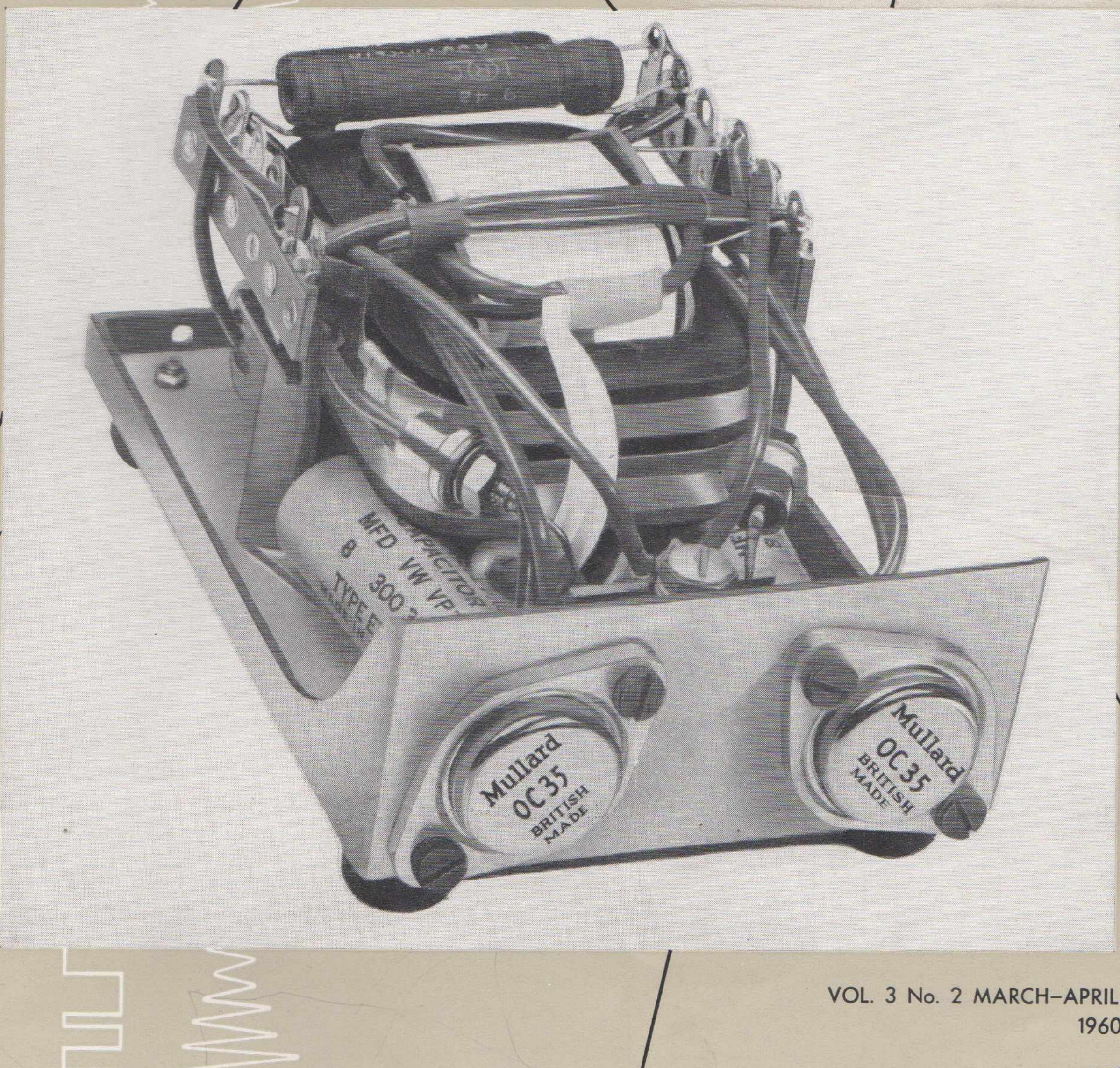
Mullard Outlook

Australian Edition



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MULLARD - AUSTRALIA PTY. LTD.

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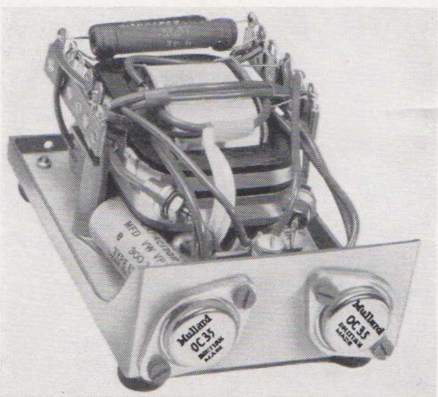

Our front cover shows a Transistorised D.C. Converter
For more detailed information *see page 19.*

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It knocks but once — opportunity, and with it new patterns, new concepts and somewhere specialists, whether creating, manufacturing, selling or servicing.

The history of mankind is one of opportunity, specialization and achievement — sometimes influenced by the highest motives, sometimes sadly misguided — but always opportunity.

Few of us are gifted with the art of thinking in terms of opportunity as the true linguist thinks in the language he is using at the time.

This issue of the "Outlook" dwells on semiconductors and, we trust, beckons to wider practical thinking and application of these devices.

Today's electronic engineers, when designing transistor operated equipment, must think in terms of transistors rather than the classical concept of thermionic valves. The vast selling orbit and public imagination has captured this theme to the extent that a transistor to some lay minds even means a complete receiver rather than a tiny device.

Finally, opportunity — opportunity for some skilled technicians to specialize now in the servicing of transistor radios, for surely here is opportunity indeed.

VIEWPOINT WITH MULLARD

"FROM US TO VIEW"

A NEW MULLARD FILM

"From Us to View" describes the immense skill and care necessary to ensure the high quality of the television pictures we see at home. It does this by tracing the progress of the picture from its beginnings in the studio to its appearance in the living room.

The first part of the film deals with the formation and transmission of the picture. The audience is introduced to the working of a television camera, the complexities of the monitoring equipment and telecine apparatus and to the functions of the "vision mixer" the director and other studio personnel. Then follow sequences showing how the picture is relayed from the studio to the transmitter; and, finally, there are scenes filmed at the Lichfield, England, transmitting station of the I.T.A.

The second part emphasises that no matter what efforts are made in the studio to ensure good quality pictures, the final result depends on the performance and reliability of the television picture tube. Sequences filmed at the Mullard plant at Simonstone, England, take the audience through the many intricate processes involved in the manufacture of Radiant Screen television picture tubes and an insight is given into the extensive research and development programmes that contribute to their long life and high performance.

Produced by the Macqueen Film Organisation, "From Us to View" runs for 23 minutes and is in black and white on 16mm. sound film. The studio sequences were filmed in Associated Rediffusion's studios at Television House, London.

The film is suitable for non-technical audiences and copies are available on loan to film societies, clubs and service organisations, etc., on application.

Conditions of loan are as follows:—

(1) *Mullard-Australia Pty. Ltd., will arrange to forward the films to you so that they will be available for screening at the required time.*

(2) *After screening, the films are to be replaced in their original containers, and details of any damage noticed whilst screening should be indicated.*

(3) *The borrower should forward the films to this Office at his own expense, so that they will be received by the due date. Where films are on loan in country districts or interstate, it is requested that the fastest transport service available be employed.*

(4) *Films are also available from Mullard-Australia Pty. Ltd., 123-129 Victoria Parade, Collingwood, N5, and borrowers are requested to return films to the office of issue.*

NEW FILM STRIPS

The new Film Strip "Thermionic Oscillators" Parts 1 and 2 were mentioned on Page 3 of the last issue of Outlook and Parts 3 and 4 are now completed. The Strips are in the following sequence:—

- Typical Waveforms—Sinusoidal; Saw-tooth; Square Wave; Periodic Pulses
- Basic Principles of the Relaxation Oscillator
- Simple Saw-tooth Generator
- Voltage and Current Waveforms
- Relaxation Oscillator with Thyratron
- Voltages in a Saw-tooth Generator with Thyratron
- Thyratron Saw-tooth Generator with Synchronisation
- Voltages in a Synchronised Thyratron Oscillator
- I_a/V_a Characteristic of a Pentode
- Thyratron Saw-tooth Generator with Pentode as Charge Resistor
- Improvement in the Linearity of a Saw-tooth Waveform
- Aperiodic Capacitive Saw-tooth Generator
- Basic Circuit of the Squegging Oscillator
- Basic Circuit of the Blocking Oscillator
- Cathode-coupled Multivibrator Saw-tooth Generator
- Puckle Hard Valve Generator
- The Transistron Saw-tooth Generator
- Basic Circuit of the Miller-Transitron Generator
- The Basic Multivibrator Circuit
- Anode Current, Anode Voltage and Grid Voltage Excursions of a Multivibrator
- Basic Circuit of a Multivibrator with Varying Pulse Width and Frequency
- Basic Circuit of a Triggered Pulse Generator.

