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CONTENTS

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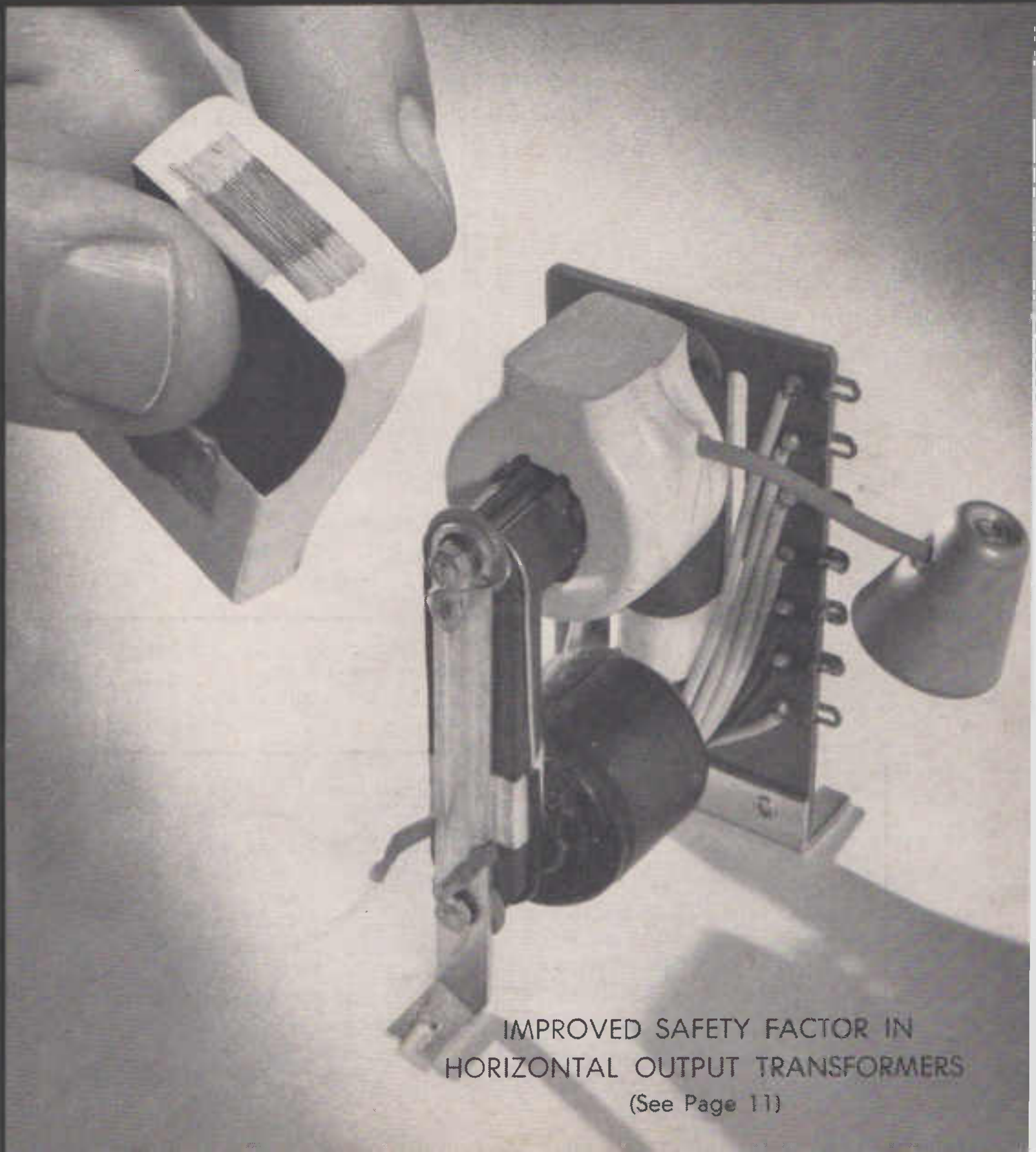


IMPROVED SAFETY FACTOR IN HORIZONTAL OUTPUT TRANSFORMERS

(See Page 11)

The "Miniwatt" Digest

to be published monthly — is offered as a concise source of technical and commercial information to those concerned with electronics in its many phases.

It will feature articles and comment of topical interest, besides providing general and application data on the extensive range of electronic components and materials manufactured and marketed by the Miniwatt Electronics Division of Philips Electrical Industries Pty. Limited. In the preparation of this information specific regard will be paid to the influence of factors peculiar to this country.

As time passes, we are confident that the Miniwatt Digest will be appreciated as a ready reference source of information, proving of value to our regular and prospective customers by limiting time spent in search and enquiry.

— *The Editor*

FUNDAMENTAL FACTS

Although power rectifiers are basic elements of most circuits, even an experienced user may misinterpret the published data. For this reason, the more important maximum ratings are discussed together with their application in circuits. In order to ensure operation within the ratings in all circumstances it could be necessary to choose a rectifier which at first sight may appear to be under-run.

A quick reference guide is given below to aid the selection of suitable silicon rectifiers and circuits for most moderate power applications.

The special case of series connection of silicon diodes is examined, showing that a diode may appear to be operated within its maximum ratings when this may not be the case. Some protective measures are outlined.

SILICON RECTIFIER SELECTION GUIDE

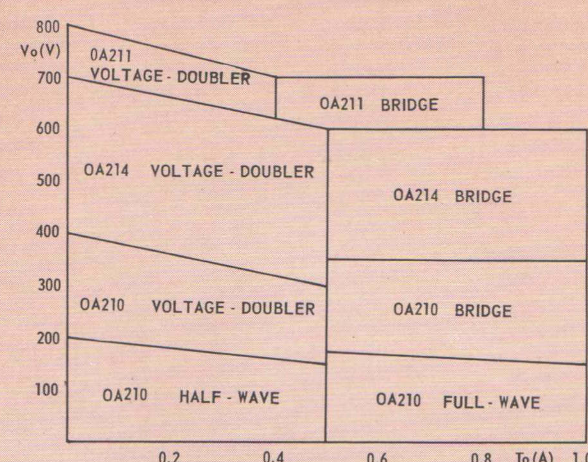


Table 1

Circuit	I_D / I_o	V_o (no load) / V_i	V_o (no load) / $-V_{DM}$	$V_{im} / -V_{DM}$	$-V_{DM} / V_i$
Half-wave	1.0	1.41	0.5	0.5	2.82
Full-wave	0.5	1.41	0.5	0.5	2.82
Voltage-doubler . .	1.0	2.82	1.0	0.5	2.82
Bridge	0.5	1.41	1.0	1.0	1.41

Table 1 lists the relationships between the various voltages and currents. The symbols used are listed below and conform with those in the published data:—

V_i = RMS value of AC input voltage (Refer Fig. 1)
V_{im} = Peak value of AC input voltage
$-V_{DM}$ = Peak inverse voltage
V_o = DC output voltage
I_o = DC load current
I_D = Forward Diode Current
I_{DM} = Peak Forward Current

ABOUT RECTIFIER RATINGS

— with particular reference to Silicon Diode Circuits

The design theory for vacuum tube and mercury vapour rectifiers has been thoroughly dealt with in textbooks[1, 2] and applies equally well to semiconductor diodes[3]. This theory will not be discussed here in detail, but the emphasis will be placed on correct application in practice.

