THE VALUE OF QUALIFICATIONS

In the absence of more reliable means, the acquisition of knowledge can only be measured by independent examination. During the next few weeks* several hundred students of radio engineering will take a test to prove their knowledge, and for those who succeed there is an immediate reward for the efforts to which they have been devoted for at least three years.

In this *Journal* a statement is published twice every year showing the entries and successes in the May and November examinations of the Institution. This information may not be of immediate interest to the majority of readers, but it is of tremendous importance to over 1,000 candidates who sit the examinations each year. The value of the efforts made by these candidates is measured by the standard set by the examiners, and those who are successful derive a qualification which should be of benefit throughout the whole of their professional lives.

Publication of the entries and successes also gives to interested observers some indication of the ability of candidates. For example, the pass lists show that in the 1955 examinations the ratio of successes to failures in all subjects was one to three.

On behalf of the unsuccessful candidate much enquiry is made into the circumstances of failure. As was expressed at a recent meeting of the Institution in South Wales, failure may be attributed to lack of a properly designed course involving laboratory training and practical experience. The shortage of suitable and properly qualified technical teachers is also advanced as a reason for failure. Even the term "unsuccessful" is, however, of limited definition: the preparation involved is not without merit and, indeed, all pass lists include candidates who were previously unsuccessful. A candidate certainly has every encouragement to seek a worthwhile qualification, especially if he has not the advantage of a university degree. In Great Britain today more attention is focused on technical education than ever before in the history of the country, and the reasons may well be found in the article written by the Rt. Hon. R. A. Butler† when he stated "I have always felt that our future as a great commercial and trading country would ultimately be decided by the quality and quantity of our scientific and technical knowledge and its wider spread in industry."

It cannot, however, be too strongly emphasized that the passing of a professional examination, or indeed any examination—university or otherwise—does not *per se* guarantee success in any walk of life. Examination efforts must be supplemented by other qualities. If a man be reasonably well possessed of qualities of initiative and leadership, then the procurement of a qualification based on a series of searching examinations can be—and usually is—an invaluable asset, particularly when opportunities for personal advancement arise.

Certainly those candidates who submit themselves to examination give promise of initiative and a sense of responsibility. In an age when there are so many *divertissements* from the pursuit of knowledge and skill, the unsuccessful candidate is deserving of every help and encouragement.

---

* The closing date for entries from candidates in the United Kingdom for the May Graduateship Examination is 31st March.

† Mr. Butler, a former Minister of Education, was writing on "Freedom and Opportunity" in *The Spectator*, 20th May 1955.

APPLICANTS FOR MEMBERSHIP

New proposals were considered by the Membership Committee at a meeting held on 19th January, 1956, as follows:—37 proposals for direct election to Graduateship or higher grade of membership, and 27 proposals for transfer to Graduateship or higher grade of membership. In addition, 43 applications for Studentship were considered. This list also contains the names of seven applicants who have subsequently agreed to accept lower grades than those for which they originally applied.

The following are the names of those who have been properly proposed and appear qualified. In accordance with a resolution of Council and in the absence of any objections being lodged, these elections will be confirmed 14 days from the date of the circulation of this list. Any objections received will be submitted to the next meeting of the Council with whom the final decision rests.

Direct Election to Member

BOOTH, Andrew Donald, D.Sc., Ph.D. *Fenny Compton.*
RHYS-JONES, John Emyr. M.B.E. *Iver.*
TAYLOR, Denis. M.Sc., Ph.D. *Abingdon.*
WALKER, Frederick Maynard, B.Sc. *Banstead.*

Transfer from Associate Member to Member

HAMER, Edward George, B.Sc.(Eng.). *Baltimore, U.S.A.*
LAURENCE, Roger Frederic. *Guildford.*

Direct Election to Associate Member

BARTLEY-DENNISS, Lt.-Col. George Arthur, R.A. *London, S.W.1.*
BOOTH, Philip Charles. *Croydon.*
HINDE, Douglas Wylie, B.Sc.(Hons). *Huddersfield.*
JONES, Captain Frederick Rees Bedlington, R.E.M.E. *Arborfield.*
LONGSTAFF, William. *Frimley.*
MESSENGER, Charles Henry. *Forest Rise**
MONTANINI, Wg. Cdr. Victor Montague. *Chingford.*
NEWSON, William Kenneth. *Little Chalfont.*
QUESTED, Derek Ray, B.Sc., M.Sc. *Brighton.*
SATYANARAYANA, Kanagala, B.A. *Jamnagar.*
SWALLOW, Major Dennis Lawrence, R. Signals. *Malvern.*
VESELY, Kurt. *Glynneath.*

Transfer from Associate to Associate Member

BRAY, Captain Donald Harry Arthur, B.E.M., R.E.M.E. *Arborfield.*
HUNT, George Ernest. *Fleet.*

Transfer from Graduate to Associate Member

HOWE, Frank Ralph. *Colchester.*
PENFOLD, Reginald Charles. *Droitwich.*
SMAILES, George. *Nairobi.*
WATSON, Dennis Hugh. *South Norwood*

Transfer from Student to Associate Member

CULLINGTON, Sidney Albert John. *South Harrow.*
HALBE, Capt. Madhusudan Narayan, B.Sc. Indian E.M.E. *New Delhi.*
SMITH, Derek Leslie A., B.Sc.(Eng.). *Potters Bar.*

Direct Election to Associate

DJOKIC, Cedomir. *Wolverhampton.*
HART, Stephen George. *Watford.*
NOLAN, Hugh Joseph. *Zomba, Nyasaland.*
REYNOLDS, Frederick Winston. *Shepperton.*

Transfer from Student to Associate

HOBBS, William Edward. *Great Baddow.*
PLUMB, Ronald Henry. *Shevington.*
RIDOUT, John Richard. *Bournemouth.*
SATHYANARAYANA IYER. *Bangalore.*

Direct Election to Graduate

ANDERSON, Ralph Taylor. *Sunderland.*
BALL, Plt. Off. Derek, B.Sc.(Hons). *R.A.F., Patrington.*
BOWLES, Mervyn Edward. *Sydney, Nova Scotia.*
DAVIES, Stanley James. *London. S.E.25*
ELLIS, Reginald Heber. *London, S.W.19.*
GRICE, Henry Richard Christopher. *Peterborough.*
HOLE, Bernard Gerard. *Thornton Heath.*
HURST, Sydney. *Feltham.*
MURRAY, Patrick Joseph. *Coventry.*
PHILLIPS, Flg. Off. Peter Clifford. B.Sc. *R.A.F., Wyton.*
SANDHAM, Peter Harold. *Malmesbury.*
SAUNDERS, Flg. Off. Donald. *R.A.F., Laarbruch.*
THOMPSON, Donald Gordon. *Morden.*
WILSON, Capt. Harold, R.E.M.E. *Salisbury.*
WONG, Wait Kheong. *Manchester.*

Transfer from Student to Graduate

BURGESS, Thomas George. *New Plymouth, New Zealand.*
GARDINER, Alan. *Wells.*
MAH, Seck Wah. *Brighton.*
RAJA, Manakkulam Krishna. *Meeret.*
SIDHU, Niranjan Singh, B.A. *Moga.*
SOOD, Ramesh Chander. *Bombay.*
SUBBARAMU, Kattepur Ramaswamy, B.Sc. *Bombay.*
THOMAS, William James. *Stafford.*
WALTERS, Henry George. *Weston-Super-Mare.*

STUDENTSHIP REGISTRATIONS

ABDUL GAFFAR, M. *Bangalore.*
AGHARKAR, Ratnakar Vinayak. *Bassein.*
BHATT, Narayan Mugatram. *Harij.*
BOWEN, Kenneth. *King's Lynn.*
BROWN, Henry Watson. *Bridgwater.*
CHEUNG SHIU HUNG. *Hong Kong.*
CHOPRA, Sushil Chander. *Ambala.*
CUTLER, George Donald. *Cambridge.*
DESHMUKH, Madhukar Raghunath, B.Sc. *Bombay.*
GREENWOOD, Frank. *Adelaide.*
GURCHARAN SINGH SURIE. *Bangalore.*
HALLEGUA, Isaac A. *Isleworth.*
HALLWORTH, Robert Philip. *Alderley Edge.*
HANDA, Jagdish Rai. *Agra.*
INKSTER, Alan Henry Peter. *London, S.W.19.*
JOSHI, Rajaram Sivasanker. *Madras.*
KERWOOD, Maurice Edward. *Coventry.*
KHAN, Mohammad Tahir. *Southampton.*
KHATRI, Dindayal Tahilram. *Bangalore.*
LEWIS, John Edward. *Frankston, Australia.*
McMILLAN, Thomas Lynn. *Glasgow.*
MADAN, Amrit Lal. *New Delhi.*

MARTLEW, Allan. *Billinge.*
MIAN, Riaz Ahmad. *London, N.7.*
MOHD ZAHIR KHAN, B.A. *Karachi.*
MUKHERJEE, Samir Kumar. *New Delhi.*
NAIR, Thekkepat Krishnankutty. *Poona.*
NAVALKAR, Vinayak Pratapchandra, B.E.(Hons). *Bombay.**
PARKER, John Leslie. *Huddersfield.*
PARKES, Eric Edwin. *Birmingham.*
PARKASH THUKRAL, S. *Bangalore.**
PRIT PAL SINGH. *Agra.*
RAJAGOPAL, A. *Bangalore.*
ROPER, Roy Frederick Allen. *Margate.*
SATYARAJ, Karagi Vasudevarao. *Bombay.*
SAVAGE, Raymond George. *Uxbridge.*
SILVERSTONE, Ralph. *London, N.16.*
SONI, Jitender Nath, B.Sc. *Bangalore.**
TOPPING, John. *Carlisle.*
TSANG CHIU SUN. *Singapore.*
WHITE, Nigel John. *Wells.*
WILKINSON, David. *Gibraltar.*
WILKINSON, George Arthur. *Hull.*
YOGINDER SINGH JASWAL. *Secunderabad.*

* Reinstatement.

NOTICES

New Year Honours List

The Council of the Institution has sent congratulations to Mr. John Hudson-Davies (*Member*) on his appointment as a Commander of the Most Excellent Order of the British Empire.

Mr. Hudson-Davies, who held the post of Chief Engineer to the Sudan Posts and Telegraph Department from 1942 to 1955, has recently been appointed Postmaster-General of Aden Colony. He was elected an Associate Member of the Institution in 1942, and was transferred to full Member in 1945.

Obituary

Advice has recently been received of the death of Dr. William Ernest Boyd (*Member*) in September, 1955, and the Council of the Institution expresses its sympathy with Dr. Boyd's relatives. He was sixty-four years of age.

Dr. Boyd, who studied medicine at Glasgow University from 1908 to 1914, subsequently obtained his M.A. degree in Natural Science in 1917. In 1919, he founded the laboratories of the Boyd Medical Research Trust in Glasgow, where he was engaged in biophysical research until the time of his death.

Dr. Boyd's interests were in electro-medical work and radiology, and he was the author of numerous medical and scientific papers. Two of his papers were published in the *Brit.I.R.E. Journal*: "The study of vibrations on an impulsed board: a new camera recording method," in September 1940, and "The wave analysis of low-frequency potentials of the human body," in March 1948.

He was elected to Membership of the Institution in 1940.

Incorporation of the Radio Trades Examination Board

As was anticipated in the Institution's Annual Report, published in the September 1955 issue of the *Journal*, the Radio Trades Examination Board has recently been granted incorporation under the Companies Act of 1948 by the Board of Trade. The R.T.E.B. was formed in 1942 jointly by the Radio Manufacturers' Association (now the Radio Industry Council), the Radio and Television Retailers' Association (R.T.R.A.) Ltd., the Scottish Radio Retailers' Association, and the Brit.I.R.E. The Institution provides secretarial facilities for the Board, and together with the representatives of the radio industry and retailers gives financial support to its activities.

The Memorandum of Association of the Board states that among its principal objects is the promotion of a high standard of skill and efficiency in the work of radio mechanics and technicians. To this end, the Board at present conducts examinations in collaboration with the City and Guilds of London Institute, leading to the award of Radio and Television Servicing Certificates. The Institution is supporting a proposal that the activities of the Board should be extended beyond the holding of the examinations in radio and television servicing.

The nature of the Board's work has been recognized by the Board of Trade, which has granted a licence for the word "Limited" to be omitted from its title.

Annual Subscription Reminders

Members are asked especially to note that reminders for subscriptions due on 1st April for the year 1956-7 will be sent out with the March issue of the *Journal*. This is being done to save postage, and the co-operation of members is requested in remitting subscriptions promptly.

International Telecommunications Award

The 1955 winners of the annual award instituted by the City of Genoa to honour the memory of Christopher Columbus have now been announced. They are the British and American engineers responsible for the transatlantic telephone cable project, Sir Gordon Radley, C.B.E., Ph.D.(Eng.), Director-General of the Post Office, and Dr. Mervin J. Kelly, President of the Bell Telephone Laboratories.

It will be recalled that the President of the Institution was invited to nominate two candidates for this "International Prize of Communications" (see *Brit.I.R.E. Journal* for July, 1955) and members will be interested to learn that the award agrees with his nominations.

Correction

In the paper "Design of Studios for Small Broadcasting Stations," by R. F. Goodsman, which appeared in the January *Journal*, it was stated on page 9, Sect. 3. 2. 1, that the air-space between double walls should be between 2 ft. and 4 ft. wide. These distances should have been 2 in. and 4 in. respectively.

# of current interest . . .

The London Audio Fair

During recent months, there has been considerable discussion in the Institution's *Journal* and in technical periodicals on the role of the specialist exhibition, and of its value to the radio and electronics engineer. It is interesting therefore to learn that a new exhibition is to take its place in the calendar of events, namely the London Audio Fair, which, although primarily aimed at the general public, should also attract the professional engineer.

Thirty-nine firms specializing in sound equipment will be exhibiting at the Fair, which is to take place on 13th, 14th and 15th April at the Washington Hotel, Curzon Street, London, W.1. The Fair will be open from 11 a.m. to 9 p.m. each day, and admission will be invitation only. Tickets are however available from the Secretary (L. H. Brooks Esq., 17 Stratton Street, W.1) on request.

Broadcasting Developments in Africa

Construction has begun on new buildings to house the Gold Coast Broadcasting Service. The main features of the building programme are the erection of a Broadcasting Village and the conversion of the existing Broadcasting House at Accra to accommodate new transmitter installations.

The Broadcasting Village, in Accra, is to consist of two fully-equipped blocks of air-conditioned studios for talks, drama and music, also blocks of offices to house broadcasting and administrative staff, and facilities for staff and visitors. The separation of the buildings, rather than their incorporation in a single block, is particularly suitable for the climatic conditions and technical problems of a tropical country; stray noise can also be more easily controlled, and there is scope for future extensions.

The present Broadcasting House will be converted to accommodate a 20-kW transmitter and two 120-kW generating sets, and will, it is hoped, be in operation by the middle of this year. The studio centre will be ready for use in the first half of 1957.

A new broadcasting station at Kisumu, Kenya, and extended coverage and longer programme hours for the Mombasa transmitter are among projects to be implemented soon to extend the Colony's radio facilities. The Kisumu station is intended to satisfy the demand for a full regional service in Nyanza Province.

Robert Blair Fellowships

It has recently been announced that the value of the Robert Blair Fellowships in the case of Canada and the U.S.A. has been increased from £1,000 to a maximum of £2,000.

The Fellowships, which are for advanced study or research abroad in Applied Science and Technology, are administered by the London County Council. They are open to suitable candidates over twenty-one years of age, preference being given to those who have completed a course of study in engineering science in a London institution. Further information may be obtained from the Education Officer, London County Council, The County Hall, London, S.E.1.

Long Range V.H.F. System Using "Forward-Scatter"

Equipment has recently been installed in Oxfordshire for direct communication between the United Kingdom and Iceland using the recently developed system of v.h.f. transmission known as "forward-scatter." This U.K.-Iceland link is one in a long-range system installed by the U.S. Air Force for communication between Britain and America. The terminal station is situated on an escarpment of the Chilterns at Kingston Blount.

Two-way communications will be carried in single-stage hops from Limestone, Maine, in the United States, to Goose Bay, Labrador, and on through Sondestrom in Greenland to Iceland and the United Kingdom. Thule, the U.S.A.F. base, and Narsarssuak in Greenland are also brought into the system by similar long-range v.h.f. links, which, throughout, will be able to carry at least eight teleprinter circuits and one speech channel.

"Forward-scatter" at v.h.f. makes use of the D-layer of the ionosphere, and the power radiated is usually about 50 kW from highly-directional arrays, diversity reception being employed.

Paris-Brussels Automatic Telephone

Starting in April, telephone subscribers in Paris and Brussels will be able to dial one another directly. It is hoped to extend this facility to other European capitals, and London may be the next city to be linked with Paris by an automatic telephone service. The British G.P.O. is also working on the installation of automatic trunk dialling, and it was recently stated that Bristol will be the first city in Great Britain to be linked thus with London.

64

DESIGN OF MAGNETIC RECORDING AND REPRODUCING EQUIPMENT FOR DOMESTIC USE

With Special Reference to Stereophonic Reproduction*

by

M. B. Martin and D. L. A. Smith, B.Sc. (Eng.) (Associate Members)†

Read before a meeting of the Institution in London on January 5th, 1956
In the Chair: Mr. H. J. Leak (Member)

SUMMARY

The design criteria of domestic recorder-reproducers and high quality tape reproducers, the problems encountered and practical results obtainable are dealt with. Reference is made to the standards for tape records. The problem of stereophonic reproduction in the home is discussed and a practical reproducer described which uses twin-channel magnetic tape.

1. Introduction

In recent years the development of magnetic tape recording has made it a more practical proposition to popularize sound recording in the home. It is probable that gramophone records will remain the most popular way of buying commercial recordings for some years, but without much doubt when it comes to home recording and to providing the highest possible fidelity reproduction from studio recordings, there is no serious competitor to magnetic tape. In addition to the considerations of distortion and signal-to-noise ratio, with tape it is a comparatively simple matter to provide commercial stereophonic recordings.

In the field of professional sound recording the use of magnetic tape has made great changes in the techniques of recording gramophone records, sound recording for films and the recording of programmes for radio broadcast. The features of a tape system which have led to its supremacy in these fields may be recapitulated briefly as follows:

(*a*) By the use of completely separate record and replay channels it is possible to monitor the actual recording and directly compare it with the incoming signal whilst the recording is being made.

(*b*) Once the original is made there is no further processing required to produce a Master Record.

(*c*) The bandwidth covers the whole of the audible range without any great difficulty, due to the absence of resonances associated with electro-mechanical transducers.

(*d*) The signal-to-noise ratios obtainable practically are high, and permanently so.

(*e*) Editing is a comparatively simple matter.

(*f*) Distortion is low and does not increase as the recording progresses, in the way that it does from a pickup scanning a disc.

(*g*) Rehearsals can be recorded and then erased; the tape can then be re-used.

(*h*) The recording does not deteriorate due to repeated playing.

(*i*) It is possible to obtain many effects by multiple "dubbing" without serious loss in quality at each operation.

(*j*) The tape is not particularly susceptible to mechanical damage.

The average user of a domestic recording machine is not in a position to make music records of, say, long-playing disc record standard, even, if only because he has not an orchestra and studio at his disposal. However, with the advent of f.m. transmissions, the source is there if he cares to run the risk of violating the copyright laws!

D

2. Mechanical Design

When the specification of a machine is published, the stress is usually laid on the electrical performance and the all-important mechanical aspects are usually to a large extent ignored, except that the wow and flutter, spooling time and the number of motors are referred to. At first glance this may appear to be all that is required, yet the most important part of any tape machine is the basic mechanical design. On it depends the whole performance of the machine, and a large part of the manufacturing cost.

Within the scope of this paper only the basic principles of the mechanical design can be discussed. This is because there are many combinations of solutions to the problems involved and in the present state of the art, each complete system requires to be assessed on its own merits. Common to all systems, however, are a capstan-controlled tape motion, some means of spooling rapidly in either direction and a braking arrangement.

The angular speed and diameter of the capstan determine the linear speed of the tape, and here is met one division of the types of mechanism: those which employ a relatively large diameter, low speed capstan, and those with a small capstan driven at a higher rate. The division is significant because low-speed capstans are nearly always driven through a speed-reduction arrangement, whereas it is often possible to drive directly a small capstan.

A reduction drive may involve any of the known methods of mechanics but, in practice, gearing, even with helical, involute toothed wheels, does not give a sufficiently constant-velocity tape motion. The result is a flutter component corresponding to the meshing speed of a given pair of gears. This effect is great when the train is new, falls to a minimum as the gear wheels become run in and then rises slowly but progressively as the wear increases. Belt drives are commonly employed and have the advantages that they are silent, inexpensive and simple to assemble.

A compliant belt is used in order to assist in the filtering of the high frequency components of the driving speed fluctuations. On the other hand, the initial tension on the belt must be high if slipping is to be minimized and this increases the side-thrust on the bearings, thus reducing the life of the bearings and belt and also increasing appreciably the load on the

driving motor; further, a non-uniform belt thickness or hardness causes the velocity ratio to change. A toothed belt followed by a mechanical filter has been used on at least one professional machine, but has not been applied to domestic versions.

The method most successfully used for indirect drives involves a return to the fore-runner of the gear-wheel—the friction disc. This takes the form of a compliantly-tyred idler wheel interposed between the motor shaft and a wheel of the requisite diameter coupled to the capstan (Fig. 1).

The points of contact between surfaces at A and B in the figure should be arranged so that the angle AOB, called the servo-angle, is obtuse; in fact there is an optimum servo angle which represents the best compromise between side-thrust on bearings and a slip-free drive. Although the drive has been shown on to a convex surface for clarity, the better arrangement is for the driven surface to be concave.
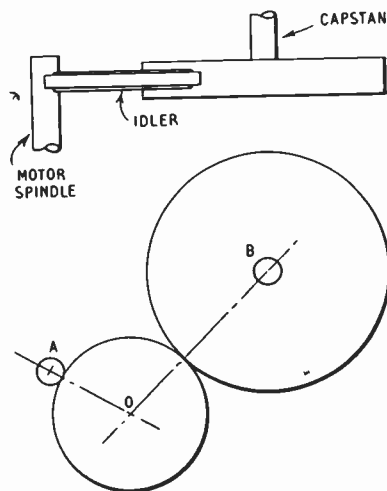

Fig. 1.—*Friction disc reduction drive.*

The detailed arrangement of the deck may vary widely, and this is especially true where the drive is indirect. A direct drive whilst superficially more attractive is much less flexible in its application: for example, multi-speed operation and reversibility are not facilitated. For a direct drive, either a specialized, slow-speed motor can be used or the capstan must
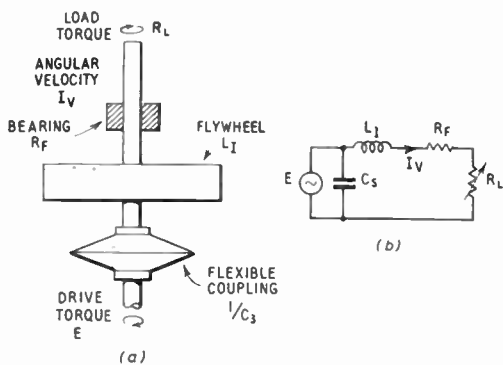
66

Fig. 2.—*Simple mechanical drive filter and its analogue.*

be reduced to a dangerously small diameter (particularly when the playing speed of the machine is low), as a little arithmetic will show: for a speed of, say, $7\frac{1}{2}$ in./sec, two typical combinations of motor speed and capstan diameter would be 375 rev/min/0·38 in. or 1500 rev/min/0·095 in. If the production rate can justify it, a specially-developed slow-speed motor is a proposition. A higher speed motor is smaller, cheaper and more efficient but the associated capstan may then require mechanical support from either a top bearing or a full-length half bearing in order that the bending moment due to pinch-wheel thrust can be resisted. Furthermore, as the flutter component due to lack of capstan circularity or of capstan concentricity is proportional to the error in these parameters, expressed as a fraction of the capstan radius, the allowable absolute error with a small capstan becomes very small indeed. With the 0·095 in. diameter capstan mentioned above, and allowing, say, 0·2 per cent. peak flutter from this source, the total instantaneous deviation from this diameter may be only $0·002 \times 0·095$ in./2 or just under 1/10,000 in.