MULLARD-AUSTRALIA PERSONALITIES

Heading the Mullard Applications Laboratory at 35-43 Clarence Street, Sydney, is Mr. John R. Goldthorp who joined the Mullard Company in 1949.

Mr. Goldthorp made a study of Valve and Semiconductor Applications with our parent Company in the United Kingdom, associate companies on the Continent and in the United States during 1954. Upon his return he became more closely concerned with technical liaison services to set-maker customers and the direction of the Applications Laboratory.

A credit diplomate of the Sydney Technical College, Mr. Goldthorp serves on both the Sydney Division Committee and the Publications Board of the Institution of Radio Engineers Australia.

A former N.S.W. championship swimmer and a keen surfer, he now takes his week-end exercise establishing a bush property on a river headwater some 20 miles out of Sydney.

SELF-OSCILLATING MIXER USING AN OC170 TRANSISTOR

The mixer stage described covers the medium- and long-wave bands and includes two short-wave bands giving continuous coverage from approximately 5.8 to 26Mc/s. It uses one transistor for the combined function of mixer and oscillator, so reducing the cost to a minimum.

For the medium- and long-wave bands and the first short-wave band, the fundamental component of the oscillator frequency is used for mixing, but for the higher-frequency short-wave band the second-harmonic component of the first short-wave oscillator is used. This system gives improved frequency stability and reduces frequency pulling caused by the aerial tuned circuit at the higher frequencies.

CIRCUIT DESCRIPTION

D.C. Conditions

The complete circuit diagram of the mixer unit is shown in Fig. 1. The supply voltage is —7V and the emitter current of the transistor is 1mA. Stabilisation of the d.c. working-point is used to control the operating current and so also to control the a.c. performance of the transistor.

Aerial Circuit

The medium- and long-wave aerial coils are mounted on a Ferroxcube rod aerial, while the short-wave coils are contained in separate coil cans. Each aerial coil has a coupling winding for connection to the base of the mixer transistor. The low-potential ends of these coupling coils are returned to the base bias potentiometer. Switching is provided to select the desired coil.

Oscillator Circuits

The oscillator circuits are orthodox and use emitter and collector coupling coils closely coupled to the tuned oscillator coils. A tracking capacitor is included in series with the common medium- and long-wave oscillator coil. This coil is switched to long waves by adding a fixed capacitor in parallel with the oscillator tuning capacitor.

One oscillator coil is used for the two short-wave bands. The fundamental oscillator frequency range of 6.27 to 13.47Mc/s provides the first short-wave band of 5.8 to 13Mc/s and the second-harmonic component of the oscillator frequency gives the high-frequency band of 12.07 to 26.47Mc/s. The design of the oscillator coils is such that they are not heavily loaded by the input and output impedances of the transistor, so giving a high working Q and good frequency stability. At the same time, the coupling is sufficient to ensure operation of the oscillator when the battery falls to half of its nominal voltage.

Choice of Emitter and Base Bypass Capacitors

Both the emitter and base bypass capacitors are chosen to ensure overall stability of the circuit. The emitter bypass capacitances must be sufficiently large to be effective at both aerial and oscillator frequencies, but not large enough to allow any possibility of oscillator squegging. At the same time, this emitter capacitance, in conjunction with the base bypass capacitance, controls the output conductance of the transistor at 470kc/s.

With some combinations of these two capacitances, the mixer output conductance can become negative and either cause oscillation at 470kc/s or impair the stability of the I.F. amplifier. With these considerations in mind, the bypass capacitances have been switched to suit the appropriate frequency ranges, thus ensuring overall stability of the mixer.

I.F. Transformer

An existing type of I.F. transformer has been employed by utilising the whole primary winding (instead of the primary tap) as the collector load. This transformer has the desired characteristics of an undamped dynamic impedance of 150kΩ and a winding ratio of 20:1. A double-tuned transformer having similar properties could be used if an improved I.F. response curve is required.

Aerial Matching

In choosing the number of turns on the aerial coupling windings, conversion gain,