Although rectifiers are sometimes used with resistive and inductive loads, the capacitive load is most common in domestic and light industrial electronic equipment where reasonable regulation is required. The four most commonly used rectifier circuits with capacitive loads (Fig. 1) will be discussed. Voltage regulation curves are included for each circuit.

Rectifier Ratings

It is the responsibility of the manufacturer to determine and publish the maximum ratings of a rectifier, and the user must take the risk of damage if he exceeds any one of these ratings. Where these are absolute maximum values, allowance must be made for mains voltage fluctuations and component tolerances so that no rating is exceeded under any circumstances. The most important ratings will be dealt with individually.

Peak Inverse Voltage ($-V_{DM}$)

This is the maximum permissible voltage which may be applied in the reverse direction across a non-conducting diode. Referring to the basic half-wave circuit of Fig. 1a, it will be seen that on each positive-going half-cycle of the input voltage, the diode will conduct and capacitor C will be charged to the peak value of the applied voltage. During the following negative half-cycle the capacitor starts to discharge at a rate determined by the value of the capacitor and the load resistance. At the same time the anode of the diode approaches a negative voltage equal to the peak value of the applied voltage, while the cathode is positive by the amount of the charge on C. The total voltage across the diode is thus the sum of these two voltages. When no load is drawn the capacitor is charged to the peak of the input voltage, so that twice this voltage appears across the diode on the negative half-cycle. It can thus be seen that the maximum allowable DC output voltage is equal to *half* the peak inverse voltage (PIV) rating. The same reasoning applies to the circuit of Fig. 1b.

In the case of the voltage doubler circuit (Fig. 1c), for a given input, the output is twice that obtained from the half-wave and full-wave circuits. Thus a no-load DC output equal to the PIV rating is allowable.

In the bridge rectifier circuit (Fig. 1d) the only voltage appearing across each of the non-conducting diodes is the DC voltage across the capacitor, so that the available no-load DC output is equal to the PIV rating.

Due to the voltage drops in the rectifier and current-limiting resistance, the voltage on load is always less than the above values. Nevertheless, it is necessary for the designer to assume that an inverse voltage equal to that on no-load may be reached. A common instance of this case occurs in a domestic TV receiver using a number of valves supplied from a silicon diode rectifier. No load current is drawn until the valves have warmed up, but the full DC supply voltage is available immediately, as the diodes have no warm-up time.

In some cases there is published a transient or surge peak inverse voltage rating, in excess of the normal recurrent rating, which may be applied for a short duration. In the absence of such a rating, care must be taken that the recurrent rating is not exceeded when transients or surges are superimposed on the normal voltage.

Forward Current (I_D)

This is the rating which determines the maximum permissible load current. For half-wave and voltage-doubler circuits the average diode forward current is equal to the load current. In full-wave and bridge circuits the average diode current is half the load current, so that here a total load of twice the diode rating may be drawn.

Peak Forward or Recurrent Peak Current (I_{DM})

At the peak of each conduction half-cycle the forward current reaches a maximum value determined by the load current, the size of the reservoir capacitor, and the current-limiting resistor. To ensure that this current does not exceed the permissible rating, it is usually necessary to estimate its value from design curves [2]. In the case of the OA 210 series of diodes a maximum value for the capacitor is specified, and provided that all other ratings are observed, the peak current will not be excessive.

The minimum possible value of peak current is 3.14 times the average diode current. This value applies with a resistive or inductive load, or with a capacitive load when the reservoir capacitor is very small. Values of about 7 to 10 times the average diode current at full load are typical in the circuits of Fig. 1.

Surge Peak Current ($I_{D\ surge\ max}$)

Before the power supply is first switched on there is no charge on the reservoir capacitor. Immediately the switch is closed, the diode is virtually working into a short circuit as it attempts to rapidly charge the capacitor. If the switch is closed at the peak of the conduction half-cycle a very heavy surge current is drawn. In order to limit this surge current to a safe value there must be sufficient resistance in series with the diode. If the minimum value of this resistance is not specified in the published data it may be calculated from the ratio $V_{im}/I_{D\ surge\ max}$, where V_{im} is the peak value of AC input voltage and $I_{D\ surge\ max}$ is the surge peak current.

When a transformer is used, this resistance will be provided wholly or in part by the source resistance of the transformer, in which case the following formula applies:

$$R_t = R_s + n^2R_p + R_1$$

where the symbols have the significance shown on the circuit diagrams.

If an auto-transformer is used, the fact that a part of the transformer winding is short-circuited by the mains should be taken into account when calculating R_1. Note that the minimum value of R_t as published for the OA 210 series of diodes is sufficient when the diodes are used in half-wave, full-wave or voltage-doubler circuits, but in the bridge circuits V_{im} may be twice as high as in the former circuits so that R_t must be doubled. This also applies in the case of diodes in series and *as a general rule* R_t should be increased in proportion to the number of diodes in series.

Maximum Ambient Temperature (T_{amb})

For semiconductor diodes all of the above-mentioned ratings are temperature dependent and such diodes must not be used in an ambient temperature above that at which the ratings apply. In some cases a derating curve is published for operation at higher temperatures.

A heat sink must be used when so specified in the published data. For example, the ratings of the OA 211 and OA 214 apply with a minimum heat sink area of 5 sq. cm:

Selection of a Rectifier

Normally, a DC power supply is specified in terms of load current and voltage, and the designer has to select a suitable circuit, components and the applied voltage to meet the specification.

Compared with valves, silicon diodes have the advantages of reduced size, longer life, lower losses, less heat dissipation and greater power output for an equivalent space. In addition, the elimination of heater voltage supplies is a major economy, particularly in bridge rectifier circuits where a separate insulated winding is required for each valve heater.

Quick selection of silicon diodes and circuits for DC voltages below 800 V and currents below 1.0 A may be made by using the ready reference chart on page 2. The configurations recommended are the most economical for the particular conditions, but a circuit recommended for high voltages and high currents may be used at lower voltages and lower currents even though there may be several less costly circuits.

Having selected the required circuit and diode, reference to the appropriate curves of Fig. 1 will give the RMS input voltage required for a given output. Note that these curves are only valid for 50 c/s mains supply and that in each case the minimum value of current-limiting resistance has been assumed. Higher values of resistance will result in poorer regulation, but a guide to the order of input voltage may still be obtained from the curves.

Values of reservoir capacitance other than those recommended may be used, provided that a capacitor at the upper limit of its tolerance does not exceed the maximum permissible value. Manufacturing tolerances of up to +100% are common for large electrolytic capacitors.

Higher output currents may be obtained by connecting diodes in parallel provided that individual limiting resistors are used in each parallel branch to ensure equal current sharing.

If higher output voltages are required it may be necessary to connect two or more diodes in series and this application is discussed below.

SERIES OPERATION OF SILICON DIODES

Before silicon diodes can safely be used in series, the designer must recognise some additional problems which arise from a large spread in some inverse characteristics, low thermal capacity and small junction capacitance of these devices.

Naturally, all the normal precautions in regard to voltages and currents, including transients, apply equally to series-connected diodes.

Forward Characteristics

The same forward current flows through all diodes of a series-connected chain and the spread in forward characteristics of individual units has very little influence on the operation of the chain as a whole as long as the maximum current ratings are observed.

There can be individual differences in forward dissipation and effective cooling, leading to differences in junction temperatures. It is usually inconvenient to employ a common heat sink for temperature equalisation of all units in a chain. Although the variation of junction temperatures is only of minor importance for the conducting condition of the diodes, it may become significant during the period of non-conduction as the inverse characteristics are strongly temperature dependent.