Whatever the means of connection between motor and capstan, it is invariably necessary to include mechanical filtering to minimize the speed fluctuations due to pole flutter, motor vibration and imperfect regulation. This can be thought of as a low-pass filter which cuts off at as low a frequency as possible and provides a constant-current source (Fig. 2). The reactive components of the filter are represented by the compliances of the coupling and the inertia of a flywheel, but it is doubtful whether

their relative magnitudes are often calculated precisely. The cut-off frequency is usually a few cycles per second.

The situation is analagous to the so-called "brute-force" smoothing filter of a domestic receiver power-supply, where the inductance and capacitance values are made as small as possible consistent with the tolerable h.t. ripple judged subjectively at the loudspeaker. In the same way the motor-to-capstan coupling components are varied empirically, with the flywheel in particular being made as light and as small as possible to the limit set by the permissible wow and flutter. Complementary to this filtering, the variations of the load must be made small, as is indicated below.

Apart from the capstan drive, tape-tensioning and a means of running the tape rapidly in either direction are required. How these things are done depends on the number of motors allowed for in the design. Successful machines have been produced where all the required torques are derived from a single motor; other machines use a capstan motor and a second motor common to both spools, whilst a large number of decks use three motors.
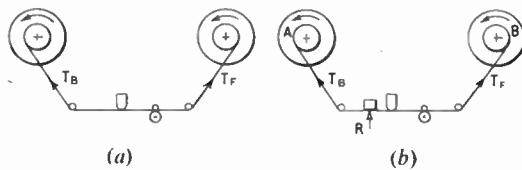

Fig. 3.—*Tape tensions.*
(a) *Ideal frictionless case* $-T_B = T_F$
(b) *Practical case* $-T_B \neq T_F$

Ideally, the tape-tension is constant throughout its length and the tape motion controlled only by the capstan. This implies that the back tension T_b shall equal the forward tension T_f and that there shall be no frictional or variable inertia resistances to the tape motion. This is illustrated in Fig. 3 *a*. In practice, especially for the domestic machine, neither of these conditions is achieved. Friction is present at all the tape-guides and across the heads and there is still a great deal of work to be done to find economical means of tensioning the tape uniformly. At present, what is done, essentially, is to put in a relatively large, and as far as practicable, constant frictional drag at a judicious point (or points) along the tape run.

In this way the effect of randomly-varying tape tension in the vicinity of the head is proportionately reduced, is "swamped" in fact. Such added drag may come from a pad placed so as to bear on the tape as it passes over the face of the head; or on the tape on the leading side of the head (as at R, Fig 3b); or from a labyrinth around which the tape is led. These and similar arrangements are the subject of various patents.

The take-up tension, T_F, comes from a small torque developed at the spool spindle, B, by the spooling motor run on under-voltage (if a separate spool is fitted) or via some form of semi-rigid transmission from the capstan motor, if this is the only motor fitted. Evidently, in both cases, the torque must be inversely proportional to the angular speed of the spool for constant tape tension and as has been remarked, this requirement leaves room for further inventiveness. The same is true of the back-tension, T_B, which may be due to an under-excited motor, a drag brake or to some kind of fluid brake.

Spooling is invariably carried out by disengaging the pinch-wheel and increasing T_F: the voltage to a take-up motor is increased or a "slipping" drive is locked. Although the random tape-speed variations are not significant when spooling, it is nevertheless desirable that here also the tape-tension shall be kept as constant as possible. This is primarily in order that the spooled tape shall unwind smoothly and uniformly when it is next used. This is especially necessary if the spool is to be stored for a period: a loosely-wound tape will slip until the whole of it rests on one cheek of the spool and also the back-tensioning device will not control the tape when it is re-used, whilst an excessively tightly wound tape will become permanently elongated. The sketches of Fig. 4 show the sort of differences between well- and poorly-wound spools. The tension in the spooled tape should be sufficient just to make it self-supporting on its hub—about 80 gm. is right for a normal-thickness, P.V.C. base tape.

One of the ways in which machines vary in performance is in that of braking. The modern tendency is for spooling speeds to be fairly high and the inertia forces to be opposed by the brakes are very large. They are also variable, according to the relative quantities of tape wound on each spool at a given time. Here again the choice of methods is wide, from

band-brakes through fluid brakes to regenerative types. In every case, the braking on each spool must be accurately synchronized and the braking torque on the trailing spool made slightly higher than that on the leading spool. If these conditions do not obtain the tape will be either stretched excessively or will form a loop which will cause it to leave its guides. The cost of perfect braking may be considered excessive for the domestic market and is often abandoned as a feature which does not directly affect the performance of a record/replay machine as such.

We will now attempt to summarize what has been said, or implied, before passing on to the electrical aspects of the design. The capstan drive may be direct or indirect, with the latter preferred for single, low tape-speed work or for multi-speed operation, the motor should have a level speed-torque characteristic and be inherently free from pole-flutter. A mechanical filter will be necessary. The spooling system should have a falling speed-torque characteristic with smooth low-speed performance; it can incorporate suitably designed motors or a semi-rigid transmission. The braking system should respect the mechanical characteristics of the tape and the convenience of the user!

(a) Poorly - spooled tape
LOOSE, IRREGULAR WINDING, TENDING TO SAG WHEN STORED HORIZONTALLY

(b) Well - spooled tape
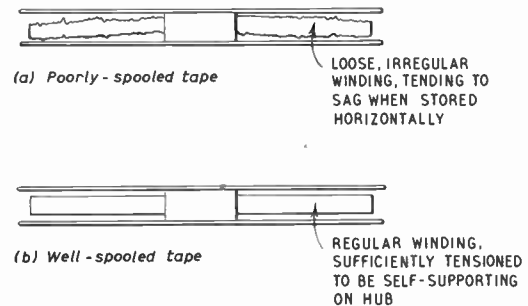REGULAR WINDING, SUFFICIENTLY TENSIONED TO BE SELF-SUPPORTING ON HUB

Fig. 4.—Showing typical influence of spooling quality of machine on spooled tape.

After a balance has been struck between these essential points, i.e. method of driving, spooling and braking, in relation to cost and desired grade of performance, the allocation of expenditure on the system of control will then define the extent of its refinement. A proportion of the cost will often have to be found for the provision of an indicator, while the loading of the machine with the tape should be an uncomplicated process for the user. Control

systems, which may be predominantly mechanical or electrical in nature, are not subject to comparatively rigid principles as is the tape-transport mechanism proper, and so there is more scope to satisfy the salesman who specifies them or the engineer who designs them.

3. Amplifiers and Heads

The electrical performance of any tape equipment depends on the amplifiers and heads. This may appear to be stating the obvious, but it does seem that it is not always appreciated that the head and amplifier must be considered as a unit. In particular, it is by no means certain that two heads, even with the same nominal (replay) gap and inductance, will give the same performance when used with the same amplifier: this is because the high-frequency losses of the heads differ with their design and the care with which they are manufactured. The degree of high-frequency output depends primarily on the magnetically-effective length of the front gap. This length is always greater than the physical thickness of the gap spacer, of course, but is also considerably modified by lack of parallelism of the edges of the gap, and by the gap boundaries not being clearly defined and linear. As secondary effects, the variations of iron losses and self-capacitance will also contribute to differing high-frequency responses.

With professional equipment it is usual to use entirely separate recording and replaying chains in order that instantaneous monitoring can be carried out. This is unnecessarily expensive for domestic use, and it is, therefore, almost universal practice to have one head and amplifier to perform both functions. The disadvantage of this is that in order to obtain maximum frequency response the head has to have a very small front gap, which is not necessarily the optimum for recording purposes, but the cost of providing a second head normally far outweighs any considerations of this kind. Moreover the actual shape of the gap is less critical when the head is not normally required to reproduce tapes recorded by another head.

The variations in performance between different brands of magnetic tape can be quite large, if the tape is not used under optimum conditions. A recorder will record on most types of tape but in general it will only reach its specified performance on the tape around which it was designed; a reproducer will

satisfactorily reproduce any tape which has been recorded to the correct frequency characteristic. The variations between tapes lie in the optimum high frequency bias field required during the recording process; the signal field which will magnetize the tape to the chosen peak distortion level; the resultant residual magnetization at this level, i.e. the sensitivity; the performance with under- or over-biasing; the ground noise and modulation noise levels; and the frequency response of the tape itself.

3.1. Head Design

As the ultimate frequency response of a machine depends upon the effective front gap of the replay head, this is normally as short as possible. On most machines on the market today the gap lies between 0·00025 in. and 0·0003 in., unless sensitivity is the prime consideration, when a gap of 0·0005 in. or 0·0001 in. is commonly used. The front gap of a replay or record head must be straight and the edges must be parallel within very fine limits, the working face should have a high degree of surface finish without the laminations being "smeared" over the gap spacer or the fine film of insulant separating the laminations.

It appears to be fairly general to employ a high impedance head with an inductance of about 500mH; this, however, has several rather major disadvantages:

(a) The bias losses are high due to the high self-capacitance of the winding and the shunting of the head by stray capacitance in the supply lead.

(b) The bias voltage required is high, making the design of the oscillator unnecessarily difficult unless a comparatively low frequency is used. If the machine is to record to 15 kc/s the bias frequency should be at least 75 kc/s and it is preferable to use 85 kc/s or higher.

(c) The self-resonance frequency (including the effect of input capacitance) on replay is usually within the audio range of the machine.

(d) The necessary gauge of wire on the coils is very fine, making winding in production difficult.

Its chief advantage over a low impedance head is the fact that no input transformer is required, which may result in a considerable saving in cost. However, if an input for

a low impedance microphone is required for recording, the head impedance can be chosen so that the transformer can be used for both purposes, e.g. the input transformer for a 4-mH head will give good results with a 20-Ω microphone. A further advantage with a low impedance head is that the record current can readily be derived from the secondary of the loudspeaker output transformer, thus avoiding any switching in the anode of the output stage, which immediately makes switching pulses with a high d.c. content less likely and thus removes a possible cause of severe head magnetization.

The authors have come to the conclusion that low impedance heads are to be preferred in most cases—this opinion arises from their experience. Whichever type of head is used, the maximum voltage output with a minimum of Johnson noise and hum is required. The first requirement is met by winding as many turns as possible on a high impedance head or by increasing the turns-ratio of the transformer with a low impedance head. In both cases, a limit is set by considerations of self and stray capacitance effects, but because of the greater scope for refinements in the winding of a transformer, there is a net gain when one is used. The Johnson noise will be somewhat higher with a transformer, but the hum can be considerably reduced because the screening can be complete and the transformer positioned as required for minimum hum pick-up. It should be noted that to achieve the highest voltage at the grid of the input valve, the external stray capacitance must be made as low as possible— it is interesting also to note that it is very easy to produce a resonance within the pass-band of the amplifier even with a pentode input valve if the anode–grid coupling is increased by careless positioning of anode and grid resistors around, say, a small B9A valve base.

When the gap and impedance of the head have been decided on, the parameters necessary to the amplifier design can be measured. These are:

(1) Frequency response on replay from a constant current recording.

(2) The bias current required for optimum performance on the chosen tape type.

(3) The output voltage at 1 kc/s from a tape recorded at peak level.

3.2. *Equalization*

It is well-known that the equalizers of tape amplifiers have to compensate for two factors— the 6 db per octave rise in replay output with increasing frequency, and the gross high frequency losses due to the loss in the record head, replay head, and tape, and to the imperfect contact between tape and head. The high-frequency correction can be divided between the record and replay processes, but the whole of the low-frequency lift must be carried out on replay because of the saturation characteristic of the tape in relation to the energy content of a normal programme.

At this juncture, the recommended equalization procedure should be referred to: this is defined, for tape speeds from 30 in./sec to 7½ in./sec, in B.S. 1568: 1953, Amendment No. 1, July 1954. This specification which embodies the C.C.I.R. recommendations for programme interchange purposes, defines the bass rise on replay (in terms of the impedance of a series CR-combination of given time constant) and requires that the high-frequency losses in the record and replay processes shall be made good in the respective amplifiers.

It is not essential that domestic equipments should follow these standards, but if they do, a tape recorded on any one manufacturer's machine will replay satisfactorily on any other: this applies particularly for the case of tape records. It is sometimes necessary in the interests of economy to use the same high-frequency equalizers for both record and replay: in this circumstance, the standard cannot be adhered to exactly, but the deviations need not be large.

If separate equalizers are used it is usual that the lift required on record is less than that on replay and on domestic equipment can be achieved by means of a comparatively simple fixed resistance-capacitance combination. The replay equalizer presents a different problem: if the optimum performance is to be reached with each machine made, this lift must be variable in order to meet the likely variations in a production run of heads. If the amplifier noise is to be kept within bounds the response curve should turn over at the highest frequency to which the machine is to replay, and if peaks in the curve are to be avoided the equalizer curve should closely match the loss curve.

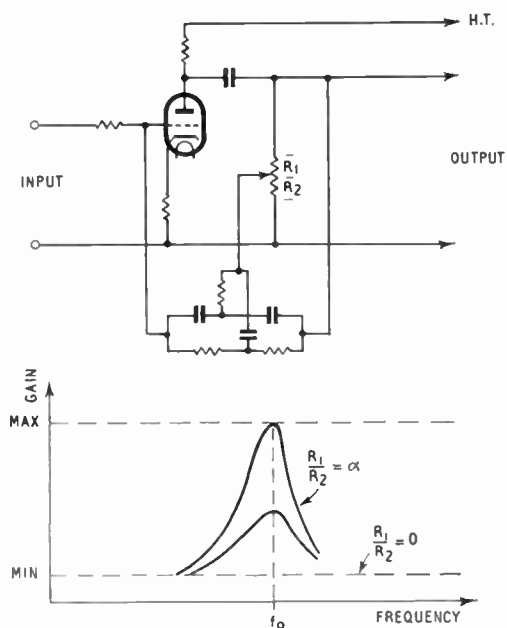There are a variety of circuits in general use for equalization, from resonant LC-circuits with

Fig. 5.—*Parallel-T feedback equalizer stage.*
(*British Patent No.* 696,827)

variable damping to selective-RC feedback circuits. It is preferable to use an RC-network because the components can be held to closer tolerances, and spurious coupling is less.

There are two ways of controlling a frequency-selective feedback network: one, by varying the feedback over the major portion of the frequency response, and the other by varying the feedback at the peak frequency of the network. The second is to be preferred, as the main gain of the amplifier varies only very slightly when the equalizer is adjusted. This point becomes of major importance when dealing with equalizers for stereophonic machines. Fig. 5 shows a simple circuit which illustrates this type of variable lift obtained by the use of a parallel-T filter in the feedback path; the curves below illustrate the range over which the response is variable and which is perfectly adequate for this application.

3.3. Switching

The record/replay amplifier presents a switching problem which can be very difficult, especially if a choice of impedance or function is carelessly made. When the amplifier is switched from record to replay, the head is taken from the output to the input; the loudspeaker is connected directly to the output, instead of through a pad which limits the power available for monitoring; the erase current and bias are switched off, usually by switching off the oscillator; all inputs for microphones etc. are disconnected; and if necessary the equalizers are changed. It needs no elaboration to make the problem apparent: instability and head magnetization through switching pulses. Fortunately, both difficulties can be largely overcome by switching at low impedance to the head. If a high impedance head is used, the magnetization problem is aggravated by having to provide a high alternating voltage to drive it; this usually necessitates shunt feed from an anode via a capacitor which introduces the danger of charging currents being passed through the head.

Instability can usually be avoided by carefully arranging the relationship between switch cards, and carefully routing all leads to the switch. The problem of magnetization usually involves very careful control of the time constants in all circuits which are switched. The control of switching sequence is not usually an easy thing to achieve if the normal type of rotary card switch is employed, and as the contacts tend to move on their rivets, this is not a solution which can be recommended. Muting systems can be included effectively to inhibit head magnetization and also to suppress the audible clicks generated by the switching process.

3.4. Noise

The greatest noise problem on either a tape recorder (or a pure reproducer) is without doubt, hum. At $7\frac{1}{2}$ in./sec the lift in a replay amplifier between 1 kc/s and 50 c/s is about 25db, and the available peak signal from a 4-mH head and 1:40 transformer at 1 kc/s is about 10mV from a half track recording, and 6db greater from a full track. The signal available at 50 c/s is thus about 600 μV, and if the signal-to-noise ratio is to be 60db, the equivalent hum at the grid of the first stage can be no greater than 0·6μV. The only way to achieve this sort of figure is to select the pre-amplifier valve type and layout with care, all earthing must return to one point, and if hum bucking is used it must be controlled in production. The chassis material plays a vital part in the hum fields and currents set up, and great care must be taken before changing the chassis of a successful amplifier from a magnetic

71

to a non-magnetic metal or vice-versa. It is essential that the first stage is a low-noise valve, usually a pentode such as Z729. By using this type of valve with careful input circuit layout, with a small hum bucking coil in the earthy side of the grid lead, this order of signal-to-hum ratio can be achieved without resorting to expensive screening of mains transformers and motors. If an input transformer is used it is advisable that it be astatically wound; if this is not so it will be found necessary to use multiple screening cans.

The other main cause of noise is the hiss resulting from the high degree of lift in the high frequency equalizers. The only ways of mitigating this are by considering the early valve type used and by a careful choice of resistor and capacitor type in any stages before and including the equalizers. The intrinsically better signal-to-noise ratio of the triode would be valuable here but its higher input capacitance is usually prohibitive.

3.5. Level Indicators

A recorder without a positive indication of recording level is almost useless. The minimum requirement is a warning that the overload point has been reached. However, a satis-factory record level indicator should cover the complete dynamic range of the recorder and the relative levels should be clearly readable; it should have as nearly instantaneous rise-time as possible; it should not suffer from under- or over-shoot; and it should return toward zero at a much slower rate than that of its rise.

There are three main types of level indicators in use: meters, "tuning indicators" of the magic eye variety, and neons. Meters are almost entirely confined to professional equipment and neons are rapidly being superseded by tuning indicators. For domestic use, neons suffer from two disadvantages: their striking voltage varies considerably between samples and they give warning of over-load only. Because a meter gives a quantitative measure of level on a steady tone, it is usually chosen for professional use, in spite of the fact that meters almost invariably undershoot on really sharp transients and need quite complicated circuitry to make them follow a dynamic range of 60db.

A "magic eye" indicator can be made to satisfy the requirements mentioned above, with the exception of the first two. The EM84 type of indicator can be made to give a usable

indication of 20db, and the EM34 type with the double eye giving two sensitivities can be made to work over about 36db. The rise time of such indicators is limited only by the forward resistance of the associated diode and the fall time can be controlled to the recommended 19db per second by a simple discharge circuit.

3.6. Oscillators and Erase Heads

It is usual with domestic machines to design the bias and erase oscillator to provide the current to the heads direct. If high impedance record/replay heads are used the oscillator has to provide a high alternating voltage in order to supply the necessary current; typical figures are 150V r.m.s. and 1mA at 50 kc/s. The impedance of the erase head is decided from this in order that the correct erase current is obtained. This unfortunately is usually a wasteful system due to the losses in leads to heads and the heads themselves, and involves the use of fairly large power valves as oscillators in order that the necessary low-distortion current can be obtained.

The current passing through an erase or record head must have negligible distortion, less than 0·5 per cent. total with the even-order harmonic content below 0·05 per cent. With single-ended oscillators this usually involves harmonic rejectors, and with high impedance circuits push-pull oscillators are too expensive.
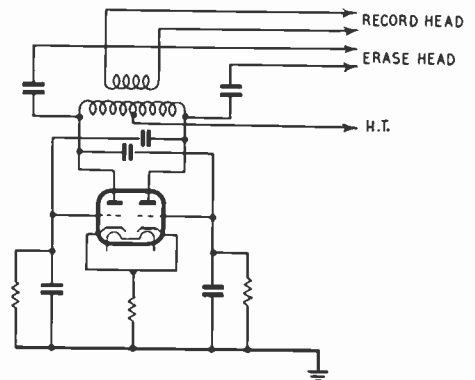

Fig. 6.—Push-pull bias and erase oscillator.

The solution to the dilemma is the use of very low loss and extremely low impedance erase heads, with cores made of ferrites or similar material if possible and an inductance of 1mH or less. If a push-pull oscillator of the type

72

shown in Fig. 6 is used with an erase head having the optimum impedance, the distortion is below the figures quoted and there is ample reserve for erasure at the lower tape speeds even when the valve is a double triode such as a 12AU7, delivering about 1 watt.

4. Tape Records

The standards to which tape records are made have been very clearly defined, and fortunately at the moment one is not faced with the bewildering mass of frequency characteristics which seem inseparable from disc records. The tape speed is $7\frac{1}{2}$ in./sec and the necessary replay characteristic is the C.C.I.R. reproducing characteristic for that speed; namely a bass rise equivalent to that of a series combination of resistance and capacitance with a time constant of 100μ sec together with high frequency lift to compensate for replay head loss.
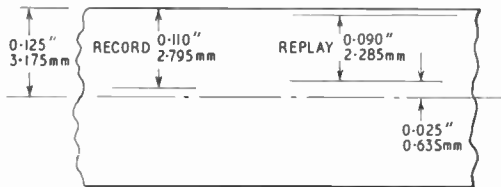

Fig. 7.—*Track dimensions for half-track recordings.*

loss. The recordings are twin-track recorded to conform with the British Standard Specification, that is when played with the active side of the tape away from the observer and the tape passing from left to right the upper track is being scanned. The track dimensions are as given in Fig. 7; the replay track is 0·020 in. narrower than the recorded track in order that no modulation is caused by scanning the edge of the recording. The maximum harmonic distortion on a tape record is approximately 2 per cent. on peak levels, a signal 3db below peak has a distortion content of about 0·4 per cent., and one 6db down, 0·1 per cent. Even if the recording is occasionally allowed to overload, the result is not as distressing to the ear as the tracing distortion produced by the excessive levels frequently found on modern long-playing records.

The basic requirements of a reproducer are, as is to be expected, the same as for any sound reproducer, but in order to obtain the best from tape records, the power amplifiers and loud-speakers should be of a very high standard. The skeleton specification of a tape reproducer can be said to be : —

(1) Tape speed $7\frac{1}{2}$ in./sec.
(2) Wow and flutter better than 0·2 per cent. r.m.s. total.
(3) Replay response to the C.C.I.R. recommendation up to 10 kc/s, with response extended to 15 kc/s if possible.
(4) Dynamic range greater than 50db.
(5) Low distortion amplifiers throughout.

5. Stereophonic Reproduction

A number of attempts have been made in the past to overcome the limitations of a single communication channel, wherein the so-called "auditory perspective" is lost to the listener. Some of these were completely artificial, in the sense that they relied on the adjustment of the frequency and phase characteristics of two or more loudspeaker channels, each of which originated from one common signal input. Demonstrations have been given of true multi-channel systems which have shown the possibilities of quasi-stereophony. However, the commercial establishment of any such system was not facilitated until the advent of tape records. When these appeared, a two-channel system of recording was developed and referred to as the E.M.I. "Stereosonic" system. The recording technique is outside the scope of this paper; we have here to consider the replaying of the Stereosonic tapes as issued.

The consideration of space and cost of a stereophonic reproducer is of some importance. The cost is considerable, being in the order of one-and-two-thirds times that of the equivalent monaural reproducer. The mechanism required by the two is the same, but the cabinets, amplifiers and loudspeakers are duplicated. If full advantage of the system is to be gained it is necessary that the reproducer covers a wide frequency range. A wide frequency range is usually associated with loudspeaker enclosures of large dimensions, of the order of 8 cubic feet; in most homes two units of this size together with a control and amplifier unit would be intolerable, however good the result. With the equipment to be described, an attempt has been made to achieve a wider frequency range within a gross volume comparable with that of a large radiogramophone.
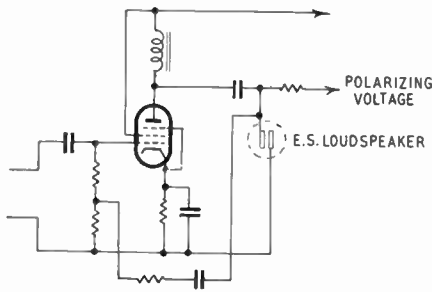
Fig. 10.—*Side amplifier for electrostatic loudspeaker supply.*
(British Patent Application No. 9507/55)

stability type in order to ensure that the gain of all production amplifiers is held within 0·5db of standard, thus obviating the use of individual gain adjustments.

The loudspeakers are discussed later: the bass unit is a 5-Ω moving coil speaker and the treble unit is an electrostatic speaker. The latter is driven from a single valve, choke loaded amplifier whose driving voltage is obtained from the secondary of the push-pull output transformer. This amplifier delivers 140V r.m.s. to the electrostatic unit when the main amplifier is delivering 10 W to the bass speaker. This high frequency amplifier response is controlled by two feedback loops and RC-networks in the grid circuit so that the output is 3db down at 6 and 16 kc/s. The circuit arrangement is shown in Fig. 10.