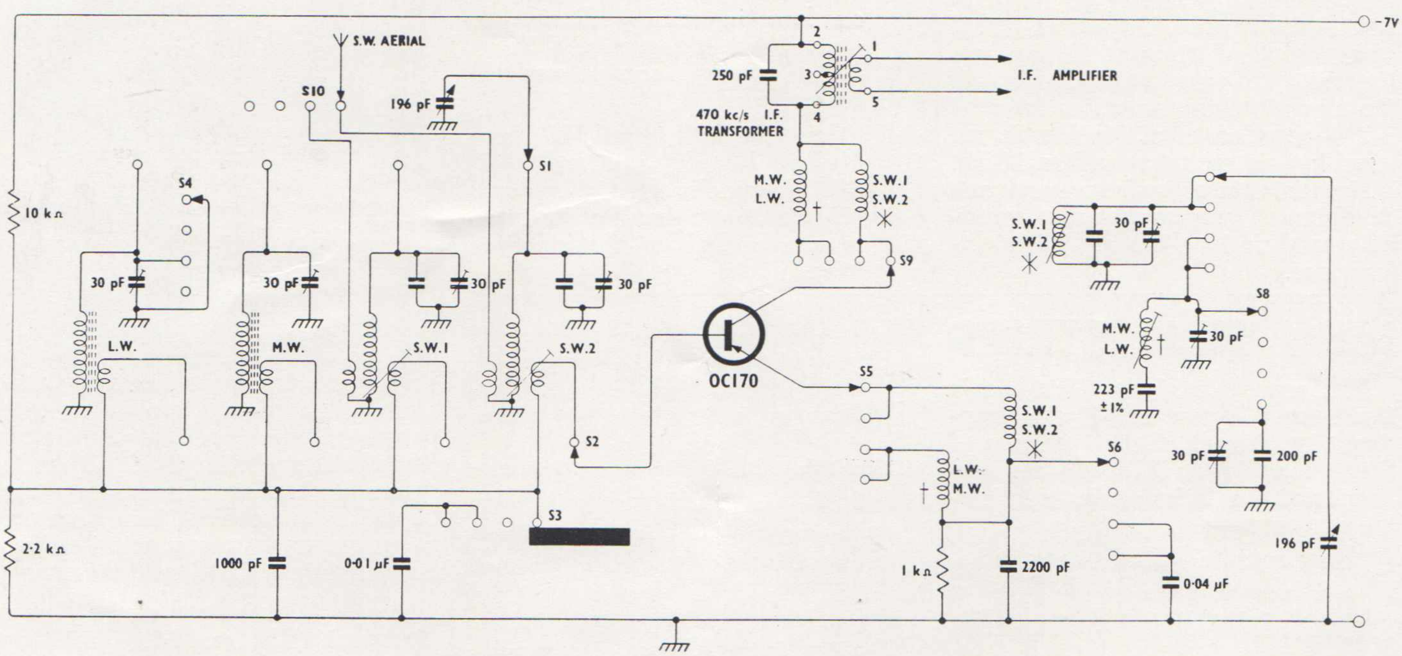

FIG. 1

SWITCHES S1 S2 S3 S4 S10: AERIAL CIRCUIT

SWITCHES S5 S6 S7 S8 S9: OSCILLATOR CIRCUIT

GANG: POLAR 196 + 196 pF SWING

⁎ COUPLED

† COUPLED

MEDIUM- AND LONG-WAVE BAND AERIAL

Mullard Ferroxcube rod, type FX1247

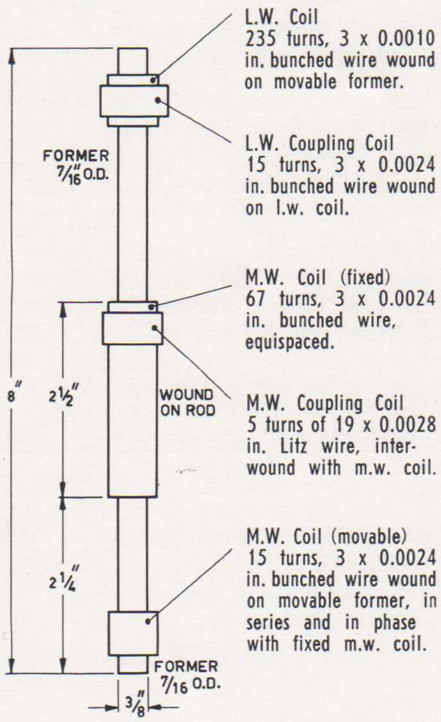

FIG. 2

SHORT-WAVE BAND AERIAL COILS

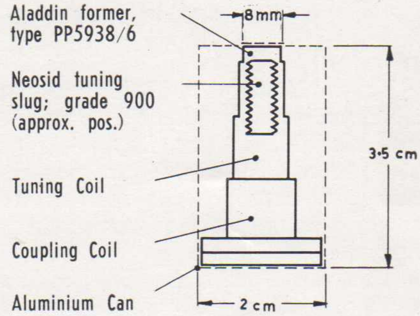

FIG. 3

Short-wave Band 1

Tuning Coil: 28 turns, 24s.w.g., e.s.s. wire closely wound.

Coupling Coil: 4 turns, 36s.w.g., e.s.s. wire closely wound over earthy end of tuning coil.

Short-wave Band 2

Tuning Coil: 12 turns, 20s.w.g., e.s.s. wire closely wound.

Coupling Coil: 1½ turns, 20s.w.g., e.s.s. wire closely wound over earthy end of tuning coil.

SHORT-WAVE BAND OSCILLATOR COIL

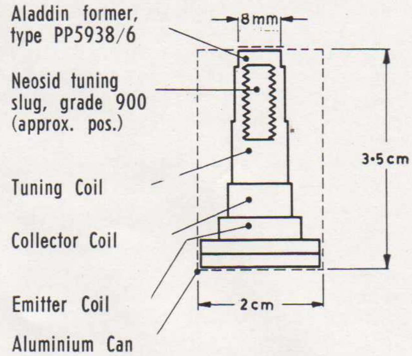

FIG. 4

Tuning coil: 28 turns, 24s.w.g., e.s.s. wire closely wound.

Collector Coil: 9 turns, 24s.w.g., e.s.s. wire closely wound over earthy end of tuning coil.

Emitter Coil: 2 turns, 24s.w.g., e.s.s. wire closely wound over earthy end of collector coil.

spread in gain from transistor to transistor, noise, stability, image rejection, oscillator-frequency pulling by the aerial tuned circuit, oscillator radiation, and aerial circuit bandwidth have to be taken into account. The choice of the number of turns must obviously be a compromise between conflicting requirements.

All the above factors, with the exception of gain (and possibly aerial bandwidth in the medium- and long-wave bands) are improved if the aerial source resistance at the input of the mixer transistor is as low as possible. On the other hand, the greatest average gain would be obtained by matching the aerial to the input resistance of an average transistor at an average frequency within a given wave band.

Non-average transistors operating at the end of a wave band would, in this case, be mismatched, thus causing loss of gain. In particular, a lower limit transistor working at the high-frequency end of a wave band would be operating from a source resistance appreciably greater than its input resistance, so reducing its effective gain and also sacrificing performance in other respects.

The aerial coupling windings have therefore been chosen to present source resistances to the mixer equal to about three-quarters of the input resistance of a nominal transistor at the highest frequency in each wave band, where the gain is lowest. This arrangement of coupling gives near-optimum matching and maximum gain with below-nominal transistors, and maximum mismatch with above-nominal transistors. The performance spread can therefore be reduced to a minimum by appropriate circuit design. The cost of this improvement in spread is a few decibels of gain with the higher-gain transistors, and the accompanying advantages are improved minimum gain, stability, frequency pulling, noise, oscillator radiation and image rejection.

An exception from this practice is made in the long-wave band, where many of the foregoing considerations do not apply. The aerial has been matched to the transistor at 200 kc/s. This allows a higher-Q aerial winding, giving maximum signal pick-up consistent with a reasonably wide working bandwidth.

PERFORMANCE

Frequency Range

Long-wave band:	160 to 280kc/s
Medium-wave band:	540 to 1640kc/s
Short-wave band 1:	5.8 to 13Mc/s
Short-wave band 2:	12.07 to 26.47Mc/s

Conversion Gain

Long-wave band:	30dB at 200kc/s
Medium-wave band:	30dB at 1Mc/s
Short-wave band 1:	24dB at 6 & 12Mc/s
Short-wave band 2:	18dB at 13Mc/s
	16dB at 25Mc/s

Conversion gain is defined as the ratio of the power in the I.F. load resistor (777Ω) to the maximum power available from the aerial circuit.

Image Rejection

Image rejection is the response of the mixer to the image frequency relative to the response at the frequency to which it is tuned.

Medium-wave band:		40dB at 1Mc/s
Short-wave band 1:	23dB at 6Mc/s;	18dB at 12Mc/s
Short-wave band 2:	20dB at 13Mc/s;	16dB at 25Mc/s

Frequency Stability

The maximum change of oscillator frequency occurring in the short-wave bands when the battery voltage is halved is approximately 12kc/s at any point in the frequency range.

Minimum Battery Voltage

The oscillator continues to function in all wave bands until the battery has fallen to less than half its nominal voltage.

Squegging Tolerances

The emitter bypass capacitance can be increased from its nominal value of 2200pF in the short-wave bands to 0.01μF before squegging occurs. This indicates a safety factor against squegging, in these terms, of 4.5. The corresponding factor for the medium- and long-wave bands is 2.

Frequency Pulling

The maximum change in oscillator frequency occurring when the aerial circuit is tuned from its normal resonant frequency to that of the oscillator is about 40kc/s for the lower-frequency short-wave band and 16kc/s for the higher-frequency band.

Sensitivity and Signal-to-Noise Ratio

The following figures apply when the mixer unit is coupled to a conventional I.F. and A.F. amplifier. The sensitivity at the base of the first I.F. transistor is 500μV for 50mW audio output.

For medium- and long-wave operation, the voltage required at the base of the mixer transistor is 5μV (30% modulation) for 50mW output power.

For short-wave operation, the input signals through 400Ω to the aerial terminal at frequencies of 6, 12, 13 and 25Mc/s are 50, 65, 80 and 55μV respectively for an output power of 50mW and 32, 32, 65 and 55μV respectively for a signal-to-noise ratio of 20dB.

COIL SPECIFICATIONS

M.W. and L.W. Aerial Coil (Fig. 2)

Core: Mullard Ferroxcube, type FX1247 (8 x $\frac{3}{8}$ in.)

Q values (aerial mounted on receiver chassis)	Medium wave (at 1Mc/s)	Long wave (at 200kc/s)
Undamped (Q_o)	133	60
Working (Q_w)	approx. 100	approx. 30

M.W. and L.W. Oscillator Coil

Inductance:	173μH
Windings:	
Main:	72 turns
Collector:	4 turns
Emitter:	1 turn
Undamped Q (Q_o):	130 at 1Mc/s
(with 132pF tuning capacitance)	

Short Wave Coils

Coil	Inductance (μH)	Undamped Q (Q_o)	
S.W.1 Aerial* (Fig. 3)	3.05	100 at 6Mc/s	121 at 12Mc/s
S.W.2 Aerial* (Fig. 3)	0.71	117 at 13Mc/s	170 at 26Mc/s
S.W. Oscillator* (Fig. 4)	2.57	120 at 6.27 Mc/s	130 at 13.47 Mc/s

*Suitable coil formers (5000A) and slugs (H 6 x 1 x 12), are available from Neosid Limited, 23-25 Percival Street, Lilyfield, N.S.W.

(Coil wound on former and inside aluminium cans. Tuning slug inserted at end remote from coupling.)

This article has been reproduced from Mullard Leaflet TP399, which was issued at the 1959 National Radio Show in London.

Intending constructors should note that the long-wave band is peculiar to European conditions and may be omitted.

