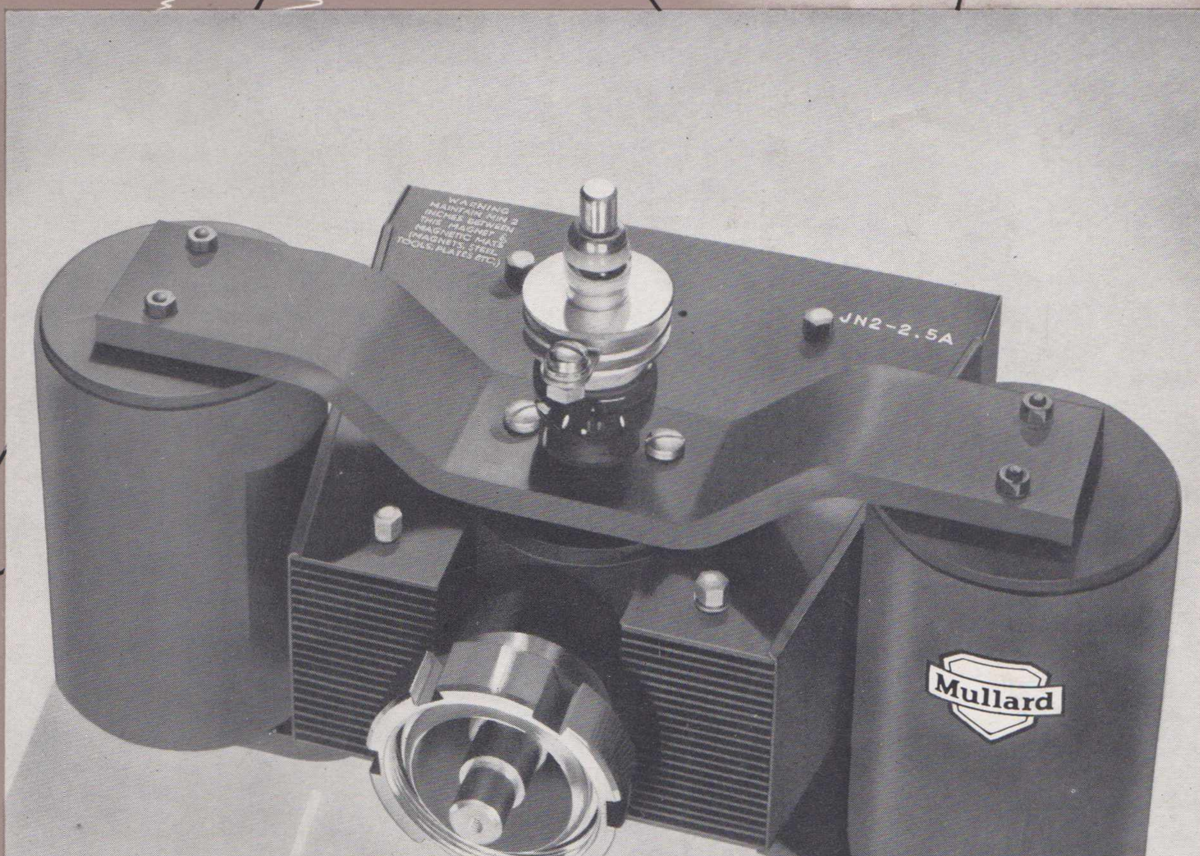
Mullard Outlook

Australian Edition



VOL. 3 No. 3, MAY-JUNE
1960

MULLARD - AUSTRALIA PTY. LTD.

VOL. 3 No. 3 MAY-JUNE, 1960

Editorial Office:
35-43 Clarence Street, Sydney
Telephone—BX 2006

Editorial Staff:
B. P. A. BERESFORD, A.M.I.R.E. (Aust.)
JOERN BORK

CONTENTS:

The JN2 - 2.5A Magnetron which is shown on the cover of this issue is primarily designed for microwave heating. The power output is 2.5kW c.w. at a fixed frequency of 2.45 ± 0.025 Gc/s.
More detailed information may be found on Pages 30 and 31.

MULLARD-AUSTRALIA PTY. LTD.
35-43 CLARENCE STREET, SYDNEY
Phone: BX 2006
123-129 VICTORIA PDE, COLLINGWOOD, N.5
VICTORIA
Phone: 41 6644
Associated with
MULLARD LTD., LONDON
MULLARD OVERSEAS LTD.
MULLARD EQUIPMENT LTD.

Gigacycles—but not at St. Trinians!

Many Outlook readers are aware that a Gigacycle is not a device stowed in the bike shed of a girls' school but is a unit of frequency for—one Gigacycle per second = 1,000 Mc/s or 10^9 c/s.

You may consider it somewhat remote from television and sound broadcast receivers, 23″ pictures tubes, frame grid valves and so on, but to the industrial and professional valve user it is a neater unit than "so many thousands of megacycles."

The professional flavour of this Outlook covers two noteworthy developments by Mullard engineers, one dealing with a method of stereophonic broadcasting, the other a new concept in tank circuit design for industrial R.F. generators.

Such aspects are generally confined to the many purely professional Mullard publications, e.g., the Industrial Valve News Letter, April, 1960, deals, among other things, with a new travelling wave tube for the 4 Gigacycle (Gc/s) band and a forward wave amplifier for the band 2.7 to 3.3 Gigacycles.

When selling television receivers equipped with Mullard Radiant Screen picture tubes and valves or when using these in your Service Department or your television transmitters, we suggest this relationship, this reputation, this pedigree of a long line of electronic valve development as your safeguard for reliability and customer confidence.

VIEWPOINT WITH MULLARD

STEREOPHONIC BROADCASTING DEMONSTRATION

Engineers and authorities in several countries have for some time been investigating the practicability of providing radio transmissions of stereophonic sound, and various methods are at present under review by different bodies. The European Broadcasting Union, meeting in Cannes, France, in February, considered a number of proposals, among them being a new system, developed by G. D. Browne of the Mullard Research Laboratories, based on a time multiplexing technique similar to that used in some point-to-point communications systems.

This system has recently been demonstrated to the Press, and representatives from industry and other interested parties including the B.B.C.

The demonstrations were of transmissions from an experimental transmitter to a receiver over a short distance. A variety of programme material showed the capabilities of the system and a further interesting demonstration was of its bi-lingual possibilities.

The system requires only one VHF carrier wave to transmit the stereophonic signals, and, in fact, one important feature is that existing VHF/FM broadcast transmitters could be used with only minor modifications.

Simple Stereo Receivers

The system's chief advantage, however, is that it would enable stereophonic receivers to be produced which are little more complex than conventional monophonic sets: apart from the second loudspeaker and audio stage necessary for stereophony, the only addition would be a circuit using at most two valves, or possibly one transistor and two crystal diodes. Existing VHF/FM radiograms with facilities for stereophonic record reproduction would, in many cases, be convertible to receiving stereo broadcasts on the system by the addition of only one valve and its associated circuit.

Compatibility

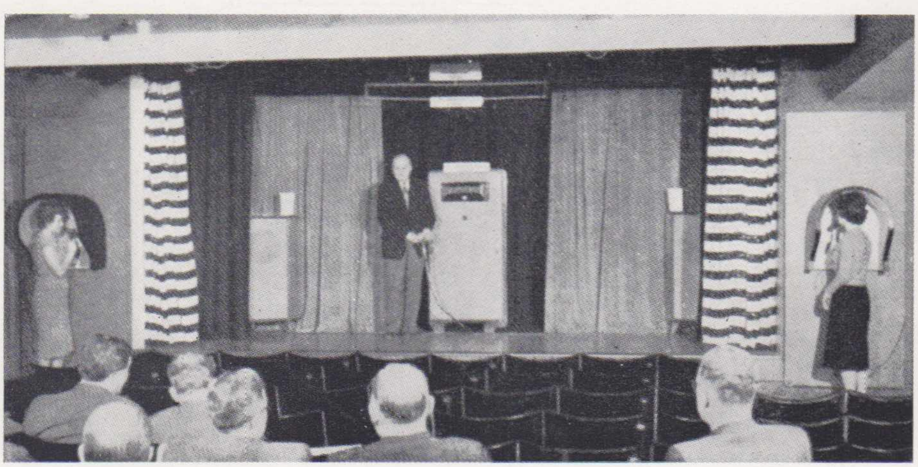
The system meets all the generally accepted performance requirements for stereophonic broadcasting. It is compatible with the existing VHF sound transmissions, so that conventional VHF sets will accept the stereo transmissions and reproduce them monophonically in the normal way; it is also reversely compatible, which means that the listener with a stereophonic receiver designed for the system will hear the monophonic signal from both channels of his receiver when tuned to a monophonic transmission. Both these facilities were demonstrated.