5.4. *Loudspeakers*

The two main problems with the loudspeaker system were the small volume available for the bass unit, and the problem of avoiding beaming of the high frequencies. After some preliminary experiments it was decided to use an electrostatic speaker from 6 kc/s upwards, this speaker being 24 in. long and $1\frac{1}{2}$ in. wide, bowed over an arc subtending a 60-deg. included angle. The dispersion of sound is very even and the response unusually smooth. Distortion measurements were carried out in comparison with a horn-loaded ribbon, the electrostatic being the better of the two by a comfortable margin.

The bass unit used consists of an elliptical loudspeaker with a $13\frac{1}{2}$-in. major axis, the surround of which is very compliant, giving a fundamental resonance of 20 to 25 c/s enclosed in an airtight box of $3\frac{1}{2}$ ft³. In the undamped

box the system resonance is 52 c/s and after damping the resonance is lowered to 47 c/s. The effect of damping was determined by using very low frequency square waves and observing the waveform of the speech coil current on an oscilloscope. The normal reproducer amplifier was used to drive the speaker, the square waves being applied to the grid of the first stage. The results obtained are shown in Fig. 11.

The speaker is designed to cut off above 5 kc/s, thus a cross-over with the electrostatic unit is achieved without having to resort to filter networks with the attendant problems of ringing under transient conditions. The elliptical unit is mounted with its major axis vertical in order to obtain the optimum horizontal polar diagram up to 5 kc/s. Great care was also taken to eliminate, as far as possible, peaks in the response produced by cone break-up phenomena.
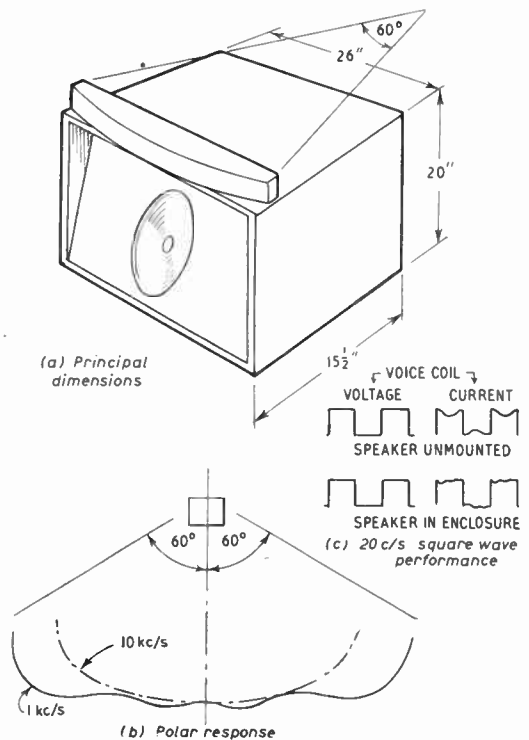

Fig. 11.—*Loudspeaker for stereosonic tape reproducer.*
(a) Principal dimensions.
(b) Polar response.
(c) 20 c/s square wave performance.

The resultant loudspeaker unit, with its associated amplifier, has a good response from 40 c/s to 15 kc/s, and is, in advertising terms, substantially flat from 30 c/s to 18 kc/s. The main drawbacks of the system are the relatively poor efficiency of the bass unit and the fact that if this were increased, the necessary increase in output from the electrostatic unit would become a major problem, bearing in mind the limitations of the existing electrostatic loudspeaker amplifier and unit.

The results obtained from this reproducer as a whole are very good from the point of view of stereophonic definition, the lack of beaming in the high frequencies playing a large part in this. The best results are obviously obtained when the listener is on the centre line between the two cabinets, but good results are obtained over quite a wide area in front of the units. The exact results depend to a large extent on the acoustics of the listening room, but there is no obvious reason why they should not be perfectly satisfactory in most houses.

6. Acknowledgments

The authors wish to thank the Directors of Electric and Musical Industries Ltd. for permission to publish this paper, and for facilities for its presentation.

DISCUSSION

L. Nelson-Jones (*Graduate*): Can the authors give any information on whether ferrite heads are suitable for purposes other than erasure. Secondly, I should be interested to know what type of microphone is suitable for recording stereo tapes, particularly with reference to the matching of their frequency response.

E. A. W. Spreadbury (*Member*): Stereophonic reproduction is often referred to as three-dimensional, but actually it is only two-dimensional, giving the impression of direction in a horizontal plane, and I wonder whether the authors could explain to us what means our ears detect direction in a vertical plane.

I was recently present at a "Stereosonic" demonstration at which the positions of the performers, who were singing an operatic extract, could be followed quite clearly as they moved across the stage, but it appeared that the orchestra was on the stage also, and not in the pit as it is in the theatre. Does this mean that for true realism at least three sources of sound are necessary in stereophonic reproduction before it can give a three-dimensional impression?

D. T. Bennett: What is the relationship between the spacing of the two loudspeakers used for the stereophonic demonstration and the spacing of the microphones for the recording?

A Visitor inquired if any enhancement of the stereophonic effect would be achieved by adding a proportion of each channel together and passing the resultant signal to a centrally-placed loudspeaker.

F. H. Brittain: The authors have made stereophonic sound appear to be very complex and mysterious, and I don't feel myself that this is in fact the case. As an example, I agree that the levels of the two channels have to be balanced to within half a decibel; on the other hand, given a suitable potentiometer and the normal complement of ears, there is no difficulty whatsoever in accomplishing this feat.

It seems to me in addition that several of the specifications of the domestic machine were more stringent than those of the 30 in. professional machine.

I entirely agree that accurate phase relationship is necessary at low frequencies of the order of 50 c/s for good stereophonic positioning in free space. I do not agree that a small living room permits this accurate phase relationship to be obtained, particularly with those frequencies which have a wavelength considerably greater than the size of the room. I do not believe that accurate phase compensation to those frequencies has any meaningful purpose when reproduction takes place in a small living room, the position of low frequencies of musical instruments being given by the medium and high frequency components in their wave-forms.

E. V. T. Taylor: The equipment as demonstrated here is described as being for domestic use but to house these quite large units comfortably would require a very large lounge or living room. Is there at the moment, therefore, any smaller twin channel equipment available for the average size home?

77

No price has been quoted but one might assume that it would be prohibitive, anyway, to the average enthusiast. Mr. G. A. Briggs in his book "Sound Reproduction" quotes certain investigators as saying that "two channel reproduction good to 5 or 6 kc/s is better than single channel reproduction to 15 kc/s". Would it not, therefore, be profitable to all, both in respect of cost and space, if the signals from the recorder were taken through an output system of narrower band-width and even, perhaps, less power?

P. J. Walker: The authors state that it is necessary to control the phase relationship of the two loudspeakers to within 15 deg. at low frequencies for stereophonic reproduction. Now although it is, of course, possible to control the phase of the voltage fed to the speech coil of a loudspeaker, the relationship of the acoustic pressure delivered by the loudspeaker to the phase of the voltage at the speech coil is not predictable. The reason for this is that the phase of the acoustic pressure from the loudspeaker is dependent upon the phase of the accoustic load presented to the cone and this again is dependent upon the position of the loudspeaker in the listening room. This being so, one is driven to the conclusion that either:

(a) 15 deg. is too close a tolerance for the requirements at low frequencies, or
(b) true stereophony at low frequencies is not possible in a normal listening room.

From the practical point of view it would seem that stereophony at low frequencies is very unimportant as compared with that at middle and higher frequencies.

A. Kraicer: I should like to raise a point concerning the use of resonant circuits in the equalizer.

If the effective Q of the resonant circuit is greater than 0.5, then the output of the circuit at its resonant frequency will continue in the form of a damped oscillation after an exciting signal has ceased. The magnitude of this "ring" will depend on the value of Q necessary for satisfactory equalization, which in turn will depend on the heads, the tape, the speed of the tape, the frequency to which one is equalizing, and the division of the equalizing function between the record and the playback circuits.

Could the authors please say if this "ringing" effect is ever noticeable in practice?

AUTHOR'S REPLY

Ferrite heads can be used for recording and reproducing, as has been suggested by Mr. Nelson Jones, but the results obtainable are not as good as those required for high-quality recording and reproduction. The limitation is in the formation of a good working gap. Ferrite heads are used for these functions in applications other than sound recording.

In reply to questions regarding microphones, this paper is concerned only with the reproduction of "Stereosonic" tape records. To give a fair and complete answer to this question would involve the authors in a field of work outside their terms of reference. This information, and data on the relative spacing of microphones and loudspeakers, will in fact be published at a later date in a paper by other authors.

As Mr. Spreadbury says, stereophonic reproduction is two-dimensional. As far as we are aware, no really satisfactory explanation has been given to account for aural detection of direction in the vertical plane, and it is by no means certain that the addition of a third channel will provide true realism. It would appear that the ears, for this function, are assisted by the eyes to a large degree. If one sits in a theatre with the eyes closed and the head still, this effect can be experienced. Under these or similar conditions most people appear to tilt or rotate their head to help in vertical direction finding.

The "Stereosonic" recording technique is essentially a two-channel one and recordings are made on the basis of a twin replay chain such as we have described. A third replay channel which has been suggested would thus be spurious. The absence of a central loudspeaker in no way implies a lack of central image, as can easily be verified by noting the very definite central image when a common signal (as from a single channel tape) is applied to each of the two side loudspeakers.

We agree with Mr. Brittain in that stereophonic sound is not complex and mysterious, we have merely tried to indicate that care has to be taken when designing the amplifiers and loudspeakers for this purpose. With reference to the balancing of gains within half a decibel, besides two ears

and a potentiometer one needs a suitable recording, preferably one with a musical instrument or some other complex sound, recorded so that the resultant image appears in the centre. This procedure is quite satisfactory for individual experiments, etc. However, we were discussing machines designed to be produced by normal factory methods and to be capable of being set up and operated simply by non-technical persons, with the express purpose of providing entertainment in the home.

As Mr. Brittain may know, the ears are unable to locate the source of a steady note (this was shown by Moir and Leslie* at the Brit. I.R.E. Audio Frequency Engineering Convention in 1951) and for this reason one cannot localize a low frequency note in any size of room, if the note is steady. If, however, one listens to a recording of music containing, say, a bass drum played so that the main component of the sound is a very low frequency, one can place the drum very accurately even in a room whose maximum dimension is less than the wavelength of the sound. What matters is the time of phase difference at each ear of the wavefront of the direct sound: thus it is obvious that a mere fraction of a wavelength is sufficient to provide the necessary information. If Mr. Brittain cares to spend a short time with an oscillator, a loudspeaker, an intelligent observer (preferably blindfolded) and a suitable means for switching off the speaker, he should be able to prove this for frequencies below about 1 kc/s; above that frequency the level differences at the ear are the determining factor.

With regard to the matter of specification of the 7½ in./sec. domestic machine and the 30 in./sec. professional machine, we would agree that certain

* J. Moir and J. A. Leslie. "The stereophonic reproduction of speech and music," *J.Brit.I.R.E.*, *12*, 360, June 1952.

results are a little more difficult to achieve when the speed is reduced four times.

Replying to Mr. Taylor, we know of no smaller "Stereosonic" equipment on the market at this time.

The price of the present machine is no greater than many single channel systems giving reproduction of comparable quality. Our experiments have shown that to increase the bandwidth on a two-channel system from the 5 kc/s mentioned to that of this machine is well worth the effort, and as a result of these experiments we are not in entire agreement with the investigators quoted by Mr. Briggs on this matter, with the qualification that noise and distortion on the wide band system are no greater than on the narrow band equipment. With regard to cheaper machines with a lower standard of reproduction, we have no doubt that if the demand exists machines will be made to satisfy it.

In our opinion, Mr. Walker's remarks concerning acoustic load apply only at frequencies at or around the bass resonance of the system, as at frequencies above this the main impedance controlling the phase of the loudspeaker cone is the mass of the cone itself, which is reasonably constant up to a frequency above which phase does not matter. Moreover, we have used a closed-back loudspeaker, hence the loading on the rear of the cone is constant and independent of the room acoustics.

The "ringing" effect referred to by Mr. Kraicer is noticeable in practice if an attempt is made to equalize for the high rate of attenuation at high frequencies caused by a poor head. A practical limit to the amount of equalization applied is about 18db per octave and a lift of about 10db referred to 1 kc/s at 7½ in./sec.

GRADUATESHIP EXAMINATION, NOVEMBER, 1955—PASS LISTS

This list contains the results for *all* successful candidates in the November 1955 Graduateship Examination. A total of 620 candidates entered for the examination. This number included 165 candidates attempting all or parts of the examination in order to complete qualification for election to Graduateship or Associate Membership of the Institution.

LIST 1—UNITED KINGDOM AND EUROPEAN CENTRES

The following candidates, having completed the requirements of the Graduateship Examination, are eligible for transfer or election to Graduateship or higher grade of membership.

ANDERSON, Ralph Taylor. *London.*
ATKINS, Peter James. (S) *London.*

BRUSH, Alberto Valera. (S) *Porthcurno.*

CHARLTON. George Peter. (S) *London.*
CLARKE, Arthur Philip Blake. (S) *London.*

DAVIES, Stanley James. *London.*
DEV, Abinash Chandra. (S) *London.*

FLENS, Jan. (S) *London.*

GREGORY, Frederick Robert. (S) *Cardiff.*
GRICE, Harry Richard Christopher. *H.M.S. Eagle.*

HAMMOND, Harold. (S) *London.*
HURST, Sydney. *London.*

JAMES, James Roderick. (S) *London.*

LEGGETT, Geoffrey Charles. (S) *London.*
LONGSTAFF, William. *London.*

MARTIN, Arthur William. (S) *London.*
MORRIS, Lionel Alfred. *Cardiff.*

PIERCEY, Alfred William. (S) *London*

SANDHAM, Peter Harold. *Bristol.*
SMITH, Edward James. (S) *London.*

THOMPSON, Donald Gordon. *London.*

WAI-CHI WONG. (S) *London.*

ZAFIROPOULOS, Peter. (S) *Athens.*

The following candidates were successful in the Parts indicated.

AHMAD, Jalal-ud-din. (IV) (S) *London.*
ARDITTI, Joseph. (IIIb) (S) *London*
AWOMOLO, George Oludaisi. (I) (S) *London.*
AYLWARD, Patrick. (II and IIIa) (S) *Dublin.*

BACON, Philip Morgan. (II) (S) *Wood Norton.*
BELL, Duncan. (II) (S) *London.*
BLEZARD, Kenneth Brian. (IIIb) (S) *R.N.A.S. Culdrose.*
BONNER, John Stafford. (IIIa) (S) *Newcastle.*
BUNTING, Derek Henry Stanley. (IIIa) (S) *London.*

CAMPBELL, Malcolm. (I and IIIa) (S) *London.*
CHAMPION, Allan Robert. (I) (S) *Manchester.*
CHMIELOWSKI, Jerzy Czeslaw. (I) (S) *London.*
COLLINS, Gerald. (IIIb) (S) *Bristol*
CRIDLAND, William Wyndham. (IIIa) (S) *Manchester.*
CROLE-REES, David George. (IIIb) (S) *London.*

DUNELL, Wilfred Maurice. (I) *London.*

EXARCHOS, Vladimir. (II and IIIa) (S) *Athens.*

FARNWORTH, Geoffrey. (I) (S) *London.*
FEEK, Derrick Townley. (II) (S) *Cardiff.*
FLORIDIS, Georges. (IIIa) (S) *Athens.*

GEORGE, Julian. (I) (S) *London.*
GILVARY, David Francis. (IIIb) (S) *Birmingham.*
GROVES, Peter. (IIIb) (S) *Bristol*
GUNN, Douglas Arnold. (I) (S) *Dublin.*

HAKIM, Maurice Isaac. (I) *Manchester.*
HARIZANOS, Constantin. (IIIa) (S) *Athens.*
HATFIELD, John Keith. (I) (S) *Glasgow*
HISCOCK, David John (I) (S) *Birmingham.*

ISMAIL, Sulman. (I) (S) *London.*

JONES, William Byron. (II) (S) *London.*
JOYCE, Arthur. (I) (S) *Bristol.*

KERSH, Cyril. (I) (S) *London.*
KITAJEWSKI, Ryszard. (I) (S) *London.*
KLIMEK, Georges Eugene. (IIIa) (S) *London.*
KONTOPIADIS, Demetrius. (II) (S) *Athens.*
KOUROUKLIS, Paul. (IIIa) (S) *Athens.*

LAKDAWALA, Homi Feroze. (IIIa) (S) *London.*
LARGE, Douglas Blake. (I) (S) *London.*
LAWSON, Bernard John. (I) (S) *London.*

MAJUMDAR, Amal Kumar. (II and IIIb) (S) *London.*
MARSHALL, Anthony Robert John. (I) (S) *Glasgow.*
MARSHALL, Laurel Everleigh. (IIIb) (S) *London.*
MASSELOS, Christos. (II and IIIa) (S) *Athens.*
MAVROMMATIS, John. (IIIa) (S) *Athens.*
MEHDI, Ch. Ghulam. (IIIa) *London.*
MEIJER, Christoffel Hendrick. (I) *Oslo.*
MOSS, Geoffrey. (I) (S) *London.*

NEIGHBOUR, Kenneth John. (IIIa and IIIb) *London.*
NICE, Richard Keith. (IIIa) (S) *Newcastle.*

PIATEK, Tadeusz Ludwick. (I) (S) *London.*

RAI THAKRAL, H. B. (IIIa) (S) *Birmingham.*
RATCLIFFE, Norman. (II) (S) *Manchester.*
REGE, Damodar Balvant. (IIIb) (S) *London.*
RICHARDS, William Milner. (IIIb) (S) *London.*
RICKERS, Denis. (IIIa) (S) *London.*
ROBSON, Alan. (I and II) (S) *Newcastle.*
RUDD, Dennis. (II) (S) *Bristol.*
RYNKIEWICZ, Arthur Edward. (I) (S) *London.*

SAHNI, Pritam Singh. (IIIb) (S) *Shrewsbury.*
SCHOFIELD, Paul. (II) (S) *London.*
SEXTON, Brian Leslie. (I) (S) *London.*
SIBBALD, John Scott. (I) (S) *Edinburgh.*
SLOOTS, Arnoldus H. J. (IIIa) (S) *Delft.*
SMITH, Charles Edward. (I) (S) *London.*
SOOD, Shiv Dutt. (IIIb) (S) *London.*
STRANGE, Norman Edward. (I) (S) *London.*

TAIT, David Adams Gilmour. (I) *London.*
TAYLOR, William John. (I) (S) *London.*
THORNE, John Frederick. (I) (S) *Bristol.*
TSONIS, Fotios. (IIIa) (S) *Athens.*
TZORTZOPOULOS, Constantin. (II and IIIa) (S) *Athens.*

WALES, Sydney Alfred (II) (S) *H.M.S. Saintes.*
WARIN, John William. (IIIb) (S) *London.*
WEBB, Frederick Kenneth. (IV) *London*

YOUNG, Alexander Robertson. (II) (S) *Edinburgh*
YOUNG, William Thomas. (I) (S) *Belfast.*

(S) denotes a registered Student.

LIST 2—OVERSEAS CENTRES

The following candidates, having completed the requirements of the Graduateship Examination, are eligible for transfer or election to Graduateship or higher grade of membership.

BHALLA, Narindar Singh. *Calcutta.*
BHAT, Manjeshwar Yashwanth. (S) *Bombay.*
BOSE, Arun Kumar. (S) *Calcutta.*
BOWLES, Mervyn Edward. *Sydney.*
BRAUN, Simon. (S) *Tel-Aviv.*

CHANDRA DUTT, Govindan. (S) *Trichinopoly.*
CHARANJIT SINGH. (S) *Agra.*

GEORGE, Panthradil Samuel. (S) *Singapore.*
GRAHAME, Jackson. (S) *Lusaka, Northern Rhodesia.*

HATTANGADI, Bansidhar Srinivas. (S) *Bangalore.*

KUNJUVAREED, V. V. (S) *Madras.*

MAHAJAN, Raj Kumar. (S) *Lucknow.*
MAHLAB-EZRA, Salim. (S) *Tel-Aviv.*

NARAYANA, Anur Krishnamachar. (S) *Bombay.*

PUNJABI, Hari Mangharam. (S) *Bombay.*

RAMACHANDRA RAO, N. (S) *Madras.*
RIESEL, Chaim Manfred. (S) *Tel-Aviv.*

STAMOPOULOS, Charalambos Denis. *Sao Paulo.*

TEJUJA, Mahonlal M. *Bombay.*
TYLER, Graham Reginald. (S) *Toronto.*

VENKATARAMAN, Anthiyur Raju. (S) *Bangalore.*
WADDINGTON, Damer Evelyn O'Neill. (S) *Johannesburg.*
WADHAWAN, Krishnan Lal. (S) *Delhi.*

The following candidates were successful in the Parts indicated.