Abridged Data OC170

Limiting Values (absolute ratings)

Collector voltage, grounded base		
$v_{c(pk)}$ max	–20	V
V_c max (d.c.)	–20	V
Collector current		
$i_{c(pk)}$ max	10	mA
I_c max (d.c.)	10	mA
Emitter current		
$i_{e(pk)}$ max	10	mA
I_e max (d.c.)	10	mA
Collector dissipation at $T_{amb} = 45°C$	60	mW
Max junction temperature	75	°C

Characteristics at $T_{amb} = 25°C$

Grounded base		
$I_{c(o)}$ at $V_c = -6V$, $I_e = O$	2.0	μA
$f_{\alpha(av)}$ at $V_c = -6V$, $I_e = 1.0mA$	70	Mc/s
$f_{\alpha(min)}$ at $V_c = -6V$, $I_e = 1.0mA$	40	Mc/s
Grounded emitter		
I_b at $V_c = -6V$, $I_e = 1.0mA$	20	μA
V_b at $V_c = 6V$, $I_e = 1.0mA$	–300	mV
α' at $V_c = -6V$, $I_e = 1.0mA$	80	
Noise figure at $V_c = -6V$, $I_e = 1.0mA$		
$R_s = 500\Omega$, f = 1.0kc/s	25	dB
$R_s = 200\Omega$, f = 450kc/s	4	dB
$R_s = 150\Omega$, f = 10.7Mc/s	5	dB

Max body length 9.5 mm   Max diam 9.1 mm

NEW SEMICONDUCTOR TYPE NOMENCLATURE

To facilitate the identification of individual semiconductor devices, Mullard Limited, London, in co-operation with several other major manufacturers in Great Britain and Europe, have introduced a new semiconductor type nomenclature.

This new system, besides providing a rapid means of identifying the individual type, also classifies the device as to the application for which it is primarily designed and the class of service for which it is intended. It will only be applicable to new types (the older types for the time being retaining their original type numbers) and differentiates between devices for professional and entertainment purposes by using a different form of letters and figures in the code. Semiconductors used in the entertainment field will be indicated by two letters and three figures whilst those in the professional field will be indicated by three letters and two figures.

From the Table shown below it may be seen that the **first letter** in the new key indicates the material (germanium, silicon) and also whether p-n-p or n-p-n construction is used. The **second letter** identifies application and classification of the service for which the device is primarily designed.

It is noteworthy that the letters A, and Y in the semiconductor type nomenclature are identical in their meaning to the same letters used in the Mullard Valve Type Numbering System.

CODING BY MEANS OF LETTERS AND FIGURES

MEANING OF THE FIRST LETTER

A. Germanium diodes and p-n-p transistors.
B. Silicon diodes and p-n-p transistors.
N. Germanium n-p-n transistors.

MEANING OF THE SECOND LETTER

A. Diodes, including diodes used as variable capacitors.
C. Transistors for low frequency applications.
D. Power transistors for low frequency applications.
F. High frequency transistors.
L. High frequency power transistors.
P. Photo semiconductors.
S. Switching transistors.
T. Thyristors, Four-layer diodes, controlled rectifiers.
Y. Power diodes (rectifiers).
Z. Reference and Zener diodes.

FIGURES

For the entertainment semiconductors, figures between 100 and 999 are used. These numbers have no technical meaning, but show the sequence of release.
For the professional section, letter and figure groups are used thus:—

Z10 .... to Z99
A10 .... to A99

which is an obvious reverse of sequence.

SOME TYPICAL EXAMPLES

ASZ15 (the former OC28)	A. Germanium p-n-p Transistor	S Switching Transistor	Z15 Professional Semiconductor Sequence Number
BCZ10	B Silicon	C Low-Frequency Transistor	Z10 Professional Semiconductor Sequence Number
BA100	B Silicon	A Diode	100 Entertainment Semiconductor Sequence Number

TRANSISTORISED INVERTERS AND D.C. CONVERTERS

Power transistors are now finding widespread application to the technique of power switching and indeed supplanting rotary machines and the vibrator cartridge.

Although much has been written on the design of inverter and converter stages using semiconductor devices, and the parameters detailed for the associated transformer, the majority of articles fail to correlate the basic factors or present the design in a form convenient for practical calculation. Since for high power requirements the push-pull saturating transformer type of supply is most suitable — this configuration implying maximum utilisation of both the transistors and the transformer — only this type will be considered in detail.

The basic operating conditions of this stage are so chosen that the saturation of the transformer enables rapid excursion of the collector voltage-current conditions so that the transistors are alternatively either "bottomed" i.e. passing considerable collector current at a low collector to emitter voltage, or "cut off" i.e. passing no current and with a high voltage impressed across the collector to emitter connections. This is illustrated in Figure 1 which depicts an

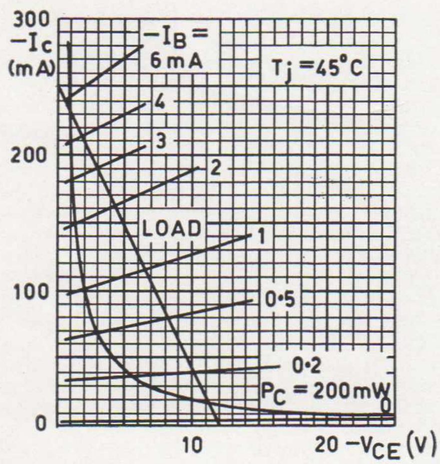

Fig. 1

idealised load line on the collector characteristics from which it may be observed that it is only in the transition or switching state that the transistors operate under conditions of high dissipation. It is obvious, therefore, that the transition period should be as small a portion of the complete cycle as practicable, whilst in the interests of symmetry each side of the push pull stage should contribute equally. The inverter current waveform is, therefore, a square wave of the shortest possible rise and fall times. Obviously if this transition period occupies an appreciable portion of the cycle, not only will the efficiency be less but the dissipation in the transistors will be greater; the latter factor seriously reducing the power handling capabilities of a design using a given pair of transistors.

Another factor which is often overlooked is that the faster the transition time the more likely the transformer to "ring" at a frequency determined by the leakage inductance between the two halves of the primary and the effective shunt capacitance; the magnitude and decay rate of the "ring" being determined by the effective "Q" of the circuit. From the practical point of view, and in cases where it is likely that the inverter/converter may be lightly loaded, it is essential that the leakage inductance between the two halves of the primary be kept to a minimum. This will insure against excessive ringing.

Bifilar winding techniques are virtually essential for the primary winding, and in the interests of tight coupling between this and the feedback or base drive windings, the latter are often bifilar wound as well. It is obvious that the lowest practical leakage inductance will occur when these two windings are bifilar wound on a core which has high permeability and low reluctance. Because of the short magnetic length and absence of gap, a ring core provides the most efficient magnetic circuit and by suitable design a toroid may be wound which produces very short transition time, saturates readily, has good efficiency, and by virtue of the bifilar windings produces but little overshoot. Besides degrading the efficiency of a design, overshoot and ring influence the peak voltage rating of the transistors selected for a specific design, and with a poorly designed transformer it is not uncommon to find that more expensive transistors, with higher collector voltage ratings, must be employed to cope with these transient phenomena which otherwise would be of considerably smaller magnitude.

The two major factors to be determined by the designer are operating frequency and core type and material. The two are interrelated and the requirements are analysed elsewhere. In principle, however, the aim is to keep the switching frequency as high as possible, since this eases the filtering requirements and reduces the cross-sectional area of the core for a given power level, consistent with the increase in losses at high frequencies and in transistor dissipation as a result of the transition time becoming an appreciable portion of the switching cycle. The practical impact of these requirements is that switching frequencies of from 1 to 3 Kc/s are common with ferrite cores and 200 to 1000 c/s for nickel and grain oriented silicon steel strip cores. Conventional power transformer type laminations of the medium resistance or stalloy grades 0.014″ thick are not satisfactory above 100 c/s due to increasing iron losses and "spiking", i.e., overshoot and ringing problems are likely to be severe due to the poor ratio of permeability to reluctance of a practical core using this material. It is true that spiking problems may be eased by the addition of a timing (buffer) capacitor as is common in vibrator power supplies, but this approach invariably leads to lower efficiency and increased dissipation in the transistors due to the increased transition time. Listed hereunder are the salient features of some core types and materials which have been used in practical inverter/converters.

TABLE 1

Saturating Transformer Core Types and Materials.