High Fidelity

Full stereophonic effects, comparable with the best provided by the direct reproduction of stereo discs or tapes, are given by the system, and since the audio bandwidth transmitted is 15,000 c/s, very high quality reproduction is possible from a well-designed receiver. The radio frequency channel bandwidth is such that the system occupies no more channel space than is normally taken up by conventional VHF transmissions.

Bi-lingual Possibilities

Apart from its primary purpose of transmitting entertainment stereophonic radio programmes the system can also be used for bi-lingual speech transmissions and other twin-signal applications. Thus a programme could be broadcast in, say, French and German simultaneously from the same transmitter; one channel of the stereophonic receiver would give monophonic reproduction of the French version and the other of the German. This facility may have an application in schools broadcasting where, by placing the two speakers of the stereo receiver in different classrooms, two language lessons, for example, could be given at one and the same time. It would also be useful for entertainment broadcasting in countries where not all the inhabitants have the same language.



On the stage at the demonstration rehearsal is Mr. G. D. Browne, behind him is the experimental stereophonic receiver, and on the left and right the two loudspeakers. Helping to demonstrate bi-lingual speech transmission were French-born Mrs. Conchita Hacking from our Blackburn factory and Miss Nina Isitt from Mullard House.

MULLARD-AUSTRALIA PERSONALITIES

With this issue we have pleasure in introducing Mr. Laurence Wade, who heads the Mullard Industrial and Professional Valve Department at 35-43 Clarence Street, Sydney.

Mr. Wade joined the Company in 1956, bringing with him an engineering background gained in both industry and Government Departments and commercial experience as a Sales Engineer with an electronic component manufacturer.

Late last year he visited our Parent Company in England and associate companies in Europe and the U.S.A., where he studied new developments in special valves, tubes and semiconductors.

He is an Associate of the Sydney Technical College, Associate Member of the Institution of Engineers Australia, and an Associate Member of the Institution of Radio Engineers Australia.

An enthusiastic amateur boat builder, Mr. Wade spends much of his leisure time boating and water skiing on Pittwater, near Sydney.

Mullard

MODULATOR DESIGN WITH OC26 TRANSISTORS

The application of transistors to low power modulators can result in smaller transmitter power supplies, thus reducing the cost of transmitters designed to operate from storage batteries. The additional increase in overall efficiency may also be considered a worthwhile improvement.

A transistor modulator requires only a fraction of the space taken up by a comparable valve unit and allows full use of the high voltage power supply for the R.F. section of the transmitter.

Among the advantages of a modulator of this type are compactness, high overall efficiency, no warm-up time and low quiescent current when not modulating. The design to be discussed will modulate a transmitter R.F. output stage having an input of up to 25W at any desired impedance level depending on the number of turns on the secondary of the modulation transformer.

The design investigated is required to modulate a P.A. stage of 25 W anode input from a 325 V rail, i.e., modulating impedance:—

$$Z = \frac{E^2}{W} = \frac{325^2}{25} = 4{,}225$$

The power output required from the modulator is, of course, 12.5 W for 100% modulation.

Peak current rating of the OC26 is 3.5 A and the "knee voltage" approximately 1 V.

Assuming a 12 V supply rail the load per transistor may be determined:—

$$R_L = \frac{Ec}{Ic} = \frac{12-1}{3.5} = \frac{11}{3.5} = 3.15\Omega$$

R_{cc} is thus 3.15 x 4 = 12.6Ω (assuming class B operation).

Output transformer impedance ratio is thus $\frac{4225}{12.6} = 335$.

Output transformer turns ratio $= \sqrt{335} = 18.3$.

The primary inductance necessary, calculated on the basis of a 6dB low frequency loss 1 octave below 300 c/s is:—

$$L = \frac{X}{2\pi f} = \frac{12.6}{6.28 \times 150} = 13.3\text{mH}$$

Selecting a core of two Z371033 "C" core loops the total core area is 0.258 square inches and the mean path length 4.26 inches. The primary turns may be calculated by reference to the peak flux density in the core, and verified by calculating the resulting inductance.

$$\text{Primary turns } N_p = \frac{E \times 10^8}{26\ fA\ B\text{max}} = \frac{11 \times 10^8}{26 \times 150 \times 0.258 \times 16 \times 10^3} = 69 \text{ turns}$$

(f = 1 octave below 300 c/s i.e. 150 c/s)

For convenience the primary winding is 60 + 60 turns bifilar wound.

$$L \text{ at } 1000 \text{ c/s} = \frac{3.2\mu A\ N^2}{10^8\ L} \text{ at Bmax}$$

= 16000 gauss at 150 c/s

i.e. 2400 gauss at 1000 c/s

μ 1000 c/s = (not quoted)

To determine the inductance it is necessary to compute the incremental permeability by Hanna's method and so in this design the magnetising force due to anode current flow in the secondary winding must be considered.

Total secondary turns = 18.3 × 120 = 2,200

D.C. secondary current = 75mA (25 W input to P.A.)

∴ Magnetising ampere turns = 165

$$\text{Thus } \frac{AT}{L} = \frac{165}{4.26} = 39 \text{ ampere turns per inch}$$

and Magnetising Force = 39 x 0.495 = 19.2 *Oersteds*

Hanna's curves for the core material indicate a gap ratio of 0.001″ per inch and an incremental permeability of approximately 500.

The cores should thus be assembled with 0.002″ paper gaps and the inductance may be calculated as:—

$$L \text{ at } 1000 \text{ c/s} = \frac{3.2\mu A\ N^2}{10^8 L} = \frac{3.2 \times 500 \times 0.258 \times 120^2}{10^8 \times 4.26} = 14 \text{ mH}$$

For a peak collector current of 3.5 A a nominal OC26 requires an input base current of 140 mA. At a collector-emitter potential of 1 V the base-emitter potential under these conditions is some 800 mV.

The OC74 transistor can supply this peak drive power for an input of 1.4 mA at a base-emitter potential of 330 mV approximately.

The peak input impedance looking into the OC74 is thus:—

$$Z = \frac{E}{I} = \frac{800 + 330}{1.4} = 800\Omega \text{ approx.}$$

i.e. an average input impedance would be almost twice this figure—say 1500Ω.

The microphone input transformer may be designed to match the 50Ω source impedance of a carbon microphone to this value:—

$$\text{Impedance ratio} = \frac{1500}{50} = 30;$$

$$\text{Turns ratio} = \sqrt{30} = 5.45$$

Inductance parameters of this transformer may be calculated by Hanna's method assuming a D.C. bias current of 50 mA, due to the carbon microphone, flowing through the primary winding. A simple butt joint (0.002″ paper) suffices for this value of induction with a 120 turn primary; the turns ratio implying a secondary winding of 650 + 650 turns.

The push-pull class B "super alpha" configuration has been chosen for this modulator because of the savings in components and ease with which the penultimate stage operates. An added advantage is that only input and modulation transformers are necessary—no separate driver transformer is required. Some 9 dB of negative feedback is applied (with nominal transistors) which not only enhances the frequency response and minimises distortion but reduces the effect of transistor characteristic spreads.

Fig. 1 details the schematic of the prototype modulator.

Neglecting input transformer losses the primary input requirements become:—

$$V_{peak} = \frac{1.13}{5.4} = 0.21 \text{ V}$$

$$I_{peak} = 1.4 \times 5.4 = 7.8\text{mA}$$

With 9 dB of negative voltage feedback applied

$$V_{peak} = 0.21 \times 2.82 = 0.59 \text{ V}$$

Assuming an input transformer efficiency = 80%

$$V_{in} \text{ peak} = 0.59 \times 1.2 = 0.71 \text{ V}$$

The input power is thus:—

$$P_{in} = \frac{0.71^2}{2 \times 50} = 5 \text{ mW}; \quad P_{out} = 15 \text{ W}$$

Therefore power gain:—

$$\frac{15 \times 10^3}{5} = 3 \times 10^3 = 35 \text{ dB approx.}$$

$$\text{Input sensitivity} = \frac{0.71}{\sqrt{2}} = 0.5 \text{ V r.m.s. (into } 50\Omega)$$

which is conveniently supplied by a carbon microphone.

