.I.R.
India rubber, vulcanized	V.I.R.
Lead covered	L.C.
Paper insulated	P.I.
Tough rubber sheathed	T.R.S.
Polyvinyl chloride covered	P.V.C.
Wire armoured, single	S.W.A.
Wire armoured, double	D.W.A.
Steel tape armoured	S.T.A.
Conductors	
Steel cored	S.C.
Single cotton covered	S.C.C.
Double cotton covered	D.C.C.
Triple cotton covered	T.C.C.
Single silk covered	S.S.C.
Double silk covered	D.S.C.
Enamelled	Enam.

Symbols Used in the Classification of Emissions

According to the International Radio Regulations (Atlantic City), emissions should be designated according to :

(1) Type of modulation.
(2) Type of transmission.
(3) Supplementary characteristics.

(1) *Types of Modulation*

A	Amplitude
F	Frequency or Phase
P	Pulse

(2) *Types of Transmission*

Absence of any modulation intended to carry information	0
Telegraphy without use of modulated audio frequency	1
Telegraphy by keying of a modulated audio frequency or by keying of the modulated emission	2
Telephony	3
Facsimile	4
Television	5
Composite transmissions and cases not covered above	9

(3) *Supplementary Characteristics*

Double sideband, full carrier	(none)
Single sideband, reduced carrier	*a*
Two independent sidebands, reduced carrier	*b*
Other emissions, reduced carrier	*c*
Pulse, amplitude-modulated	*d*
Pulse, width-modulated	*e*
Pulse, phase- or position-modulated	*f*

For the full designation, the group given by the above coding systems should be prefixed by a number indicating the width in kc/s of the total frequency band occupied by the emission.

Examples

Twenty-five-word-per-minute telegraphy (no tone modulation)	0·1A1
Amplitude-modulated telephony (max. freq. 3,000 c/s)	6A3
Frequency-modulated telephony (max. freq. 3,000 c/s, 20,000 c/s deviation)	46F3

TABLE 6.—NOMENCLATURE OF RADIO WAVES

Below 30 kc/s	Very low frequency (V.L.F.)	Myriametric waves	Above 10,000 m.
30–300 kc/s	Low frequency (L.F.)	Kilometric waves	10,000–1,000 m.
300–3,000 kc/s	Medium frequency (M.F.)	Hectometric waves	1,000–100 m.
3,000–30,000 kc/s	High frequency (H.F.)	Decametric waves	100–10 m.
30–300 Mc/s	Very high frequency (V.H.F.)	Metric waves	10–1 m.
300–3,000 Mc/s	Ultra-high frequency (U.H.F.)	Decimetric waves	100–10 cm.
3,000–30,000 Mc/s	Super-high frequency (S.H.F.)	Centimetric waves	10–1 cm.
30,000–300,000 Mc/s	Extremely high frequency (E.H.F.)	Millimetric waves	10–1 mm.

Valve Symbols (B.S. 1409 : 1947)

Valve elements are designated as follows:

Anode	a
Cathode	k
Grid	g
Heater	h
Filament (emitting)	f
Target	t
External metallizing	M
Internal conducting coating	m
Deflector electrode	x or y
Internal shield	s

Where confusion may arise, subscripts or primes should be added:

(*a*) In multiple valves the respective assemblies are distinguished by adding the following subscripts:

Diode	d
Triode	t
Tetrode	q
Pentode	p
Hexode, heptode	h
Rectifier	r

(*b*) If there is more than one grid in the same section, a subscript is added consisting of a figure denoting the sequence of grids counting from the cathode.

(*c*) Where there are two or more similar electrode systems, it may be necessary to add primes.

Examples

Triode (directly heated)	$f\ g\ a$
Pentode (indirectly heated)	$h\ k\ g_1\ g_2\ g_3\ a$
Triode-pentode (indirectly heated)	$h\ k\ g_t\ a_t\ s$ (where applicable) $g_{1p}\ g_{2p}\ g_{3p}\ a_p$
Full-wave rectifier	$h\ k\ a'\ a''$

TABLE 7.—MATHEMATICAL SYMBOLS

Term	*Symbol or Abbreviation*
Antilogarithm	antilog
Approaches limit	$\rightarrow$
Approximately equal to	$\doteqdot$
Base of natural logarithms	e
Because	$\because$
Brackets	$[\{(\quad)\}]$
Co-ordinates, Cartesian	$x,\ y,\ z$
Co-ordinates, Polar	$r,\ \theta,\ \phi$
Cosecant	cosec
Cosine	cos
Cotangent	cot
Decimal point	· (middle point)

TABLE 7 (*contd.*)