(a) Spiral Tape Cores (0.002″ tape in Permalloy "C") B sat = 8,000 gauss

Type No.	O.D. (inches)	I.D. (inches)	Width (inches)	Core Area (sq. ins.)	Path Length (inches)	Rating (VA)
WP2950	2.25	1.50	0.625	0.234	5.72	120
WP2951	1.53	1.01	0.438	0.114	4.0	60
WP3120	0.86	0.53	0.275	0.031	2.02	10

(b) "C" Core Loops (0.004″ tape in grain oriented silicon steel) B sat = 16,400 gauss

Type No.	Window Dimensions (inches)	Width (inches)	Core Area (sq. ins.)	Path Length (inches)	Rating (VA)
Z371033	1.125 x 0.438	0.375	0.129	4.26	15
Z371032	0.875 x 0.375	0.375	0.129	3.63	15

Note: Z371033 is more suitable for multiple stacking.

(c) Ferroxcube T.V. "U" Cores (in A2 material) B sat = 3,500 gauss

Type No.	Window Dimensions (inches)	Core Area (sq. ins.)	Path Length (inches)	Rating (VA)
FX1770/1412	1.50 x 1.50	0.263	7.9	30
FX1788/1452	1.455 x 1.406	0.217	7.4	30

Suitable switching transistor types in the Mullard range are listed hereunder, together with their relevant characteristics:—

TABLE 2

Power Transistor Characteristics—Inverter Service.

Type	$-V_{ce\ peak}$	$I_{e\ peak}$ (A)	$I_{c\ peak}$ (A)	V_{knee}	V_{be}	α'	$\Theta_{m\ max.}$ (°C/W)
OC28	60	7.2	6.0	0.5	0.9	55-22	1.5
OC35	48	7.2	6.0	0.5	0.9	70-27	1.5
OC26	32	—	3.5	0.6	0.8	42-25	1.2
OC77	60	0.25	0.25	0.6	0.6	45-15 min	0.3
OC76	32	0.25	0.25	0.6	0.6	45-15 min	0.3

The basic process of design is initially to determine the number of primary turns for a given core so that the core will be driven into saturation at a frequency previously selected for the particular application. This is achieved by substituting in the fundamental formula:—

$$N = \frac{E \times 10^8}{4 \text{ f.A. } B_{sat}}$$

where N = number of turns
f = frequency in cycles per second
A = cross-sectional area of core in square cm.
B_{sat} = saturation flux density in gauss

For convenience of calculation this may be expressed as:—

$$N_p = \frac{E \times 10^8}{26.0 \text{ f.A. } B_{sat}}$$

where N_p = primary magnetising turns
f = frequency in cycles per second
A = cross-sectional area of core in sq. *inches*
B_{sat} = saturation flux density in *Gauss*

It must also be determined that sufficient current is available from the transistors to saturate the core, consistent with the number of turns and the magnetic path length, and this may be verified by the following equation:—

$$H = \frac{4\pi \text{ NI}}{10 l}$$

which for the saturation case becomes:—

$$H_{sat} = \frac{4\pi N_p I_L}{10 l}$$

where H_{sat} = magnetising force for saturation
I_L = the inductive current in Amps.
l = the magnetic path length in cm.

For convenience of calculation this formula may be rearranged to solve for I_L and converted to accommodate the path length in inches and becomes:—

$$I_L = \frac{2 l \ H_{sat}}{N_p}$$

where H_{sat} = magnetising force for saturation in *Oersteds*
l = magnetic path length in *inches*
N_p = primary magnetising turns

With the practical cores detailed in Table 1 this requirement is easily satisfied.

A third factor is that the core cross-section and path length be such that at the frequency selected it is possible to transform the amount of energy required and although this can be verified by reference to the iron loss curves for the material, a more forthright approach is often to measure the core temperature rise and so establish suitable VA ratings for the core. Typical figures are included for the core types and materials given in Table 1.

In practice the design of an inverter/converter for a specific task may be considerably simplified by the adoption of the information set out in the two tables, the choice of the common collector bi-phase switching circuit and an estimate of the feedback factor based on previous experience. The schematic of a converter of the type discussed is shown in Figure 2. Basically it consists of a cross coupled push pull oscillator, the primary winding of the saturating transformer being in the emitter circuit rather than the more obvious collector circuit. The following two advantages accrue from this method of connection:—

1. The collectors of the transistors are now common and hence the transistors may be mounted directly on a heat sink which can be connected directly to one pole of the battery supply i.e. the negative side, with pnp transistors.

2. Within the limiting ratings of the transistors this connection permits an increase in primary input power and hence secondary output power over the common emitter circuit because the base input current for the transistors is additive in the emitter circuit resulting in increased primary current to the transformer for the same primary voltage swing.

Although somewhat modified by diverse requirements in a practical design, a third advantage is that this arrangement is less critical of feedback factor due to emitter follower action tending to limit base input current.

The base divider network applies a small forward bias to ensure satisfactory starting with lower gain transistors at the lowest ambient temperature and together with the ratio of feedback to primary turns (i.e. the feedback factor) should ensure that on overload the unit ceases to oscillate and primary input current drops to a low value. There is only one value of forward bias and feedback factor for a given pair of transistors in a specific design which will meet these requirements without seriously degrading the conversion efficiency and it should be the aim of the designer to strive to achieve this mode of operation as besides providing self protection for the transistors in the event of overload it also results in lower quiescent dissipation in the overload condition. In practice it is found that transistor spreads do not markedly affect performance for a given feedback factor in the common collector configuration whilst variations in Ico′ and Vbe may to a large extent be swamped by suitably proportioning the base divider. Practical values are shown on the schematic in Figure 2 which result in the desired performance with transistor types OC35 and OC28 in converters of 40 to 60W rating. Besides ensuring that the supply is self-protecting and will reliably start under load at the lowest operating ambient temperature, it is necessary to check with an oscilloscope the knee voltage of the transistors by monitoring the emitter to emitter voltage waveform and the base input current waveform at no load and full load. Small adjustments to the designed feedback factor and base voltage divider values will ensure that the transistors operate within their ratings and result in an efficient design.

So far we have considered only the switching of the transistors and the method by which switching is achieved, i.e. a saturating transformer, and in brief the power handling capabilities of a saturating core. Before proceeding with the rectification circuits which can be used in D.C. to D.C. converters a worked example may serve to illustrate the factors so far discussed.

Let us assume that we require a converter to operate from a nominal 12V D.C. source to power equipment requiring 325V D.C. at 125mA. Reference to the table of core sizes and materials indicates the possibility of the Permalloy "C" toroid WP2951, four Z371033 "C" core loops or two ferrite "U" cores. For a piece of communications equipment where space could be a premium the designer's choice would lean towards the smaller toroid or "C" core loops but, due to the difficulty of winding toroids, the commercial solution may well lie with the "C" cores, and for this design four loops of Z371033 have been chosen.

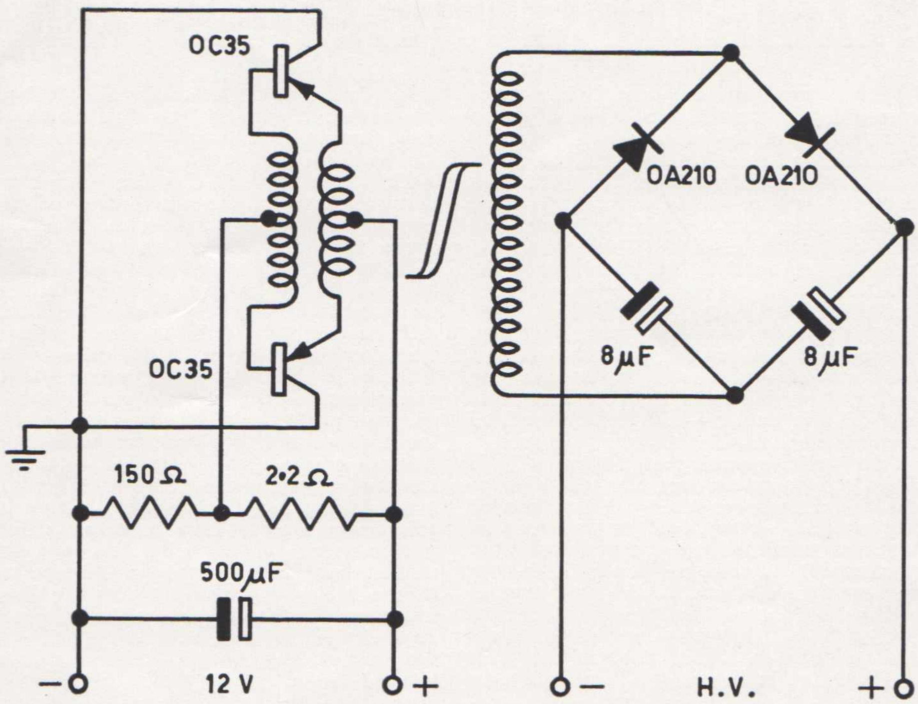

Fig. 2

The primary turns may now be determined by substitution.

$$N_p = \frac{E \times 10^8}{26.0 \text{ f.A. } B_{sat}}$$

Allowing 1 volt drop for V_{knee} of the transistor plus primary copper loss

$$N_p = \frac{11 \times 10^8}{26.0 \times 400 \times 4 \times 0.129 \times 16{,}400}$$
$$= 12.5 \text{ turns}$$

say 12 + 12 turns bifilar wound.