SILICON DIODE RECTIFIER CIRCUITS WITH CAPACITIVE LOADS
Operating from Single-Phase 50 c/s Mains Supply

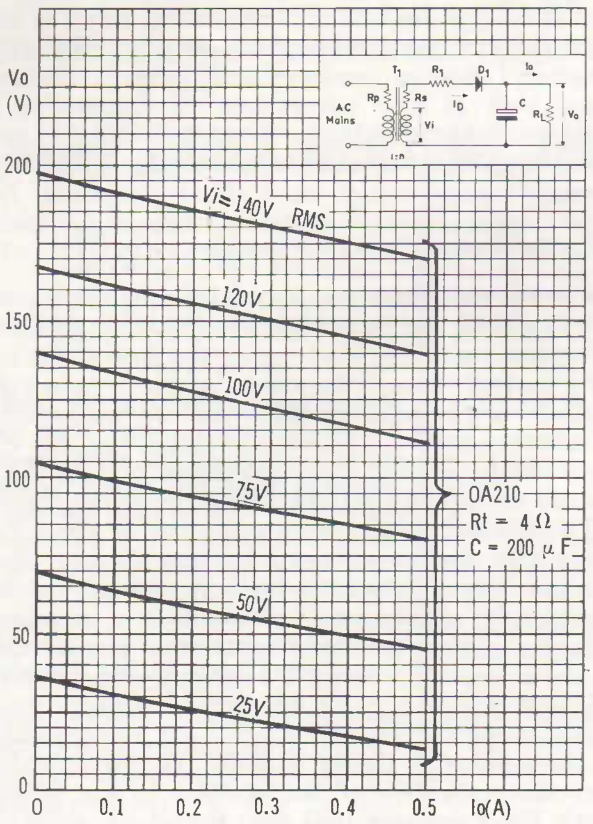


Fig. 1a. Half-wave.

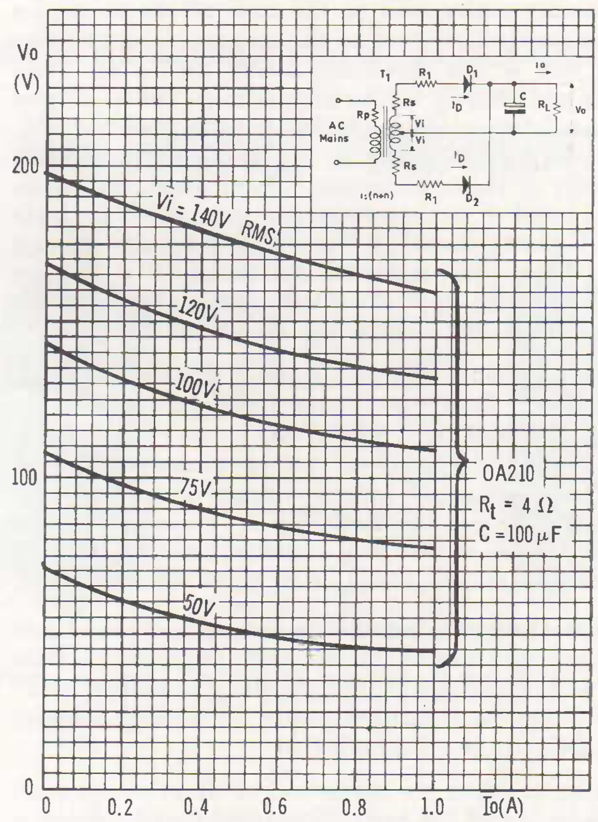


Fig. 1b. Full-wave (C.T. Transformer).

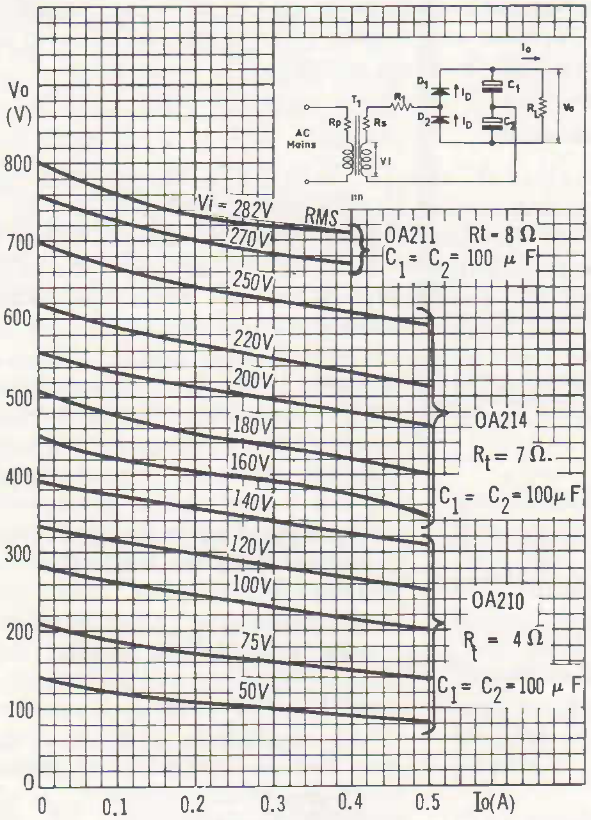


Fig. 1c. Voltage doubler.

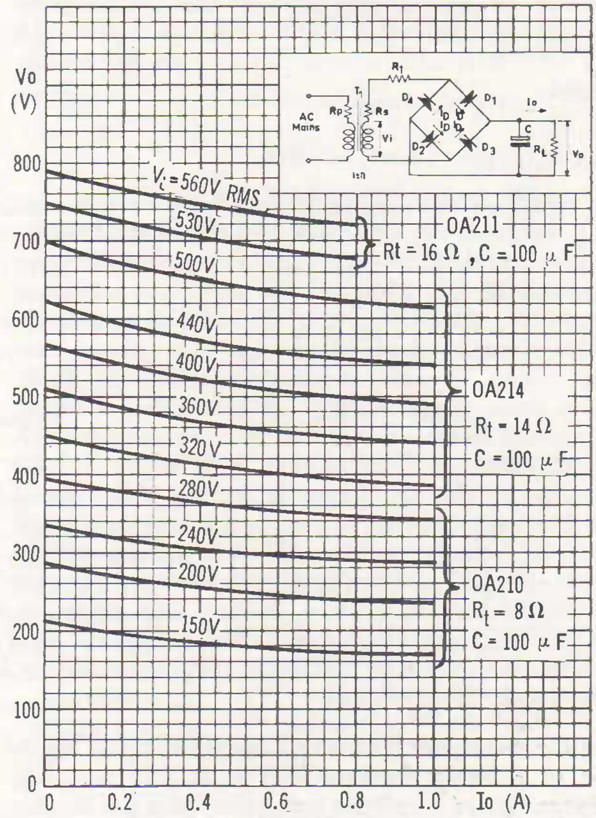


Fig. 1d. Bridge.

Static Inverse Characteristics

At the present state of the art, the user of semiconductors is faced with problems arising from spread of characteristics; and in the case of silicon diodes this is particularly true of the inverse characteristics. This fact is not significant where diodes are used singly and entirely within their ratings, but it assumes great importance where units are connected in series, since the inverse resistance of each diode determines its share of the total inverse voltage across the chain. The spread of inverse resistance from diode to diode may be more than 100:1, and it is therefore obvious that when the sum of individual PIV ratings approaches the applied inverse voltage, the PIV rating of individual diodes may be exceeded.