Differential coefficient of y with respect to x	$\frac{dy}{dx}$ or dy/dx
Differential coefficient, nth	$\frac{d^n y}{dx^n}$ or $d^n y/dx^n$
Differential coefficient, Partial	$\frac{\partial y}{\partial x}$ or $\partial y/\partial x$
Divided by	$\div$ or /
Equal to	$=$
Equal to or greater than	$\geqslant$
Equal to or less than	$\leqslant$
Factorial	! or $\lfloor$
Function	$f(\ )$, $F(\ )$, etc.
Gamma function	Γ
Greater than	$>$
Hyperbolic sine (and similarly for other hyperbolic functions)	sinh
Identical with	$\equiv$
Increment	δ or Δ
Infinity	∞
Integration	$\int$
Inverse sine (and similarly for other functions)	$\sin^{-1}$ or arc sin
Less than	$<$
Logarithm of x to base 10	$\log x$ or $\log_{10} x$
Logarithm of x to base e	$\log_e x$
Magnitude of	$\|\ldots\|$
Minus	$-$
Much greater than	$\gg$
Much less than	$\ll$
Multiplied by	$\times$ or . (low point)
Not equal to	$\neq$
Not greater than	$\ngtr$
Not less than	$\nless$
nth root	$\sqrt[n]{\ }$
Operator $\frac{\partial}{\partial x}$	D
Operator $\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2}\right)$	∇^2
Parallel	$\parallel$
Perpendicular	$\perp$
Plus	$+$
Plus or minus	$\pm$
Ratio of circumference to diameter of circle	π
Root mean square	r.m.s.
Secant	sec
Sine	sin
Square root	$\sqrt{\ }$
Square root of minus one	i or j (Symbol j preferred in electrical engineering)

TABLE 7 (*contd.*)

Summation	Σ
Tangent	tan
Therefore	$\therefore$
Varies as	$\propto$
Versine	versin

TABLE 8.—THE GREEK ALPHABET

Capital	*Small*	*Name*	*Capital*	*Small*	*Name*
A	α	alpha	N	ν	nu
B	β	beta	Ξ	ξ	xi
Γ	γ	gamma	O	o	omicron
Δ	δ	delta	Π	π	pi
E	ϵ	epsilon	P	ρ	rho
Z	ζ	zeta	Σ	σ	sigma
H	η	eta	T	τ	tau
Θ	θ	theta	Υ	υ	upsilon
I	ι	iota	Φ	ϕ	phi
K	κ	kappa	X	χ	chi
Λ	λ	lambda	Ψ	ψ	psi
M	μ	mu	Ω	ω	omega

47. PROGRESS AND DEVELOPMENTS

47. PROGRESS AND DEVELOPMENTS

Colour Television Equipment (Marconi's Wireless Telegraph Co. Ltd.)

Marconi colour television camera channels are being produced both for live pick-up and for use with film and slides. The equipment can be supplied to operate on 405, 525 or 625 interlaced scanning lines per frame, generating the American N.T.S.C. type of signal, modified as necessary with 405- and 625-line systems. For normal broadcast use the simultaneous systems are encoded to produce a compatible signal, but for most closed-circuit applications the encoding system may be dispensed with.

Two basic colour cameras are available. The studio camera employs three 3-in. image orthicon tubes, enabling good pictures to be obtained at light levels of about 200 ft. candles. A three vidicon camera is available for use where a scene illumination of 1,000–1,500 ft. candles is no problem; the longer storage characteristic of the vidicon tube also makes this camera suitable in telecine applications by eliminating the need for fast pull-down projectors and synchronous running.

The two cameras employ similar optical systems. Light enters the camera through one of four turret-mounted lenses to form an image at a field lens mounted inside the turret. An optical relay system follows, and the final image is formed at a distance which allows a light-splitting system of filters and mirrors to be interposed. The light division is performed by two dichroic filters mounted as a vee, green light passing

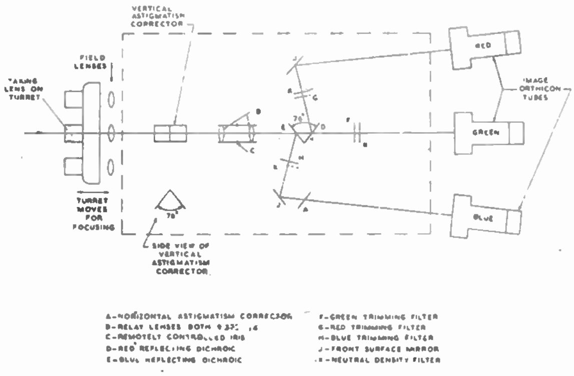

FIG. 1.—COLOUR CAMERA OPTICAL SYSTEM.
(*Marconi's Wireless Telegraph Co. Ltd.*)

straight through while blue and red are deflected to opposite sides, the light paths then being brought back into line by front-silvered mirrors. Further shaping of the spectral response of each channel is performed by a combination of dichroic and conventional colour filters, with astigmatism correctors.

Three pick-up tubes are used, one for each colour, and neutral density filters are inserted in the individual light paths to ensure that the tubes are operated over similar parts of their characteristics. The close-tolerance focusing and deflection yokes provide under normal working conditions a degree of registration equivalent to 500-line definition at the centre and 350 lines in t e corners.

Except for the camera and the camera remote-control panel housed in the operating console, the various units of the camera channel are designed to operate both with the studio and with the telecine camera. All those units which handle the simultaneous red, green and blue signals are housed in the operating console together with a black-and-white picture and waveform monitor.

Colour Television Receiver Developments (The General Electric Co. Ltd.)

Research engineers at The General Electric Company's Wembley Laboratories have reduced the size of their experimental colour television receiver to that of a domestic 21-in. monochrome set. This represents a scaling-down of nearly 50 per cent on their original model.

This receiver (Type TT4) is fitted with an R.C.A. 21-in. shadow-mask tube Type 21AXP22A. It has a complement of thirty-five valves and a total power consumption of 450 watts. By using G.E.C. metal rectifiers and "series run" techniques it has been possible to accommodate the receiver in a cabinet measuring 21¾ in. × 28½ in. × 31¾ in. Further developments are expected to enable a yet smaller cabinet to be used. In designing the receiver, a good-quality picture has been the goal rather than extreme economy in valves and circuits.