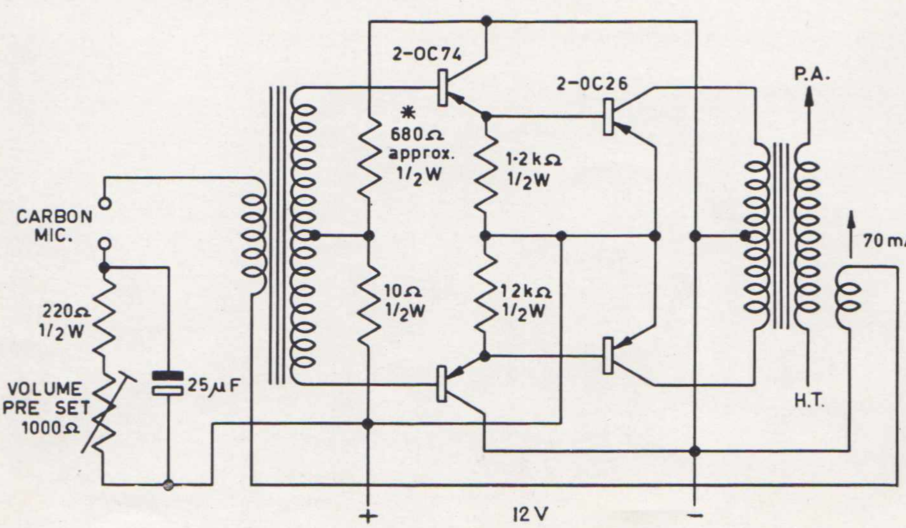


FIG. 1

* Adjust base bias for total quiescent current of 40mA, not including microphone current.

The tertiary feedback turns may be determined by considering the voltage gain from input to the primary of the modulation transformer.

Input voltage (without feedback) $= \dfrac{0.71}{2.82} = 0.25$ V peak

Output (Primary) voltage at 15 W into 12.6Ω

$= \sqrt{WR} = \sqrt{15 \times 12.6} = 13.8$ V r.m.s.

$= 19.4$ V peak, say 20 V peak

Voltage Gain (without feedback) $= \dfrac{20}{0.25} = 80$ times

$$\text{Feedback Fraction } -B = \frac{A - M}{AM} = \frac{80 - 28.5}{80 \times 28.5} = \frac{51.5}{2280} = \frac{1}{40}$$

The tertiary turns are thus $\dfrac{120}{40} = 3$ turns

No allowance has been made in this calculation for the loading of the tertiary winding or for coupling losses. In practice it was found that 4 turns were necessary to obtain 9 dB of feedback but at least the approximate calculation performed above serves as a guide.

CONSTRUCTION

The complete transformer specifications are given in the table below.

TRANSFORMER DETAILS
MODULATION TRANSFORMER

Core: 2 loops of Z371033 "C" butt joint 0.002" paper.

Half Secondary: 1,100 turns of 34 B & S enamelled copper wire wound in 9 layers.

Primary: 60 + 60 turns bifilar wound of 21 B & S enamelled copper wire wound in 4 layers.

Half Secondary: 1,100 turns of 34 B & S enamelled copper wire wound in 9 layers.

Tertiary: 4 turns of 27½ B & S enamelled copper wire.

Insulation: Between windings 0.005 Presspahn.
Between secondary layers 0.002 Kraft.
Over tertiary 0.005 Presspahn.

INPUT TRANSFORMER

Core: ½" stack of 1½" x 0.014" scrapless EI grain oriented lamination, butt joint 0.002" paper.

Secondary: 650 + 650 turns bifilar wound of 34 B & S enamelled copper wire wound in 16 layers.

Primary: 120 turns of 30 B & S enamelled copper wire wound in 2 layers.

Insulation: Between windings 2 x 0.002" Kraft.
Between layers 0.002" Kraft.
Over Primary 2 x 0.002" Kraft.

Since the collector is common with the transistor case, mica spacers (supplied with the unit) must be inserted between the transistor and chassis. In order to reduce the possibility of the mica being accidentally punctured, the transistor mounting holes in the chassis should be carefully deburred. Component layout is not critical, although care should be taken to ensure that the chassis is large enough to provide an efficient heat sink for the output transistors. Please note that the OC74 driver transistors should also be clamped to the chassis with the clips provided.

A standard carbon microphone is suitable for use with this modulator and although not shown in the circuit diagram should be connected so that the "press-to-talk" button provides power to the modulator only when in the "on" position.

TESTING

The measured performance of the prototype modulator is listed in the table below.

MEASURED PERFORMANCE OF PROTOTYPE MODULATOR

Maximum power output at onset of clipping	= 15.6 W
Power frequency response at 10 W. (150 c/s – 18 k/cs)	= ±3 dB
Total harmonic distortion at 1000 c/s and 10 W	= <0.85%
Total harmonic distortion at 1000 c/s and 15 W	= <1.25%
Total current drain at 15 W (sine wave)	= 2.1 A
Total current drain at 15 W (Average speech, peaks)	= 0.6 A approx.
Power efficiency	= 60% approx.
Total quiescent current at 25° C	= 60 mA
Stability Factor 25 – 55°C	= 6 approx.

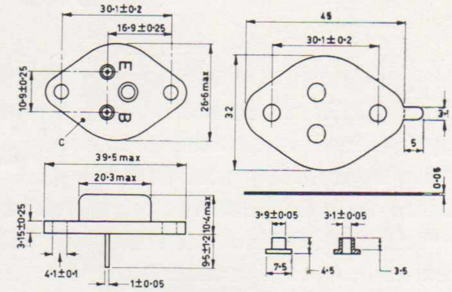
It should be noted that the OC26 output transistors employed have a fα*e* of 4.5 k/cs. The phase shift and drop in α' gives rise to a decline in transistor efficiency which causes an elevation of junction temperature. To help stabilise this runaway condition at the higher frequencies a base leak of 1.2 kΩ has been used with the OC26's. Nevertheless, when checking for maximum power output at frequencies above 1 k/cs a current meter should be inserted in series with the supply for visual indication of runaway, and drive power only applied for a brief period.

There is not sufficient sustained high frequency power in normal speech to precipitate this form of instability and hence the performance of the modulator does not suffer as the power level in speech (even female) declines rapidly as the frequency increases above some 800 c/s. Similarly the response of the carbon microphone falls rapidly above 2.5 k/cs. Nevertheless, as an added precaution when testing the modulator an 0.47μF condenser in parallel with the 50Ω source impedance audio generator is recommended to synthesise the high frequency roll-off of the microphone.

After construction the unit should be tested for proper operation by connecting a dummy load to the output of the modulation transformer and checking the current variation whilst speaking into the microphone. Do not, under any circumstances, try to operate the unit without a load as this may damage the output transistors. A modulation percentage test may be made when the modulator has been connected to the transmitter. It is desirable to use an oscilloscope for this purpose in order to achieve the optimum setting for the pre-set volume control.

J. R. GOLDTHORP

OC 26 GERMANIUM TRANSISTOR



All dimensions in mm

ABSOLUTE MAXIMUM RATINGS

Collector

$-V_{CB}$	= max. 32V
$-V_{CEM}$	= max. 32V
$-V_{CE}$	= max. 16V

Current

$-I_C$	= max. 3.5A
P_C	= max. 12.5W

Emitter

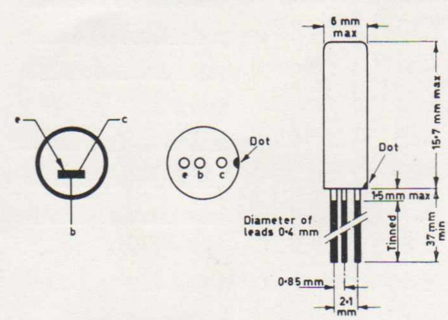
$-V_{EB}$	= max. 10V
T_j continuous operation	= max. 90°C