ABU EL HAJ, Akram. (I) (S) *Amman.*
ABDUL GAFFER, M. (I) *Bangalore.*
ACHUTAN, Madras Gopalan. (IIIb) (S) *Bangalore.*
AGHARKAR, Ratnakar Vinayak. (I and II) *Bombay.*
AHUJA Gojindar Singh. (IIIb) (S) *Bangalore.*
AZAR, Yoram. (I) (S) *Tel-Aviv.*

BALARAMAN, Venwkataraman. (IIIb) (S) *Madras.*
BALBIR PARKASH. (IIIa) (S) *Delhi.*
BANERJI, Sanat Kumar. (I and II) (S) *Delhi.*
BASHYAM, R. (IIIb) (S) *Trichinopoly.*
BENERAGAMA, Don Kingsley Wimalasiri. (I) (S) *Colombo.*
BERLOWITZ, Julius Gad. (I) (S) *Tel-Aviv.*
BHACKA, Sam Jamshadji. (I, II and IIIa) (S) *Bangalore.*
BHAGAT, Shiv Raj Huria. (I) (S) *Bombay.*
BHATIA, Madan Mohan Sugnomal. (IIIb) (S) *Bombay.*
BHATTACHARYA, Dilip Kumar. (IIIa) (S) *Delhi.*
BHOWMICK, Saurendra Nath. (III) (S) *Bangalore.*
BUEHLER, Walter. (II) (S) *Toronto.*
BUTA SINGH. (IIIb) (S) *Delhi.*

CHATTERJEE, Bhabatosh. (IIIb) *Calcutta.*
CHOPRA, Sushil Chander. (I) *Delhi.*
CHUGH, Jyoti Prakash. (I) (S) *Lucknow.*
CORKETT, Raymond John. (IIIa) (S) *R.A.F. Luqa.*

DHALL, Raj Kumar. (II) (S) *Calcutta.*
DHINDAW, Rabindra Nath. (IIIa) (S) *Delhi.*
DIAS, Cyril Francis. (II) (S) *Bombay.*

FITCH, Lawrence William. (I) (S) *Melbourne.*

GARAI, Profulla Kumar. (IIIb) (S) *Calcutta.*
GITTINS, Leonard. (IIIb) (S) *Cairo.*
GOGATE, Balchandra Damodar. (IIIb) (S) *Delhi.*
GUHA ROY, Jayanta Kumar. (IIIa) (S) *Calcutta.*
GURCHARAN SINGH SURIE. (I) *Bangalore.*

HALTOVSKY, Afraim David. (IIIa) (S) *Tel-Aviv.*
HAMILTON, James Henry. (I) (S) *Zomba.*
HANDA, Jagdish Rai. (I) *Agra.*
HERLEKAR, Balvant Vishnu. (II and IIIa) (S) *Delhi.*

JAYARAMAN, Krishnamurthy. (II and IIIa) *Bangalore.*
JEYASINGH, Rajamoncy Daniel. (IIIa) (S) *Bangalore.*
JOSHI, Sivasanker Rajaram. (I) *Madras.*

KAPOOR, M. R. (IIIa) (S) *Calcutta.*
KAR, Saroj Kumar. (IIIb) (S) *Calcutta.*
KHATRI, Dindayal Tamilram. (I) *Bangalore.*
KRISHNAMOORTHY, T. S. (I) (S) *Madras.*
KRISHNAMURTHY, K. S. (IIIa) (S) *Bangalore.*
KRISHNASWAMY, T. L. (IIIb) (S) *Bangalore.*
KUMAR DEVA, Shanti. (I) (S) *Bangalore.*
KUNDU, Sati Nandan. (IIIa) (S) *Lucknow.*

LAKSHMANA RAO, Motamarry. (I and II) (S) *Bombay.*
LALL, Brij Behari. (IIIb) (S) *Delhi.*
LAWSON, Oswald Edward. (I) (S) *Lagos.*

LESSING, Zvi. (IIIa) (S) *Tel-Aviv.*
LUND, Hugh Forsyth. (I) (S) *Durban.*

MADAN, Amrit Lal. (I and II) *Calcutta.*
MALHOTRA, Chaman Lal. (II) (S) *Bombay.*
MANVENDRA SINGH CHAUHAN. (I and IIIa) (S) *Agra.*
MEHTA, Adarsh Kumar. (IIIa) (S) *Delhi.*
MEHTA, Gunvant Jagjivandas. (I) (S) *Bombay.*
MITRA, Somendra Nath. (II and IIIa) (S) *Bangalore.*
MOHAMMAD IBRAHIM. (IIIb) (S) *Madras.*
MUKHERJEE, Parimal Kumar. (I) (S) *Calcutta.*

NAIR, Thekkeppat Krishnankutty. (I) *Bangalore.*
NAMBIAR, Kunhi Kannan. (IV) (S) *Madras.*
NEUMARK, Nathan. (I and IV) (S) *Tel-Aviv.*

OM PRAKASH CHOPRA. (IIIa) (S) *Delhi.*

PANIKER, K. Viswanatha. (I) (S) *Bombay.*
PISHARODY, Unni Krishna. (II) (S) *Bombay.*
PRIT PAL SINGH. (I) *Agra.*

QAZI, Mohammad Akram. (I) (S) *Lahore.*

RAGHAVAN, Narayan. (I and IV) (S) *Bangalore.*
RAJAGOPAL, A. (I) *Delhi.*
RAJENDRA NATH. (IIIa) (S) *Bangalore*
ROBINS, Robin Noel. (II) (S) *Christchurch.*
ROY, Biman Bihari. (IIIa) (S) *Calcutta.*

SACHDEV, Ved Parkash. (I) (S) *Delhi.*
SAMBASIVAN, Krishnamurthy. (IIIa) (S) *Hyderabad.*
SANGHOI, Chunilal V. (IIIa) (S) *Bombay.*
SANKARA RAO, Nagaraja. (IIIa) (S) *Madras.*
SANKER NARAYAN, P. V. (IV) (S) *Madras.*
SAPRU, Kanhaiya Lal. (I) (S) *Delhi.*
SATYANARAYANA, Madabhushi. (IV) *Madras.*
SATYARAJ, Karjagi Vasudeva Rao. (I) *Bombay.*
SEKHRI, Guatam Dev. (I) (S) *Bangalore.*
SEN GUPTA, Rabindra Nath. (IIIa) (S) *Calcutta.*
SESHA, Venkatarama Adi. (I and IIIa) (S) *Delhi.*
SIKARWAR, R. S. (IIIa) (S) *Jabalpur.*
SIVASANKARA PILLAI, Anandan. (IIIa) (S) *Trichinopoly.*
SMOOHA, Menashe. (IIIa and IIIb) (S) *Tel-Aviv.*
SOBTI, Mahinder Nath. (IIIa) (S) *Bangalore.*
SOOD, Parkash Chand. (II) (S) *Delhi.*
SRINIVASAN, Rama Desikachari. (IIIa) (S) *Madras.*
STEYN, Marthinus Jacobus Dewald. (I and II) (S) *Capetown.*
SWAMINATHAN, Kashi Ramamurthi. (IIIb) (S) *Madras.*

TIWARI, Madnukar Eknath. (I) (S) *Bombay.*
THADANI, Hiro K. (I) *Delhi.*

UPPAL, Kanwal Krishan. (I) (S) *Delhi.*

VAISHNAV, P. M. (IIIa) (S) *Bombay.*
VENKATESWARAN, V. V. (I) (S) *Bangalore.*
VISWANATHAN, S. (IIIb) (S) *Madras.*

YOGINDER SINGH JASWAL. (I and IIIa) *Hyderabad.*

(S) denotes a registered Student.

· ATOMIC ENERGY DEVELOPMENT IN GREAT BRITAIN

Atomic Energy Authority's First Report

In July, 1954, the Government decided to transfer responsibility for atomic energy matters from the Ministry of Supply to a non-departmental body, and most of the staff who had contributed to the success of the project in the past continued to serve the new organization. Thus the United Kingdom Atomic Energy Authority came into being with the important asset of many years' experience and achievement.

Much of the work done on atomic energy is subject to strict security, hence the Authority's first Annual Report* issued recently, does not contain a full account of its activities. The period under review has, however, seen the publication of the Government's important Nuclear Power Programme, and a prototype power station is being built at Calder Hall in Cumberland.

Since the period covered by the Report, it has been announced that two more reactors are to be built on this site, and four at Annan in Dumfriesshire, while the Central Electricity Authority has stated that other stations are eventually to be built in Southern England. The Dounreay establishment in the extreme north of Scotland is intended to house several reactors of experimental or advanced types, together with the plants for fabricating their fuel elements and treating them after irradiation. Several reactors are also being built at Harwell for research purposes.

The Authority has at A.E.R.E. Harwell the major establishment for pure and applied research. In the Industrial Group is maintained a Research and Development Branch mainly devoted to applied investigations, and at A.W.R.E. research, both fundamental and applied, for the weapons programme, is carried out. The activities of these laboratories are closely co-ordinated and research on specific projects is frequently shared among all three groups.

In the field of nuclear physics, Harwell is developing an extremely powerful source of neutrons for use in the laboratory, namely a 110 in. synchro-cyclotron. The first stage of a working equipment is to be installed in the autumn of 1956. Design studies have been undertaken for the development of even more powerful accelerators; the machine most exhaustively studied has been a 600-million-volt proton linear accelerator of which a 50-million-volt pilot section is under construction.

In the development of electronic instrumentation for nuclear reactors and processing plant, for ore-prospecting and health protection, the Authority has collaborated with members of the Scientific Instrument Manufacturers' Association, and this has led to a growing export trade. An example of one of the more elaborate applications of electronics is afforded by the electronic analogue computer being developed at A.E.R.E. to simulate the operation of a large atomic power station by showing directly, and so without the need for elaborate calculation, how the essential factors involved change with time and interact with one another. With this machine it is hoped to be able to predict the performance of such a station under very varied conditions, and to study the effects of alterations in design.

Passing to the question of technical staff, the Report says that the Authority is in competition with industry for staff to carry out the development programmes, and, as a body dependent upon public funds, has not the same freedom of manœuvre as industry in the matter of salaries. Although the shortage of technological skill is nation-wide, the Report states that it is essential to stress this point, because the development of atomic energy is unique in the variety of specialized skills which it demands.

Information on Atomic Energy

Unclassified atomic energy information from the U.K.A.E.A. is now supplied to certain libraries which are designated depository libraries. Until recently there were only two such libraries in the United Kingdom, the Science Museum Library in London and the Sheffield Central Library. To provide for the increasing demand for the documents, particularly from industry, the Central Libraries of Birmingham, Liverpool, Manchester and Newcastle, and the Mitchell Library in Glasgow, are now included in the depository library scheme.

Documents deposited in these libraries are freely available to the public, and photostat copies may be supplied at the usual rates on request. The documents are also sent to the Copyright Libraries and to the Patent Office Library, and some of them are on sale through Her Majesty's Stationery Office.

NEW POINTS OF VIEW IN THE DESIGN OF ELECTRON GUNS FOR CYLINDRICAL BEAMS OF HIGH SPACE CHARGE *

by

M. Müller, Dr. Ing.†

SUMMARY

The paper outlines the influences of the anode aperture, disregarded in the theory of the Pierce gun, which lead to an unsatisfactory functioning of such electron guns, particularly in the case of a high perveance. These influences are taken into consideration by imposing a rectilinear boundary on the beam, so that the space charge potential can be simply determined in the electrolytic tank. For this purpose, the anode is made tubular and sharp edged, and besides the Wehnelt electrode, an additional electrode of substantially the same potential is introduced, being mounted in the immediate vicinity of the anode aperture. Simple electrode forms and their favourable positions are given for a wide range of dimensioning. By the introduction of corrections, the dimensioning rules are extended to high perveance values. The magnitude of the current leaving the assumed boundary of the beam because of its initial thermal velocity is determined.

1. Introduction

By the term electron gun is understood an electrode arrangement which accelerates the electrons from an emitting cathode and "shoots" them in the form of a beam having the sharpest possible lateral boundaries, and frequently of very high current density, into a space which is normally free from d.c. fields. Such devices are required in all electron tubes which operate with a flow of electrons in the form of a ray or beam, for example cathode ray tubes, image tubes, electron microscopes and, more recently, tubes for generating and amplifying very high frequency oscillations.

In the last mentioned field of application in particular, the requirement is to produce high currents at relatively low voltages and at the same time a high factor of concentration of the emission current which can be obtained continuously from the normal cathode.

For some time it has been usual to employ, for concentrating the current, concave cathodes, generally having the form of spherical caps from which the electrons start on convergent paths, provided, as shown by Wehnelt, there is arranged around the cathode an additional non-emitting electrode (known as the "Wehnelt cylinder") at or below the cathode potential. The positive electrode situated opposite the cathode has an aperture through which the electron beam emerges from the accelerating space. With these three essential components: concave cathode, Wehnelt cylinder and perforated anode, provided they are appropriately arranged, the construction of current-concentrating electron guns does not offer any considerable difficulty in the absence of specific requirements. Methods are also known which, by extended numerical treatment of existing electrode or field configurations, permit one to make predictions regarding the functioning of guns of any construction.

The object of the project engineer, however, is to design the gun immediately so as to satisfy as far as possible the requirements concerning cathode loading, perveance,‡ concentration ratio, position of the point of maximum current density, etc. This is possible only for very special arrangements, the field configuration of which can be made amenable to calculation.

J. R. Pierce [1,2] made an important step in solving this problem by introducing a stable

‡ "Perveance" is the constant K in the equation for space charge limited emission: $i = KE^{3/2}$.

arrangement, known as the Pierce gun, which whilst applying the principle of "rectilinear flow" is based on the spherical diode of Langmuir and Blodgett.[3] The dimensioning rules developed take into account the spherical flow and the approximate field distortion due to the hole in the anode necessary for the passage of the electrons, in so far as these exert a lens effect on the final electron paths. The field distortions, unavoidably occurring in the acceleration space, are disregarded however. Electrode forms of cylinder segment systems for flat beams by the application of conformal-imaging to approximate functions of the potential are given by Pierce[1] with the same disregard for the field distortions. Likewise Helm, Spangenberg and Field[4] have determined electrode forms for cylindrical beams over a wide range of variations of the parameters by extensive model investigations in the electrolytic tank.

Electron guns constructed on these principles, however, exhibit considerable departures from theory in their behaviour, especially if the high perveance values necessary in modern velocity modulated tubes are to be attained. The difficulties are principally lower currents than expected and high current losses. The cause of this is generally considered to be the initial thermal velocity of the electrons and the inadequate accuracy of the technical construction of such guns. A calculation of the thermal loss current (see Appendix), however, leads to a much lower order of magnitude as does the estimation of the influence of tolerances in a fairly careful construction. In explanation of this discrepancy, therefore, the points of the theory of the spherical electron gun will be investigated in detail, and new dimensioning methods will be given. In doing this, attention will be paid more particularly to all the influences of the aperture in the anode.

2. The Influence of the Anode Aperture

In a spherical capacitor, the field lines are radii of a sphere and the equipotential surfaces are spherical shells (shown in dashed lines in Fig. 1). These conditions also persist when the external shell emits electrons, except that different voltage values, corresponding to the space charge potential, are associated with the equipotential surfaces.

If we now imagine the inner electrode (anode) to be perforated in such a manner that a

spherical sector on angle aperture θ, with respect to the axis of the hole, just projects into the latter, it will be seen at once that the equipotential surfaces, of which one is of course the surface of the perforated inner sphere, are distorted in a manner indicated qualitively in Fig. 1 (solid lines). Since the field lines remain orthogonal to the potential surfaces, some of the field lines previously projecting in the sector θ on the inner sphere will now have been displaced out of it, whilst the others must penetrate the interior of the anode aperture and hence must have become longer. This will have a twofold consequence on the electron flow in the diode: due to the total capacitance of the arrangement having become smaller, the perveance will be smaller, and furthermore a larger proportion of the current outside the aperture will impinge on the anode than before.
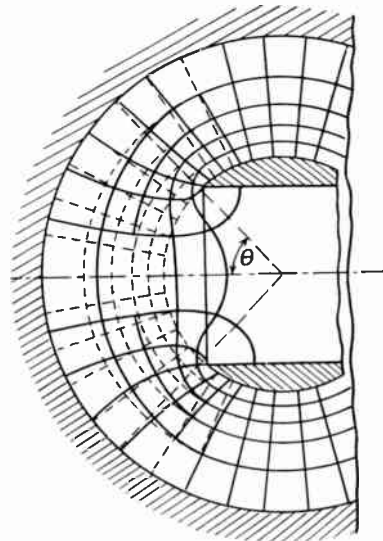

Fig. 1.—*Field distortion in a spherical capacitor due to a hole in the inner electrode.*

In an actual gun, only the cathode surface enclosed by the cone angle θ emits. If, therefore, we construct a gun strictly according to the known dimensioning rules,[2,4] the total current must necessarily be less than that predicted by calculation, and furthermore a considerablt proportion (up to 25%) of the current will be lost by impinging outside the aperture instead of passing through the latter as is desired.

Conditions will become somewhat more favourable if the anode aperture is made larger than corresponds to the spherical sector, as proposed by Huber and Kleen[5] (for avoiding loss of electrons with initial thermal velocity transverse to the beam). Such a step, however, will
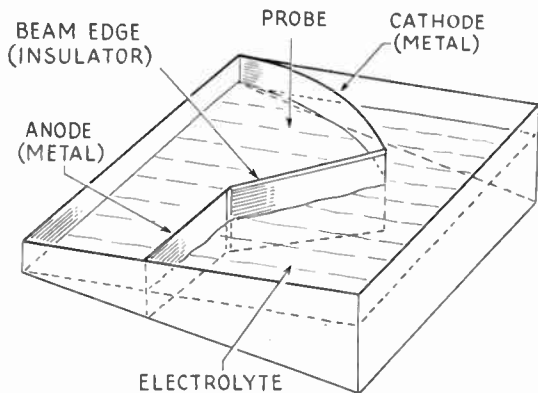

Fig. 2.—Electrolytic tank arrangement for measuring the potential field of a perforated spherical diode.

reduce the perveance still further and result in an increase in the beam diameter, so that the functioning of the gun will depart from the result of calculation. Other correcting steps, however, are possible which re-establish calculable conditions and which, to a good approximation, permit a very simple consideration of the space charge in the electrolytic tank.

3. Compensation of the Influence of the Anode Aperture

The object of the aperture corrections to be employed is to permit the beam boundary, even in the presence of the anode aperture, to extend along the straight side of the conical shell on which the calculation of focusing has been based. This is ensured when the generators of the conical surface are electrical field lines in the diode. Such a field pattern, independently of whether or not it can be realized in actual diodes, can be produced and measured in the well-known manner in the inclined electrolytic tank by the use of a wedge shaped model having a rotational symmetrical arrangement; the boundary of the beam (conical surface) is represented by an insulator, the surface of which in the electrolyte always represents a line of flow (see Fig. 2). If the field is measured in the interior of this model arrangement, potential

lines substantially as shown in Fig. 3 (solid lines) are obtained. It will be seen that in the vicinity of the cathode, the potential lines are distorted in lens fashion as expected but pass into circles near the cathode (and three-dimensionally therefore into spherical shells), just as in the undisturbed spherical diode.

This configuration represents the field pattern of the spherical sector diode with cylindrical anode hole, which in future will be referred to as "perforated spherical diode," in the case of no space charge. If, now, we compare the potential in the vicinity of the cathode with the potential φ_{KK} of the spherical capacitor, which has the radial dependency

$$\varphi_{KK} = U\,\frac{(R_k/R) - 1}{(R_k/R_a) - 1}\ \dots\dots\dots\dots(1)$$

(where $U =$ voltage, $R =$ running co-ordinates, $R_k =$ cathode radius, $R_a =$ anode radius) we find that the field strength in this region has been considerably reduced, just as if we had a spherical capacitor with a smaller inner electrode. This implies also a reduction in perveance in the case of space charge.

Since a complete analytical treatment of the present potential problem presents difficulties which can scarcely be surmounted, the author, following the systematic procedure of Helm, Spangenberg and Field[4] carried out with the
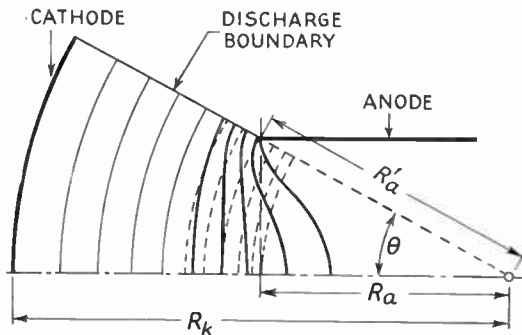

Fig. 3.—Potential field of a perforated spherical diode (indicated by solid lines) and the equivalent spherical segment diode (dotted lines).

above arrangement a series of model measurements of perforated spherical capacitors in the electrolytic tank. The (half) aperture angle θ was varied between $10°$ and $40°$, the radius ratio R_a/R_k was varied between 0·4 and 0·6. The field strength was measured at the cathode

of the perforated spherical capacitor model, the value of which, as is well known, determines the current drawn from the cathode in the case of emission limited by space charge. It was desired to ascertain the dimensions of the equivalent unperforated spherical capacitor possessing the same cathode field strength, that is to say also the same perveance. It was found that, contrary to what has previously been assumed, the decisive factor is not the real distance R'_a of the anode from the centre of curvature of the cathode, but only its projection R_a on the axis of the hole (see Fig. 3). If we denote the perveance of the perforated diode by P_{LD}, and that of the equivalent spherical diode P_{KD}, we may write

$$P_{LD}\left(\frac{R'_a}{R_k}\right) = P_{KD}\left(\frac{R_a}{R_k}\right) \dots\dots\dots\dots(2)$$

where $R_a = R'_a \cos\theta$. It is here assumed that the inner wall of the anode which now acts as a real electrode, is of cylindrical shape, and has an axial length at least equal to the aperture diameter. If a conical hole is selected, the perveance alters; it is larger if the hole becomes narrower from the entrance onwards and smaller when it becomes wider.

Equation (2), according to experimental results, is applicable for $R_a/R_k \leqq 0.6$ and $\theta \leqq 30°$, except for errors in measuring accuracy of about 2 per cent. If these corrections are disregarded, however, errors of almost 30 per cent. are obtained for the perveance. If the aperture angle is made still larger, appreciable field distortions occur at the cathode which cause uneven current loading of the cathode (increasing from the centre of the cathode to the edge), these being greater as R_a/R_k and θ become greater. The use of the correction according to Equation (2) is rational up to about $\theta = 40°$, when the emission perveance becomes about 10 per cent. greater than that expressed by equation (2).

4. Space Charge Potential of the Spherical Diode

By the steps described above, the spherical sector-shaped discharge space is rendered independent of external space. This means merely that a potential is assumed at that place which renders possible an exactly rectilinear beam boundary. With this and from the knowledge of the capacitor field, which can be obtained by tank measurements, it is possible to give in a very simple manner, at least

approximately, the potential field with space charge.

In arrangements in which the field lines are rectilinear, as in plane cylindrical or spherical systems, the electrons follow the field lines (rectilinear flow). In this case, the equipotential surfaces in the absence of space charge will necessarily also be equipotential surfaces in the discharge bounded by space charge. In the latter case, they merely have another lower potential value according to the space charge law.

The perveance of the spherical diode is a function only of the ratio $R_a/R_k = \rho_a$ of anode to cathode radius of curvature and was determined by Langmuir and Blodgett.[3] With the auxiliary magnitude α determined there we have

$$\frac{I}{U^{3/2}} = \frac{16}{9}\pi\varepsilon_0\sqrt{2\eta}\frac{1}{(-\alpha)^2} = \frac{29.35}{(-\alpha)^2}10^{-6}. \dots(3)$$

The dimensionless magnitude $(-\alpha)^2$ is tabulated in Ref. 3 as a function of R_k/R_a. The space charge potential is therefore proportional to the magnitude $\varphi^* = (-\alpha)^{4/3}$, which in view of its importance has been recalculated in the following Table 1 as $\varphi^* = f(\rho)$ in the pertinent range of $\rho \geqq 0.4$. The potential φ_{KD} of the spherical diode is obtained from this as

$$\varphi_{KD} = U\frac{\varphi^*(\rho)}{\varphi^*(\rho_a)} \dots\dots\dots\dots(4)$$

where U is the anode voltage, and $\varphi^*(\rho)$ is the continuous Table value corresponding to R/R_k and $\varphi^*(\rho_a)$ is the fixed Table value corresponding to R_a/R_k.

Table 1

Normalized Space Charge Potential of the Spherical Diode

(According to Langmuir and Blodgett[3])

ρ	φ^*	ρ	φ^*	ρ	φ^*	ρ	φ^*
0.4	1.328	0.5	0.828	0.6	0.507	0.85	0.096
0.42	1.208	0.52	0.750	0.65	0.390	0.9	0.053
0.44	1.098	0.54	0.684	0.7	0.295	0.95	0.019
0.46	0.995	0.56	0.622	0.75	0.215	0.98	0.0055
0.48	0.915	0.58	0.563	0.8	0.149	1.00	0

In the perforated spherical diode, the electrons now depart appreciably from the field lines in the vicinity of the anode, that is to say where their space charge, in accordance with the velocity already attained, can exert only a

86

slight deviating influence on the field pattern. It must therefore be possible to construct, to a useful approximation, the space charge potential field of the perforated spherical diode from the capacitor potential field, if the potential lines are given a voltage which is lower by the difference of the space charge potential according to Equation (4) and Table 1 from the capacitor potential according to Equation (1) of the spherical diode.
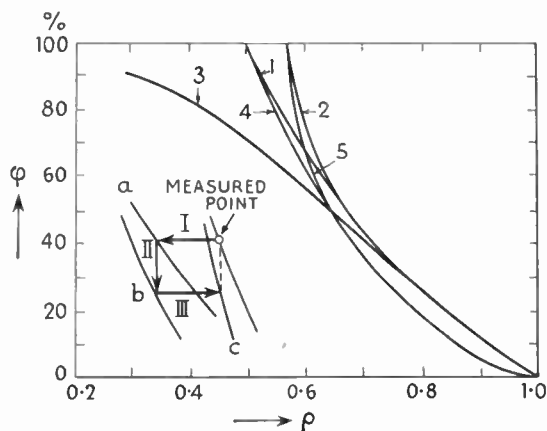

Fig. 4.—*Construction of the potential variation with space charge of the perforated spherical diode:—*

a, 1. *Potential of equivalent spherical capacitor (theoretical).*

2. *Potential at edge* ⎫ *of perforated spherical*
3. *Potential on axis* ⎰ *capacitor.*

b, 4. *Space charge potential of the spherical diode (theoretical).*

c, 5. *Constructed space charge potential at the discharge edge of perforated spherical diode.*

This method is applied and explained in Fig. 4. This figure shows by way of example the potential variation in the perforated spherical diode at the edge and on the axis of the spherical sector, as a function of the radius ratio ρ.

5. Configuration of the Electrode Outside the Beam

In Section 2 it was assumed that the boundary of the beam, corresponding to the tank model in Fig. 2, was formed of rectilinear peripheral lines of the spherical sector. This is the case if along this desired beam pattern the electrical field strength transversely to it disappears, or what amounts to the same, if the equipotential surfaces pass through the beam surface at right-angles thereto.

The potential in the beam at this surface as a solution of the Poisson differential equation, in the presence of space charge, equals that in the outer space, as a solution of the Laplace equation without space charge. The potential function of the inner space is given by the electrodes confining the space (spherical cap cathode, inner wall of the anode aperture) and by the space charge. The potential function of the outer space must now be selected so that at the beam boundary it is identical with that of the inner space and its gradient is radial. This boundary condition can only be satisfied by suitable choice of the other boundaries, namely the electrode form in the outer space. The decisive factor, therefore, is the space charge potential at the beam boundary which is shown in Fig. 4 for one example.

This potential variation, as a function of the radial co-ordinate has at the cathode, as in all space charge diodes, an initial slope of $(1-\rho)^{4/3}$. Bearing in mind that curved arrangements can be regarded as plane over small ranges, and that plane potential problems are solved by the laws of conformal transformation, it is possible to state at once that the initial gradient of the potential can be produced by an electrode at cathode potential, if this adjoins the cathode directly and makes with the beam boundary an angle which is three-quarters of a right-angle.