There are three ways of avoiding this difficulty: (a) by keeping the applied inverse voltage considerably below the total rated PIV of the chain, (b) by selecting diodes with "matched" inverse characteristics, (c) by enforcing equal inverse voltage distribution by external means. Method (a) is uneconomic and method (b) is at best only a partial remedy, since all inverse parameters cannot be matched simultaneously.

Inverse characteristics change throughout life, a fact which should not be overlooked. Miniwatt declines to select diodes since true matching is neither technically nor economically feasible. Method (c) is therefore the only practical solution.

If equal resistors are connected in shunt across each diode such that the external resistive bleed current is at least 5 times the maximum diode inverse current at the maximum junction temperature, then substantially equal voltage distribution will be enforced across each unit. The increased reverse power loss is usually negligible.

Dynamic Inverse Characteristics

Since power diodes are almost invariably used in rectifying circuits, the dynamic inverse characteristics must also be taken into account. With a diode inverse capacitance of only a few pF, the capacitive reactance at **50 c/s** is generally higher by an order of magnitude than the inverse resistance, so externally enforced resistive voltage division is generally sufficient. However, in some circumstances the hole storage effects may warrant instead the use of equal external capacitors connected in shunt across each diode. These capacitors should be chosen to swamp the normal diode inverse capacitance, and thus achieve enforced equality of voltage division. Minority carrier storage or "hole storage," as it has come to be known, assumes major importance only when dealing with high forward currents, long chains of diodes and steep rises of inverse voltage (inductive loads, polyphase circuits, higher supply frequencies).

When the rectifying "barrier" is lowered by the application of a heavy forward bias to a P/N junction, minority carriers (electrons and holes) pass the barrier (electrons from N to P and holes from P to N), constituting a heavy forward current. These minority carriers have only a finite lifetime in the material of the opposite polarity before undergoing recombination. Because of this finite lifetime, a fast reversal of bias at the junction will not immediately result in the static reversed bias current flow. Instead, there is a substantial flow of unrecombined minority carriers swept back to their parent material constituting a large transient current. The time interval before the low static value of reverse current is reached is termed "reverse storage time". Thus appreciable reverse voltage can only be established across the diode after the transient has died somewhat.

Since recovery times vary from diode to diode, one unit in a chain will be the first to assume a high value of inverse resistance, interrupting the flow of the reverse current through the stack. As the reverse current is now very small, the other units are not capable of assuming appreciable inverse voltage until the excess minority carriers have recombined with majority carriers of opposite polarity. This allows the almost full inverse voltage to appear as a transient across the diode with the shortest recovery time, leading to a possible destruction of this unit. A remedy can be provided by shunting each diode with a capacitor. Since the current through a capacitor is proportional to the change of voltage, a relatively large current will bypass the unit with the shortest recovery time, thus allowing the other units to recover more quickly. As the number of diodes in series increases, so also does the probability of one of them having significantly faster recovery time than the others, resulting in greater possibility of failure unless protected. In case parasitic oscillations appear after commutation, a low value of damping resistor may be inserted in series with the shunt capacitor.

Conclusion

Not all of the factors involved in series operation are important in any specific case and this has led to a variety of manufacturers' recommendations. Thus a degree of judgment must be exercised by the circuit designer. Should any doubt still exist, the manufacturer should be consulted.

(This survey has been compiled by Messrs. D. J. Hancock and P. Heins of the Miniwatt Electronic Applications Laboratory.)

References

1. Terman, F. E. — Radio Engineers' Handbook, McGraw-Hill (First Ed., 1943), Section 8.
2. Gray, T. S. — Applied Electronics, John Wiley & Sons (Second Ed., 1953), Chapter VI.
3. H.T. Supply Units for Television Receivers equipped with Silicon Diodes. Electronic Applications, Vol. 21, No. 1, 1960-61.

Professional Tubes

NEW QUICK-HEATING TUBES FOR MOBILE EQUIPMENT

QQC03/14(7983)
QC05/35(8042)

These directly-heated tubes enable the rapid attainment of full output power after the transmitter "press-to-talk" switch has been operated. Their use can greatly reduce battery drain in the "stand-by" condition. Special attention to filament design provides the additional advantages of robustness, reduced drive power requirements and low effective cathode lead inductance.

In a modern mobile transceiver with a fully transistorised receiver section, more than 90% of the standby current is due to tube heater drain. If steps are taken to reduce this drain, the life of the battery can be greatly prolonged. This is especially the case if the set is installed in a vehicle which is often stationary, and if the transmit duty cycle is low. Furthermore, cabinet heating due to continuous tube heater operation is detrimental to the performance and life of the transistors and other components.

As the full power output has to be available almost immediately the "press-to-talk" is operated, directly-heated tubes must be used. Both the QQC03/14 (7983) and QC05/35 (8042) possess specially designed filament structures which are superior to those of the usual directly-heated cathode tubes. Besides being more robust, they provide larger effective emissive areas, and this reduces drive power requirements. This has resulted generally in lower impedance filament structures, which in addition to contributing to increased mutual conductance (more cathode area being effective because voltage gradient is reduced), also reduces the effective cathode lead inductance. This inductance would otherwise reduce the RF power output at higher frequencies. The QQC03/14 has a coated nickel ribbon filament, whilst the QC05/35 uses a "harp" construction in which a large number of short oxide-coated tungsten wires are connected in parallel.

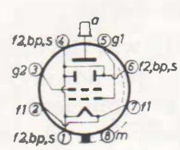
The required filament voltage could be obtained from a separate winding of the DC/DC converter supplying the transmitter.

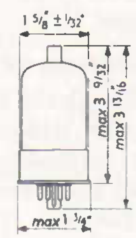
Although the QQC03/14 and QC05/35 are more suited to mobile applications, their indirectly-heated counterparts, QQE03/12 (6360) and QE05/40 (6146) respectively, are to be preferred, for example, in base stations operating with a high transmit duty cycle.

Application details will follow in a later issue.

PHILIPS
QC 05/35 (8042)
Directly heated Beam Power Tetrode for use as RF Output Amplifier for Intermittent Service; 60 W at 60 Mc/s, 30 W at 175 Mc/s.

General Electrical Data
Filament, oxide-coated
Heater voltage, 1.6 ± 15% V
Heater current, 3.2 A
Heating time after 0.4 secs. power output, —3dB max.



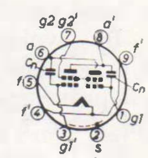


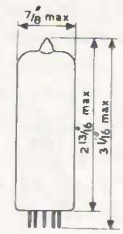
Direct Interelectrode Capacitances
Plate to Grid No. 1, max. 0.24 pF
Plate to filament, 8.5 pF
Grid No. 1 to filament, 13.5 pF

Typical Characteristics
Plate voltage, 200 V
Grid No. 2 voltage, 200 V
Plate current, 100 mA
Amplification factor, 4.5
Mutual conductance, 7 mA/V

PHILIPS QQC 03/14 (7983)
Directly heated Double Tetrode for use as RF Output Amplifier for Intermittent Service. Useful Output power, 11 Watts at 200 Mc/s.

General Electrical Data
Filament, oxide-coated
Heater voltage, 3.15 ± 10% V
Heater current, 1.65 A
Heating time after 0.8 secs. power output, —3db max.