TEMPERATURE

Temperature rise from the mounting base of the transistor to the junction

K = 1.0°C/W = max. 1.2°C/W

OC 74 GERMANIUM TRANSISTOR



All dimensions in mm

ABSOLUTE MAXIMUM RATINGS

Collector

$-V_{CBM}$	= 20 V
$-V_{CB}$	= 20 V
$-V_{CEM}$	= 20 V
$-V_{CE}$	= 20 V

Current

$-I_{CM}$	= 300mA
$-I_C$	= 300mA
P_c (at 25° Camb)	= 550mW

Emitter

$-V_{EBM}$	= 6 V
$-V_{EB}$	= 6 V
I_{EM}	= 310mA
I_E	= 310mA

MULLARD INDUSTRIAL VALVES FOR RADIO FREQU

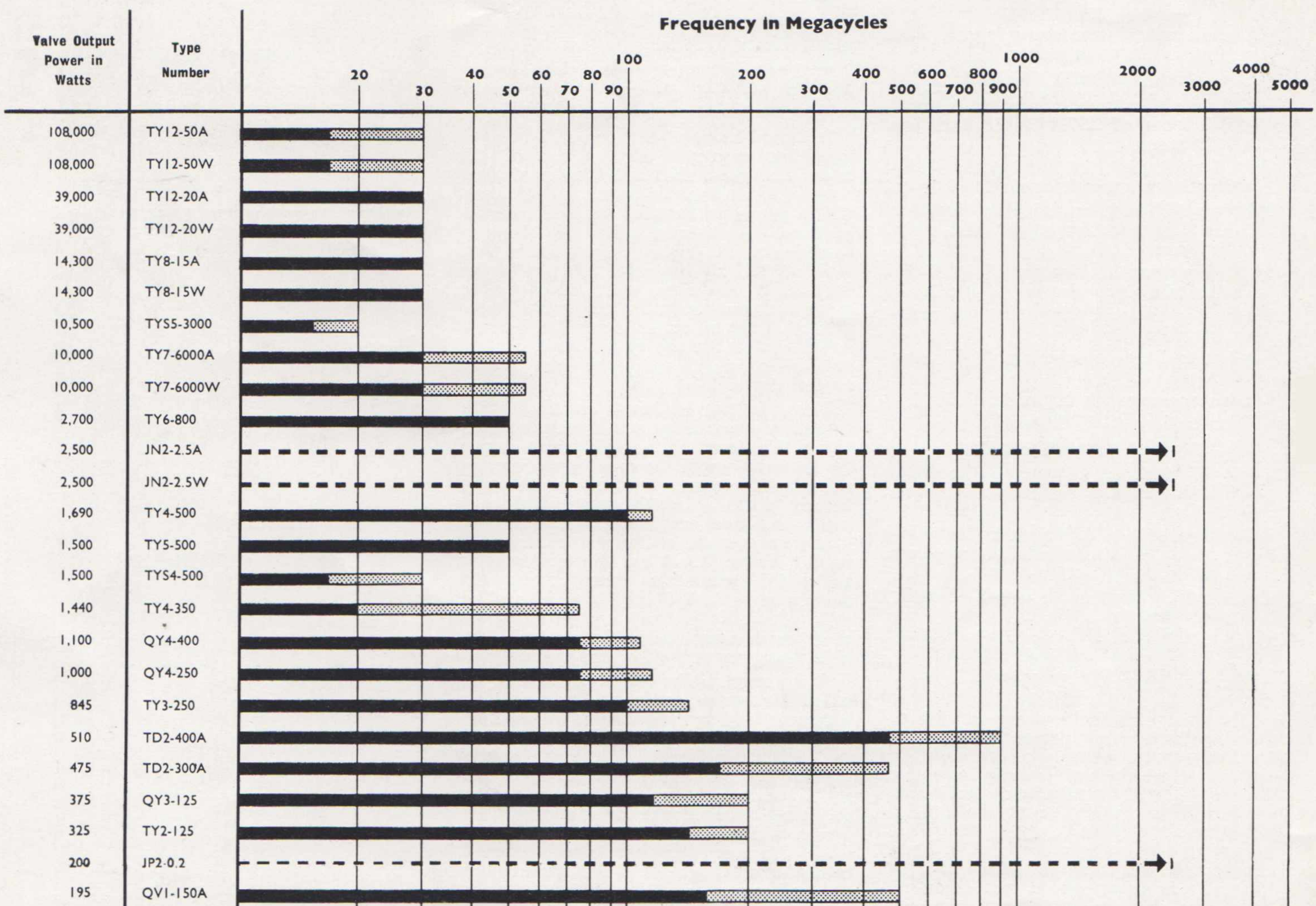
The demand for oscillator valves for use as R.F. generators in industrial heating and electromedical diathermy applications is increasing rapidly. Transmitting valves have been widely used for this purpose for some years. More recently, several new valves have been developed specifically for industrial service rather than for normal transmitting and communications requirements. This new conception of design permits an increased degree of overload to be sustained in certain directions; and it allows the valve characteristics to be so arranged that the R.F. output maintains a more constant level with changes of load impedance. Reliability and long life under rigorous industrial conditions have been regarded as major design considerations in these new valves.

The present Mullard range of triodes, tetrodes and magnetrons for the various types of R.F. heating are given in the Table.

Full data sheets for all types, and for the Mullard Thyratrons and mercury and xenon rectifiers which are required in the associated power supplies may be had on application.

Technical advice on valves for R.F. heating circuits at all power levels is freely available from the Mullard Industrial and Professional Valve Department.

The power quoted here is the maximum valve output available in industrial equipment with continuous class C operation at frequencies up to the junction of the solid and shaded lines. The extent to which a valve may be used at higher frequencies with reduced ratings is shown by the shaded line.



Increased Ratings The ratings shown here are for **continuous** operation. In many industrial applications where valves are operated **intermittently** ratings may be considerably increased. Details are available on request.

NCY HEATING AND ELECTROMEDICAL DIATHERMY

TETRODES AND TRIODES

Type No.	USA/ EIA No.	CV No.	Description	Cathode Type	Cathode V_f or V_h (V)	Cathode I_f or I_h (A)	gm (mA /V)	CLASS 'C' CONTINUOUS WAVE OPERATION — Limiting Values (Max. Ratings) P_a (W)	V_a (kV)	I_k (mA)	V_{g2} (V)	Freq. (Mc/s)	Typical Operating Conditions — Full Ratings Freq. (Mc/s)	Full Ratings P_{out} (W)	Reduced Ratings Freq. (Mc/s)	Reduced Ratings P_{out} (W)
QVI–150A	4X150A	2519	Tetrode for U.H.F. therapy	IH	6·0	2·6	12	150	1·25	250	300	500	165	195	500	140
QY3–125	6155	2130	Radiation cooled beam power tetrodes for diathermy equipment.	DH	5·0	6·5	2·2	125	3·0	300	400	200	120	375	200	225
QY4–250	6156	2131		DH	5·0	14·1	4·0	250	4·0	420	600	120	75	1000	120	500
QY4–400	4–400A	—	General purpose tetrode	DH	5·0	14·1	4·0	400	4·0	420	600	110	75	1100	110	700
TY2–125	5866	1924	R.F. power triodes for general purpose industrial heating applications. Suffix "A" and "W" to the type number indicates "forced air cooled" and "water cooled" respectively.	DH	6·3	5·4	2·8	135	2·5	250	—	200	150	325	—	—
TY3–250	5867	1350		DH	5·0	14	5·0	250	3·0	480	—	150	100	845	150	350
TY4–350	833A	635		DH	10	10	4·0	400	4·0	600	—	75	20	1440	—	—
TY4–500	5868	1351		DH	10	9·9	4·5	450	4·0	650	—	120	100	1690	120	875
TY5–500	—	—		DH	5·0	32·5	3·3	500	5·0	770	—	50	50	1500	—	—
TY6–800	7092	—		DH	6·3	32·5	5·1	800	6·0	1050	—	50	50	2700	—	—
TY7–6000A	6961	—		DH	12·6	33	15	6000	7·0	2300	—	55	30	10000	50	6000
TY7–6000W	6960	—		DH	12·6	33	15	6000	7·0	2300	—	55	30	10000	50	6000
TY8–15A	—	—		DH	6·3	130	23	10000	8·0	5000	—	30	30	14300	—	—
TY8–15W	—	—		DH	6·3	130	23	15000	8·0	5000	—	30	30	14300	—	—
TY12–20A	—	—		DH	8·0	130	25	15000	13	6300	—	30	30	39000	—	—
TY12–20W	—	—		DH	8·0	130	25	20000	13	6300	—	30	30	39000	—	—
TY12–50A	6078	—		DH	17·5	196	50	45000	15	15000	—	30	15	108000	30	50000
TY12–50W	6077	—		DH	17·5	196	50	50000	15	15000	—	30	15	108000	30	50000
TYS4–500	—	1889	R.F. power triodes with silica envelopes	DH	10	10	6·0	500	4·0	750	—	30	15	1500	18	1250
TYS5–3000	—	—		DH	20·5	26	15	3500	6·0	2800	—	20	12	10500	20	6300