The receiver power pack incorporates direct rectification of the mains supply to provide one H.T. line of 250 volts at 200 mA, and a

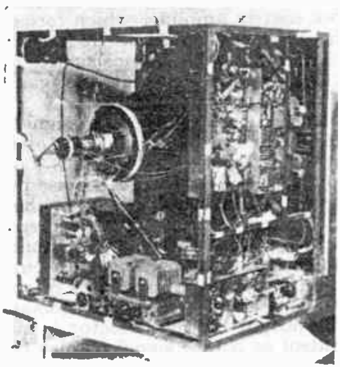

FIG. 2.—EXPERIMENTAL COLOUR TELEVISION RECEIVER.

The main and sub-chassis of a recent G.E.C. exper---ntal receiver.
The General Electric Co. Ltd.)

second H.T. line of 450 volts at 400 mA is obtained by a voltage-doubling circuit. A stabilized negative line of −150 volts is also available for bias supplies. G.E.C. metal rectifiers are used, and the whole power-pack unit measures only 5½ in. × 7 in. × 11 in. Most of the valves in the receiver are series run, but two small transformers supply the higher-current valve heaters and line-time-base valves.

The tube E.H.T. supply of 23 kV at 1 mA is obtained directly from the line flyback without voltage doubling, and a triode shunt regulator is used to stabilize the E.H.T. and thereby minimize the effect of E.H.T. changes on the convergence of the three electron beams.

Passive convergence circuits are fed from the line and frame time-bases and supply the appropriate current waveforms to the dynamic convergence coils mounted round the neck of the tube. D.C. currents are also applied to these coils for static convergence adjustments, and all the pre-set convergence controls are easily accessible at the front of the receiver. Line and frame raster shift is controlled by D.C. injection into the scanning coils.

On the signal side, a standard R.F. switchable tuner feeds a slightly modified production type I.F. deck. Sound rejection and sound I.F. take-off conform to usual monochrome practice, but in the vision circuits two separate crystal detectors (GEX54's) are used to provide isolation between the luminance and chrominance channels. This arrangement enables a high-definition luminance signal to be maintained, and when this is fed to the ferrite-loaded delay line (to provide time coincidence of luminance and chrominance) small reflections in the line are prevented from disturbing the smooth and symmetrical chrominance-channel frequency response.

The luminance signal is amplified by two video stages in a "bootstrap" circuit and is ultimately fed in the correct drive ratios to the three cathodes of the tube. Master brightness is controlled by a bias arrangement in one of the video stages, and the D.C. component of the signal is maintained to within about 1 dB.

Two chrominance amplifiers with a 6-dB band-width of ±500 kc/s supply two "clamping" triodes for high-level demodulation along the red and green difference axes, $R-Y$ and $G-Y$. After filtering, these difference signals are fed to the red and green grids of the tube, and also to a matrix amplifier which forms the $B-Y$ signal for the blue grid of the tube. A reference frequency amplifier provides the two appropriately phased reference signals for the triode demodulators.

A front-panel chrominance-gain control is provided for adjustment of the colour saturation.

The continuous reference signal required for synchronous demodulation is obtained from an L.C. oscillator, which is frequency and phase locked by a two-mode A.P.C. loop, or D.C. quadri-correlator, so that both fast pull-in and good noise performance are achieved.

One of the two detectors in the A.P.C. loop provides a D.C. voltage only when a burst signal is present and when the oscillator is phase locked. Since this is a synchronous detector, it provides an accurate indication of the presence of a burst even under adverse signal-to-noise conditions, and is used as a colour "killer" control for switching off the chrominance channel automatically when a burst signal is absent. The output of this detector is also fed as an automatic chrominance control or A.C.C. signal to bias the chrominance amplifier from which the burst output is taken.

The colour burst is separated from the chrominance waveform and amplified before being applied to the A.P.C. detectors. The separation is carried out by a gating circuit which switches on the burst amplifier only during the burst period of about 4 microseconds. In order to be quite independent of line time-base synchronization, the gating circuit operates from the back edge of the line-synchronizing pulses appearing in the synchronizing separator output.

FIG. 3.—PYE TELEVISION MICROWAVE LINK EQUIPMENT.

This long-range microwave television link is suitable for the transmission of N.T.S.C. colour or monochrome video signals simultaneously with high-fidelity audio signals. One watt of output power on 6875 to 7425 Mc/s provides transmission distances of 50 miles or more under favourable path conditions. The equipment can be used back-to-back as a demodulator repeater for long multi-stage transmission distances. All equipment can be fitted into four luggage-type cases.

(*Pye Telecommunications, Ltd.*)

B.B.C. Television Translator at Folkestone

The B.B.C. began transmitting programmes in 1958 from a new type of low-power television transmitter, known as a " translator ", at Folkestone. This town is typical of small populated areas which are within or adjacent to the service areas of the main B.B.C. stations, but are prevented by surrounding hills from obtaining satisfactory reception.

A translator converts the sound and vision transmission frequencies from one channel to another without demodulation to audio and video frequencies, which occurs when a normal receiver and transmitter relay installation is employed. This simplification increases the reliability of the equipment, which can therefore be arranged for automatic operation without attendant staff. Because the equipment is small it can conveniently be housed in weather-proof and insect-proof cabinets, thus dispensing with the need for a station building.