The lines of the constant, real and imaginary parts of the complex variable z are orthogonal to each other and can be represented alternatively as potential or field lines of a plane potential field. The potential function property is retained by any other analytical configuration function $w = f(z)$. If the lines $\text{Re}(z) = \text{constant}$ are denoted as potential lines, the real axis represents a field line and an advance along it by the amount dz signifies a definite potential step. To this potential step there corresponds on the mapped curve of the real axis in the w plane an advance by the amount dw. The differential quotient dz/dw is thus a measure of the field strength. If in an image curve x a potential distribution is produced according to the power law $\varphi(x) = k_1 x^n$, a useful transform function is $w = k_2 z^{1/n}$. This function leaves the zero point, maps the real axis on the real axis and converts the imaginary axis which was a potential line, into a likewise rectilinear image curve, namely the radius vector $r.\exp(j\pi/2n)$,

which forms the angle $\gamma = \pi/2n$ with the real axis.

At the anode, the angle γ of the equipotential surface (inner wall of the anode hole), which is decisive for the beam flow, is given by the predetermined beam boundary. For the case of the cylindrical hole is just $\gamma = 180° + 0$ If similar considerations to those given above are applied to the anode vicinity, we see that the potential in the anode vicinity can be described by

$$\varphi(R)_{R \to Ra} = \varphi_a \left[1 - \left(\frac{R}{R_a} - 1 \right)^{\pi/2\gamma} \right].$$

Thus, with the infinity tangent, there tends to be just as high a power against the anode potential as the inner wall of the anode hole is inclined by right-angles away from the beam edge. Anode forms, shown in existing publications (without information on the configuration in the interior of the aperture) as having a potential surface impinging from the outside perpendicularly on the assumed beam edge and producing a potential variation with finite tangent, thus appear to be quite unsuitable for compelling the rectilinear flow on which the calculation is based. In order in fact to obtain potential surfaces in the vicinity of the anode which are orthogonal to the assumed rectilinear beam edge, the anode electrode in the outer space should be inclined backwardly to the beam edge by the same angle γ as it is inclined internally in mirror image fashion.

Unfortunately, this theoretical requirement cannot be realized, because for this purpose a perforated anode electrode of considerable negative wall thickness would have to be used. The outer wall of the electrode, however, will be turned back at as great an angle as possible to the beam edge, which necessitates the use of a sharpened anode, shaped like the cutting edge of a punch.

The extreme slope of the potential variation obtainable in this way will, however, be less adequate for larger aperture angles in the diode. There remains in this case only to select the potential variation in the outer space to be as close an approximation as possible to that given in the beam edge, by making the two agree at individual points. In principle, for this purpose, a number of other electrodes could be introduced between cathode and anode, with a suitable potential imposed on them. In practice, it will generally be sufficient to provide a single

electrode of suitable form substantially at cathode potential at a suitable place in the vicinity of the anode aperture.

This additional electrode close to the anode has a similar object as the additional electrode in the vicinity of the cathode first introduced by Wehnelt. As is well known, by producing a mirror image of the vanishing field strength at the cathode, the Wehnelt electrode prevents spreading of the beam in the vicinity of the cathode; the "focusing or concentrating electrode" in the vicinity of the anode is intended to prevent spreading of the beam before it passes the aperture by approximation of the field strength which is increasing beyond all limits at the anode aperture. The action of this additional electrode may also be described by stating that, due to its close position, it absorbs the field lines starting from the outer surface of the anode electrode. Hence the cathode only receives field lines which come from the interior of the anode and therefore none of the electrons starting from the cathode (without initial velocity) can impinge externally on the anode.
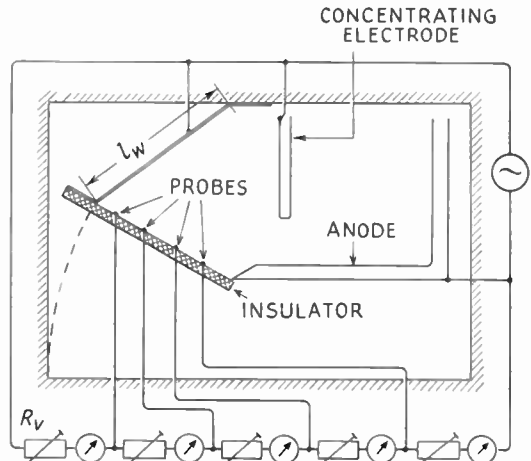

Fig. 5.—Adjustment of the concentrating electrode in the electrolytic tank.

We thus obtain electrode forms shown as a tank model in Fig. 5. The Wehnelt electrode at cathode potential has the form of a frustrum surface of aperture angle $67.5° - 0$. The length of the surface l_w in practice need not be greater than the distance between anode and cathode. The focusing electrode in the vicinity of the anode, which may conveniently be also adapted

for cathode potential and which if desired may be integral with the Wehnelt electrode, has in this case for reasons of simplicity the form of a plane perforated disk; the aperture of the disk is approximately twice the diameter of the anode aperture and is set back from the anode aperture by about one third of the diameter of the latter. These numerical data are applicable
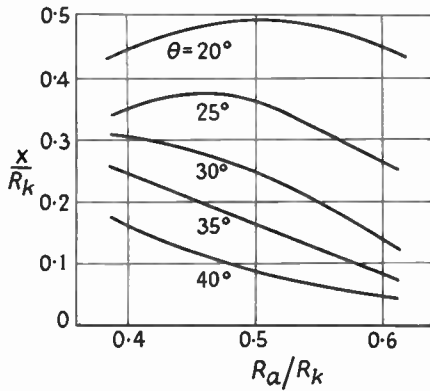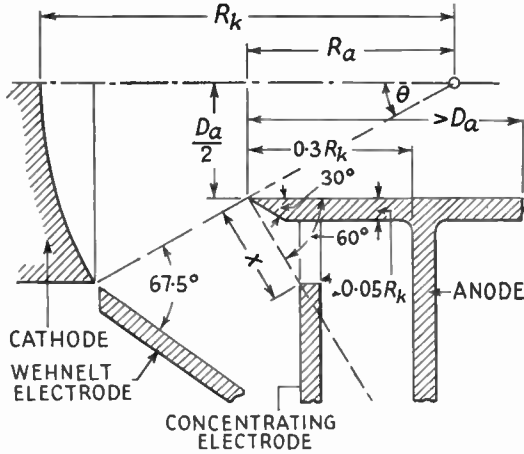

Fig. 6.—*Positions of the concentrating electrode for* $20° \leqq \theta \leqq 40°$ *and* $0.4 \leqq \rho_a \leqq 0.6$.

for the examples shown; generally speaking, the closer the focusing electrode is mounted to the anode aperture, the larger will be the gun aperture angle θ. If the focusing electrode is electrically separated from the Wehnelt electrode and cathode, it is possible at the same time to vary its position in the completed gun, the potential imposed on it being varied.

Useful positions of a focusing electrode at cathode potential were determined in the electrolytic tank over the range of variation of the parameters necessary in practice. The result is shown and explained in Fig. 6. The method employed by the author is shown in Fig. 5. Measurement takes place in the well-known manner[4] in an inclined electrolytic tank. The insulating partition representing the desired rectilinear beam edge is provided on the outside with fixed probes, the measuring instruments are connected by a multiple bridge circuit, as distinct from the method described[4], so that in equilibrium they do not take any current from the probe. In this case, they are not connected in the zero branch of the bridge because a.c. instruments are insensitive at the zero point due to their square law readings. Instead of very high resistance valve voltmeters it is possible to use commercial moving coil instruments, which have merely to satisfy the condition of giving the same deflections when the same current passes through them. Their series resistances R_v are adjusted so that they give a predetermined deflection, the same for all instruments, on applying to each pair of adjacent probes the potential difference obtained from measurement and the construction of the space charge potential in the interior (see Section 3). On applying the measuring voltage to the tank electrodes (Wehnelt electrode, anode) without electrolyte, all the instruments should therefore give the same reading. If the electrolyte is now poured in, the instruments in general will show different readings, the difference being greater, according to the amount the probe potential differs from the predetermined potential. The form and position of the "concentrating" electrode at cathode potential are now varied until all the instruments again show the predetermined deflection, at least approximately.

This requirement is most difficult to satisfy in the immediate vicinity of the cathode (theoretically not at all, see above). It is possible, however, to arrange matters so that an unavoidable deviation of the potential of one probe is compensated by an intentional opposite potential at the adjacent probe so as to ensure in the finished gun minimum deviation of the edge electrons from the desired rectilinear path.

If now the insulating partition is removed from the tank, the entire potential pattern can be measured in the absence of space charge and

there are obtained families of curves as shown in Fig. 7, characteristic of the electron gun for the production of rectilinear flow.
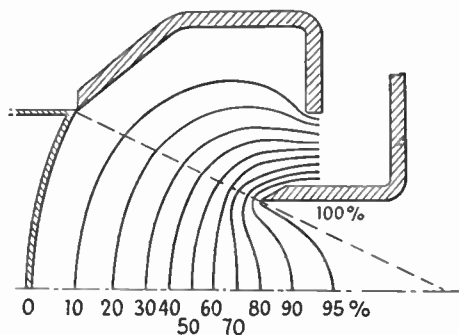

Fig. 7.—Characteristic potential field (without space charge) of an electron gun for producing a nearly rectilinear discharge edge.

6. The Lens Effect of the Anode Aperture

On the assumption of freedom from space charge of a plane-parallel electrode system, one electrode of which has a circular hole, in the field disturbance range of which the potential varies little, it is possible to derive in the range of applicability of the Gaussian dioptrics (rays in the vicinity of the axis) the refractive power of the field disturbance (see for example Rothe and Kleen[6]). If in this case the diaphragm at the potential U_2 separates an accelerating space of the field strength E_1 from a field-free space, the diaphragm acts like a diverging lens with the focal length

$$f_0 = -4 \frac{U_2}{E_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots(5)$$

This focal formula, despite the many limitations on which its derivation is based, is usually also employed for the anode aperture of the spherical diode for want of better, the anode voltage being substituted for U_2 and the field strength (with space charge) directly at the anode of the spherical diode, now assumed to be undisturbed, being substituted for E_1. It will be seen that such a procedure can only be justified if it allows the real state of affairs to be adequately described, at least for small aperture angles θ.

This fact can be confirmed if electron path constructions are carried out in the usual manner in the space potential field obtained as described (see Section 3). Such path constructions carried out systematically by the

author showed that equation (5) may be regarded as valid up to approximately $\theta = 20°$. In this range, lens errors are also still insignificant. If larger aperture angles are employed, it is found that the refracting power becomes stronger and aperture errors of the lens occur. From a number of investigations extending up to $\theta = 40°$, it was possible to show that the beam centre, in which at least 50 per cent. of the current flows, is subjected to scattering which may be well described by the corrected focal length formula

$$f = f_c \cos 0 = -4 \frac{U_a}{E_a} \cos 0 = 4 R_k \frac{\varphi^*(\rho_a)}{\left(\dfrac{d\varphi^*}{d\rho}\right)_{\rho = \rho_a}} \cos \theta \quad \ldots(6)$$

while the rays near the edge are refracted less strongly. This is valid for an ideal rectilinear beam edge in the accelerating space, which according to Section 5 cannot be realized in practice. The unavoidable deviations from this ideal case may, however, have the effect of reducing the lens errors. (The question as to whether the lens errors can be substantially eliminated by a slight widening of the anode aperture as compared with the case shown in Fig. 3 or Fig. 6 cannot be pursued here.)

7. The Space Charge Spreading of the Beam

The refraction produced by field distortion in the vicinity of the anode occurs to some extent in the cathode-anode space but also occurs, especially for rays near the edge, after passing the plane of the anode aperture, in the region therefore where the space charge potential, as determined according to Section 3, is certainly subject to errors. A consequence of this approximation is that the electron paths appear to leave the accelerating field rectilinearly towards a point C on the axis. The space charge effect in this region can, however, be obtained by superposition of the space charge forces already considered in the accelerating field and those of a beam of circular cross-section and homocentric initial velocity in an otherwise field-free space.

For the tangent at the beam edge of radius r, we obtain, after certain limiting assumptions which normally can be regarded as satisfied (see ref. 7, equations 11 or 21-23)

$$\frac{dr}{dz} = K \sqrt{\log_e \frac{r}{r_0}}, \text{ where } K = \sqrt{0.0303 \frac{I}{U^{3/2}} \cdot 10^6} \quad \ldots\ldots\ldots\ldots(7)$$

90
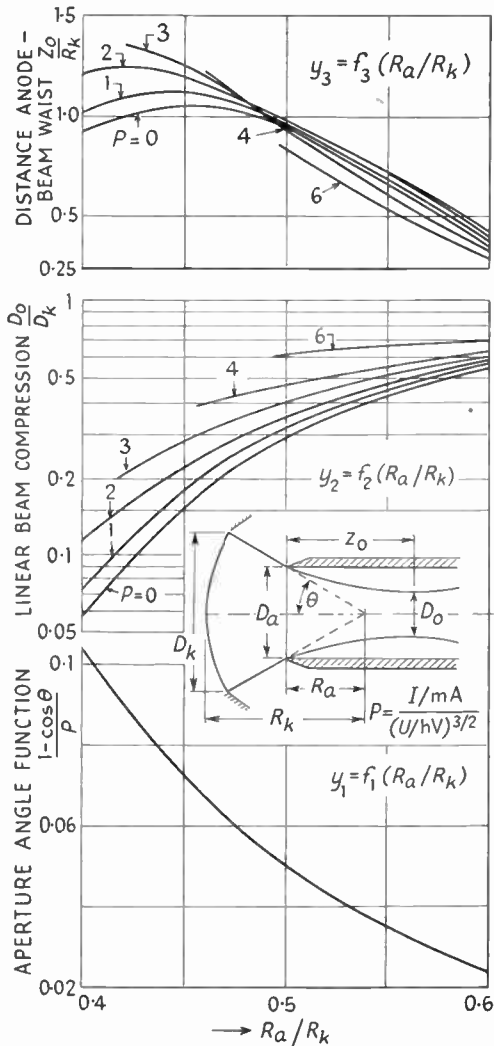
Fig. 8.—*Basis for dimensioning electron guns with aperture angles θ to $40°$.*

(r_0 represents the minimum radius, or "beam waist," which the beam can assume in the presence of space charge spread.)

At the anode aperture $r = r_a$ and we have

$$\left(\frac{\mathrm{d}r}{\mathrm{d}z}\right)_{r=r_a} = \frac{r_a}{c} = K\sqrt{\log_e\frac{r_a}{r_0}}\quad\ldots\ldots(8)$$

where c is the distance of the "crossover" from the anode aperture, i.e. the point at which the electrons intersect the beam axis if they carry

no charge. We thus obtain the linear beam compression from the anode to the beam waist

$$r_0/r_a = \exp\left[-(r_a/cK)^2\right]\quad\ldots\ldots\ldots(9)$$

In order to apply the conditions discussed elsewhere[7] for magnetic focusing, it is necessary to know the position of the beam waist. In order to obtain the distance z_0 of this point from the anode aperture, equation (7) must be reintegrated

$$\frac{z_0}{r_0} = \frac{1}{K}\int_1^{r_a/r_0}\frac{1}{\sqrt{\log_e(r/r_0)}}\mathrm{d}\frac{r}{r_0}\qquad(10)$$

If in addition we write with equation (8)

$$\sqrt{\log_e(r/r_0)} = r/cK = x$$

we have

$$r/r_0 = \exp x^2 \text{ and } \mathrm{d}\frac{r}{r_0} = 2x\exp x^2\,\mathrm{d}x\quad\ldots(11)$$

Integration has to be performed from 0 to $x = r_a/cK$ and we obtain:

$$\frac{z_0 K}{r_a} = 2\exp(-x_a^2)\int_0^{x_a}\exp x^2\,\mathrm{d}x\ldots\ldots(12)$$

The right hand side of equation (12) can be calculated by series expansion.

8. Dimensioning Rule

The results so far obtained, taking into consideration the above-mentioned corrections, can be comprised by transformation in a single graphical representation depending upon the radius ratio $R_a/R_k = \rho_a$ of the perforated spherical diode and its perveance. The perveance p of the spherical sector of angle θ is $\frac{1}{2}(1-\cos\theta)$ times that of the entire spherical diode. With equation (3) therefore

$$p_\theta 10^6 = P_0 = 14\cdot 67\frac{1-\cos\theta}{(-\alpha)^2}\ldots\ldots(13)$$

$(-\alpha)^2$ is a function of ρ_a and is tabulated in ref. 3;

$$y_1 = \frac{1-\cos\theta}{P} = \frac{(-\alpha)^2}{14\cdot 67} \cong 1\cdot 89\exp(-7\cdot 25\,\rho_a)$$
$$\ldots\ldots\ldots\ldots(14)$$

is plotted in Fig. 8 as a function of R_a/R_k.

Curves of the linear beam compression are shown as ratio of the beam waist diameter D_0 to the cathode diameter D_k as y_2 in Fig 8

(centre); curves are also shown of the distance z_0 of the beam waist from the anode aperture, over the cathode radius of curvature R_k, with the beam perveance as parameter as y_3 in Fig. 8 (top).

These values are obtained from equation (9) or (12) and (13) by

$$r_a = R_k \rho_a \tan \theta = R_k \rho_a \frac{\sqrt{1-\cos^2 \theta}}{\cos \theta} \ldots (15)$$

According to equations (13) and (14)

$$\cos \theta = 1 - y_1 P$$

and

$$\sqrt{1 - \cos^2\theta} = \sqrt{2y_1 P} \cdot \sqrt{1 - \tfrac{1}{2}y_1 P}$$
$$\cong \sqrt{2y_1 P}(1 - \tfrac{1}{4}y_1 P) \qquad \ldots\ldots\ldots(16)$$

The magnitude x_a is thus

$$x_a = \frac{r_a}{cK} \cong \sqrt{\frac{2y_1 P}{0 \cdot 0303\ P}} \frac{\rho_a R_k}{c} \frac{1 - \tfrac{1}{4}y_1 P}{1 - y_1 P}$$
$$\ldots\ldots\ldots\ldots(17)$$
$$\cong 8 \cdot 14 \sqrt{y_1} \frac{\rho_a R_k}{c} \frac{1}{1 - \tfrac{3}{4}y_1 P}$$

if K is substituted according to equation (8).

For the image distance c of the diverging lens of the anode aperture we obtain according to the laws of geometric optics from the (negative) object distance $\rho_a R_k$ and the (negative) focal length f

$$\frac{1}{c} = \frac{1}{\rho_a R_k} - \frac{1}{|f|} \ldots\ldots\ldots\ldots(18)$$

The magnitude f can be determined graphically according to equation (6) from a representation of φ^* according to Table 1. A value of $|f_0/R_k|$ between 0·8 and 0·85 will be obtained, which is only slightly dependent upon ρ_a in the pertinent range $0 \cdot 4 \leq \rho_a \leq 0 \cdot 6$.

It is possible to obtain immediately from Fig. 8 the most important dimensions of the gun if, as is usually the case, the cathode current density j_k is expressed in milliamperes per square centimetre, the beam current I in milliamperes, the beam voltage U in hundreds of volts (hV or hectovolts) and the beam diameter D_0 in millimetres. We have for the cathode diameter:

$$D_k = 11 \cdot 3 \sqrt{\frac{I/\text{mA}}{j_k / \frac{\text{mA}}{\text{cm}^2}}} \text{ mm.}$$

(The substitution of the cathode surface projected on to the plane instead of the curved

92

surface allows for uneven cathode loading for large angles of aperture.)

The beam perveance in milliamperes per hectovolt to the $\tfrac{3}{2}$ power is

$$P = \frac{I/\text{mA}}{(U/\text{hV})^{3/2}}$$

From the necessary beam compression D_0/D_k we find in Fig. 8, centre, at the parameter P a certain abscissa value $R_a R_k$; vertically above it is the corresponding beam waist distance; while vertically below, at the point of intersection with the curve y in the ordinate of the latter, the (half) aperture angle θ from $\theta = \arccos(1 - y_1 P)$. The radius of curvature of the cathode is then

$$R_k = \frac{D_k}{2 \sin \theta}$$

which gives the anode-cathode distance and the anode diameter in an elementary manner.

9. Discussion.

Figure 8 shows that the first approximation ($P \to 0$) can only be considered to be valid for very small space charge constants. The evaluations for Fig. 8 were performed up to $\theta \cong 40°$. It will be seen that with such angles of aperture useful current concentrations can be obtained up to perveances of 4×10^{-6} (i.e. 4 mA at 100 V or 125 mA at 1,000 V). It should here be pointed out that Fig. 8 gives a somewhat pessimistic notion of the beam compression at large angles of aperture, because the anode lens was assumed to be free from error in performing the evaluation. The above-mentioned lens errors have the effect that the beam may become thinner in the waist in some cases without the paths in the beam waist departing from parallelism at such values of R_a/R_k at which z_0 passes through a maximum and is independent of the magnitude c. The "optimum" gun proposed by Huber and Kleen[5], suitably corrected, has thus this advantage of independence of any lens errors. It should furthermore be mentioned that with large angles of aperture, the beam necessarily assumes an extensively annular cross-section with the maximum current density at the edge, due to the uneven cathode load, the peculiarity of refraction in the anode lens, and the errors of the latter.

The range $\theta \leq 40°$ is fully adequate for the majority of cases. If still greater perveance and current concentrations are required, the

principles of "rectilinear flow" on which the present considerations are based no longer appear to provide the best results and the design of electron guns must be left more to the intuition of the gun constructor. However, Fig. 8 indicates that for this purpose θ must be made very large and the effective R_a/R_k must be relatively small.

By means of the dimensioning rules just developed, electron guns have been made for travelling wave tubes which at 1,000 volts had to shoot currents of 50 mA from cathodes of 6·5 mm diameter through helices of 3 and 2 mm diameter and 150 mm lengths, a beam diameter of about 60 per cent. of the internal diameter of the helix being required. This corresponds to current concentrations of 1:12 or 1:30 and beam current densities of about 2 and 5 A/cm². In the first case a current efficiency of 99 per cent. was obtained and in the second 97–98 per cent., the loss of current on the helix being less than 1 per cent., in each case, whilst initial attempts to dimension guns with aperture angles of 23° and 27° by the rules given in literature for strict requirements led to failure. In both cases, it was possible to increase the beam current to 80 mA and above without substantially impairing the current distribution. This result was obtained with careful application of the magnetic method already carried out elsewhere.[7]

Only a very small proportion of the current loss of 1–3 per cent. could be explained as being due to initial thermal velocity of the electrons, as is shown by a rough calculation (see Appendix). For the current escaping beyond the ideal beam edge between cathode and anode we obtain in A, V, and cm:

$$I_q \simeq 0·03 \times 10^{-6} \frac{1}{\sqrt{(\log_e I_s/I)}} \frac{D_k}{R_k} \times$$
$$\times \; U_a \sqrt{U_T} \cdot \rho_a^{-3/2} \; \exp(7·25)\rho_a$$

(I_s = saturation current, I = cathode current, U_T = mean temperature velocity in volts.) Substituting $I_s = 54$, $U_T = 0·1$ V (corresponding to the cathode temperature of 1160° K), U_a in kV and normalizing to $\rho_a = 0·5$, we have

$$I_q = 0·5 \frac{D_k}{R_k} U_a \cdot f(\rho_a) \text{ mA,}$$

where $f(0·4) = 0·67$, $f(0·5) = 1$ and $f(0·6) = 1·55$.

These values are usually less than 1 per cent. of the total cathode current and can only

increase immaterially up to the beam waist, that is to say, to the commencement of the guiding magnetic field. Slight errors in focusing, due to inaccuracies in construction and possibly the result of collisions of electrons with unavoidable gas residues in the tube appear to be responsible for the current losses.