DISC SEAL TRIODES

Type No.	USA/ EIA No.	CV No.	Description	Cathode Type	V_f or V_h (V)	I_f or I_h (A)	gm (mA/V)	P_a (W)	V_a (kV)	I_k (mA)	V_{g2} (V)	Freq. (Mc/s)	Full Freq. (Mc/s)	Full P_{out} (W)	Reduced Freq. (Mc/s)	Reduced P_{out} (W)
TD2–300A	7004	—	Forced air cooled glass/metal triode for U.H.F. therapy	DH	3·4	19	20	300	2·0	520	—	470	175	425	470	385
TD2–400A	—	—	Forced air cooled ceramic/metal triode for U.H.F. therapy.	DH	3·4	19	10	400	2·0	520	—	900	470	405	810	214

MAGNETRONS

Type No.	USA/ EIA No.	CV No.	Description	CATHODE Type	CATHODE V_h (start) (V)	CATHODE V_h (run) (V)	CATHODE I_h (A)	FREQUENCY RANGE (Mc/s)	TYPICAL OPERATION C.W. V_a (kV)	I_a (A)	P_{out} (kW)
JN2–2·5A	7091	—	Forced air cooled, high power microwave generator for industrial processing and heating	1H	5·0	1·5	32	2425 to 2475	4·5	0·85	2·5
JN2–2·5W	7292	—	Water cooled, high power microwave generator for industrial processing and heating	1H	5·0	1·5	32	2425 to 2475	4·5	0·85	2·5
JP2–0·2	7090	—	Radiation cooled generator for microwave therapy	1H	5·3	4·5	3·3	2425 to 2475	1·6	0·2	0·2

HIGH-SPEED DECIMAL STEPPING TUBES

OPERATING PRINCIPLE OF THE TROCHOTRON

Cold cathode stepping tubes have the advantages which are shared by all cold cathode devices—not the least being the absence of a heater, which means simpler supply arrangements and no temperature problem. However, any gas-filled device has a maximum frequency limit which, broadly speaking, is set by the deionisation time of the gas.

In a cold cathode stepping tube it is not possible to switch the glow discharge from one output electrode to the next faster than a few thousand times per second. Thus, although the Mullard Z302C, Z303C, and Z502S are widely used in a great variety of applications where the maximum frequency is a few kc/s, there are certain jobs where a still faster tube is necessary. For example, in nucleonic scalers and in some decimal computers or electronic switching circuits, an operating speed of 1.0 Mc/s and above may be required, and the stepping tube must be able to handle successive input pulses which arrive almost simultaneously.

For such high speeds and good resolution it is necessary to use a vacuum tube, in which the switched element is not a relatively sluggish glow discharge but a highly mobile electron beam similar to the beam of a television tube. The "trochotron" is a device of this kind.

Trochotron Principle

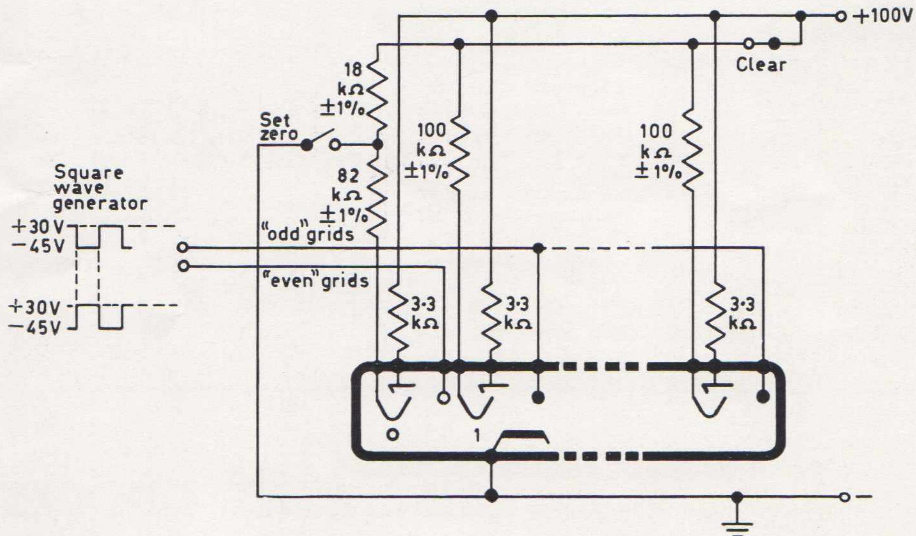
The cold-cathode technique of manipulating the breakdown voltages of a series of guides is obviously not possible in a vacuum tube. The trochotron makes use of a quite different principle which resembles that of a simple magnetron.

If a suitable electric field is set up between a cathode and an anode, any electron leaving the cathode will travel straight to the anode. If, however, a magnetic field is applied as well, at right-angles to the electric field, the electron can be forced into a helical or trochoidal path somewhere between cathode and anode. If the relative strengths of the electric and magnetic fields are suitably adjusted, no electrons will reach the anode at all.

In a tube in which the cathode is surrounded by, say, ten anodes, permanent circulation of electrons in this way will go on so long as the electric field is uniform and the magnetic field is uniform. No electrons will reach any anode. If, however, either the electric or magnetic field is deliberately distorted, the electrons can be directed to whichever anode one wishes. If some means is then provided for rapid transfer from anode to anode, a high-speed stepping tube has been achieved.

In a trochotron it is simplest to use a permanent magnet to provide the magnetic field; and since the magnet could not very well be moved physically at the high speeds envisaged, the *magnetic* field is fixed in strength and direction. The required "distortion" must therefore be introduced in the *electric* field.

The electric field is set up by voltages applied to the anodes *(targets)* and the *spades*. In the circuit diagram the targets are the electrodes connected through 3.3kΩ to the h.t. line, and the spades are the U-shaped electrodes. To the right of each spade is a *grid,* indicated by an open circle or black dot, which is used for switching from target to target (a process which will be outlined later). In the Mullard ET51 there are ten of these target-spade-grid systems arranged in a circle round the cathode. The whole tube is cemented inside a cylindrical permanent magnet.



Starting

Now, with say 100V applied to all the targets and spades, the electric field is uniform, and no current will flow. But if the "set zero" switch is momentarily closed, spade "0" will fall to cathode potential. The field will therefore be distorted in the "0" region, and electrons will be able to reach the "0" target. When the "set zero" switch is then opened, a small part of the current will flow to the "0" spade and through its 100kΩ resistor, so that this spade will be held down near cathode potential, the distortion will be maintained, and the electron beam will be effectively "locked" on the "0" target.

All the ten output systems in a trochotron are precisely the same, therefore the "set zero" circuit could be connected to any spade. However, it is usual to start at the "0" position.

Resetting

At the end of a count, the electron beam may of course be resting on any target, say "5". To get it to the "0" position for the start of another count the tube is first put into the "clear" state by momentarily opening the switch in the spade supply, thus restoring the uniformity of the electric field and setting up once more the circulatory electron path. Operation of the "set zero" switch then initiates conduction to target "0", as already described.

Switching

Some means now has to be introduced to transfer the electron beam from one target to the next. This is done by moving the distortion of the electric field onward one position. The action is provided by the grid.

If the beam is on target "0", and is therefore locked there by the voltage drop in the spade "0" resistor, the distortion of the electric field can be extended to the right by applying a negative voltage to grid "0" (which is the open circle in the circuit diagram). This local distortion starts conduction to spade "1". The spade characteristic is such that the spade "1" current rises rapidly and spade "0" ceases to conduct. The electron beam has therefore been transferred to target "1" and locked there by spade "1".

If all the grids were connected together the beam would obviously be transferred immediately to spade "2" and target "2". To prevent this happening, the "even" grids and "odd" grids are separately connected. To one set is applied the negative switching voltage (–45V), and to the other set a positive voltage (+30V) which blocks unwanted transfer of the beam.