The translator is on high ground where good reception is possible from the Dover B.B.C. television station, and the transmitting aerial overlooks Folkestone. The receiving-aerial system consists of a double three-element array, and the transmitting aerial has four tiers of single, folded dipoles.

If the sound and vision signals of a television system are amplitude modulated, difficulties are introduced if they share a common amplifier in the translator because of the greater possibility of intermodulation.

These difficulties are eased in the systems such as those used on the continent, in which the sound signal is frequency modulated. In this equipment separate channels have been provided for the amplification of the sound and vision signals using common frequency-changing oscillators. The separate channel arrangement reduces the risk of intermodulation and enables separate automatic gain control to be employed to combat the effect of differential fading between the sound and vision signals, which could occur if a translator were dependent upon reception from a really remote B.B.C. station. The automatic-gain-control voltages are also used to initiate an automatic changeover to reserve translator equipment should the normal unit become faulty.

The double frequency-changing process facilitates the rejection of spurious signals and provides additional protection against " in band " feedback. The first frequency-changing process resembles that in a normal television receiver, producing vision and sound intermediate frequencies of 34·65 and 38·15 Mc/s respectively, and the second frequency-changing stage produces vision and sound signal frequencies in the required channel.

The Folkestone translator peak white vision power output is 1·5 watts, and in conjunction with the type of transmitting aerial used gives an effective radiated power of 7 watts in the direction of maximum radiation.

V.H.F./U.H.F. Scatter Amplifiers (Marconi's Wireless Telegraph Co. Ltd.)

High-power amplifiers for V.H.F. and U.H.F. tropospheric and ionospheric forward scatter techniques are now in production; these include the HS315 a 1-kW amplifier for operation within the range 680–970 Mc/s with a three-cavity klystron output stage; the HS201 a 20-kW amplifier, designed for operation in parallel to provide 40 kW output within the range 35–55 Mc/s and with a final stage comprising a pair of high-power tetrode valves with cross-neutralization inductively coupled to an unbalanced 50-ohm feeder; and the tropospheric scatter amplifier Type HS313, which will provide an output of 10 kW within the range 400–525 and/or 680–970 Mc/s using a four-cavity power klystron.

Tropospheric scatter links are now providing reliable high-quality U.H.F. communication services over distances of up to 250 miles in a single hop. One arrangement is to use a Type HS313 amplifier in conjunction with a single drive unit (for example, Type HD313). By the use of spaced receiving aerials and a two-path receiver, dual diversity operation may be achieved on such a system. Where higher reliability is required, or the path conditions dictate a system of greater gain, two HS313 amplifiers may be associated with two drive units and the outputs fed to separate spaced aerials, either on different frequencies or in different polarizations.

Employing reception from two aerials, and a four-path receiver, a quadruple diversity system results. This means that: (1) The transmitters, while normally working together, can operate alone and thus form an active stand-by to each other if either should fail. (2) With full operation, a diversity gain equivalent to some 5 or 6 dB improvement over the dual diversity system is achieved. The HS315 1-kW amplifier can be used on similar services where lower radiated power is permissible.

FIG. 4.—TROPOSPHERIC SCATTER AERIAL.

The 30 ft. diameter dish aerial and transmitter building at Start Point, Devon for a 200-mile tropospheric scatter multichannel radio link with Galleywood, Essex. The link is capable of carrying up to 60 simultaneous telephone channels.

(*Marconi's Wireless Telegraph Co. Ltd.*)

Aerials for Tropospheric Scatter Systems (Marconi's Wireless Telegraph Co. Ltd.)

To meet the requirements for tropospheric scatter communication special aerial equipment has been developed. The system comprises a reflector rigidly mounted and illuminated by a horn feed which can be adjusted in order to vary the deviation of beam by 2 degrees above and below, and 1 degree either side of, the reflector axis; there is also adjustment for focus of the horn.

The aerials at each end of the link are identical and can be used for transmitting and receiving simultaneously; separation of the two fields is effected by polarizing one vertically and the other horizontally. The horn feed effects the polarization in that it is a combination of horn and diplexer, termed " dual-polarized ".

The paraboloid reflectors are 30 or 60 ft. in diameter (a 120-ft. version is also made) and the lower rim may be about 5, 25 or 50 ft. from ground level. The aerials are suitable for frequencies 400–1,000 Mc/s and the forward gain at 1,010 Mc/s is 37 dB (30 ft.); 43 dB (60 ft.); 49 dB (120 ft.), the gain at 384 Mc/s being 30, 36 and 42 dB respectively. The beam-widths between −3-dB points at 1,010 Mc/s are 2, 1·2 and 0·25 degrees and at 384 Mc/s are 5·4, 4·7 and 0·75 degrees.

B.B.C. Vision Electronic Recording Apparatus (" VERA ")

A magnetic-tape equipment for the recording and reproduction of the vision and sound signals of television programmes has been designed and installed by the B.B.C. Research Department. The machine employs a three-track system of recording, two of the tracks being devoted to the storing of the video signal and one to the sound signal. Separate recording and reproducing head-stacks are employed, each stack containing three identical heads separated from each other by copper screens and carefully aligned with respect to one another in manufacture. An operational channel consists of two machines, controlled from a central control desk. Magnetic tape 0·5-in. wide is