10. Appendix: The Current Escaping Beyond the Beam Edge in Spherical Diodes.

The number of electrons escaping per unit time in the velocity interval dv from a cathode at temperature T (in degrees K) is (see, e.g., ref. 6):

$$dN = N_{total} \cdot 2 \frac{v}{v_0} \exp[-(v/v_0)^2] d\frac{v}{v_0}$$
$$\text{where } v_0 = \sqrt{\frac{2kT}{m_0}} \qquad \ldots\ldots\ldots\ldots(19)$$

Only those electrons reach the anode whose exit velocity is higher than v_1, v_1 being the velocity with which an electron is accelerated by the voltage U_1 of the potential minimum relatively to the cathode. With I as the current flowing in the cathode lead and I_s the saturation current of the cathode, we have

$$\frac{N_1}{N_{total}} = \frac{I}{I_s} = \int_{v_1/v_0}^{\infty} \frac{dN}{N_{total}} = \exp[-(v_1/v_0)^2] \ldots\ldots(20)$$

$$\text{i.e. } \left(\frac{v_1}{v_0}\right)^2 = \frac{|U_1|}{U_T} = \log_e \frac{I_s}{I}$$

$U_T = kT/q_0$ being the mean square of the volt velocity of all the electrons escaping from the cathode. In the case of oxide cathodes ($T = 1160°$ K), U_T has approximately the value 0·1 V. The mean square of the initial velocity $v_{1,0}$ of the electrons arriving at the counter electrode is

$$v^2_{1,0} = \frac{1}{N_1} \int_{v_1}^{\infty} v^2 dN$$
$$= \frac{N_{total}}{N_1} v_0^2 \int_{v_1/v_0}^{\infty} 2\left(\frac{v}{v_0}\right)^3 \exp[-(v_1/v_0)^2] d\frac{v}{v_0},$$

$$\left(\frac{v_{1,0}}{v_0}\right)^2 = \frac{I_s}{I}\left[\left(\frac{v_1}{v_0}\right)^2 + 1\right] \exp[-(v_1/v_0)]$$
$$= \log_e \frac{I_s}{I} + 1 \qquad \ldots\ldots\ldots\ldots(20)$$

These electrons thus leave the potential minimum with the mean initial velocity

$$v_{eff} = v_{1,\,0} - v_1$$

$$= v_0 \sqrt{\log_e \frac{I_s}{I}} \left[\sqrt{1 + \frac{1}{\log_e(I_s/I)}} - 1 \right]$$

$$\cong \left[\frac{v_0}{2} \frac{1}{\sqrt{\log_e(I_s/I)}} \right]_{I_s/I \gg 1} \quad \ldots\ldots(21)$$

The time taken by an electron starting from the potential minimum without initial velocity and reaching the anode is

$$\tau = -\int_{z=R_k}^{R_a} \frac{dz}{v_z} = \frac{R_k}{\sqrt{2\eta U}} \sqrt{\varphi_a^*} \int_{\rho_a}^{1} \frac{d\rho}{\sqrt{\varphi^*}},$$

$$\text{since} \quad v_z = \sqrt{2\eta\varphi} \quad \ldots\ldots(22)$$

The integral can be solved graphically by means of a curve representation of Table 1. In the range $0.4 \leq \rho_a \leq 0.6$, the result may be approximated to

$$\sqrt{\varphi_a^*} \int_{\rho_a}^{1} \frac{d\rho}{\sqrt{\varphi^*}} = f_1(\rho) \cong 0.72\rho_a^{-3/2}\ldots\ldots(23)$$

In this time, an electron moving at the velocity $v_q = v_{eff}$ according to equation (21) transversely to the beam flow in a field-free space (in the direction q) travels the distance

$$s = v_q \cdot \tau \quad \ldots\ldots\ldots(24)$$

Assuming that all electrons leave the potential minimum at the velocity v_{eff} and that all the exit directions $0 \leq \beta \leq 180°$ are of equal probability, it is possible to give the current escaping beyond the ideal beam edge. (This assumption is made on account of the uncertainty of exact knowledge of the form of the potential minimum at the cathode edge. Actually electrons escaping obliquely from the cathode must have considerably higher initial velocities than v_1, according to the angle, in order to pass through the potential minimum. A calculation according to the assumption would therefore in any case give too high a value in the transverse current.) If q is the running meridian co-ordinate on the potential minimum, from which a current of the density j_e escapes, the current proportion

$$dI_q = \frac{\beta}{\pi} j_k \pi D_k \, dq \quad \ldots\ldots\ldots(25)$$

passes beyond the beam boundary from the annular element of surface πD_k, where

$$q = s \cos \beta, \quad \text{i.e. } dq = s \sin \beta \, d\beta$$

Thus

$$I_q = j_k D_k s \int_0^{\pi/2} \beta \sin \beta \, d\beta = j_k D_k \, s \quad \ldots\ldots(26)$$

The current density j_k is $\frac{1}{4}\pi R^2$ of the total current of the complete spherical diode, so that according to equation (14) with the approximation applicable to the range $0.4 \leq \rho_a \leq 0.6$ we can write

$$j_k \cong U_0^{3/2} \frac{2 \times 10^{-6}}{1.89 \times 4\pi R^2} \exp(7.25)\rho_a \ldots\ldots(27)$$

The combination of all the results, equation (21) to (27), gives

$$I_q \cong 0.03 \frac{10^{-6}}{\sqrt{\log_e(I_s/I)}} \times$$
$$\times \frac{D_k}{R_k} U_a \sqrt{U_T} \, \rho_a^{-3/2} \exp 7.25\rho_a$$
$$\ldots\ldots\ldots\ldots(28)$$

11. References and Bibliography

1. J. R. Pierce, "Rectilinear flow in beams." *Journal of Applied Physics*, **11**, pp. 548-554, 1940.

2. J. R. Pierce, "Theory and Design of Electron Beams." (Van Nostrand, New York, 1949.)

3. I. Langmuir and K. B. Blodgett, "Currents limited by space charge between concentric spheres." *Physical Review*, **24**, pp. 49-59, 1924.

4. R. Helm, K. Spangenberg, and L. M. Field, "Cathode design procedure for electron beam tubes." *Electrical Communication*, **24**, pp. 101-107, 1947.

5. H. Huber and W. Kleen, "Electron beams of high current density in electrode static fields." *Arch. Elektrotechnik*, **39**, pp. 394-414, 1949.

6. H. Rothe and W. Kleen, "Principles and Characteristics of Electron Tubes." pp. 5-6 and 105-111. (Akad. Verlagsges., Leipzig, 1943.)

7. M. Müller, "Magnetic electron optics in longitudinal beam amplifier tubes." *Telefunken-Zeit.*, **26**, pp. 95-101, 1953.

94

DISC-SEAL CIRCUIT TECHNIQUES
Part 2 — Microwave Disc-Seal Oscillators *

by

John Swift, B.Sc.†

SUMMARY

A simple, basic oscillator circuit is described and various design problems are discussed. Subjects which are important in both oscillator and amplifier design are covered, including electronic efficiency, tuning devices for resonators, tuning curves, r.f. by-pass elements and d.c. circuit requirements.

1. Introduction

Although various kinds of power sources are available at microwave frequencies, disc-seal valve oscillators are frequently used up to frequencies of about 4,000 Mc/s.

Despite the fact that the performance at high frequencies is limited by transit time effects, the disc-seal oscillator has the advantages of ease of operation together with relatively simple circuit and power supply requirements.

These oscillators may be used in c.w., modulated or pulsed operation and therefore are frequently used in the transmitters of communication systems. They may also be used in the receivers of the same systems to provide the local oscillator signal to the mixer.

Disc-seal valves may often be used in oscillators operating at frequencies as low as 400 Mc/s since transit time effects are small and excellent frequency stability is possible.

In the laboratory, disc-seal oscillators provide a useful power output over a large frequency range.

A variety of disc-seal valves are manufactured and the type used will depend on availability, r.f. output power requirements and the upper frequency limit desired.

The various types of valve include lighthouse and inverted lighthouse or "oilcan" valves[1], pencil triodes[2], triodes with exceptionally small electrode spacings[3] and tetrodes[4].

Tetrodes will theoretically give a better performance than triodes at microwave frequencies (Section 4), but their development has been retarded due mainly to the difficulty of providing adequate screen dissipation. Several microwave tetrodes are available, however, and experimental types have been operated at over 10,000 Mc/s.

Most disc-seal valves employ metal to glass seals but some have been made with a ceramic envelope[4]. The high melting point of ceramic insulation enables higher outgassing and operating temperatures to be used, losses tend to be less and a more accurate valve construction is possible.

2. The Oscillator Circuit

2.1. *The Common-Grid Oscillator*

This circuit is similar to that of the microwave amplifier of Part 1, Section 2, except that no input connection is used and the resonators are coupled by a feedback system.

The common-grid circuit may take various forms depending upon the valve used and the application[5]. A very common form (the "folded back" circuit) is shown in Figs. 1 and 2.

The oscillator has two resonators. The output (anode-grid) line mainly determines the oscillator frequency and the second (grid-cathode) line applies the correct phase and magnitude of r.f. feedback voltage to the valve to produce steady oscillation.

As with the amplifier previously described, power is usually coupled from the output resonator to the load by means of a loop or probe (Part 1, Section 5) and the two resonators are normally tuned by movable bridges.

For some applications the coaxial resonators may be replaced by radial cavities, annular resonators or elements of rectangular wave guide (Part 1, Sections 9, 10 and 11) or at lower frequencies by butterfly circuits (Ref. 6, Vol. 2, p. 749).

In all these forms the microwave oscillator is essentially the same as a low-frequency one, consisting basically of a valve, two resonators coupled by a feedback system and an output load.

2.2. *Other Oscillator Circuits*

Less frequently used is the re-entrant oscillator (Ref. 5, Ref. 7, p. 179) which uses a radically different circuit construction to the oscillator above. Tuning over a wide frequency range is possible with a single control and the oscillator has a very flexible feedback system. The frequency stability, however, is usually inferior to that obtained with the circuit described above and it will not be discussed further.

An oscillator has been constructed that uses transit time effects (Part 1, Section 7) to advantage in obtaining efficient operation. A single resonator is used between grid and anode and the cathode and grid are operated at the same r.f. potential. (Ref. 8, p. 597.)

3. Elementary Oscillator Theory

Many types of low or high frequency oscillators depend on the same principles of operation. They consist of one or more resonators coupled to the electron beam of a valve. If the electron beam were not present, oscillations started in a resonator would soon die away because of resonator losses and dissipation in the external load. The motion of the electrons inside the valve is controlled in such a way that power removed from the output resonator is continually replenished by the electron beam, the energy coming ultimately from the d.c. power supply.

In low frequency oscillators the electron beam of the valve and the resonators are usually quite separate but in the microwave region they are intimately related. In klystrons and magnetrons both electron beam and circuit may be inside the vacuum envelope and in disc-seal oscillators, although the major part of the circuit is usually external to the valve, the two are closely related.

The frequency of oscillation in disc-seal oscillators is mainly determined by the output (anode-grid) resonator. The grid-cathode line together with the feedback system controls the relative phasing of r.f. input and output voltages of the valve.

It can be shown[9] that for correct phasing and hence efficient oscillation, the input and output resonators must present reactances of opposite sign at their respective valve gaps. Thus, if the grid-cathode line presents a capacitive reactance at the operating frequency (i.e., is resonant at a lower frequency) then the output line must present an inductive reactance (i.e., resonate at a higher frequency than the operating one).

Oscillation will cease if the input line is made so short that it presents an inductive reactance or if the capacitive reactance presented is too small to provide a r.f. voltage of adequate magnitude. For a detailed discussion of oscillator theory see Ref. 6, Vol. 1, p. 426 and Ref. 8, p. 504.
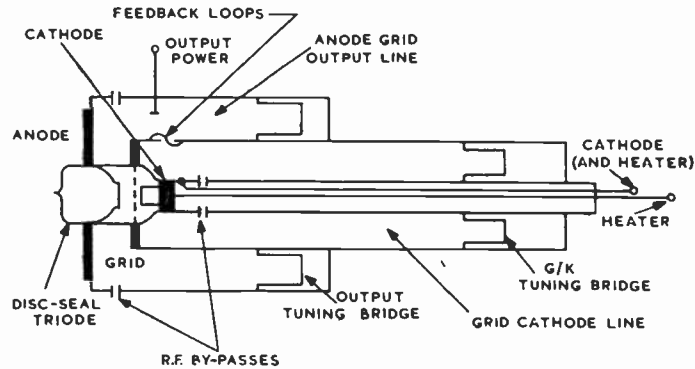

Fig. 1.—*Common-grid coaxial oscillator.*

4. Circuit and Electronic Efficiency

It is often convenient to consider separately the efficiency of two processes in oscillators (and amplifiers). The circuit efficiency relates to the proportion of available r.f. power from the valve that reaches the external load. The term electronic efficiency is applied to the conversion of d.c. power (from the supply) to r.f. power due to the interaction between the electrons and field within the valve.

96

4.1. Circuit Efficiency

This subject has already been discussed in connection with the output resonator of an amplifier (Part 1, Section 3) and the results are equally applicable to oscillators.

Thus for high circuit efficiency the power dissipated in resonator losses should be as small as possible compared with that reaching the external load. This requires a high ratio of unloaded to loaded Q in the output resonator which should be used in the $\frac{1}{2}\lambda$ mode if resonator losses are mainly surface ones. (If on the other hand resonator losses are mainly due to physical shunt losses such as from poor valve glass or faulty tuning bridges, the circuit efficiency will be about the same in the $\frac{1}{4}\lambda$ and higher modes.)

For reasonably high values of Q, a quantitative relation between the circuit efficiency (τ_i) and the unloaded and loaded Q's (Q_u and Q_l respectively) is
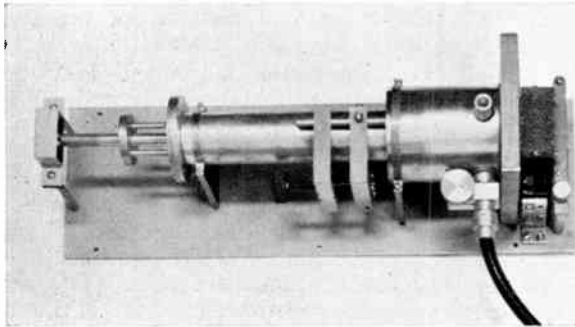
$$\tau_i = (Q_u - Q_l)/Q_u.$$

Fig. 2.—1,000 Mc/s 2C39 oscillator.

4.2. Electronic Efficiency under Class C Conditions

For high electronic efficiency at frequencies where transit time effects are not important (Part 1, Section 7), amplifiers and oscillators are operated under class C conditions. When triodes are used in this type of operation, the d.c. grid bias applied is greater than the cut-off value and anode current only flows during the short portion of each cycle when the r.f. input voltage is near its peak positive value.

For best electronic efficiency, the electron bunches produced should enter the grid-anode region when the instantaneous anode voltage is close to its minimum value. At any instant, the anode voltage is the resultant of the d.c. supply voltage and the instantaneous peak r.f. voltage developed between anode and grid by the output resonator. Thus for best electronic efficiency the r.f. voltage is nearly equal to the d.c. anode supply during the transit of electron bunches.

The reason for keeping the resultant anode voltage and hence the accelerating field to a small value during electron transit is that the electron bunches deliver r.f. energy to the output resonator most efficiently when they arrive at the anode with little or no kinetic energy.

There is an optimum value of shunt (load) impedance presented by the resonator to the output gap of the valve. At resonance, the r.f. voltage and current across the (real) load impedance are in phase and the optimum shunt resistance is such that when multiplied by the fundamental component of anode current, the correct r.f. voltage variation on the anode is obtained. If the shunt load is too low, the r.f. voltage swing is less, the accelerating field during electron transit is stronger and electrons reach the anode with more kinetic energy. As this energy is dissipated as heat the electronic efficiency is poorer. On the other hand, too large a value of shunt load produces an excessive r.f. voltage swing on the anode, and the accelerating field is too weak during electron transit. In this case the electronic efficiency is not necessarily reduced, but as the anode current is less, the power output falls.

A qualitative picture of how the electrons react with the fields inside the valve is most simply given for the case of a tetrode. The screen and anode are assumed to be at the same d.c. potential and the output resonator is assumed to be terminated by these electrodes.

As in the triode case, electrons pass through the grid in bunches centred at the short portions of a cycle when the grid-cathode field has its peak accelerating value. In the grid-screen region they are strongly accelerated by the d.c.

97

F

potential between these electrodes and finally enter the screen-anode output gap. For best electronic efficiency the electrons meet a retarding field in this region due to the r.f. voltage developed across the screen-anode space by the output resonator. Retardation of the electrons results in energy being delivered to the output resonator and ideally the retarding field should have such a magnitude as to bring all the electrons in a bunch to rest at the anode.

The r.f. field between screen and anode is produced initially and is sustained by the movement of electron bunches across the screen-anode gap. A r.f. current is induced in the output resonator and due to the definite impedance presented by the resonator to the screen-anode gap a r.f. voltage and field is produced. The r.f. voltage can only be sustained if the r.f. field it produces retards electrons in the output region so that they deliver most of their energy to the resonator. The system will only operate correctly if the relative phasing between the r.f. voltages at the input (grid-cathode) and output (screen-anode) gaps is correct. In oscillators, the phasing is obtained by using correct feedback adjustment (Section 5) but in an amplifier no (deliberate) feedback path is provided. The phase of the r.f. voltage across the screen-anode gap therefore adjusts itself to retard electrons entering this region and give best energy transfer to the output resonator.

In triodes, the action is similar to that described above except that the d.c. accelerating and r.f. retarding fields act together in the grid-anode output region.

4.3. *Transit Time and Electronic Efficiency*

When the time taken for electrons to cross two electrodes in a valve is large compared with one cycle of the r.f. voltage, various transit time effects occur. In Part 1, Section 7, most of the discussion dealt with transit time in the grid-cathode space. In considering electronic efficiency, however, at high frequencies, transit time in the grid-anode or screen-anode region is also important.

For good electronic efficiency electrons should traverse the output region when the r.f. retarding field is near its greatest value. Thus the transit time in this space should be short so that the field cannot change appreciably from the value that results in best transfer of d.c. to r.f. energy.

In a triode, the d.c. accelerating potential is almost cancelled by the r.f. retarding voltage which acts simultaneously in the grid-anode space. Electrons enter the output region with zero or small velocities and because they encounter a weak accelerating field during most of their transit, the time of transit is long. Further, differences in transit time acquired in the grid-cathode region result in differences in grid-anode transit time of individual electrons.

In tetrodes, however, electrons enter a strong d.c. accelerating field between grid and screen and are subsequently retarded between screen and anode. Thus transit time is less in the case of tetrodes and individual electrons tend to have similar times of transit.

Consequently the power output and electronic efficiency of tetrodes are generally higher than for triodes of similar emission capabilities and in equivalent operating conditions.

For both valves, the electronic efficiency can be improved by reducing the portion of a cycle during which electrons leave the cathode, i.e., by using extreme class C conditions. However, the power output is reduced since the effective anode current is less and cathode back-heating (Part 1, Section 7) becomes more serious.

The interaction of electrons and fields inside the valve is discussed quantitatively in Ref. 7, p. 31, and Ref. 8, p. 504.

5. Feedback in Oscillators

The purpose of the oscillator feedback system is to couple energy from the output to the input resonator so that with correct grid-cathode line tuning, sustained oscillation occurs. The r.f. feedback voltage applied between the grid and cathode of the valve must be correct in both magnitude and phase so that the electron beam inside the valve delivers optimum r.f. energy to the output resonator and external load. Various feedback systems can be used and the type chosen will depend on the oscillator application and the frequency range to be covered. No really satisfactory feedback system has yet been developed and as a result, microwave oscillators are usually more difficult to design than amplifiers. Some common feedback arrangements are discussed below.

5.1. *Amplifier with Feedback*

An amplifier can be converted to an oscillator by linking the output and input connections with a coaxial line of variable length. An extra

98

output connection is then provided to couple power to the external load. The feedback system in this case is the coaxial link together with the coupling devices to the two resonators. In theory, this arrangement is very flexible and both the phase and magnitude of the feedback voltage can be controlled. In practice, however, it is seldom used because of the number of adjustments needed, particularly in wide-band oscillators, and because it is difficult to reduce losses in the relatively long feedback path.

5.2. *Feedback Inside the Valve*

The unwanted coupling of energy between the resonators of an amplifier (regeneration) has already been discussed (Part 1, Section 8). This coupling occurs mainly through the anode-cathode capacitance of the valve and suggests a useful feedback system for oscillators. Usually the capacitance is too small to cause actual oscillation when the oscillator is loaded, so its value is enlarged by using a small rod that projects either side of the grid plane[9]. At a given frequency there is an optimum value of total anode-cathode capacitance for best oscillator performance but this value is not extremely frequency sensitive. In oscillators

Fig. 3.—*Double probe and loop feedback.*

that are required to cover a wide frequency range, this fixed feedback system is very convenient despite the fact that optimum performance will not generally be obtained over the entire operating range. The anode-cathode capacitance is usually chosen to be of optimum value at the high-frequency end of the band. At lower frequencies, the feedback is then less efficient but this is partly compensated by better circuit performance due to lower losses. If the feedback capacitance were chosen for optimum performance at the middle or lower end of the tuning range, the feedback

would be excessive at the high end and "squegging" (Sec. 14) and instability might occur.

This feedback system is convenient in many cases but is not suitable when optimum performance is required of an oscillator over a wide frequency range.

With exceptional disc-seal valves that can operate at very high frequencies, the anode-cathode valve capacitance alone may provide sufficient coupling to produce oscillation. The author has obtained reliable oscillation at over 7,000 Mc/s in this way. The feedback could not be varied, however, and optimum performance was probably not obtained.

Related to anode-cathode capacitive feedback is the possibility of strong coupling occurring through the grid seal-to-circuit connection. In amplifiers, strong regeneration and even oscillation may occur because of energy leakage through the joint. In other cases, the phase may be such that an amplifier is neutralized or an oscillator will not operate. A suitable connection to the grid is described in Section 13.

5.3. *Wide-Band Feedback Systems*

In wide-range oscillators, optimum performance is often sacrificed for the convenience of having a fixed feedback system. One such system has already been described above. A more common arrangement is to couple the input and output resonators through the common grid tube of the oscillator, placing the coupling close to the grid so as to be clear of the tuning bridges. The nature of the coupling used will depend on the field near the valve in both resonators. At relatively low frequencies (about 1,000 Mc/s) where the $\frac{1}{4}\lambda$ mode may be used for both lines and foreshortening due to the valve capacitances is small, a strong electric field will be present near the valve in both resonators. The double probe system shown in Fig. 3 can therefore be used over a moderate frequency range, two or three units being placed symmetrically around the grid tube.

At higher frequencies the electric fields near the valve may be weak in both lines and the double loop system shown in Fig. 3 should be used to couple to the strong magnetic fields. Often the fields near the valve in the two lines will be different due to different valve terminating structures and a combination probe-loop system is more suitable. This is shown in Fig. 4 for the case of a strong electric field near the
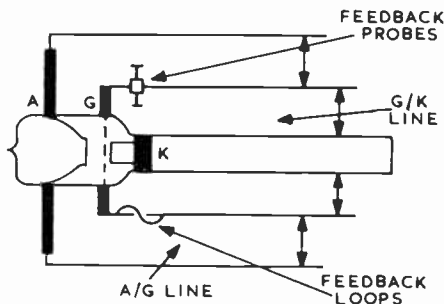
99

valve in the output line and a strong magnetic field in the input resonator.

The feedback design is obviously complicated since the field patterns will change with frequency and poor coupling may occur at some frequencies in the tuning range. Further information is given in Ref. 6, Vol. 1, p. 365. In general, the sizes of the loops and probes are best found by experiment.
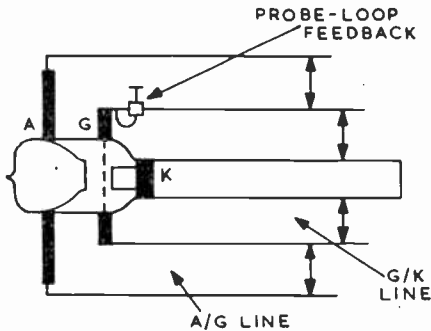

Fig. 4.—Probe-loop feedback.

5.4. *Probe Feedback*

A simple and effective feedback system can be made by placing a small metal screw across the output line so that it projects through a small hole in the grid tube into the input line (Fig. 5). The feedback can be controlled by changing the screw penetration.

This system is only satisfactory if a reasonably strong electric field is present near the probe in both resonators. The author has used this method in an oscillator tuning from 1,700 to 2,300 Mc/s and in oscillators operating at higher frequencies.

An alternative arrangement is to mount the probe on the fingers of the output tuning bridge, allowing it to project into the input line through a slot in the grid cylinder. Again, it is necessary that an electric field should be present near the probe in both lines, over the tuning range.

5.5. *Phase-shift of the Feedback System*

All the feedback methods described above with the exception of the first give little control over the phase of the feedback voltage which is usually of the same or opposite phase to the r.f. voltage in the output resonator. A practical feedback system which gives good control over the phase and magnitude of the feedback voltage is a feature of the re-entrant oscillator[5]. This type of circuit, however, is

not widely used because of its relatively poor frequency stability.