Drive

The incoming pulses which are to be counted are used to trigger a bistable gate circuit (not shown in the diagram) which feeds negative pulses alternately to the two sets of grids. Thus the first incoming pulse will make the "even" grids (including grid "0") negative; the beam will switch to position "1", but will be stopped from further movement by the positive voltage on the "odd" grid "1". The second incoming pulse will reverse this state of affairs, making grid "1" negative—thus stepping the beam to position "2", but making grid "2" positive—thus blocking transfer to position "3".

The incoming pulses may come in only at long intervals; but the bistable trigger circuit will of course stay in its last position however long the interval, and the beam will not be moved again until the next pulse arrives. On the other hand, pulses may come in rapidly—say at a repetition rate of 1.0 Mc/s—and the negative pulses from the trigger circuit will then be of short duration.

Unlike the cold cathode stepping tube, the trochotron will circulate in only one direction, which is determined by the direction of the magnetic field and by the design of the electrode system.

Output

When the electron beam is locked on a particular target by means of the voltage drop in the spade resistor, most of the output current flows to the target and through its series resistor. The target characteristic is of pentode type, and the tube must be operated "above the knee" to obtain a substantially constant output of 5.5mA. This current (which compares favourably with the 400μA or so of a cold cathode stepping tube) can be used to operate a direct read-out digital indicator; it can provide some required action after a predetermined number of input pulses has occurred (that is, it can perform "selector" functions); or it can drive a further trochotron in a units, tens, hundreds . . . system by means of a simple bistable coupling circuit.

There is no visible indication of the position of the electron beam. If this is required, then the ET51 can be connected directly to a Z503M or Z510M without valve coupling circuits.

Performance

In the above description the need for uniformity of the electric and magnetic fields has been stressed. If the magnetic field is distorted by stray fields, by nearby magnetic materials, or by another trochotron mounted within say four inches, then the tube will not work reliably. Also, the magnet should not be knocked, and it must not be touched by steel tools.

Close tolerance resistors (1%) should be used in the spade circuit, and they should be mounted directly on the tube socket, because any stray capacitances would slow down the switching action. The supply voltage for the spades must be within the absolute limits 90 to 110V.

If these precautions are observed, the ET51 will work reliably and accurately at switching speeds up to at least 1.0Mc/s. The chief uses of the tube are in nucleonic scalers, decimal computing equipment, and electronic switching circuits. In these applications one ET51, with its comparatively simple circuitry, can take the place of a number of conventional valves or transistors and their associated components.

ET51

ABRIDGED DATA

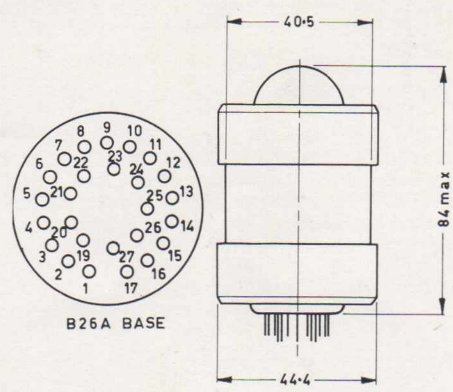
HEATER

Vh	6.3	V
Ih	300	mA

OPERATING CONDITIONS

Target supply voltage	100 ± 10%	V
Target load resistor	3.3	kΩ
Target output current (each target)	5.5	mA
Spade supply voltage	100 ± 10%	V
Spade current	1.0	mA
Spade load resistor	100 ± 1%	kΩ
Grid bias voltage	+25	V
*Minimum grid input pulse amplitude	–60	V
Grid current (approximate)	50	μA
Cathode current	6.5	mA

*The grid input capacity of each group of grids is approximately 10pF.



All dimensions in mm.

Pin No.	Connection	Pin No.	Connection
1	Spade 0	14	Spade 2
2	Target 9	15	Target 1
3	Target 8	16	Grid "even"
4	Grid "odd"	17	Target 0
5	Target 7	19	Spade 9
6	Spade 7	20	Spade 8
7	Target 6	21	Heater
8	Target 5	22	Spade 6
9	Spade 5	23	Spade 4
10	Target 4	24	Spade 3
11	I.C.	25	Heater
12	Target 3	26	Spade 1
13	Target 2	27	Cathode

Mullard

DESIGN OF LAMINATED CIRCUITS

FOR INDUSTRIAL R.F. GENERATORS

With conventionally designed tank circuits the combination of high output with low loss involves very high cost. The laminated circuit described here offers a good compromise between efficiency and cost. Manufacturing costs for quantity production are low enough to offset the extra cost of the separate screening necessary. The physical size of the circuit is such that it can be incorporated in existing equipment.
It is envisaged that, with standardisation to a few preferred sizes of lamina, a manufacturer will be able to assemble tank circuits covering a wide range of supply voltage, operational frequency and loaded-Q.

An industrial r.f. generator is often subjected to a load incorporating a reactive component, the magnitude of which can vary considerably during operation. This varying reactance will be, at least partially, reflected into the tank circuit and could cause frequency drift of the oscillator outside the desired limits.

Although this drift cannot be completely eliminated, it can be kept within predetermined limits if the capacitance of the tank circuit is made sufficiently large, so that the effect of the reflected reactance is small compared with the reactance presented by the tank capacitance. A large tank-circuit capacitance implies large circulating currents and the resistance of the circuit must be very small to avoid wastage of power. The ideal circuit could be very costly.

The circuit described in this article is an attempt to achieve good transfer efficiency in a construction which is suitable for mass production and hence relatively inexpensive.

CIRCUIT CONSIDERATIONS

Past experiments suggest that, with normal working voltages, loaded-Q values should be of the order of 100 to 400, if the stability conditions for various frequency bands are to be satisfied. With these high values, circuit design and construction become increasingly difficult.

If a set of valve conditions gives an r.m.s. value for the fundamental component I_1, the circulating current set up in the tank circuit,

$$I_c = Q_L . I_1$$

The power loss in the circuit,

$$P_c = I_c^2 . R_s$$

where R_s is the total surface resistance of the tank circuit. If R_s is too high, an uneconomically large share of the total available output power will be dissipated within the tank circuit. In many conventional layouts most of the surface resistance arises from the inevitable joints between the tank capacitor and tank inductance.

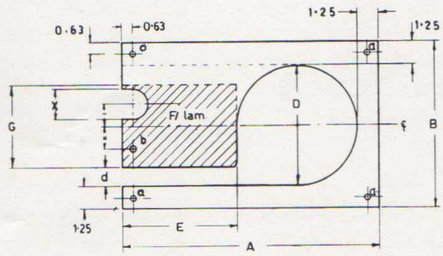
It was thought necessary to aim at a solution giving an integral L and C with a minimum of joints or sharp corners at right angles to the current flow, but to depart from the coaxial and cavity types of construction which are expensive.

LAMINATED CIRCUIT

Construction

A circuit was evolved having, as building elements, rectangular plates or laminae, each of which represents a multiple of the inductive path as well as a proportionate fraction of the tank capacity. From Figs. 1 and 2 it can be seen that if these laminae are stacked with the desired dielectric spacing so that their outlines and circular cut-

This article is based on a report by F. Dittrich of the Mullard Applications Research Laboratory. British Patent Application (15711/59) has been made covering the construction described.



X : 2d + diam of stabilising spacers

(all dimensions in cm)

FIG. 1

outs match but the slots (equal in width to the plate spacing, d) lie on alternate sides of the stack, a multi-plate capacitor shunted by a single-turn inductance is formed. The tank capacitance increases and the inductance decreases with increasing number of plates, therefore the frequency is little changed and Q_L rises. Additional inductive paths in parallel are automatically provided for the increased circulating currents.

To reduce still further the surface resistance, the inside edges (which are the areas of highest current density) of the inductive cut-outs should be bevelled and finished as smoothly as possible.

All parts of the laminated circuit except the areas F/lam (Fig. 1) are unipotential and may therefore be clamped together with metal spacers and bolts, thus obviating the use of expensive insulating materials. If the thickness of the dielectric is small or the capacitive area F/lam large, additional stabilising spacers and some insulators might be required at the capacitive end of the stack to provide the necessary mechanical rigidity (Fig. 2b).

The bolts clamping the stabilising spacers and insulators can also be used to connect the valve into the circuit, as well as to carry two adjustable flaps for final frequency trimming (Fig. 2a and Fig. 7).