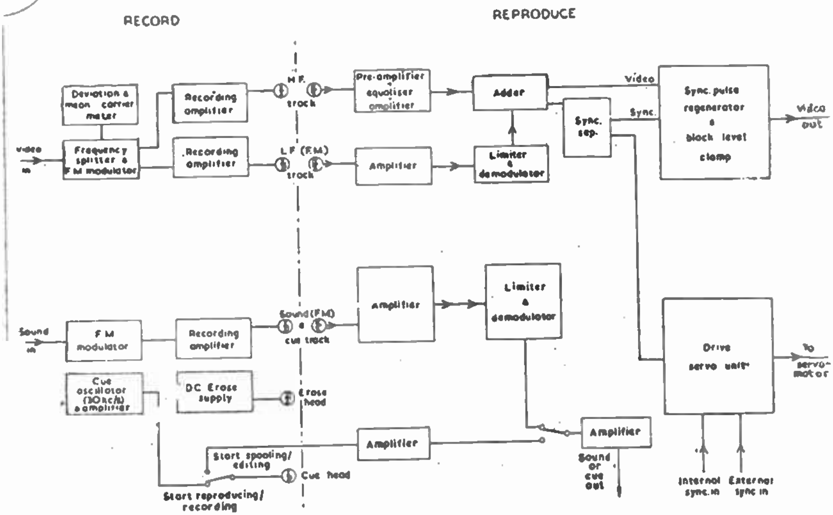

FIG. 5.—BLOCK DIAGRAM OF THE PRINCIPAL ELECTRONIC UNITS OF THE B.B.C. VISION ELECTRONIC RECORDING APPARATUS.

(B.B.C.)

used and a 20½-in.-diameter reel will accommodate 15 minutes of programme. The tape speed used is 200 in./sec., and conventional thin-base sound-recording tape is suitable. Continuous monitoring during recording is possible.

For storing the video signal the two video tracks are associated, on the recording side, with a band-splitting system to divide the signal into two bands, approximately 0–100 kc/s and 100 kc/s–3 Mc/s. The 0–100-kc/s video band is made to frequency-modulate a carrier, which is then recorded on one track. The bottom of the synchronizing pulses corresponds to 1 Mc/s and the peak-white level of the picture to 500 kc/s. This system overcomes the low-frequency and long-wavelength difficulties inherent in conventional magnetic recording and also enables limiting to be used to overcome most of the "drop out" difficulties.

On reproduction the output from the frequency-modulated video track is limited, demodulated and added to the output from the higher-frequency track to reform the composite television waveform. Before transmission the synchronizing information is extracted, reconstituted and added to the video signal.

The development of heads having sufficiently low losses at high frequencies is a major problem. Ferrites are used for the head cores, but as their mechanical properties are unsuitable for the gap edges and working surfaces, these are faced with magnetic material of superior mechanical properties, such as "Mumetal" or "Alfenol". To provide the necessary resolution gap lengths of the order of 2×10^{-5} in. are required. For this purpose mica spacers may be used or, alternatively, silicon monoxide, can be evaporated *in vacuo* on the pole faces of the head to form a durable, stable, gap spacer.

In the tape-transport mechanism, most of the power is supplied by spooling motors, which are arranged to move the tape past the heads at a speed just below 200 in./sec. even when the drive motor is not engaged. The drive motor is then required to supply only a limited amount of power. A "Velodyne" system of speed control and correction of the driving capstan is used.

Mechanical Filters (Collins Radio Company)

Mechanical filters, which can replace quartz crystal filters in many applications, are available in units as small as 0·3 cu. in. and provide a flat-topped frequency-response characteristic; they have been built with a 60–to–6 dB shape factor as low as 1·2 to 1.

Filters consist of three basic elements: (1) transducers which convert electrical oscillations into mechanical oscillations, or vice versa; (2) metal discs which are mechanically resonant; and (3) disc-coupling rods.

The transducer, which converts electrical and mechanical energy, is a magnetostrictive device based on the principle that certain materials elongate or shorten when in the presence of a magnetic field. Therefore, if an electrical signal is sent through a coil which contains the magnetostrictive material as the core, the electrical oscillation will be converted into a mechanical oscillation. The mechanical oscillation can then be used to drive the mechanical elements of the filter. In addition to electrical and mechanical conversion, the transducer also provides proper termination for the mechanical network.

The centre frequency of a mechanical filter is determined by the metal discs, which may be represented in an electrical equivalent circuit as a series resonant circuit. In practice, filters between 60 and 600 kc/s are being manufactured. Since each disc represents a series-resonant circuit, increasing the number of discs increases skirt selectivity of the filter; skirt selectivity is normally specified as shape factor, which is the ratio: (band-width 60 dB below peak)/(band-width 6 dB below peak).

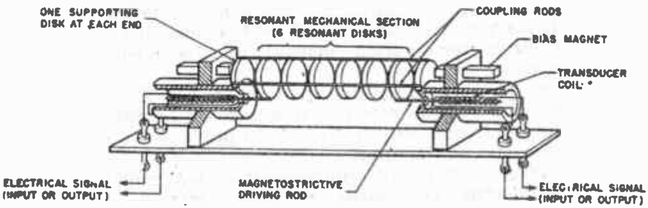

FIG. 6.—FUNCTIONAL DIAGRAM OF A MECHANICAL FILTER. (*Collins Radio Company*)

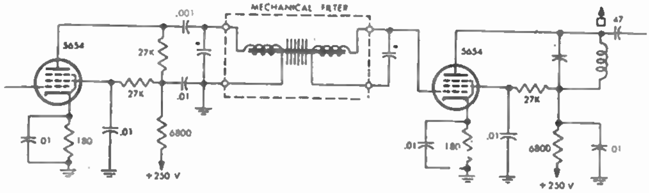

Fig. 7.—Typical Constant Gain Amplifier using Mechanical Filter.