Verifying H:—

$$I_L = \frac{2L\ H_{sat}}{N_p}$$
$$= \frac{2 \times 4.26 \times 1.95}{12}$$
$$= 1.38 \text{ Amps.}$$

i.e. at currents above 1.38 Amps the core will be driven well into saturation.

The feedback turns may now be determined and practical experience indicates a feedback factor of 1.2 to 1.3.

Choosing 1.25 for this design
$N_b = 12 \times 1.25 = 15$ turns bifilar wound.

Wire gauges are selected on the usual window area and copper current density requirements and hence the design becomes:—

Primary = 12 + 12 turns of 18 B&S enamelled copper wire bifilar wound.
Feed Back = 15 + 15 turns of 25 B&S enamelled copper wire bifilar wound over primary.

Six basic rectification circuits are practicable for use with D.C. to D.C. converters. They are:—

(a) The bi-phase or full wave rectifier circuit.
(b) The normal or voltage tapped bridge circuits.
(c) The conventional or full wave voltage doubler.
(d) The cascade or Cockcroft Walton half wave voltage doubler.
(e) The bridge voltage doubler.
(f) Any even integer cascade doubler.

It should be noted that half wave rectifier (i.e. single phase) and odd order voltage multiplication circuits (i.e. treblers, etc.) cannot be used because they load unequally the alternate half cycles of the generated square wave. This would result in uneven load sharing by the transistors, loss of symmetry and, with a converter of this type, loss of efficiency and high dissipation in the power transistors. It is beyond the scope of this article to analyse the performance of the six possible rectifier circuits with square wave input, but it is obvious that the regulation of all circuits will be markedly superior to that experienced with sine wave input, whilst the D.C. output voltage will bear an almost integral relationship to the peak value of the square wave voltage generated by the secondary winding. Experience gained in the construction of a number of converters has shown that the desired D.C. secondary voltage may be computed within a 5% accuracy at nominal load by considering only this integral relationship and computing the desired secondary turns accordingly. An example of this calculation, which is a continuation of the design investigated earlier, is detailed hereunder:—

$$\text{Desired Secondary Voltage} = \frac{V_{out}}{2}$$

as we have elected to use a full wave voltage doubling circuit

$$\text{i.e. } V = \frac{325}{2} = 162.5\text{V}$$

$$\text{Transformer turns per volt} = \frac{12}{11} = 1.09$$

$$\text{Therefore secondary turns} = 162.5 \times 1.09$$
$$= 176 \text{ turns.}$$

Selecting wire gauges as before, the secondary may be wound as 5 layers of 35 turns per layer with 21 B&S enamelled copper wire.

In the interest of increased coupling between secondary and primary it is advisable to interleave the primary with the secondary and the final specification for the transformer becomes:—

Core:—
4 loops of Z371033 "C" cores.

Former:—
$\frac{13}{16}$" x $\frac{7}{8}$" S.R.B.P. tube x $1\frac{1}{8}$" similar pattern 50/8983.

3/5th Secondary:—
3 layers of 21 B&S enamelled copper wire wound 35 turns per layer. Insulation 0.002" Glassine between layers.

Primary:—
1 layer of 12 + 12 turns bifilar wound with 18 B&S enamelled copper wire.
Insulation 0.005 Presspahn over 3/5th secondary.

Feedback:—
1 layer distributed wound of 15 + 15 turns bifilar of 25 B&S enamelled copper wire. Insulation 0.002 Glassine over primary.

2/5th Secondary:—
2 layers of 21 B&S enamelled copper wire wound 35 turns per layer. Insulation 0.005 Presspahn over feedback and two wraps of 0.005 Presspahn over 2/5th secondary to finish winding.

The performance of a practical converter using this transformer and to the schematic of Figure 2 is detailed hereunder:—

D.C. output	325V at 125mA
D.C. Input	12V at 4.15A
Switching Frequency	400 c/s
Efficiency	80.5%
Core Temperature Rise	30°C. approx.
Transistor Junction Temperature for 25°C ambient	55°C
No load input current	1.65A
No load overshoot	13%

It should be noted that where it is desired to have the supply self-protecting, in the event of a high tension short circuit, the bridge or bi-phase rectifier arrangements are to be preferred as other configurations have a somewhat looser coupling between the D.C. output terminals and the secondary winding on the transformer. As a consequence the transistor reflected load is capacitive which usually results in the converter oscillating at a lower frequency when the D.C. output terminals are shorted. This oscillation invariably occurs with the transistors "unbottomed" and hence transistor dissipation is excessive. A fuse in the secondary circuit can, of course, be used to prevent operation under these conditions, whilst the primary circuit should also be fused. A useful refinement to the basic circuit when transistors of diverse characteristics are employed is some means of ensuring load sharing. In units constructed to date it has not been found necessary to insert series or shunt base resistors to ensure this but load sharing should be checked in every practical unit constructed. Similarly the heat sink design for the transistors should be checked by monitoring the mounting base temperature and attention is drawn to the relevant article in the bibliography.

It is hoped that this article will stimulate interest in the practical adoption of transistorised inverter/converters by more forthright design procedures. The very motive renders simplification imperative and, in certain instances, remarks tend to be qualitative rather than quantitative. Those designers who prefer a more rigorous and analytical approach to the design of a variety of inverter/converters are referred to the bibliography.

J. R. GOLDTHORP

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CAUSES OF LOW OUTPUTS FROM AUDIO OUTPUT STAGES

The power output figures given in the Mullard Technical Handbook valve data represent the power available at the valve with the values of voltages, current and external anode resistance quoted. Consequently, the values of output power actually obtained in practical equipment are often lower than those which seem, from the data, to be available.

SPEECH-OR-MUSIC OR CONTINUOUS SINE-WAVE SIGNALS

The values of output power quoted in the handbook are usually given for fixed bias and screen-grid voltages because these closely represent the values actually available for speech or music reproduction. Where a cathode-bias resistor and/or a series screen-grid resistor are used, measurements with a continuous sine wave will show lower output powers than those obtained with fixed voltages. At full drive, the screen-grid current will be appreciably higher than without the signal. Therefore, if the signal is a sustained sine wave, the valve operating conditions will readjust themselves to an increased bias voltage and/or a reduced screen-grid voltage. During the reproduction of speech or music, the waveforms are complex and the sine waves are never sustained at full-drive amplitudes for long enough to affect the valve operating conditions.

The output power measured with a sustained sine wave under cathode-bias conditions is approximately 10% less than that measured with a fixed bias voltage. A simple correction allowing for the effect of a screen-grid resistor cannot be given—it depends both on the value of the resistor and on the ratio of screen-grid current at zero signal to that at full drive.

CONFUSION BETWEEN H.T. LINE VOLTAGE AND ANODE VOLTAGE

The voltages quoted in the handbook are given with respect to the cathode, and should not be confused with the voltage between the h.t. line and the chassis. Usually, the actual anode voltage will be the h.t. line voltage less the voltages dropped across the primary winding of the output transformer and the cathode resistor.