Size of the Circuit

The physical size of a laminated circuit is determined by f_o, the operating frequency, the C/L ratio (Q_L) and the proposed working voltage. The graphs in Figs. 3, 4 and 5 were calculated and checked by measure-

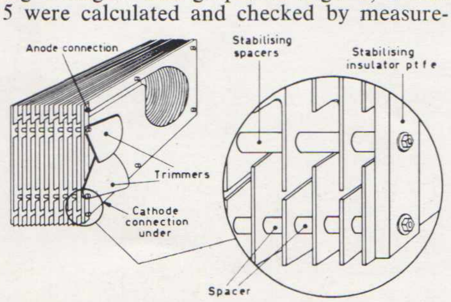


FIG. 2

ments with stacks of experimental laminae of various dimensions, and it is found (Fig. 3) that calculated capacitance graphs are linear and come very close to the measured check points. This confirms that stray effects are small and the parallel plate capacitor formula can be applied to this configuration. Fig. 3 also includes the relation of frequency to a varying number of plates for a given size of F/lam and plate spacing. The frequency scale given is made linear to show more clearly that with a large enough number of plates, say more than ten, the variation of f_o becomes small. Thus the value of f_o might still remain within the acceptable range of design tolerance even when the number of plates is varied slightly. Another set of curves (Fig. 4) shows the change of inductance for a set of plates with varied spacing. It is also shown that with a sufficiently large number of plates, the resulting change of inductance is small when the plate spacing is varied. This set of curves is also useful if a circuit of minimum inductance is required. Because the inductance is also a function of the plate spacing when the number of plates is small, this design facility is only applicable where dielectric spacing is a secondary consideration.

A third set of curves (Fig. 5) shows the required inductive cut-out diameter D for a given inductance with the number of plates and the spacing (d) as parameters.

Assessment of all these curves shows that the laminated circuits will be most practical in the region 10 to 100Mc/s, and with d.c. supplies to the anode of the valve between 500 and 5000V. The overall physical size will be acceptable even at very high C/L values.

DESIGN EXAMPLE

Specification

The design required is an oscillator unit, suitable for diathermy, working at the recommended frequency (27.12Mc/s±0.6%), and capable of delivering more than 200W into a matched load. The anode supply is to be 1.2kV d.c. A Faraday screen should be included between the tank and the coupling coil, to minimise harmonic power transfer.

Valve Conditions

The TY3-250 triode valve appears suitable for the experiment. With the anode supply at 1.2kV, consideration of cathode current will limit the obtainable output. The calculated conditions are

V_a	$= 1.2$kV	η_{load}	$= 53\%$
I_a	$= 375$mA	P_{load}	$= 237$W
P_{in}	$= 450$W	I_g	$= 90$mA
p_a	$= 144$W	R_g	$= 1.6$ kΩ
η	$= 68\%$	P_{drive}	$= 27$W
P_{out}	$= 306$W		

The instantaneous values will then be

$$v_{a(pk)} = 950\text{V and } v_{min} = V_a - v_{a(pk)} = 250\text{V}.$$

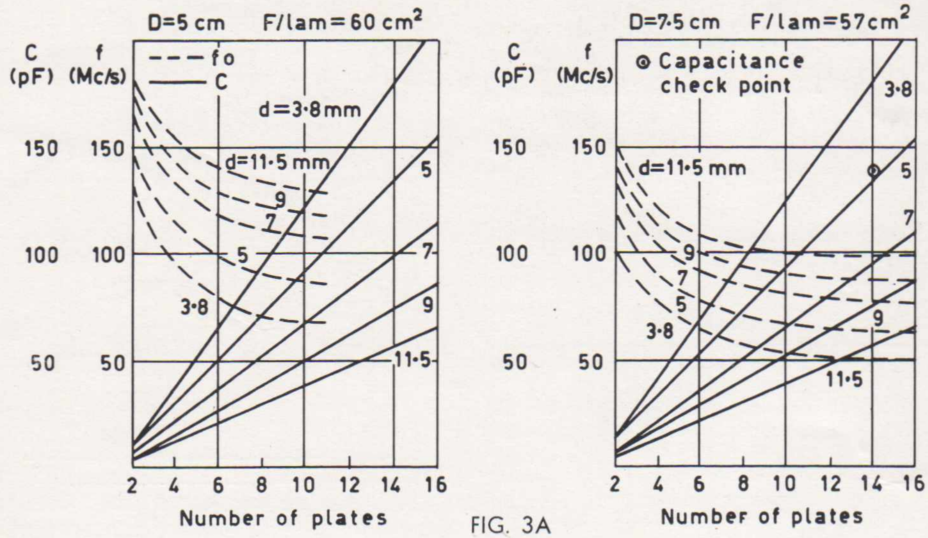


FIG. 3A

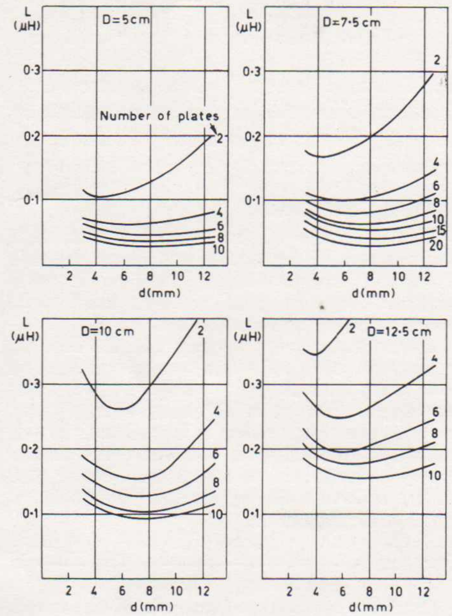


FIG. 4

Tank Circuit Constants

As $Q_L = (V_{(r.m.s.)})^2 / X_L.P_{out}$, and since at resonance $X_L = X_c$, for a tentative Q_L of 150 the reactance must be

$$X_c = \frac{(V_{r.m.s.})^2}{Q_L.P_{out}} = \frac{675^2}{150 \times 306} = 9.95\Omega$$

Hence, for a frequency of 27.12Mc/s

$C = 580pF$ and $L = 0.059\mu H$.

The circulating current will then be

$$I_c = Q_L.I_{r.m.s.}$$

and with

$$I_{r.m.s.} = \frac{P_{out}}{V_{r.m.s.}} = \frac{306}{675} = 0.454A,$$

$$I_c = 150 \times 0.454 = 68A.$$

Dimensions of the Tank Circuit

The material for the laminations should be non-ferrous metal at least 16 s.w.g. (1.5mm) thick, for good mechanical stability. Aluminium was used because of its good conductivity but low density. With this gauge of metal a circulating current of up to 5A per lamina is allowable. If in this example 20 laminae are used, the circulating current will be 3.4A per lamina, which is safe.

For a peak to peak voltage of 1.9kV, a dielectric spacing (d) of 4mm between laminae and for the slot will be sufficient. The nearest curves shown (Fig. 5) are for d equals 3.8mm, but by reference to the 5mm curves and extrapolation to 20 plates a sufficiently accurate value for D may be obtained, especially as for a large number of plates (more than 10) any error will be small. Thus D = 10cm.

If the cut-out is to be situated 1.25cm from three sides of the lamina, the short sides (B) of the rectangular plates will measure 12.5cm.

The total required capacitive area

$$F = C \times 4\pi d$$

where F is in cm², and C and d are in cm.

Dimensionally capacitance can be expressed as length. If one picofarad equals 0.9 cm, 580pF equal 522 cm and

$$F = 522 \times 12.5 \times 0.4$$
$$= 2610cm^2.$$

With 20 laminations there will be 19 dielectric elements, the capacitive area per lamina

$$F/lam = \frac{2610}{19} = 137cm^2.$$

The slot parallel to the length of the stack should be as wide as the dielectric spacing. The length of side G of the area F/lam is given by

$$G = D - 2d = 9.2cm.$$

Side E of F/lam should then be

$$E = \frac{137}{9.2} = 15cm.$$

Because the area of the semicircular cut-out, clearing the stabilising spacers, must be subtracted from F/lam, side E must be lengthened to compensate for this loss of 3.8cm². This loss is already partly made good by the surfaces bounding the slot; these have an area of 2.2cm² so that the remaining area of 1.6cm² can be recovered by slightly lengthening E.