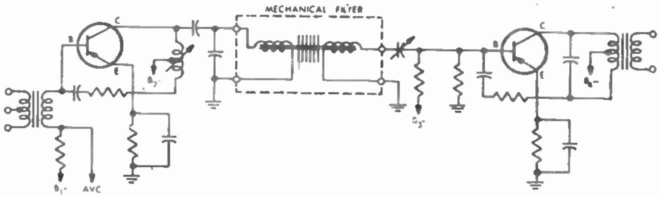

Fig. 8.—Transistor I.F. Amplifier with Mechanical Filter.
(*Collins Radio Company.*)

In an equivalent circuit, the coupling capacitors represent the rods which couple the discs. By varying the mechanical coupling between the discs, i.e., making the coupling rods larger or smaller, the band-width of the filter is varied. Because the band-width varies approximately as the total area of the coupling wires, the band-width can be increased by either using larger or more coupling rods. Standard available band-widths range from 500 c/s to 35 kc/s, and special units have been built with band-widths as narrow as 30 c/s and as wide as 60 kc/s.

The design and construction of circuits using mechanical filters is relatively simple, since no special matching networks are required. Being internally terminated, the filters need only a high-resistance termination of uncritical value at either end together with the specified resonating capacitance. They may be used simply in inter-valve couplings with the D.C. blocked out, and can also be matched into the low-impedance circuits encountered with transistors, by using a series-resonant filter termination instead of the normal parallel-resonant condition.

When used in bandpass circuits, such as an I.F. filter in a high-performance communications receiver, there should be effective shielding between the input and output circuits to prevent the input signal from partially by-passing the filter.

Mechanical filters find many applications in transmitting and receiving communications equipment, S.S.B. generation, multiplexing equipment, missile-guidance systems, doppler radar, etc.

Information on these mechanical filters has been made available by *Collins Radio Company of England Ltd.*

H.F. Crystal Filters (Cathodeon Crystals Ltd.)

Band-pass 10·7-Mc/s crystal filters BP50, BP25 and BP12 have been designed for use in V.H.F. communication receivers where channel spacing is based on 50, 25 and possibly 12·5 kc/s separation. The filter unit consists of a lattice network using high-stability quartz crystal units and associated components mounted on a printed circuit, enclosed in an hermetically sealed metal case. The three types are interchangeable permitting simple adaptation of receivers for different channel-spacing systems. It is claimed that the use of these filters makes possible single-conversion V.H.F. and U.H.F. receivers with the selectivity normally associated only with multiple-frequency conversion equipment, reducing the possibility of cross modulation, spurious signals and mixer overload and eliminating the need for critically designed high-stability I.F. transformers. Operating characteristics include: centre frequency stability ±0·0012 per cent; band-width at −6 dB 25 kc/s (BP50), 12·5 kc/s (BP25), 6 kc/s (BP12); band-width at −60 dB 45 kc/s (BP50), 22 kc/s (BP25), 11 kc/s (BP12).

Triple Conversion Tuning System (Racal Engineering Ltd.)

An interesting triple-mixing technique, based on a single high-stability crystal, is used in the Racal RA17 H.F. communications receiver to provide an easily set tuning range of 0·5–30 Mc/s without mechanical switching and with automatic drift correction. Tuning is carried out in two steps: first, the desired megacycle range is selected (1–30 Mc/s) without need for critical setting due to the error-balancing circuit; secondly, the kilocycles control adjusts the set and the 60-in. film-scale dial to the desired frequency. With the aid of a built-in crystal calibrator, a setting accuracy within 200 c/s is provided throughout the range.

The triple-conversion circuit employs a very high first intermediate frequency. This enables an untuned, low-noise, wide-band amplifier to be used in the signal-input stage, dispensing with tuned signal stages and their switching problems. The input signal, after amplification, is combined in the first mixer with the output of the first variable frequency oscillator. The first I.F. (40 Mc/s) takes the form of a spectrum of 1·3 Mc/s band-width. The 1-Mc/s crystal oscillator drives a harmonic generator to provide a range of frequencies 1 Mc/s apart up to 32 Mc/s.

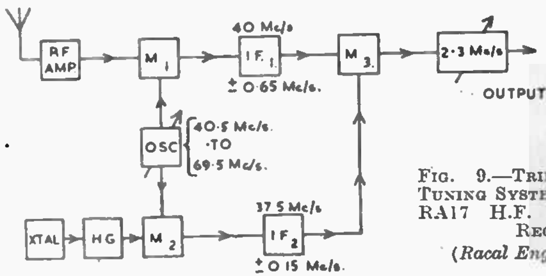

FIG. 9.—TRIPLE CONVERSION TUNING SYSTEM FITTED TO THE RA17 H.F. COMMUNICATIONS RECEIVER. (*Racal Engineering, Ltd.*)

These harmonics are combined with the output of VFO1 in mixer M4. The presence of a 37·5-Mc/s filter of 300 kc/s band-width in the output circuit limits the effective settings of VFO1 to intervals of 1 Mc/s (i.e., when one of the harmonics will combine with VFO1 to produce an output in the filter passband). This 37·5-Mc/s output is combined with the first I.F. in mixer M2 to produce the second I.F. comprising a spectrum from 2 to 3 Mc/s. The succeeding stages consist of a conventional superhet circuit tuning 2–3 Mc/s and acting as an interpolation receiver.