6. The Oscillator Resonators

6.1. *The Output Line*

The output (anode-grid) resonator has already been discussed in connection with the circuit efficiency and bandwidth of disc-seal amplifiers (Part 1, Section 3). The conclusions made are directly applicable to oscillators, although it should be noted that in this case the line resonates at a somewhat higher frequency than the operating one (Section 3).

A further design problem in oscillator and amplifier resonators is to find the line lengths necessary for resonance at a given frequency. This section will discuss theoretical tuning curves and the errors that occur in them at higher frequencies.

6.2. *The Low-Frequency Case*

At frequencies of up to about 2,000 Mc/s the coaxial output resonator and valve termination may be approximately represented by Fig. 6. The frequency is low enough for a definite valve output gap (AG) to be defined and the susceptance across this gap is mainly due to the
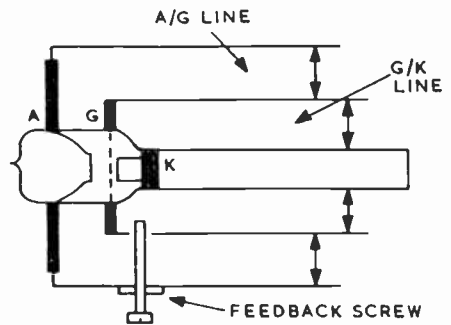

Fig. 5.—Probe feedback.

lumped anode-grid valve capacitance C. The lead resistance and shunt circuit losses are ignored.

It has already been shown (Part 1, Section 3) that the resonant line length l is related to the valve terminating capacitance C at a frequency f and wavelength by the expression

$$\tan (2\pi l/\lambda) = 1/\omega C Z_0 \quad \ldots\ldots\ldots\ldots(1)$$

where Z_0 is the characteristic impedance of the line.

Even at considerably higher frequencies, the

100

terminating susceptance is mainly capacitive even if no definite output gap can be defined. Thus the development of the theory of Part 1, Section 3, by representing circuits in the form shown in Figure 6, was justified.
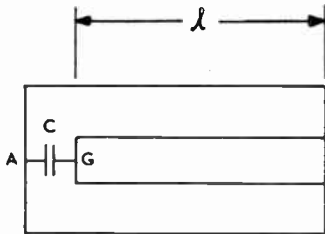

Fig. 6.—*Equivalent circuit valid up to about 2,000 Mc/s.*

However, in predicting tuning curves of λ against l for the oscillator, equation (1) is seldom useful above about 2,000 Mc/s because of the complex nature of the valve termination.

6.3. End Effects

It is evident from Fig. 6 that there is some difficulty in deciding where l should be measured to at the valve termination. For example, l could be measured to the valve anode or grid seals (A and G) or to some intermediate point. Further, near the valve, the resonator field pattern at higher frequencies is usually not that of an unloaded coaxial line. In fact, as the resonator length becomes shorter with increasing frequency, the entire resonator
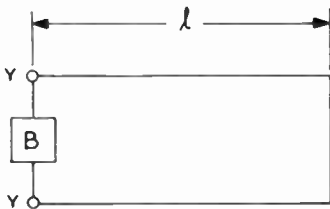

Fig. 7.—*Corner susceptance representing end effects.*

may resemble a radial cavity more than a coaxial line.

At high frequencies, the valve output gap which has up to now been considered as a simple lumped capacitance may become more complex due to various discontinuities at the termination and to inductance of valve seals.

These "end effects" are best considered by introducing the idea of corner susceptance. The resonator and valve termination are now represented by Fig. 7 and the corner susceptance B represents the complex effect of the entire end-structure. The length l may be measured now from the short circuit to any suitable reference point such as the valve grid flange.

The condition for resonance is given by

$$\tan (2\pi l/\lambda) = 1/BZ_0 \quad \ldots\ldots\ldots\ldots(2)$$

The corner susceptance B is an extremely complex function of frequency since it has several components. At frequencies below about 2,000 Mc/s it is mainly the susceptance of the valve capacitance and eqns. (1) and (2) are identical, but at higher frequencies B may be capacitive or inductive.
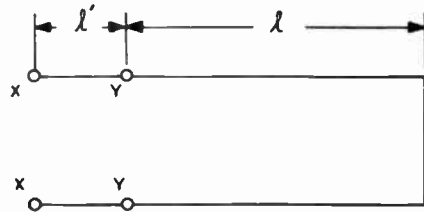

Fig. 8.—*Corner susceptance replaced by equivalent corner length.*

It is often convenient to replace the corner susceptance B by an equivalent corner length l' as shown in Fig. 8. The length l' of this open-circuited coaxial line is such that it presents the same susceptance B at YY as does the valve end-structure. The corner length is given by

$$\tan (2\pi l'/\lambda) = 1/BZ_0 \quad \ldots\ldots\ldots\ldots(3)$$

The line of Fig. 8 will resonate when the input impedance at XX is infinite. As already discussed (Part 1, Section 3) the requirement is

$$(l + l') = (2n + 1)\lambda/4\ldots\ldots\ldots\ldots(4)$$

where $n = 0, 1, 2$, etc.

This representation is useful because a resonator operating in the $\frac{1}{4}\lambda$ mode, for example, now has an equivalent electrical length of exactly $\frac{1}{4}\lambda$. However, in Fig. 6, a $\frac{1}{4}\lambda$ mode resonator has an electrical length of less than $\frac{1}{4}\lambda$ due to the foreshortening effect of the valve capacitance.

The behaviour of the valve termination with increasing frequency can be found experimentally by measuring the tuning length l at

101

various wavelengths λ. Equation (4) can then be used to find l' and hence the corner susceptance, at various frequencies.

Conversely, at lower frequencies the nature of B is known (being mainly the suspectance of the valve capacitance) and the tuning length l can be found theoretically as a function of λ using eqn. (1).

For an excellent, more detailed discussion of this subject, see Ref. 6; Vol. 2, p. 878.

6.4. *Practical Tuning Curves*

From the above discussion it is evident that over a large frequency range it is not practical to attempt calculating a theoretical tuning relation between l and λ. As well as the complex end-effects mentioned, the valve capacitance usually has large tolerances and heavy coupling to an external load may detune the resonator considerably. Further, in some non-contacting tuning bridges the r.f. short-circuit may not coincide with the front face of the plunger.

In general, it is preferable to find the relation between l and λ experimentally. To do this, an initial test oscillator is constructed with long input and output resonators, large tuning-bridge movements being provided in each line.

From these experiments the tuning movement required in each resonator to cover a specific frequency range, can be accurately found. The tests should of course be made with a number of valves to allow for capacitance variation.

6.5. *The Oscillator Input Line*

The input (grid-cathode) line has already been discussed in connection with disc-seal amplifiers (Part 1, Section 3). In the amplifier case, the grid-cathode line is heavily loaded by the valve input conductance which greatly modifies its properties as a resonator. When a driving source is matched to the input line, the total line length required does not in general obey eqn. (1) above.

In an oscillator, however, the question of matching a driving source to the valve input impedance does not arise. The feedback system couples power to the input of the valve but this connection is usually near the valve input gap where the loading conductance is located. The general effect is to make the input line similar to the output one as far as tuning properties are concerned. Most of the discussion given above for the output line can thus be applied directly to the grid-cathode line.

Over a given frequency range, the tuning lengths of the two lines are in general different and the difference between the lengths may not be constant. However, the tuning of the input line is usually quite broad due to the loading of the valve conductance and often both lines can be ganged together and tuned with a single control.

7. Mode Separation

It is a property of a coaxial line resonator terminated by a capacitance, that it will resonate at an infinite number of frequencies not harmonically related. Thus a given resonator may be used in the $\frac{1}{4}\lambda_1$ mode at the working frequency but unwanted resonances may occur at higher frequencies corresponding to the $\frac{3}{4}\lambda_2$, $\frac{5}{4}\lambda_3$ and higher modes.

In disc-seal oscillators, which have two resonators of this type, a choice of oscillator frequency is presented if both lines tune to more than one frequency simultaneously with given tuning bridge positions. In this case, oscillation may cease at some points in the desired frequency range or the oscillator may jump suddenly from one mode to another.

For frequencies of up to about 2,000 Mc/s, the line length l of a coaxial resonator is related to the terminating capacitance C at a frequency f and wavelength λ by the equation

$$\tan (2\pi l/\lambda) = 1/\omega C Z_0$$

where Z_0 is the characteristic impedance of the line. Thus the oscillator input and output lines will simultaneously tune to a number of frequencies if their respective values of CZ_0 are identical. Note that it is not sufficient that the input and output terminating capacitances should be different for good mode separation; the correct procedure is to use different values of CZ_0 for the two lines.

If the oscillator loading is correct at the desired frequency and very heavy at frequencies corresponding to higher modes, the problem of mode separation may not arise even if the two lines have identical CZ_0 values. This condition does not always occur in practice however.

At higher frequencies, the tuning equation above is not valid because of the complex nature of the valve termination in both resonators. (See Section 6.) Further, at these frequencies (above about 2,000 Mc/s), the $\frac{3}{4}\lambda$ or a higher mode has to be used because of capacitive foreshortening of the lines. The

interfering modes that give most trouble are usually those occurring at lower frequencies than the operating one corresponding, for example, to the $\frac{1}{4}\lambda_1$ mode.

The oscillator efficiency may be very high at the lower frequencies and mode separation may be difficult. A solution to this problem is to provide a frequency-sensitive feedback system, which gives optimum oscillator performance only in the desired tuning range.

The author has used this method at 4,000 Mc/s, coupling the input and output resonators of an amplifier (Section 5) via a filter that suppressed lower frequencies.

The problem of mode separation normally does not occur in an amplifier which is tuned to an input signal of a definite frequency. However, high-frequency amplifiers which use the $\frac{3}{4}\lambda$ or higher modes may tend to oscillate at a lower frequency corresponding to the fundamental mode, where the efficiency is higher. In this case, it may be necessary to modify the input matching arrangements (Part 1, Section 5) to suppress the oscillation. The subject is discussed in more detail in Ref. 6, Vol. 1, p. 344, Refs. 10, 11.

8. Interfering Modes

When operating normally, coaxial line resonators support (except perhaps in the region near the valve) electro-magnetic waves which have no component of electrical or magnetic field in the axial direction of propagation. These are called transverse-electromagnetic or TEM waves.

The various modes previously mentioned ($\frac{1}{4}\lambda$, $\frac{3}{4}\lambda$, etc.) are all of the TEM type, but a number of interfering modes which have a different field pattern can also occur in coaxial line resonators. These are objectionable because even if they cannot be supported directly by the valve, coupling between the desired (TEM) mode and an unwanted mode can occur and cause poor oscillator or amplifier performance.

The main interfering mode encountered is the transverse-electric or TE wave which has a magnetic but no electric field component in the axial direction. The most common of these, the TE_{11} wave, cannot be supported by the valve since there is no electric field at the valve gap. However, if the normal TEM and the interfering TE_{11} modes have the same resonant frequency at any part of the tuning range, coupling

between the modes will often cause poor performance. There is a critical wavelength λ_c for the TE_{11} mode and if the desired operating wavelength is longer than λ_c, TE_{11} waves cannot occur.

The cut-off wave-length λ_c is equal to the mean circumference of the coaxial line conductors and if these are reduced in diameter, the value of λ_c can be made so small that it does not lie in the required tuning range. The TE_{11} mode cannot occur in a $\frac{1}{4}\lambda$ resonator (Ref. 6, Vol. 1, p. 350) and use of the lowest mode is the best method of preventing interference.

An interesting property of TE_{11} waves is that their resonant frequency increases with larger valve capacitances. The frequency of TEM waves, on the other hand, is reduced by increasing the valve capacitance (Ref. 6, Vol. 2, p. 911).

Various other types of interfering modes are possible (Ref. 6, Vol. 2, p. 878, and Ref. 8, p. 473).

9. Tuning of Resonators

Coaxial line resonators are usually tuned by varying the electrical line length through changing the position of the r.f. short-circuit. A number of tuning methods have been developed and the one chosen will depend on the tuning range desired and on whether the oscillator (or amplifier) is of high or low power. Up to the present time, no really satisfactory general-purpose tuning method has been developed. Some typical tuning arrangements are given below.

9.1. *Short-Circuiting Bridges*

In this commonly used type of plunger, flexible metal fingers make direct contact with the inner and outer line conductors (Fig. 9).

The fingers are best made by slotting thin strips of beryllium-copper or phosphor-bronze which are then soft-soldered to the brass disc forming the base of the bridge. The assembly should then be plated.

A finger width of from $\frac{1}{16}$ to $\frac{1}{8}$ in. is convenient since a large number of contact points are possible and the fingers are flexible without being too fragile. The slots between adjacent fingers should be as small as possible to minimize leakage from the resonator to the region behind the bridge. The ends of the fingers may be curved to help in providing a smooth sliding contact or small punch marks can be made on the inside finger surfaces.

The length of the fingers should be a quarter-wavelength in the centre of the tuning range so that the ends of the fingers are near a voltage maximum. The high voltage helps in breaking down any insulating film that may be present on the contacts and since the current here is low, power losses are small.
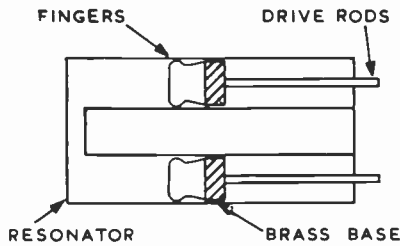

Fig. 9.—*Short-circuiting bridge.*

At low frequencies this procedure may not be practical as a quarter-wavelength is too long. For example, at 1,000 Mc/s, the ideal finger length is about 3 in. but in a practical design would be about 1 in.

In Fig. 9, it is evident that the shortest line length possible is that obtained when the tips of the fingers reach the end of the resonator. This limitation can be removed by adding a shoulder to the front of the brass base to bring the front face of the bridge close to the ends of the fingers. In this case, it is not possible to locate the finger contacts at a high voltage point.

To reduce leakage through a short-circuiting bridge, sets of fingers are sometimes soldered to both sides of the brass base. A less suitable method but one that is useful in preventing back-space resonances is to cover the back of the tuning bridge with a lossy material.

Short-circuiting bridges can be used over very wide tuning ranges and in high or low power devices. They are most suitable in equipment where only a few tuning operations are needed during use. When tuned frequently these bridges become erratic due to surface wear and may become extremely lossy.

9.2. *Non-Contacting Plungers*
When it is necessary for a resonator to be tuned many times, non-contacting plungers are often used. With correct design these tuning devices can cover a frequency range of at least 3 to 1 but the low frequency limit is about

1,000 Mc/s because the plunger length becomes excessive.

A very simple type of non-contacting plunger is shown in Fig. 10. It consists merely of a cylindrical metal slug that terminates the resonator through low impedance lines formed by the small air gaps between the slug and the line conductors. The design procedure is such as to produce an effective r.f. short-circuit or small reactance at the front face of the plunger. In practice, such plungers will always have losses due to dissipation of power in the low impedance lines and due to leakage of energy to the region behind the plunger.

If the plunger and hence the low impedance air lines are a quarter-wavelength long at the centre of the tuning range, the high impedance presented at the back face of the plunger is transformed to a low impedance at the front face. The characteristic impedance of both air lines should be made as small as possible since this reduces the effective input impedance to the plunger. The rate at which the input impedance increases as the plunger length departs from a quarter-wavelength with change in operating frequency is also decreased.
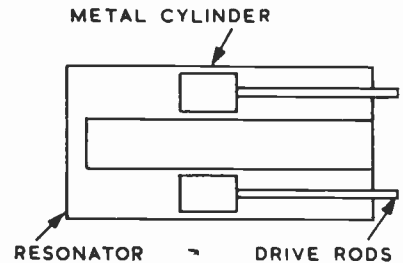

Fig. 10.—*Non-contact bridge.*

The simple plunger of Fig. 10 is seldom used because of the large leakage of energy from the resonator to the region behind the plunger. For this reason, more complicated bucket and choke non-contacting plungers have been developed. (Ref. 6, Vol. 2, p. 920; Ref. 8, p. 491, Ref. 12.)

Although non-contacting plungers are useful in low-power, wide-band oscillators and amplifiers, they are not very suitable for high power equipment because of leakage and break-down problems. Further, in any design, they are mechanically difficult to construct and interfering modes excited by them may have to be suppressed.

9.3. *Tuning Screws*

If only a small tuning range is needed a coaxial resonator may be tuned by a small, variable lumped capacitance placed across the line at a high voltage (i.e., high impedance) point (Fig. 11).

It is necessary to place the tuning screw at a high impedance point so that the relatively large capacitive reactance it presents across the line sufficiently modifies the field pattern to alter the resonator frequency. For this reason, a tuning screw is most satisfactory in the output line of oscillators and amplifiers, but its use is not recommended in the heavily loaded input line where the impedance along the entire line is often low.
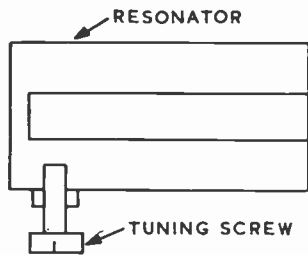

Fig. 11.—*Capacitive screw tuner.*

If the $\frac{1}{4}\lambda$ mode is used the tuning screw should be placed as close to the valve as possible, but if the line has been greatly foreshortened by the valve capacitance, the line impedance near the valve may be too low for the screw to have much effect. If the $\frac{3}{4}\lambda$ or higher modes are used, the screw should be placed $\frac{1}{4}\lambda$ distant from the r.f. short-circuit of the resonator.

Up to 100 Mc/s of tuning range is possible with this method. If excessive capacitive loading is used, however, the loaded Q of the resonator is increased, which may be undesirable in amplifiers or modulated oscillators where a large bandwidth is required. (Part I, Section 3.)

In high power equipment, care is needed to prevent flashover between the screw and resonator inner conductor.

A tuning screw has been used in a waveguide resonator to provide a wide-frequency coverage by altering the cut-off frequency of the waveguide.[13] The resonator suggested is simple and avoids lossy tuning elements.

9.4. *Movable Dielectric Slug Tuner*

A less commonly used tuning device, suitable for small frequency ranges, is made by placing a low-loss dielectric slug between the inner and outer resonator conductors. (Fig. 12.) The slug is moved along the line by means of dielectric handles extending through two or more longitudinal slots in the outer conductor. The slots should be small to reduce radiation losses.

The effect of moving the slug is to increase the effective electrical line length by a variable amount and thus tune the resonator. The region behind the slug is not "dead" but remains part of the resonator. This tuning device is useful in low Q resonators such as the grid-cathode lines of oscillators and amplifiers. In the output line the performance is often poor, probably due to dielectric losses.

One difficulty in a tuner of this type is to ensure that resonance is actually taking place. Moving the slug along the line will, for example, increase the effective electrical length to a maximum and with further slug movement it will decrease again. If the maximum effective length is less than that required for correct tuning, only an apparent resonance is obtained.
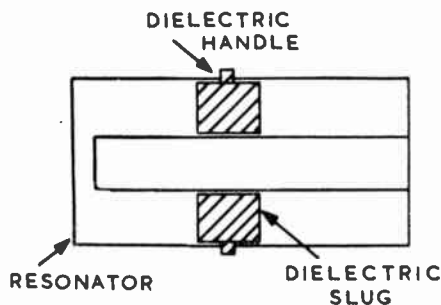

Fig. 12.—*Dielectric slug tuner.*

Design information for the dielectric tuner is given in Ref. 14.

Variations of the above method can be used. For example, the slug can slide over the inner conductor only and need not fill the space between resonator conductors. This may be useful as a fine-tuning adjustment in conjunction with a short-circuiting bridge.

A further method for tuning a resonator employs two movable dielectric slugs (Ref. 15, p. 466).

105

10. Stable Disc-Seal Oscillators

10.1. *High Stability Oscillators*

It is possible to design disc-seal oscillators that have excellent short-term frequency stability and reasonably good stability over long periods. The design procedure differs greatly from that used for power oscillators where frequent stability is often less important than obtaining a large output power.[16]

Good stability is obtained by careful attention to many factors in the oscillator design. The main requirements are given below.

(1) The valve should be capable of delivering considerably more power than is actually needed so that light coupling is possible between the external load and output resonator of the oscillator. If the desired output power from the stable oscillator is less than a few hundred milliwatts, this requirement can be met over a wide frequency range with available disc-seal valves. The low output power used, often allows the dissipation of the anode and heater to be reduced with resultant reduction in thermal effects (see below).

The valve used should be capable of operating well above the highest frequency in the required tuning range so that transit time effects are reduced and the oscillator frequency is not greatly dependent on the anode voltage.

(2) The circuit should preferably be of the common-grid type (Section 2). The re-entrant oscillator[5] is less stable and is not recommended for this application.

In the common-grid circuit, the output (anode-grid) resonator mainly determines the oscillator frequency and for good frequency stability this line must have a high value of loaded Q (Part 1, Section 3). The loaded Q (Q_l) is defined by

$$Q_l = 2\pi \times \frac{\text{stored resonator energy}}{\text{total power dissipated per cycle}}$$

The r.f. power from the valve is dissipated in the external load and in resonator losses. The requirement of light coupling to the external load ensures that Q_l will be larger than the value usually obtained in power oscillators. The valve must, however, have an ample reserve of power still to couple sufficient output to the external load.

To further increase Q_l, circuit losses must be reduced to a minimum, particular care being taken in the design of the tuning bridges. Q_l may be greatly increased by increasing the stored energy of the output resonator. This is accomplished by using the $\frac{3}{4}\lambda$ or $\frac{5}{4}\lambda$ mode in the output line.

The relevant theory has already been discussed in Part 1, Section 3, but it should be noted that previously the requirement has been for a low value of Q_l to give a large amplifier bandwidth. The requirements for large bandwidth (low Q_l) and good frequency stability (high Q_l) are therefore conflicting ones.

It has also been shown (Part 1, Section 3) that the use of the $\frac{3}{4}\lambda$ and higher modes increased the equivalent shunt loss presented to the valve output gap, if this were mainly due to surface losses in the resonator. (If actual parallel losses such as from poor valve glass or faulty tuning bridges are present and are larger than series surface losses, this effect is less pronounced in going to higher-mode operation. Resonator surface losses are usually the main factor, however.)

The poorer circuit efficiency resulting from increased shunt losses in higher-mode operation must be tolerated in oscillators that are designed primarily for good frequency stability.

Modes higher than the $\frac{5}{4}\lambda$ one are normally not considered because the circuit becomes too large, the construction less rigid and resonator losses steadily increase.

The dependence of good frequency stability on a high loaded Q in the output resonator can be explained in a simple manner. The output line may be replaced, near resonance, by a parallel-resonant lumped LCR circuit (Part 1, Appendix 2) which is located across the valve output gap. The equivalent lumped capacitance C is larger than the anode-grid valve capacitance since it includes the effect of the distributed line capacitance.

The use of higher modes, in effect, "swamps" the frequency sensitive valve capacitance with a much larger and more stable equivalent line capacitance. If, however, the $\frac{1}{4}\lambda$ mode is used, the value of C is not much larger than the valve capacitance and the "swamping" effect is less.

(3) If the oscillator is to be a tunable one, careful bridge design is vital and non-contacting plungers are preferable if many tuning operations are required. (Section 9).

(4) For good stability, the feedback system (Section 5) must produce strong oscillation over the desired tuning range. Provided this condition is satisfied the system used is not important.

(5) Thermal effects are among the main causes of poor frequency stability. Heat dissipation in the oscillator causes a temperature rise in both valve electrodes and external circuit, resulting in a long-term drift. It appears that the resonator material should have a high thermal conductivity rather than a low coefficient of expansion for best stability. This applies because a temperature rise in the valve causes a larger frequency change than does expansion of the resonators.

(6) Many other factors effect the stability. Thus freedom from vibration, stable power supplies, ample decoupling and freedom from "squegging" all contribute to better stability.

10.2. *Power Oscillators of Moderate Stability*

It is evident from the above discussion that very stable oscillators require a different design procedure to power oscillators. Frequently, an oscillator is needed that gives a moderate power output and has good long-term frequency stability. The power output required may be less than the valve is capable of delivering but considerably more than that obtained from the very stable oscillator discussed above. In addition, the oscillator may be amplitude modulated if it forms part of a communication transmitter. The design of this type of oscillator is difficult due to conflicting requirements.

If a stable, amplitude-modulated oscillator is required, the stability requirement (light coupling to the load and higher-mode operation) conflicts with the modulation requirement of a large resonator bandwidth (Section 11). More precisely, for good stability, a high loaded Q is required in the output resonator but a low loaded Q is needed to provide adequate band-width for modulation. In addition, the use of higher modes to obtain good stability results in poorer circuit efficiency (Part 1, Section 3).