Thus the overall dimensions of one lamina are

A = 26.2cm B = 12.5cm.

If additional adjustable flaps are used (see Fig. 2), these can be made large enough to recover the 1.6cm² referred to above. The main dimensions may then be simplified to, say, A = 26cm B = 12.5cm.

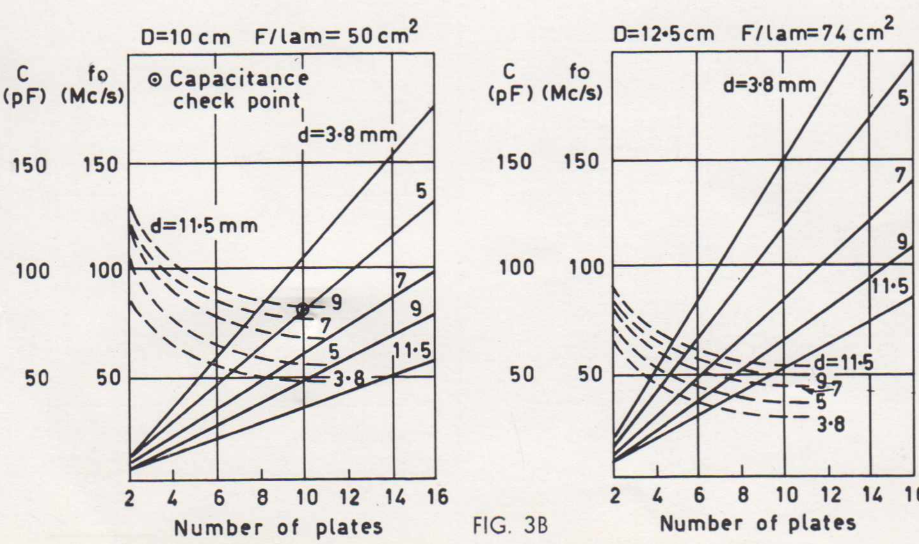


FIG. 3B

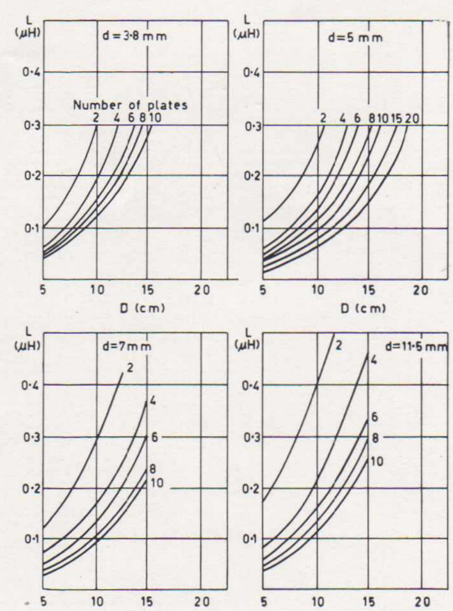


FIG. 5

A few enlarged and detailed copies of Fig. 5 are available on application.

(To be continued)

SIMPLE VALVE MEASUREMENTS

This article is the fourth of a series now being published in Outlook dealing with experiments for the examination of the properties and behaviour of thermionic valves. These experiments include measurements from which the characteristic curves of various types of valves may be plotted.

Mutual Conductance

The influence of the control grid voltage on the anode current is measured by the change in anode current resulting from a small change in grid voltage, *the anode voltage remaining constant.*

From the curves reproduced in Fig. 8, it is seen that with an anode voltage of 200 volts and a grid voltage of –4 volts the anode current is 5.0mA (Point A) and that if the grid voltage is reduced to –2 volts (V_a remaining at 200 volts) the anode current increases to 12.0mA (Point C). Thus, a 2-volt change in V_g results in a 7.0 mA change in I_a.

The ratio $\dfrac{\text{change in anode current}}{\text{change in grid voltage}} = \dfrac{\triangle I_a}{\triangle V_g}$ is termed the *mutual conductance* of the valve. It is denoted by the symbol g_m and is usually expressed in milli-amperes per volt. In the example

$$g_m = \frac{(12—5)\text{ mA}}{(4—2)\text{ V}} = \frac{7\text{ mA}}{2\text{V}} \doteq 3.5\text{ mA/V}$$

the mutual conductance is numerically equal to the slope of the I_a/V_g curve at the operating point. i.e., to tan Θ in Fig. 8.

The mutual conductance, like the internal resistance, is not constant, but owing to the curvature of the valve characteristic, changes with the operating conditions.

Amplification Factor

Referring once more to Fig. 8 it has been seen that, at $V_a = 200$, a change of V_g from –4 volts to –2 volts produces a change in anode current from 5.0mA (Point A) to 12.0mA (Point C). The same change in anode current could have been obtained by keeping V_g constant at –4 volts and increasing V_a to 270 volts (Point D).

Thus, a change of 70 volts in anode voltage has the same effect upon the anode current as a change of 2 volts in grid voltage.

The ratio

$$\frac{\text{change of } V_a}{\text{change of } V_g \text{ which would produce the same change of } I_a}$$

is termed the *amplification factor of the valve,* and is represented by the symbol μ.

In the example,

$$\mu = \frac{270\text{ V}—200\text{ V}}{4\text{ V}—2\text{ V}} = \frac{70\text{ V}}{2\text{ V}} = 35$$

The amplification factor represents the maximum *theoretical* voltage gain obtainable from the valve but, as will be shown later, this theoretical maximum can never be achieved in practice.

Like the internal resistance and mutual conductance, the amplification factor varies with the operating conditions. For this reason, when quoting values for these quantities in published valve data, the conditions under which they are measured are always specified.

Static and Dynamic Characteristics

When making the measurements from which the published characteristics of a valve are plotted, only one quantity is varied. For the I_a/V_g curves the grid voltage is varied and the anode voltage is held constant; and for the I_a/V_a curves the anode voltage is varied and the grid voltage is held constant.

Curves obtained from such measurements are termed *static characteristics.*

In practical applications, however, conditions differ from those in the laboratory where accurate adjustments can be made before taking each reading.

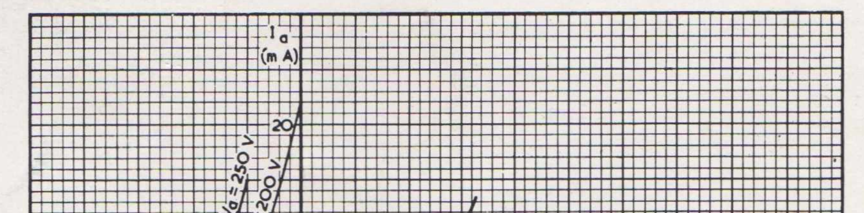
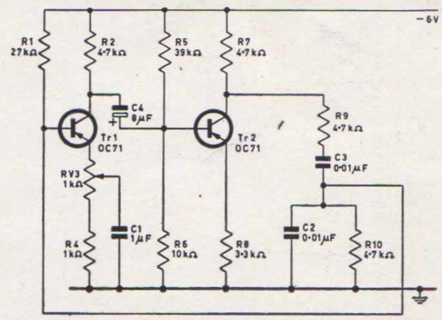


Fig. 8—I_a/V_a and I_a/V_g characteristics of a typical triode, showing the method of calculating the internal resistance, mutual conductance and amplification factor.

AMATEUR EXPERIMENTERS COLUMN

R-C SINUSOIDAL OSCILLATOR



This simple two-stage R-C oscillator will provide an audio output of reasonable waveform at a frequency of approximately 2.5kc/s. Temperature stability is good and supply voltages from –3V to –9V can be used, providing RV3 is suitably adjusted. This control provides a convenient means of adjusting the circuit for optimum performance and compensates for changes of ambient temperature. The audio output may be taken from R7 or R8.

The frequency determining components form a Wien bridge network consisting of C3, R9, C2 and R10. This configuration has been chosen because it produces zero phase shift at the frequency of oscillation f_{osc} whilst the attenuation of three times compares favourably with that of other arrangements. The phase shift across the two transistors is of the order of 360° and may be neglected. Oscillation therefore occurs at the frequency at which the network gives zero phase shift.

CIRCULATION

Please circulate this copy of OUTLOOK and sign after reading.

☐	☐
☐	☐
☐	☐
☐	☐
☐	☐

Return copy to office for filing

Printed in Australia
H. C. DURANT PTY LTD PRINTERS · SYDNEY