The tuning control of VFO1 behaves, in effect, as a twenty-nine-position electronic switch providing band selection in equal steps, starting from 1 Mc/s up to 30 Mc/s. Drift in VFO1 over fairly wide limits has no effect on the tuning of the receiver, for a slight change in the frequency of VFO1 will alter the frequency from the first I.F. to the same extent and in the same direction as the frequency received from the 37·5-Mc/s amplifier and filter; the output frequency from M2 will therefore remain unchanged.

Fig. 10.—Cossor Kit-constructed Double Beam Oscilloscope.

The home construction of reliable servicing and test equipment has been much simplified by the introduction of printed wiring panels. This largely overcomes the problem of slight variations in lay-out and construction leading to considerable differences in results and in calibration. Modern kits, as supplied by a number of makers, contain all parts together with detailed step-by-step assembly instructions. Signal generators, valve voltmeters and multi-range testmeters are among the instruments available in kit form.

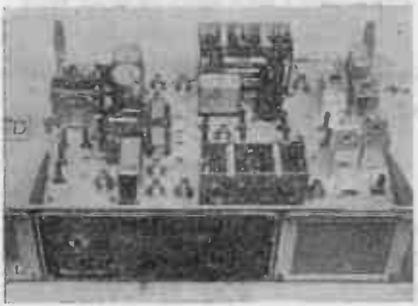

Fig. 11 (*left*).—Transistorized "Third Method" S.S.B. Generator.

The S.S.B. generator and demodulator of the Redifon GR400 60-watt S.S.B. transmitter–receiver, together with transistorized microphone and loudspeaker amplifiers in one 19-in. wide by 3½-in. high unit. Information on this type of S.S.B. generator is given in Section 7.

FIG. 12.—" SETTING UP " PICTURE TUBE FOR SERVICING.

A 6-in. magnetically-deflected tube developed as an aid to television servicing. The tube will plug into the majority of receivers using 14, 17 or 21-in. tubes and avoids the handling of full-sized tubes during field and bench work. It also increases the accessibility of the components on the chassis.

(*The General Electric Co. Ltd.*)

FIG. 13.—CARRIER DEVIATION METER.

A direct reading F.M. deviation meter with a calibrated frequency range of 4 to 1024 Mc/s in eight bands and four deviation ranges of 5, 25, 75 and 125 kc/s. The instrument is basically an F.M. receiver using a counter-type discriminator of which the output energizes a meter calibrated directly in deviation.

(*Marconi Instruments Ltd.*)

Scan Magnification (Mullard Ltd.)

A new system of scan magnification—invented by D. R. Skoyles—which overcomes the limitation of scan power available in an experimental transistor television receiver, was first described by B. R. Overton, B.Sc., A.M.I.E.E., of the Mullard Research Laboratories in the *Journal of the Television Society*, Vol. 8, No. 11, 1958. A pair of focusing quadrupole magnetic lenses and a magnetic magnifying lens are placed along the neck of the cathode-ray tube having a wide-beam triode gun. In the receiver described this system enabled the power requirements for the time-base to be reduced to approximately one-fortieth of those of a conventional receiver: a power saving of 100 : 1 being achieved in the line time-base and approximately 4 : 1 in the field time-base. It enabled a fully transistorized receiver to be operated from a 12-volt supply with a consumption of approximately 12 watts.

If an electron path is through a fixed magnetic field having opposite sense on either side of the cathode-ray-tube axis between the deflection assembly and the screen, then the deflection angle of the beam can be increased in a manner analogous to an optical lens. Normally, however, with such an arrangement the deflection in a plane at right angles to the plane in which there is magnification will be decreased. However, if the field strength of the magnifying lens is increased beyond a certain point, the scan in the second plane is also magnified, though with a reversal of sense.

Simply interposing a magnifying lens would, however, increase spot size as well as the deflection angle, so that little benefit would be obtained. This difficulty can be overcome by the use of a quadrupole focusing system, comprising a pair of quadrupole magnet assemblies spaced a short distance apart along the tube neck, and having diverging planes set at 90 degrees relative to each other.

Adjustable magnifying lenses have been developed which overcome deflection defocusing by the use of poles shaped as equilateral rectangular hyperbolæ, and which retain a good raster shape. To maintain focus in the horizontal direction when the beam is deflected vertically, there are additional correction windings, energized by the frame timebase, on the deflection yoke. It should be noted that the final focus is more sensitive to E.H.T. variations than a conventional system.

The application of scan magnification to other forms of display units, such as radar, has already been demonstrated. It seems likely that scan magnification will eventually be an essential part of television-like display units designed to be economical in power.

Percival System of Stereophonic Broadcasting (Electric and Musical Industries Ltd.)

Stereophonic broadcasting experiments have been carried out by the B.B.C. and in other countries in which the signal intended for the left loudspeaker is transmitted from one station, while that intended for the right loudspeaker is transmitted from another station. While useful information can be obtained from such experiments, this system is uneconomical and is incompatible in so far as listeners with only a single standard receiver could receive only the left or the right signal. In the U.S.A. single F.M. multiplex transmitters have been used to transmit the sum of the left and right signals in place of the ordinary audio signal, together with the difference of the two signals on a subcarrier; this normally involves a loss of transmission range equivalent to a power loss of the order of 6 dB, due to the band-width required.