In practical modulated oscillators, the band-width necessary for satisfactory modulation may be small in some cases, so the use of the $\frac{3}{4}\lambda$ or $\frac{5}{4}\lambda$ mode in the output resonator is feasible. In all other cases the best approach is to obtain the desired output power and bandwidth first (using the $\frac{1}{4}\lambda$ mode if necessary) and ensure that all other factors effecting stability are as favourable as possible (see 10.1).

11. Modulated and Pulsed Oscillators
11.1. *Amplitude Modulation*

Because of transit time effects (Part 1, Section 7), amplitude modulation of microwave disc-seal

oscillators is less successful than at lower frequencies. For various reasons, anode, grid or cathode modulation is difficult to accomplish in practice, although simultaneous anode and grid modulation appears more satisfactory.

In modulated oscillators, the loaded Q of the output resonator should be low to obtain a large bandwidth (Part 1, Section 3) and thus present a sufficiently high impedance to the r.f. sidebands produced. The $\frac{1}{4}\lambda$ mode should preferably be used together with a high value of characteristic impedance. This procedure will also give good circuit efficiency.

The problem of obtaining good frequency stability in an amplitude-modulated oscillator is discussed in Section 10.

A survey of various methods of amplitude modulation is given in Ref. 6, Vol. 1, p. 391.

11.2. *Amplitude Modulation Using Ferrites*

The problem of amplitude modulating disc-seal oscillators has been partly solved by using absorption ferrite modulators. Used alone, these ferromagnetic materials have large losses at microwave frequencies but the losses are not mainly due to hysteresis and eddy currents but to changes in the internal magnetic structure[17] The losses in these materials can be reduced and controlled by applying a d.c. field which is usually produced by an electromagnet.

Thus, among other things, an electronic attenuator can be made[18] which provides an excellent method for amplitude modulating a c.w. oscillator. Unmodulated r.f. power is sent through the attenuator and modulation is superimposed upon the d.c. field by varying the electromagnet current about a fixed value or applied by using an additional modulating coil and electromagnet. Such a device can operate satisfactorily with modulating frequencies of up to 10,000 c/s. A practical single-sideband modulator is described in Ref. 19.

11.3. *Frequency Modulation*

Frequency modulation of disc-seal oscillators is difficult to accomplish practically. Mechanical frequency variation of up to 20 Mc/s is possible using a rotating paddle of low-loss dielectric or a motor-driven bridge placed in the output resonator or in an auxiliary coupled line[20]. This type of system has been used in mechanical automatic frequency control of oscillators.

Electronic frequency modulation is difficult to obtain without incidental amplitude modulation.

At low frequencies, a reactance valve can be used to frequency modulate an oscillator. This device acts as a variable reactance when the d.c. potential on one of its electrodes is altered. With increasing difficulty, this method can be extended to the microwave region (Ref. 6, Vol. 1, p. 408). The author has used a disc-seal reactance valve to frequency modulate an oscillator at 4,000 Mc/s but incidental amplitude modulation was present and circuit adjustments were critical.

In theory, a coaxial line loaded at one end by a crystal or thermistor and a multiple of an eighth-wavelength long, could be used as a variable reactance to frequency modulate an oscillator. Small frequency deviations of up to 100 kc/s are possible by applying a small voltage in series with the grid bias.

A further, theoretically possible system is to place a special dielectric material in the oscillator output resonator whose dielectric constant can be changed by an external electric field.

11.4. *Pulsed Oscillators*

Pulsed, microwave disc-seal oscillators are used in various radar applications, in communication systems and in signal generators. A disc-seal oscillator may be pulsed by applying positive pulses to the anode, or by biasing the valve beyond cut-off and applying positive pulses to the grid or negative pulses to the cathode. The second method has the advantage that the pulsing device need not supply the anode power but the oscillator is more difficult to adjust and the valve life is poorer.

Very high anode voltages may be applied to the valve because the low-duty cycle maintains the average dissipation and emission at a reasonable value. For this reason, transit time effects (Section 4 and Part I, Section 7) are less serious and the efficiency is higher with pulse than with c.w. operation.

The peak power output depends mainly on the available cathode emission of the valve used and the anode dissipation. To ensure sharp r.f. pulses the build-up of oscillations must be rapid and strong feedback is vital. For this reason the re-entrant circuit[5] is often used for pulsed oscillators.

When pulses are applied to the oscillator there will usually be a r.f. by-pass (Section 12) shunted across the output of the pulsing device. If the pulses are applied between anode and grid, for example, the anode by-pass capacitance

will tend to distort the pulses as its value is usually large. Further, with pulse operation, the danger of breakdown across the anode by-pass is more serious because of the high voltages involved. If it is inconvenient to reduce the internal impedance of the pulsing device, the best procedure is to replace the usual r.f. by-pass with some form of choke which has less effective capacitance. When extremely sharp r.f. pulses are desired, a microwave amplifier can be pulsed and driven with the steady signal from a c.w. oscillator. Various aspects of the subject of pulsed disc-seal oscillators are discussed in Ref. 7, p. 187, Ref. 8, p. 594, and in Refs. 21, 22.

12. R.F. By-Pass Elements

In disc-seal oscillators and amplifiers it is necessary to isolate the valve electrodes for d.c. without interfering with the r.f. circuit. Thus elements are required that are open circuits for d.c. but effective short circuits for r.f.
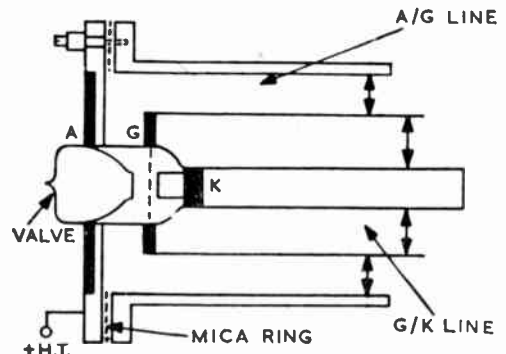

Fig. 13.—Anode by-pass.

At microwave frequencies ordinary lumped capacitances are seldom suitable and distributed r.f. by-pass elements are used.

12.1. *The Anode By-Pass*

The anode by-pass is usually located close to the valve as shown in Fig. 13. The by-pass consists of two brass flanges separated by one or more thin sheets of mica and tightly clamped together. The main part of the outer cylinder of the circuit is therefore at grid potential for d.c.

While this arrangement acts in a similar manner to a lumped by-pass capacitance at low frequencies, it is actually a radial line. At a given frequency there are optimum dimensions
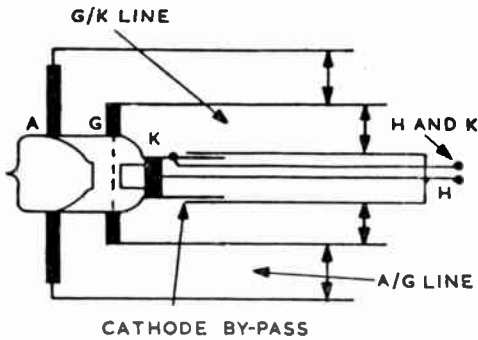
CATHODE BY-PASS

Fig. 14.—Cathode by-pass.

for the two flanges so that an approximate r.f. short circuit is presented at the inner face.

In practice, it is not always possible (due to large physical size), or necessary to use the optimum dimensions. The mica sheets should be thin enough to prevent excess radiation but not so thin that d.c. breakdown can occur between anode and grid. The flanges should have flat surfaces and must be tightly clamped together. If modulation is applied between anode and grid, the low frequency capacitance of the by-pass element should be reduced to a minimum (Section 11).

An excellent account of radial line by-pass elements, including the theory and design procedure, is given in Ref. 23.

12.2. *The Cathode By-Pass*

The cathode and grid must be isolated for d.c. so that grid bias may be applied. One method is to isolate the inner and outer fingers of the grid-cathode line tuning bridge. This is accomplished by inserting a thin wedge of mica through the brass base of the plunger. The end of the cathode line remote from the valve must also be d.c. isolated from the grid cylinder, but need not by-pass r.f. since it is behind the tuning bridge.

A more convenient method is to use the coaxial by-pass element shown in Fig. 14. This is placed as close to the valve as possible to be clear of the tuning bridge, so that most of the cathode line is at grid potential for d.c.

The by-pass element in this case is a section of coaxial line of very low characteristic impedance, open-circuited at both ends. A thin cylinder of mica is placed between the conductors to provide d.c. isolation. If the characteristic impedance of the by-pass element

is small (i.e., the mica is very thin), energy leakage is negligible and the reactance presented to the main line is small and changes very slowly with frequency. The operation is similar in many ways to that of non-contacting tuning plungers (Section 9 and Ref. 6, Vol. 2, p. 920). It is not necessary that the length of the by-pass be a quarter-wavelength provided the characteristic impedance of the element is low. For example, if the characteristic impedance of the by-pass element is 0·5 ohms, the series reactance presented to the grid-cathode resonator is zero if the by-pass length is $\frac{1}{4}\lambda$. If the length is reduced to $\frac{1}{8}\lambda$ the reactance presented is only $-0·5j$.

In cases where space is limited a wedge-shaped by-pass may be used.

13. Circuit to Valve Connections

The mechanical system used to connect the valve electrodes to the circuit is a vital part of oscillator and amplifier design.

The method used should preferably allow easy insertion and removal of the valve, be sufficiently flexible to allow for valve tolerances and should not strain the valve seals. A number of positive contacts should be made with each seal as high currents may flow at the connections.

The anode may be clamped to the circuit by some suitable locking device or a wire helix contact (Fig. 15) may be used. It is essential that good thermal contact is made with the anode as well as positive electrical contact.

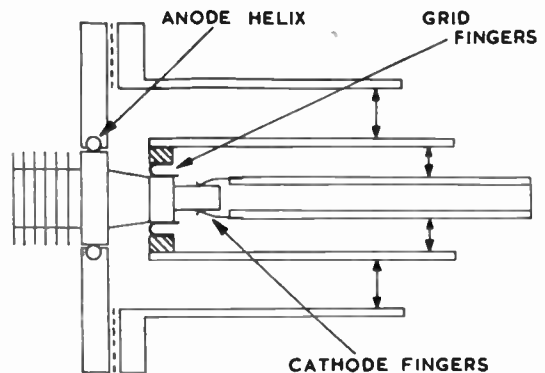The grid flange is best connected to the circuit with beryllium-copper or phosphor-

Fig. 15.—Circuit to valve connections.

bronze fingers of a similar type to those used in tuning bridges (Section 9). The spaces between adjacent fingers should be as small as possible and good electrical contact between the grid and circuit is vital. The wire helix is not recommended for the grid connection as considerable energy leakage between resonators can occur, particularly at high frequencies.

Connection to the cathode of the valve is conveniently made by slotting the end of the brass cathode tube or soldering a set of flexible fingers to the end. The valve to circuit connections should all be plated (Section 16).

14. The D.C. Circuit (Fig. 16)

Unless a high degree of stability is required[16], it is not necessary to regulate the power supply which can therefore be of conventional design.

Grid bias may be provided with a cathode bias, or grid leak resistor, or a combination of the two for high power devices. Cathode bias provides a convenient means of controlling the valve current and protects the circuit when there are no oscillations. Grid current metering gives a valuable guide to circuit operation particularly in the case of oscillators. In amplifiers it is helpful in tuning the input line to the frequency of the driving source.

Intermittent operation or "squegging" is common in microwave oscillators and self-pulsing may occur at a frequency of up to several megacycles. The effect may show itself in erratic oscillator tuning, "live" d.c. supply leads and circuit structure, as a sudden drop in anode current and as difficulty in measuring the oscillator frequency with a wavemeter. "Squegging" is usually caused by excess feedback but often the feedback system cannot be altered in a wide-band oscillator. The trouble can often be cured by increasing the anode current, lowering the anode voltage or by reducing the grid circuit time constant.

Referring to Figure 16, it is evident that if the negative h.t. lead is grounded to the mounting frame, the oscillator (or amplifier) circuit should be placed on insulated pillars to prevent short-circuiting the grid-current meter.

When two circuits are connected together, for example, an amplifier and driving oscillator, separate power supplies should be used if possible. If only one supply were provided, it is evident from Fig. 16 that the grid current meters would be placed in parallel because the grid electrodes would be linked for d.c. by the

connecting coaxial cable. If one supply only can be used the grid current meters have to be dispensed with or a r.f. by-pass (Section 12) can be placed in the connecting cable to isolate the grids for d.c. Separate heater windings must be used to prevent placing the two cathode bias resistors in parallel. (Most disc-seal valves have a common heater and cathode connection.)

15. Valve Cooling

Disc-seal valves are usually small in relation to their power-handling ability and adequate cooling must be provided.

In low-dissipation valves, it is sufficient that the anode flange makes good thermal as well as electrical contract with its clamping assembly. The clamp should be as large as possible and have a large thermal conductivity to minimize thermal drift (Section 10). With larger valves,
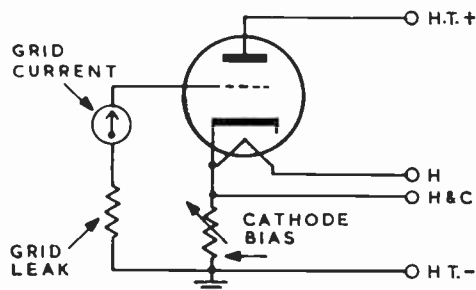

Fig. 16.—D.c. connections to a disc-seal triode.

forced-air or water cooling is used and the valve is provided with cooling fins.

Cooling systems can easily be added to the "folded-back" form of common-grid circuit shown in Fig. 1 since the valve is readily accessible. The problem is more difficult in the case of circuits using lighthouse valves[5] since the anode of the valve is enclosed by the resonators.

16. Electro-Plating of Circuits

The components of microwave oscillators and amplifiers should be plated to minimize losses and preserve their appearance, particularly if the units are to be used out of doors.

The plating material should have a high electrical conductivity and strongly resist corrosion. The plating thickness usually need not be more than about 0·0005 in. because of skin effect (Ref. 15, p. 125). The most suitable

plating materials are silver and copper but as these usually tarnish rapidly they in turn should be covered with a thin "flash" of gold, rhodium or palladium which strongly resist corrosion. As the last two metals are relatively poor conductors the "flash" thickness should be much less than the skin depth.

When components such as bridge fingers have to slide over the surface of another part of the circuit, the components should preferably be plated with different metals to prevent "welding." Silver contacts sliding on a silver surface should be carefully avoided.

A more detailed discussion of the subject is given in Ref. 15, p. 125, and Ref. 24.

17. Acknowledgments

18. References

1. E. D. McArthur, "Disc-seal tubes," *Electronics*, **18**, pp. 98-102, Feb. 1945.
2. G. M. Rose, D. W. Power and W. A. Harris, "Pencil-type u.h.f. triodes," *R.C.A. Review*, **10**, pp. 321-338, 1949.
3. J. A. Morton and R. M. Ryder, "Design factors of the Bell Telephone Laboratories 1553 triode," *Bell Syst. Tech J.*, **29**, pp. 496-530, Oct. 1950.
4. W. P. Bennett and H. F. Kazanowski, "One-kilowatt tetrode for u.h.f. transmitters," *Proc. Inst. Radio Engrs*, **41**, pp. 13-19, Jan. 1953.
5. A. M. Gurewitsch and J. R. Whinnery. "Microwave oscillators using disc-seal tubes," *Proc. Inst. Radio Engrs*, **35**, pp. 462-473, May 1947.
6. Harvard University Staff, "Very High-frequency Techniques," 2 vols. (McGraw-Hill, New York, 1947).
7. D. R. Hamilton, J. K. Knipp and J. B. Kuper, "Klystrons and Microwave Triodes." M.I.T. Radiation Laboratory Series No. 7 (McGraw-Hill, New York, 1948).
8. H. J. Reich, P. F. Ordung, H. L. Krauss and J. G. Skalnik, "Microwave Theory and Techniques" (Van Nostrand, New York, 1953).
9. J. Bell, M. R. Gavin, E. G. James and G. W. Warren, "Triodes for very short waves—oscillators," *J. Instn. Elect. Engrs*, **93**, Part 3A, pp. 833-846, 1946.
10. P. J. Sutro, "Theory of mode separation in a coaxial oscillator," *Proc. Inst. Radio Engrs*, **34**, pp. 960-962, 1946.
11. H. J. Reich, "Mode separation in oscillators with two coaxial line resonators," *Proc. Inst. Radio Engrs*, **36**, pp. 1252-1255, Oct. 1948.
12. W. H. Huggins, "Broad-band noncontacting short circuits for coaxial lines," *Proc. Inst. Radio Engrs*, **35**, Sept., Oct. and Nov. 1947.
13. N. E. Colby, "Components for 2,000 Mc/s oscillator and amplifier circuits," *Proc. Electronic Components Symposium*, pp. 159-162, 1954.
14. H. R. L. Lamont, "Theory of resonance in microwave transmission lines with discontinuous dielectric," *Phil. Mag.*, **29**, p. 521, June 1940.
15. F. L. Ragan, "Microwave Transmission Circuits," M.I.T. Radiation Laboratory Series No. 9 (McGraw Hill, New York, 1948).
16. J. G. Stephenson, "Design of stable tunable microwave oscillators," *Convention Record of the I.R.E.*, **2**, Part 8, pp. 104-112, 1954.
17. R. E. Alley, Jr., "A review of new magnetic phenomena," *Bell Syst. Tech. J.*, **32**, pp. 1155-1172, Sept. 1953.
18. F. Reggia and R. W. Beatty, "Characteristics of the magnetic attenuator of u.h.f.," *Proc. Inst. Radio Engrs*, **41**, pp. 93-100, Jan. 1953.
19. J. Cacheris, "Microwave single-sideband modulator using ferrites." *Proc. Inst. Radio Engrs*, **42**, pp. 1242-1247, Aug. 1954.
20. M. E. Hines, "A wide range sweeping oscillator," *Bell Syst. Tech J.*, **29**, pp. 553-559, Oct. 1950.
21. C. E. Fay and J. E. Wolfe, "A tunable vacuum-contained triode oscillator for pulse service," *Proc. Inst. Radio Engrs*, **36**, pp. 234-239, Feb. 1948.
22. L. S. Nergaard, D. G. Burnside and R. P. Stone, "A development pulse triode for 200 kW output at 600 Mc/s," *Proc. Inst. Radio Engrs*, **36**, pp. 412-416, March 1948.
23. R. N. Bracewell, "Applications of disc transmission lines in microwave apparatus," C.S.I.R.O. Report, Sydney, Australia, RP202, Feb. 1944.
24. R. F. Proctor, "High-frequency resistance of plated conductors," *Wireless Engr*, **20**, p. 56, 1943.
25. W. G. Dow, "Transit time effects in u.h.f. class C operation," *Proc. Inst. Radio Engrs*, **35**, pp. 35-42, Jan. 1947.
26. H. Rothe and E. Gundert, "The effects of electron transit time on the efficiency of tetrodes," *Telefunken-Zeit*, **25**, pp. 75-87, June 1952.
27. L. J. Giacoletto, "U.h.f. triode design in terms of operating parameters and electrode spacings." *Proc. Inst. Radio Engrs*, **41**, pp. 51-58, Jan. 1953.
28. S. Ramo and J. R. Whinnery, "Fields and waves in modern radio" (Wiley, New York, 1944).
29. R. N. Bracewell, "A Microwave Triode Oscillator," C.S.I.R.O. Radiophysics Report, RP261, Sydney, Australia, Sept. 1945.
30. W. P. Bennett, E. A. Eshbach. C. E. Haller and W. R. Keye, "A new 100-watt triode for 1,000 megacycles," *Proc. Inst. Radio Engrs*, **36**, pp. 1296-1302, Oct. 1948.
31. F. C. Isely, "A new approach to tunable resonant circuits for the 300 to 3,000 Mc/s frequency range," *Proc. Inst. Radio Engrs*, **36**, pp. 1017-1022, Aug., 1948.
32. C. L. Cuccia, "Certain aspects of triode reactance-tube performance for frequency modulation at u.h.f.," *R.C.A. Review*, **10**, pp. 74-98, March 1949.
33. C. A. Helber, "Improving stability of u.h.f. oscillators." *Electronics*, **20**, pp. 103-105, May 1947.

. . . Radio Engineering Overseas

338: 620.1: 621.37/9

Control in manufacturing processes.—J. DE LIGNY and A. H. SCHAAFSMA, *Onde Electrique*, **35**, pp. 1171-74, December 1955.

This is the second of a series of three articles on quality control in manufacture in the electronics industry, and deals with the practical applications of statistical control.

621.317.34: 621.397.5

Circuitry for the generation of television test-pictures. —F. PILZ, *Archiv der Elektrischen Übertragung*, **9**, pp. 547-558, December 1955.

The paper deals with the generation of television test pictures by suitable combinations of electrical pulse groups (synthetic test patterns), i.e. without using optical or electron-optical means. The generation of suitable test signals is discussed first. For the combination of these signals it is important that a pleasing picture results and that the evaluation of the single tests is simple. The circuitry involved is dealt with, and a test picture generator is described, as developed in the Rundfunk-Technisches Institut, Nuremberg.

621.373,421

Some new aspects of the Wien bridge oscillator.— M. K. ACHUTHAN, *J. Instn Telecom. Engrs*, **2**, pp. 19-27, December 1955.

This paper deals with the factors influencing the extension of the use of the Wien bridge oscillator to the radio frequency range and its use as a means of producing wide band frequency modulation in which the frequency deviation at any instant is proportional to the instantaneous value of the modulating voltage. It is shown that, even if all other conditions are favourable for the production of high frequencies, the phase shift in the amplifier which is always considerable at high frequencies reduces the frequency of oscillations to a very low value. The production of wide band frequency modulation as mentioned above, without any superimposed amplitude modulation, is also discussed briefly.

621.376.55: 621.397.5

A multiple-pulse time modulation system.—H. J. GRIESE, *Archiv der Elektrischen Übertragung*, **9**, pp. 571-572, December 1955.

Earlier studies concerning transmission of the sound accompanying television programmes over the picture transmitter during the line flyback led to phase-modulated multiple pulses for which a simple method of demodulation was found. With the same bandwidth, the same average power, but a considerably lower peak power of the transmitter, better signal-to-noise ratios can be obtained under certain circumstances with multiple pulses than with single pulses. With multichannel transmission the subdivision into individual channels at the receiving end may be effected, apart from time selection, also by frequency selection.

621.383.2

Some characteristics of the antimony-caesium photocathode.—E. KANSKY, *Elektrotehniski Vestnik*, **23**, pp. 294-297, September-October 1955.

Some properties of the Sb-Cs cathode are described. The composition of antimony and caesium which is photoelectrically sensitive and approximates to the formula $SbCs_3$ is a typical electronic semiconductor. The greater number of photoelectrons originate within the inside of the cathode, from a depth of several tens of atomic layers. The cathode is sensitive to light waves at ranges from 2,000 to 6,500 A.U., its total sensitivity lying between 10 and 150 μ A/lumen. The secondary emission causes anomalies in current characteristics, while the photo emission is dependent upon temperature. The thermal current is small. The effective life of the cathode and the causes of photoelectric fatigue are also discussed.

621.396.677.833.2

Computation and construction of double curvature reflectors.—L. THOUREL, *Onde Electrique*, **35**, pp. 1153-63, December 1955.

After re-stating the theory of surface determination by means of geometric optics, the author puts forward a practical method of calculation and examines the selection of the location of the illuminating source and the possibilities of realising the surface. Diffraction formulæ for calculating the radiation diagram of antenna are then established. The results may be used to give a final correction to the shape of the surface deduced by geometric optics. Experimental results are given which demonstrate the validity of the methods advanced.

621.396.822.1

Non-linear crosstalk in multi-carrier multichannel systems.—P. GÜTTINGER, *Archiv der Elektrischen Übertragung*, **9**, pp. 573-7, December 1955.

The problem of crosstalk in multi-carrier multichannel systems due to non-linear transmission characteristics is discussed analytically. Particular consideration is given to the case of phase-modulated individual channels where the radio frequency carriers are within one octave of frequency. At first a characteristic is assumed that can be represented by an analytical function, and the interference factor is calculated for the case of a single modulated channel, giving results which may be used with modulation of all channels. Finally, the mixing problem is discussed, assuming additive mixing and an angular mixing characteristic. It appears that by a suitable choice of operating point and heterodyning voltage, certain interference products can be considerably reduced.