Direct Interelectrode Capacitances

	Each Unit	Both Units in Push-Pull
Output capacitance	3.2	1.7 pF
Input capacitance	6.8	5.4 pF
Plate to Grid No. 1 max. (internally neutralised up to 200 Mc/s)	0.08	— pF

Typical Characteristics (each unit)
Plate voltage, 200 V
Grid No. 2 voltage, 200 V
Plate current, 30 mA
Amplification factor, 7.5
Mutual conductance, 3 mA/V

THE 6GV8

New Triode Pentode for Vertical Deflection

The 6GV8 is the new preferred Miniwatt valve for use as vertical oscillator and amplifier to drive any type of deflection yoke currently available. Because the power output capabilities of the 6GV8 are greater than those of the familiar 6BM8, it offers economies in the vertical output transformer whilst retaining a reserve of power. Type 6GV8 is manufactured in Australia and is readily available.

Since the introduction of TV to this country the triode-power pentode 6BM8 has proved very popular in vertical oscillator and output service. The 6BM8 was developed with both vertical deflection and audio service in view, so that certain compromises had to be made in the choice of its base pinning and characteristics. By avoiding such compromises it is possible to develop valves with improved performance in either of these types of service.

Miniwatt has now released two new triode power pentodes of greater power output capability than the 6BM8. Type 6GV8 has been developed exclusively for vertical deflection and so could be given optimum characteristics for this particular application. Its companion, the 6GW8, has been specifically designed for audio use and further information on this type will be given in a later issue of the Digest.

Limitations of the 6BM8

Although the 6BM8 was developed for 90° deflection, its power output capability is sufficient for 110°/114° toroidally-wound coils. In this last type of service, however, the pentode is operated near its maximum ratings as far as plate current availability, screen and plate dissipation are concerned.

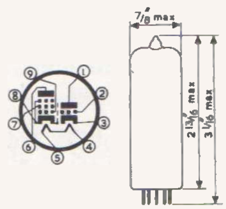
This fact requires careful design of the output transformer. Such factors as minimum primary inductance, turns ratio, DC resistance of windings and efficiency become increasingly important, resulting generally in a transformer with ⅞″ tongue width and 1″ stack if the amplifier is to be operated from a reasonably low supply voltage of, say, 220 V.

Operation from a low supply voltage (about 220 V after decoupling from the main H.T. rail) is particularly advantageous if the required H.T. for the horizontal amplifier is

Miniwatt 6GV8 Triode-Pentode for Vertical Deflection

General Electrical Data

Heater voltage	6.3 V
Heater current	0.9 A



Typical Characteristics	Pentode Section	Triode Section	
Plate voltage	170	100	V
Grid No. 2 voltage	170	—	V
Plate current	41	10	mA
Grid No. 2 current	2.7	—	mA
Grid No. 1 voltage	—15	0	V
Mutual conductance	7500	5500	μmho
Plate resistance	25	9	K Ω
Amplification factor	—	50	

Comparison of 6GV8 and 6BM8 Characteristics

Pentode Section	6GV8		6BM8		
Dynamic characteristics of average new valve (end of scan)					
Plate voltage	50	60	50	60	V
Screen voltage	170	190	170	190	V
Control grid voltage	—1	—1	—1	—1	V
Available peak plate current	200	238	135	173	mA
Peak screen current	35	—	40	—	mA
Recommended end of scan operating conditions, including allowance for valve spread, life deterioration and 10% mains drop.					
Plate voltage	50	60	50	60	V
Screen voltage	170	190	170	190	V
Control grid voltage	—1	—1	—1	—1	V
Peak plate current	120	143	70	85	mA
Max. plate dissipation	7	—	5	—	W
Max. screen dissipation	1.5	—	1.8	—	W

Triode Section	6GV8	6BM8	
Plate voltage	100	100	V
Grid voltage	0	0	V
Plate current	10	3.5	mA
Mutual conductance	5500	2500	μmho
Amplification factor	50	70	—
Max. peak cathode current	100	100	mA

also 220 V. The power supply can then be designed to supply 220 V, this being the highest H.T. rail required in the receiver proper.

Advantages of the 6GV8

Inspection of the comparison table shows that the peak plate current availability of the pentode section of the 6GV8 is almost 50% higher than that of the 6BM8, with the margin even higher for the recommended operating conditions. The increase in available power has been achieved by lengthening the pentode system. On the other hand, the triode system has been shortened to reduce microphony. Screen dissipation is kept low by the use of aligned grid techniques for grids 1 and 2.

The higher permissible peak plate current in the pentode permits the use of a much smaller output transformer than hitherto possible, with attendant saving in cost and weight. Alternatively, if transformer size is not the major concern, then increased valve life may be expected.

Examples of Two Circuits

Two vertical oscillator/output circuits follow: these use the 6GV8 with a small output transformer of ¾″ tongue width and ¾″ stack.

The first design features superior performance with economy.

Economy is achieved by employing the triode section as a multivibrator to avoid the expense of a blocking oscillator transformer and by using a small output transformer. Quality is ensured by using cathode bias in the output stage and heavy voltage feedback with two controls for linearity correction; 5% linearity can be achieved.

In the second design, economy is the main consideration; it uses the absolute minimum of components consistent with acceptable operation. The performance expected of this amplifier would not be of the same high standard as the previous one.

CIRCUIT 1: QUALITY CIRCUIT

Circuit Description

The oscillator is a very stable feedback multivibrator; and because it has some unorthodox features, the following description of its operation is included.

Assume that at the start of scan, C_2 is charged negatively and C_3 positively with respect to ground. VIA is therefore cut off during scan until the capacitors have discharged sufficiently, through R_2 and $R_3 + R_5$ respectively, for VIA to start conducting. The resultant drop in plate voltage of VIA is transmitted to the control grid of VIB, cutting this valve off. The resultant positive inductive plate-voltage pulse is fed back to the grid of VIA, making this valve conduct even more strongly. This regenerative process takes place in a very short time and constitutes the vertical retrace. C_2 and C_3 are again charged by the grid and cathode currents of VIA. The decay of the flyback pulse from VIB cuts off VIA rapidly. The cycle is then repeated.

Frequency control is achieved by varying R_3. The oscillator is synchronised by applying a negative sync. pulse to the cathode of VIA, thus initiating conduction just prior to the start of conduction in the free-running state. A two-stage integration network (R_4, C_1 and R_1, C_2) in the oscillator feedback path prevents horizontal flyback pulses induced in the vertical deflection coils from reaching the oscillator grid and interfering with interlace.