INCORRECT LOAD MATCHING

The data in the handbook gives an optimum value of effective external anode load resistance, and the output power quoted is for this optimum value. At all other values of resistance, the output power will be lower. For single-valve operation, if the effective anode load resistance R_a is greater than the optimum value $R_{a(opt)}$, the anode *voltage* swing at a given distortion is almost independent of the value of R_a. For a resistance less than the optimum, the anode *current* swing is roughly independent of R_a.

Therefore, for $R_a > R_{a(opt)}$:

$$P_{out} \simeq \frac{R_{a(opt)}}{R_a} \times P_{out(opt)},$$

and for $R_a < R_{a(opt)}$:

$$P_{out} \simeq \frac{R_a}{R_{a(opt)}} \times P_{out(opt)},$$

where P_{out} and $P_{out(opt)}$ are the values of output power corresponding to R_a and $R_{a(opt)}$ respectively. One of the most common causes of mismatching is that the resistance of the primary and secondary windings of the output transformer and of the leads to the loudspeaker have been neglected. The effect of the resistance of the secondary winding and the speaker leads is to increase the secondary load resistance. The effect of the primary resistance R_p is twofold: it increases the external anode load resistance, and it influences the valve operating conditions in that it lowers the anode voltage and, hence, the optimum anode load resistance. The effective external anode load resistance R_a is given by:

$$R_a = R_p + n^2(R_s + R_L),$$

where n is the transformer turns ratio, R_s is the resistance of the secondary winding and R_L is the secondary load resistance (including the resistance of the leads). Corrections to the optimum value of anode resistance can be made if it is assumed that the optimum value is roughly proportional to the anode voltage and the reciprocal of the anode current.

ILLUSTRATIVE EXAMPLE

Some time ago, it was found that, in an amplifier which incorporated a single-ended 6BQ5/EL84 audio output stage, the anode current was low with many samples of the valve; and the output power delivered to a 7.5Ω secondary load was only 2W instead of the 4.2W indicated in the handbook. The h.t. line voltage in the amplifier was 250V, and the current in the output stage was 36mA. The relevant data, abstracted from the handbook, are given in the table below.

It was found that a cathode resistance of 210Ω was used in the output stage of the amplifier. The actual screen-grid voltage (with reference to the cathode) was therefore only about 242V, which explained why the anode current was often low.

However, the loss in power resulted mainly from mismatching and the D.C. resistance of the windings of the output transformer. Measurements showed that the turns ratio of the transformer was 30.5 : 1, which transforms 7.5Ω connected to the secondary winding into 7kΩ across the primary. However, the primary resistance was 700Ω and the resistance of the secondary winding was 0.9Ω.

The current of 36mA through the primary winding caused a voltage drop of 25V, so that the actual anode-to-cathode voltage was only 217V. At this voltage, the optimum anode resistance for a 6BQ5/EL84 is approximately (217/250)×7, or 6.1kΩ, and at this optimum value, the output power would be (217/250)×4.2, or 3.65W. However, the transformer, with its winding resistances and a secondary load of 7.5Ω, presented to the valve an effective anode resistance given by:

$$R_a = 700 + (30.5)^2(7.5+0.9)\Omega.$$

That is, the effective anode resistance in the amplifier was 8.5kΩ. The output power available from the valve at this optimum value is approximately (6.1/8.5)×3.65, or 2.63W. There is, however, a loss of power of 0.47W in the resistances of the primary and secondary windings, so that the useful power delivered to the load is about 2.2W instead of the expected 4.2W.

Because the calculation of output power at an anode resistance different from the optimum is only approximate, and also because the transformer resistances were measured on a cold transformer, this value of 2.2W is in reasonable agreement with the output of 2W obtained with the amplifier. A small reduction in the cathode resistance, and the use of a different, though somewhat larger, output transformer (R_p=305Ω, R_s=0.2Ω, n=28.3) resulted in an increase in output power to 3.5W delivered to a secondary load of 7.5Ω.

Table of Relevant Data
6BQ5/EL84

V_a	= 250	V
V_{g2}	= 250	V
R_a	= 7	kΩ
R_k	= 210	Ω
V_{g1}	= −8.4	V
I_a	= 36	mA
$P_{out}(D_{tot}=10\%)$	= 4.2	W

This article is based on a report prepared by A. J. Heins, Head of the Mullard Applications Research Laboratory.

PRACTICAL NOTES ON VALVE MEASUREMENTS

This article is the third of a series now being published in the "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.

TRIODE

If a source of direct voltage is connected between the anode and cathode of a triode, the anode being positive (see Fig. 4) the value of the anode current is governed by two factors —the anode voltage (V_a) and the grid voltage (V_g).

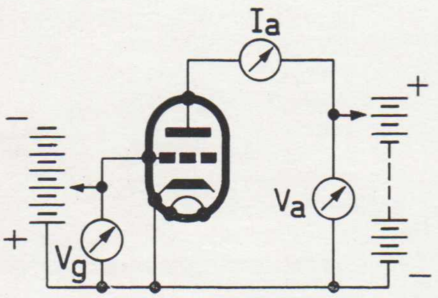


Fig. 4—Basic connections of a triode.

I_a/V_a Characteristics

If the grid is maintained at cathode potential ($V_g=0$) the curve relating anode current to anode voltage is of the form shown in full line in Fig. 5. This graph is similar to that for a diode (see Fig. 2, Vol. 2, No. 6 Outlook), and shows the saturation effect.

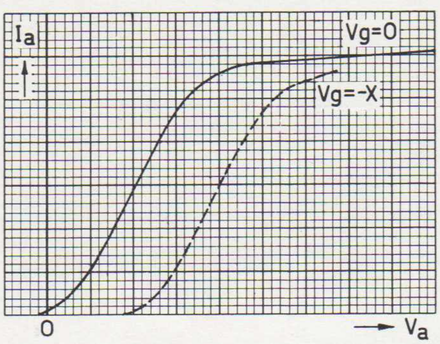


Fig. 5—General form of I_a/V_a characteristics for a triode at $V_g=0$ (full line) and $V_g=X$ (dotted line).

If, now, the grid is given a small negative potential and the measurements are repeated, the curve, although of similar form, is displaced towards the right as indicated by the broken line. This is because, with a negative potential applied to the grid, the anode must attain a certain minimum positive potential before its attractive force can overcome the repulsive force exercised by the grid on the electrons emitted by the cathode.

The higher the negative voltage applied to the grid, the further is the I_a/V_a curve displaced to the right. Fig. 6 shows a complete family of I_a/V_a curves for a typical triode at various negative values of grid voltage.

Note.—These curves are not taken into the saturation region since in normal applications valves are not operated over this part of their characteristic.

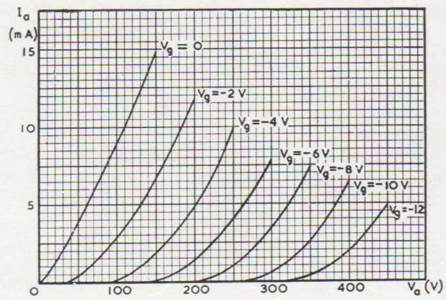


Fig. 6—Family of I_a/V_a curves for a typical triode taken at various values of V_g.

I_a/V_g Characteristics

A second type of curve can be plotted, to show the variation of anode current with variation of grid voltage, the anode voltage being maintained constant.

Fig. 7 shows a family of such curves for different values of anode voltage.

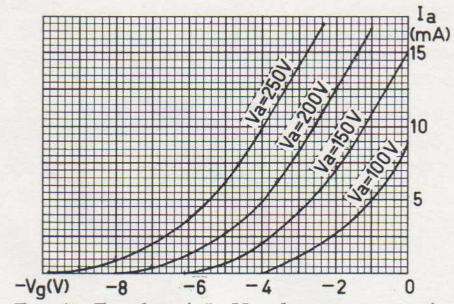


Fig. 7—Family of I_a/V_g characteristics of a typical triode taken at various values of V_a.

Internal Resistance of a Triode

The internal resistance of a triode at a particular point on its characteristic may be calculated either from the I_a/V_a curves or from the I_a/V_g curves.

The I_a/V_a and I_a/V_g curves for a typical triode, with a common I_a axis are shown in Fig. 8. Assuming a grid potential of –4 volts, and an anode voltage of 200 volts, the anode current will be 5mA (Point A). Increase of anode voltage to 250 volts causes the anode current to increase to 10mA (Point B).