In an ideal system the listener with a standard receiver would be unaware that stereo sound was being transmitted, while the listener with the necessary equipment would enjoy stereo reproduction over the same transmission range as the ordinary listener.

The transition from ordinary to stereo reproduction involves the transmission of additional information as to direction. It can be shown that the quantity of additional information is small compared with that necessary for the audio content. Hence, if the directional information could be separated and transmitted as an independent signal, the extra power and band-width required would be very small. Moreover, only the single normal audio signal would be required, so that the system would be completely compatible.

If there were only a single moving source it would be sufficient to transmit an auxiliary signal indicating the position of the source at any instant. In practice, it is necessary to deal with a number of independent sources, and it might appear that a separate signal would be required for each: however, it has been found that this is not necessary.

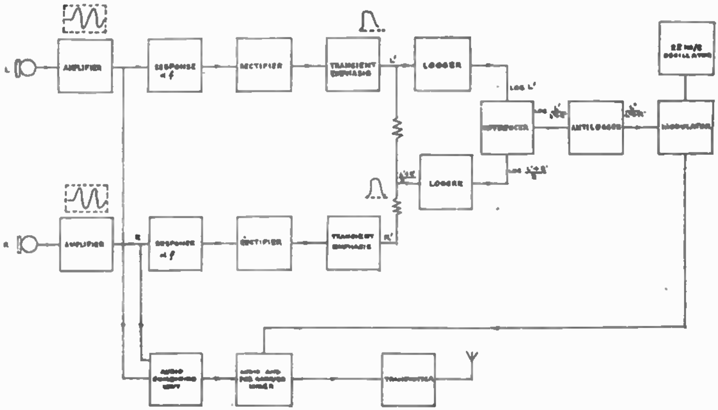

FIG. 14.—ENCODER AND TRANSMITTER FOR PERCIVAL SYSTEM.
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In the earliest experiments the directional information was given by a voltage depending on the ratio of the envelopes of the normal left and right signals. This voltage was employed to control the relative gains of the amplifiers supplying the left and right speakers. The optimum band-width for the control signal was found to be about 100 c/s. Some true stereo effects were obtained, but, as would be expected, were inadequate.

Nevertheless, the results showed that successive sounds emanating from different directions were interleaving in such a way as to give the impression of separate simultaneous sources.

Further development involved a recognition of the fact tnat the ears are able to determine the directions of certain types or parts of sounds more readily than the directions of others. Indeed, experiments showed that some sounds could completely mask the directions of certain

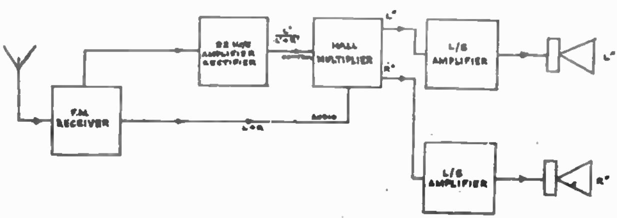

FIG. 15.—RECEIVER AND DECODER.
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other sounds. Accordingly, it was necessary to process the original left and right signals in such a way that the directional information was weighted in accordance with the ability of the ears to detect the directions of the various sounds. In other words, an attempt was made to match the channel carrying the directional information to the channel constituted by the ears and the brain.

The additional processing required to form the directional signal naturally involved some complications at the transmitter end. However, it involved no modifications to the radio transmitter itself and, very fortunately, it required no changes whatever in the very simple process of decoding in the receiver.

To test the system over radio, the low-power R.F. amplifier of a prototype of a B.B.C. Band II F.M. transmitter was utilized to transmit the signal over a distance of about ¼ mile with a power of the order of a watt.

The left and right audio signals were derived from standard commercial stereo tape, the two signals being combined to give a single compatible audio signal together with a control voltage which was modulated on to a sub-carrier at a frequency above the audio band. The audio and directional signals were then added and caused to frequency modulate the main carrier, the transmitter itself being unaltered. The reduction in level of the audio signal required to accommodate the sub-carrier would have caused a reduction of range equivalent to a loss of power of about 2 dB. However, it is hoped to do better than this.

The single compatible audio signal can be received in the same way as for a monophonic transmission, the additional directional signal being obtained on stereo receivers by demodulating the sub-carrier, which can be just above the highest audio frequency, for example at about 22 kc/s.

If L is the original left audio signal and R the original right audio signal the combined audio signal may be written as $L + R$, although it may in fact differ from the simple sum in certain respects. The direction signal is $L'/(L' + R')$, and it is required to employ this as a control voltage to give the new left and right signals L'' and R'' where

$$L'' = \frac{L'}{L' + R'}(L + R); \quad R'' = \frac{R'}{L' + R'}(L + R) \quad . \quad . \quad . \quad (1)$$

so that

$$R'' = (L + R) - L'' \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Hence L'' can be obtained by multiplying the single audio signal by the directional signal, while R'' is simply the audio signal with L'' subtracted from it.

The multiplier unit must satisfy two conditions: first, that there shall be no break-through of the control voltage into the audio output, and second, that it shall be sufficiently cheap to incorporate in a commercial receiver. Both of these conditions are satisfied by a compact form of Hall multiplier in which the control signal energizes the field, while the audio signal is passed through the crystal. The output from the Hall multiplier appears as a voltage across the Hall crystal and, after amplification, provides the input to the left loudspeaker. The input to the right loudspeaker is obtained by amplifying the difference between the audio input and the output from the Hall multiplier.

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