During the period of non-conduction of VIA, capacitor C_6 is charged through R_8 from a potentiometer (R_7, R_9, R_{10}) across the boost supply and discharged when VIA conducts. The resulting sawtooth voltage is applied to the control grid of VIB through C_5 and R_{14}. At the C_5-R_{14} junction, voltage feedback is applied from the

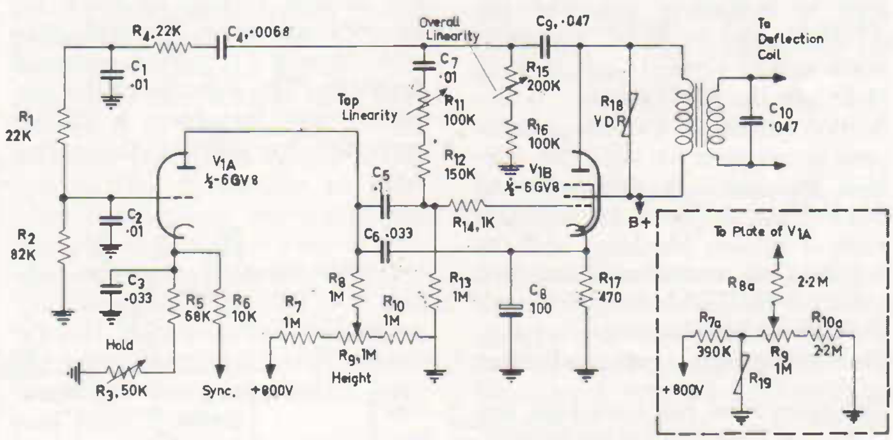


Diagram and Parts List for Circuit 1

R_1, R_4	22 K	R_{11}	100 K, lin.	R_{18}	VDR Philips
R_2	82 K	R_{12}	150 K		type E298GD/A260
R_3	50 K, lin.	R_{13}	1 M	R_{7a}	390 K, 1 W
R_5	68 K	R_{14}	1 K	R_{8a}	2.2 M, 1 W
R_6	10K	R_{15}	200 K, lin.	R_{19}	VDR Philips
$R_{7, 8, 10}$	1 M, 1 W	R_{16}	100 K		type E298ED/A265
R_9	1 M, lin.	R_{17}	470 K, 1 W		

All resistors are Philips insulated cracked carbon type B8 305 05B (½ W) and type B8 305 06B (1 W).

C_1, C_7	.01/400 V	C_4	.0068/400 V	C_8	100/25 VW electrolytic
C_2	.01/125 V	C_5	.1/400 V	C_9	.047/1000 V polystyrene
C_3	.033/125 V	C_6	.033/400 V	C_{10}	.047/125 V

All capacitors are Philips polyester type C296AA/A (125 V) and C296 AC/A (400 V) unless otherwise stated.

TRANSFORMER DATA

Wasteless E.I. laminations ¾in. tongue width, square stack, annealed grain-oriented steel.
Primary winding: 2700 turns, 34 B & S, D.T., 230 Ω, 16.5H (40 mA DC).
Secondary winding: 465 turns, 27 B & S, 9 Ω. 2.5 mil gap paper.

plate of VIB, the amount being controlled by R_{11}. This pre-setting control will mainly influence the wave-shape at the start of scan and is therefore labelled "top linearity." A shaping network (C_9, R_{15}, R_{16}) in the overall feedback path also influences the linearity, R_{15} being the overall "linearity control." The voltage-dependent resistor (VDR) R_{18} limits the flyback pulse amplitude.

Operating Conditions

Pentode			
Supply voltage	200	220	V
Deflection coil current	450	450	mA
Peak plate voltage	900	960	V
Plate voltage end of scan	56	67	V
Peak plate current	96	106	mA
Average cathode current	40	45	mA
Screen dissipation	1.17	1.35	W
Plate dissipation	4.85	6.4	W
Triode			
Peak cathode current	21	—	mA

Output Transformer

The transformer is designed to feed a deflection yoke of the AT1009 series in which the resistance of the vertical coils is 49 Ω, including the NTC resistor. A deflection current of 450 mA peak-to-peak is required for about 5% vertical overscan. A double-wound transformer is used to facilitate vertical retrace blanking, and the windings are wound on a moulded former with double-tough enamelled wire without layer-interleaving, thus aiding speed of production.

The turns ratio has been kept low (5.8:1) for the output stage to operate satisfactorily from supply voltages as low as 200 V. Due to the restricted winding space, the low primary inductance of 16.5 H at 40 mA DC requires a plate current with negative initial slope. To minimise this requirement and achieve maximum efficiency, grain-oriented steel should be used.

Stabilisation of Oscillator Supply Voltage

By adding a single VDR, it is possible to stabilise the oscillator supply voltage derived from the boosted HT. This helps considerably in maintaining correct aspect ratio in the presence of mains voltage fluctuations. The necessary modifications to circuit 1 are shown as an insert to the diagram. Improvements in performance are indicated below:—

Mains Variation	Height variation Unstabilised	Stabilised
—10%	—6.2%	—3.4%
+10%	+5.3%	+3.1%

N.B. The boosted H.T. voltage is derived from a stabilised horizontal output stage.

CIRCUIT 2: ECONOMY CIRCUIT

The oscillator is again a feedback multivibrator, but the hold control is R_2 which varies the DC bias on the grid of VIA. The negative-going sync. pulse applied to the control grid of VIB is amplified and reaches the grid of VIA through R_{10} and C_2, initiating conduction in that valve. The sawtooth appearing across C_3 is applied to the control grid of VIB via C_4 and R_{12}. Since this circuit does not employ voltage feedback for linearity correction, the grid drive waveshape is a slightly exponential sawtooth. The linearity of the output current waveform is further corrected by utilizing the curvature of the I_a/V_g characteristic of the output pentode ("conductance linearising"). This is achieved by applying a variable amount of bias to this stage in order to help produce the required parabolic component of plate current. Bias is tapped off the negatively-charged capacitor C_1 by means of R_5 ("linearity control") and applied to the grid circuit of VIB through R_{11}. With this circuit it is difficult to prevent some interaction between the controls. The output transformer is the same as that used in Circuit 1.

Operating Conditions of Circuit 2

Pentode			
Supply voltage	200	220	V
Deflection coil current	450	450	mA
Peak plate voltage	920	980	V
Plate voltage end of scan	85	100	V
Peak plate current	95	90	mA
Average cathode current	38	41	mA
Control grid voltage	—21	—21.5	V
Screen grid dissipation	1.18	1.36	W
Plate dissipation	5.7	5.9	W
Triode			
Grid voltage	—37	—38	V
Peak cathode current	32	32	mA

(This article is based on work carried out in the Miniwatt Electronic Applications Laboratory by P. Heins.)

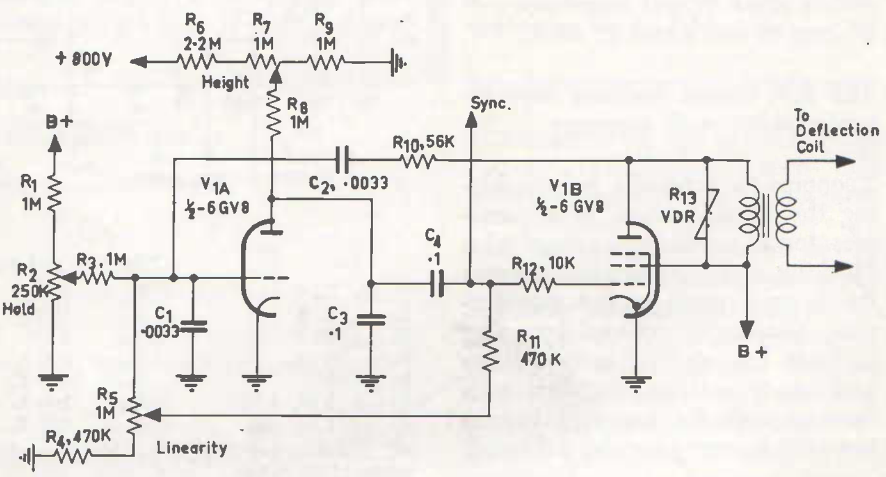


Diagram and Parts List for Circuit 2

R_1, R_8, R_9	1 M, 1 W	R_5, R_7	1 M, lin.	R_{12}	10K, ½W
R_2	250 K, lin.	R_6	2.2 M, 1 W	R_{13}	VDR Philips type E298GD/A260
R_3	1 M, ½ W	R_{10}	56 K, 1 W		
R_4	470 K, ½W	R_{11}	470 K, ½W		

All resistors Philips insulated Cracked Carbon type B8 305 05B (½ W)
B8 305 06B (1 W)

C_1 .0033/125 V C_2 .0033/600 V polystyrene C_3, C_4 .1/400 V
All capacitors Philips polyester types C296AA/A (125 V), C296AC/A (400 V).