The internal resistance is thus:

$$r_a = \frac{250\text{V}-200\text{V}}{10\text{mA}-5\text{mA}} = \frac{50\text{V}}{5\text{mA}} = 10{,}000 \text{ ohms}$$

It should be noted that, because of the curved form of the graph, the internal resistance is not constant, but depends upon the "operating point", i.e., upon V_a and V_g.

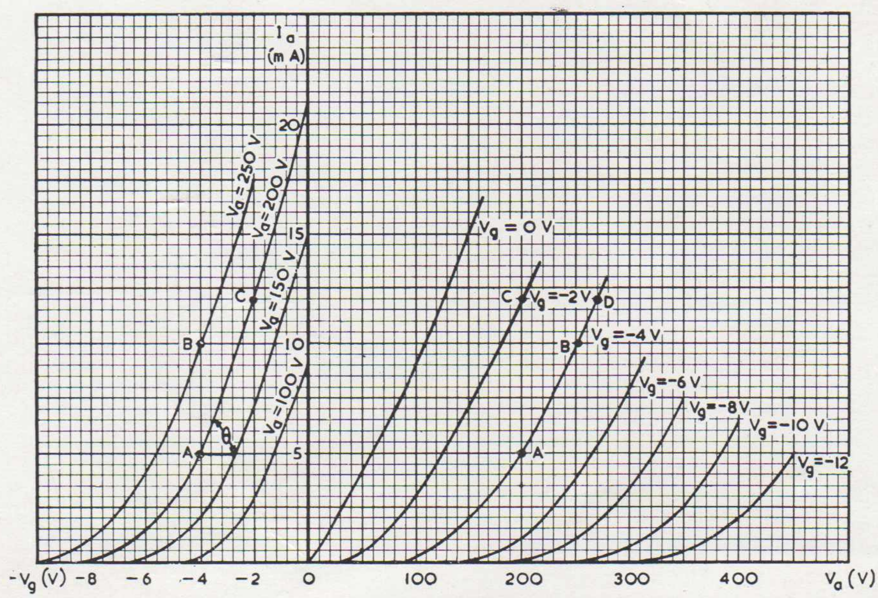
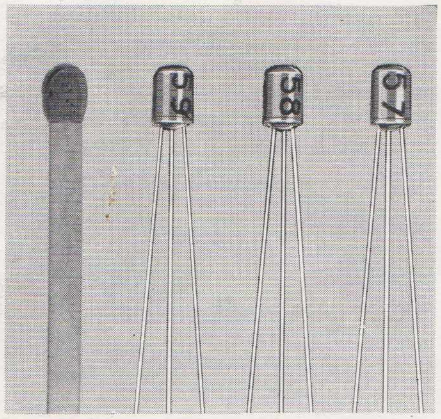


Fig. 8—I_a/V_a and I_a/V_g characteristics of a typical triode, showing the method of calculating the internal resistance, mutual conductance and amplification factor.

DIRECTLY COUPLED TRANSISTOR HEARING AID

The circuit shown in the diagram uses three subminiature germanium transistors, types OC57, OC58 and OC59, in cascaded common-emitter stages. The total current drain from the battery is 2.8mA (at 25°C), of which 2.2mA are used in the output stage to obtain the required output power. Each transistor in the first two stages consumes 0.3mA, thus giving the minimum total battery drain consistent with reasonable gain. The current in the output stage is adjusted to the correct value by selection of the base bias resistor of the first stage.



Stabilisation of the operating currents is achieved by the use of D.C. negative feedback. A.C. feedback is partially suppressed by the single capacitor in the feedback loop.

Variations of the operating conditions with changes of temperature occur because of changes in the transistor parameters I_{co}, α' and V_{be}. Although the effect of these changes is minimised by D.C. feedback, both gain and maximum output power tend to fall at the extremes of the temperature range. The performance is acceptable however over a wide range of temperatures.

Two A.C. feedback paths of importance exist in the circuit. The first is by way of the D.C. feedback loop and gives negative feedback. The magnitude of this feedback is approximately constant with frequency because of the combined effect of the inductive load and the capacitive bypass. The choice of the transistor with the highest value of α' for the first stage allows the use of a high value of base bias resistor, thus permitting the use of a relatively low value of decoupling capacitor. The value of this capacitor is chosen so that the loss of gain resulting from feedback is limited to 6dB.

The second A.C. feedback path, via the impedance of the battery, gives rise to positive feedback into the base of the second transistor. This feedback is minimised by using the transistor having the lowest value of α' in the second stage. Measurements showed that, with high-gain transistors in each stage, the cell impedance could be artificially increased to 4Ω before oscillation occurred.

PERFORMANCE

Gain.

A microphone e.m.f. of 200μV produces an output of 500mV across a 600Ω earpiece. With a conventional microphone and earpiece, this corresponds to an air-to-air gain of about 48dB.

Power Output.

The maximum r.m.s. output is 500mV across a 600Ω earpiece. With a conventional earpiece, this corresponds to a pressure 46dB above 1dyne/cm².

Temperature Range.

The loss of gain at the extremes of the temperature range 32 to 100°F is about 6dB and the corresponding reduction of maximum power output is 2.5dB.

Battery Life.

The total current consumption at 77°F (25°C) is 2.8mA at 1.3V, giving a life of about 90 hours from a Mallory mercury cell, type RM625.

COMPONENTS

Transistors: OC57, OC58, OC59.

Earpiece characteristics:
- D.C. resistance = 175Ω
- Impedance at 1kc/s = 600Ω
- Input voltage at 1kc/s for output 40dB above 1 dyne/cm² = 250mV

Microphone specification:
- D.C. resistance = 1kΩ
- Impedance at 1kc/s = 2.5kΩ
- Output at 1kc/s = 72dB below 1V per dyne/cm²

Capacitor: 6μF, 1.5V, d.c.(wkg.)

Resistors:
- 2 × 3.9kΩ, 0.1W
- 1 pre-selected (see below) 0.1W

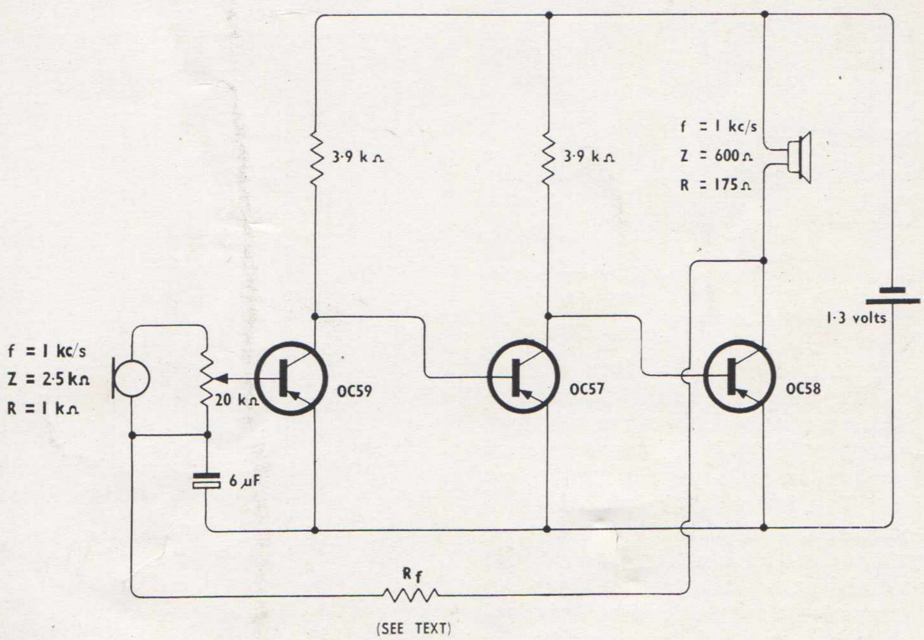
Volume control:
- 20kΩ semi-logarithmic carbon potentiometer with single-pole switch

Cell:
- 1.3V mercury type (Mallory RM625)

Setting-up Procedure.

The resistor R_f may be replaced by a 500kΩ potentiometer which should be adjusted to set the OC58 collector current to 2.2mA at an ambient temperature of 25°C.

The resistance in the circuit may then be measured and replaced with a fixed resistor. For other temperatures the value of 2.2mA is increased or decreased for higher or lower temperatures respectively by 66.67μA/°C.



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