Components

NEW CHANNELS FOR OLD TUNERS

The latest information concerning the establishment of the 26 new TV stations in country areas suggests that some of these may begin experimental transmission towards the end of this year, with permanent services being established early in 1962.

The modification of tuners in older TV receivers in the affected areas presents service organisations with an ideal opportunity for additional and profitable business. It is suggested that these modifications should form part of all current service calls, so that impossible demands will not be made on service personnel when the stations come on the air.

Servicemen and retailers are reminded that the coil assemblies necessary to modify the earlier 10-channel 12-position Miniwatt tuners, types AT7580 and NT3001, are readily available. In any one area only those coils for the additional channels, or channels with changed frequencies, need be added but coils for all channels are available if required.

Being of turret-type construction, it is a comparatively simple operation to add or change coil assemblies in the Miniwatt tuner. The coils are held in position on the rotor by spring clips, and any channel may be readily changed without affecting the others.

The new coil assemblies are factory prealigned, and normally require no further adjustment. However, for optimum results, the oscillator frequency should be reset to suit each individual receiver installation when a test signal is available. Variation of oscillator frequency can be made by adjustment of the oscillator slug which is readily accessible through a hole in the front of the tuner housing.

When ordering coil assemblies, care should be taken to specify the tuner for which they are required, as differences exist in the coils due to the different valves used in the AT7580 and NT3001 tuners.

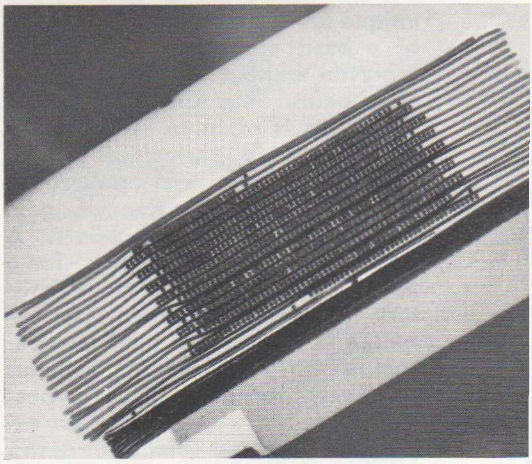
Whilst the coil assemblies bear the same channel identification marking, assemblies for type AT7580 tuner are identified by a red paint marking.

Supplies of all coil assemblies are available from Philips branches throughout the Commonwealth.

POLYESTER ENCAPSULATED EHT WINDING

Improves Factors of Safety in New Horizontal Output Transformers

Research in recent years has produced a wide variety of plastic materials, many of which offer unique characteristics making them far superior to materials previously available. Of particular importance to the electronics industry are the so-called Polyesters. Some of these have extremely high dielectric strengths, are particularly homogeneous, have negligible moisture absorption and are chemically inert up to high temperatures—making them ideal as insulating and encapsulating agents for electronic components.



Enlarged cross-section of the EHT winding of the new encapsulated horizontal output transformers. The remarkable interlayer penetration of this material can be clearly seen.

The EHT winding of TV horizontal output transformers has always presented design problems with regard to corona and flashover and many different methods have been employed in attempts to provide adequate protection against breakdown. The availability of polyester encapsulants has completely changed the situation and, as a result, two new Horizontal Output Transformers, types NT3100 and NT3101, incorporating techniques not previously used in the manufacture of these components, have now been released.

Wound on a high dielectric strength Plaskonalkyd moulded former with Poly-carbonate foil interlayer insulation, the entire winding is vacuum impregnated and completely encapsulated in a polyester. This contains a specially blended filler which provides increased mechanical strength and fire protection. EHT coils so constructed provide greater factors of safety than with previous methods. Special measures are taken in the moulding process to ensure the centrality of the winding, control the external physical dimensions of the assembly and correctly position the connecting leads.

The two transformers are identical except that the NT3101 is fitted with a resistor-capacitor network to suppress ringing phenomena. In the case of non-stabilised horizontal output circuits, variations in mains-voltage and brightness-control settings can cause picture shrinkage, making the ringing obvious to the viewer. Stabilised horizontal output stages do not suffer from these disadvantages and ringing suppression networks are not required.

Except for the new EHT coil construction incorporated in the horizontal output transformers, types NT3100 and NT3101, they are physically and electrically identical with the AT2016T series of transformers, which they supersede and can directly replace.

Semiconductors

SOME ADVANTAGES OF SILICON TRANSISTORS

— in Industrial Applications and in Stable DC Amplifier Design

Silicon transistors can operate with junction temperatures of the order of 150°C, have exceedingly low leakage currents, and possess the unique feature that thermal stability of a basic amplifier stage can be achieved without altering the operating point. Thus the use of silicon transistors is often to be preferred to comparable germanium types, especially in industrial applications and in the construction of gain stable drift-free DC amplifiers. A table is included which lists, for a preferred range of silicon transistors, the more important data required in such applications, and for comparison purposes, this includes "comparable" germanium types, that is, types which might be expected to perform similar functions.

It is generally recognised that the maximum permissible junction temperature of silicon transistors is higher than that of germanium types by a factor of approximately 2:1. Miniwatt silicon transistors can be continuously operated at junction temperatures up to 150°C. This higher T_j rating is ideally suited to industrial applications where elevated ambients are often encountered in excess of the rated junction temperatures of germanium transistors. However, even in the lower temperature range, where germanium may be used, there are still the advantages of greatly reduced leakage currents and higher power and current ratings for a given size. For example, in the case of the BCZ11, I_{CBO} is only 0.1 μA at 100°C, compared with 225 μA for the OC71 germanium transistor at 70°C.

In considering another field, that of DC amplification of threshold signals, silicon transistors again possess some advantages. The design of stable DC amplifiers of small input signals has presented problems arising from variation of quantities V_{BE}, I_{CBO}, h_{FE} and h_{fe} with temperature. For silicon transistors the leakage current can usually be neglected. However, since DC amplifiers cannot be AC coupled, a shift in operating point cannot be distinguished from a drift in input signal. *Thus both operating point and gain have to be maintained constant with changes in temperature.* Differential amplifier stages have been used in conjunction with "matched" transistors, to reduce the drift to the differences

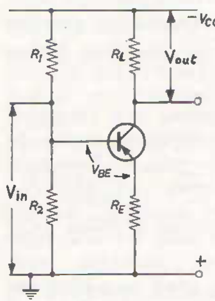


Fig. 1. Basic DC Voltage Amplifier Stage.

$R_L << r_c$ where r_c = intrinsic collector resistance.

Material	Type Number	Use	$-V_{CB}$ (V)	$-V_{CE}$ (V)	$-I_C$ (mA)	$-I_{CM}$ (mA)	P_{TOT} at 25° C (mW)	$T_{j\,max}$ (°C)	h_{fe}	h_{FE}	$-I_{CBO}$ (μA)	r_{bb}' (Ω)	F (dB)
			Absolute Maximum Ratings						Typical Charactertics at 25°C				
Si	BCY10 (OC204)	medium voltage, medium current, industrial	32	32	250	500	310	150	40	24	0.02	100	7.0
Ge	OC72		32	32	125	250	165*	75	—	70	4.5	—	<15
Si	BCY11 (OC205)	high voltage, medium current, industrial	60	60	250	500	310	150	40	24	0.02	100	7.0
Ge	OC77		60	60	125	250	165*	75	70	>45	4.5	—	<15
Si	BCY12 (OC206)	medium voltage, medium current, industrial	32	32	250	500	310	150	40	40	0.02	100	7.0
Ge	OC76		32	32	125	250	165*	75		>45	4.5		<15
Si	BCZ10 (OC200)	AF, general industrial	25	25	50	50	250	150	20		0.001	125	8.0
Ge	OC70		—	30	10	50	125	75	30	—	5.0	—	10.0
Si	BCZ11 (OC201F)	AF, general industrial	25	25	50	50	250	150	35	—	0.001	125	6.0
Ge	OC71		—	30	10	50	125	75	47		4.5		10.0
Si	BCZ13	LF, general purpose, Sub-min.	20	—	10	—	65	125	—	>10			
Ge	OC57		7	3	5	10	10	55	35		1.5		<10
Si	OC202	MF, general purpose	15	15	50	50	250	150	70		0.001	300	8.0
Ge	OC75		—	30	10	50	125	75	90		4.5		10.0
There is also a Silicon transistor which has no comparable Germaniun type but which could be considered as a high-voltage Silicon version of the OC70:—													
Si	BCZ12 (OC203)	AF, general purpose	60	60	50	50	250	150	15		0.01	125	8.0

* With cooling fin type 56200.

r_{bb}' = extrinsic base resistance. $-I_{CBO}$ = collector to base leakage current. h_{FE} = large-signal or DC short-circuit current gain. F = noise factor. h_{fe} = small-signal short-circuit current gain. P_{TOT} = allowable total power dissipation in transistor.

in the parameters of two transistors[1]. However, another problem is then introduced in that it is almost impossible to maintain equality of transistor junction temperatures.

If one considers the basic DC voltage amplifier circuit of Fig. 1, and develops the thermal stability equation[2] (condition for gain stability with temperature change), some interesting points arise. This equation is:—

$$\frac{\Delta I_e}{I_e} = \frac{\Delta T}{T}(1+\gamma\chi) - \frac{\Delta h_{fe}}{h_{fe}}\left\{\frac{1+\chi+(R_1/r_e)}{h_{fe}} + \chi\right\}$$

where ΔI_e is change in emitter current for a change in junction temperature ΔT; γ being 1.6 and 2.6 for germanium and silicon p-n-p alloy junction transistors respectively; $\chi \approx r_{bb}'/r_e$, where $r_e \approx 25/I_e\,\Omega$ for germanium transistors, and both the OC202 and the BCZ range of silicon transistors tabulated.

As h_{fe} of germanium transistors is approximately constant over the operating range, the term in Δh_{fe} reduces to zero; but the term in ΔT will never be zero. This means that gain can only be maintained stable at the expense of a change in operating point, which cannot be tolerated in a DC amplifier working with very low signals. However, for silicon transistors, h_{fe} is not constant over its wide working range, and the term in Δh_{fe} cannot be neglected. It then follows if the terms in Δh_{fe} and ΔT are equalized, that ΔI_e will be zero, and no change in operating point will result.

Thus, besides extremely low leakage current—which may be decisive in itself—there is another basic reason why silicon transistors should be considered in drift-free stable DC amplifier design. As the result of improved manufacturing techniques and quantity production of Miniwatt silicon transistors, the above advantages can be obtained at little additional expense.

References

1. "The Design of High Stability DC Amplifiers," P. Beneteau — Semiconductor Products 4, No. 2, Feb. 1961.
2. "The Thermal Stability Equation for Different Types of Transistor Voltage Amplifiers," Iu. R. Nosov and B. I. Khazanov. Radiotekhnika 15, No. 3, 38-44, 1960. (As translated in Radio Engineering 15, No. 3, 53-62, 1960.)

Ferrites

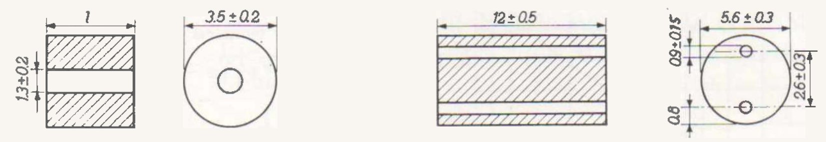
FERROXCUBE SCREENING BEADS AND FERROXCUBE-CORED CHOKES

— Inexpensive Components for VHF Rejection

Small Ferroxcube beads and chokes, simply inserted in supply leads, offer a quicker and less expensive means of damping out unwanted RF energy entering via these leads.

Supply leads in radio, TV and general electronic equipment often form easy paths along which unwanted RF energy can be carried from one circuit to another or from one stage to another. Capacitive decoupling of the leads will not always be satisfactory due to possible parasitic resonances, etc. For the same reason, addition of series inductance will not always be successful. Although the application of Ferroxcube screening beads and Ferroxcube-cored chokes to this problem is not new, it is felt that full advantage has not yet been taken of these small, inexpensive, but exceedingly useful components. Concise information on their use and the forms in which they are available is given, and the advantages of these units are contrasted with the several short-

BEADS



56 590/65/3B (or 4B). $l = 3$ mm.
Single Bead

56 390 31/4B
Double Bead

WIDE-BAND CHOKES

Single Choke

Double Choke

Type No.	No. of turns	Optimum Impedance Z max. (KΩ)	Approx. freq. (Mc/s) at which Z max. occurs
VK200 09/4B	1½	0.45 ± 20%	250
VK200 10/3B*	2½	0.75 ± 20%	50
VK200 10/4B**	2½	0.85 ± 20%	180
VK200 11/4B**	2 × 1½	1 ± 20%	110

Core without wire can be supplied under
* type No. VK211 07/3B
** type No. VK211 07/4B

All dimensions in mm.

Fig. 1. Beads and chokes available in Grades 3B and 4B Ferroxcube.

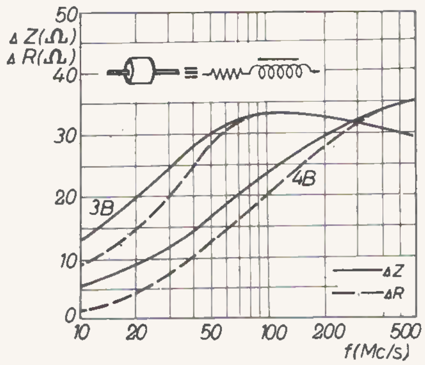
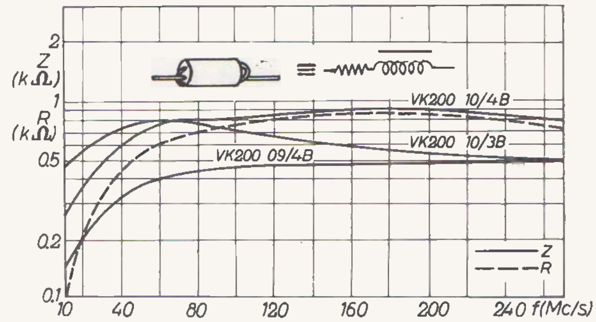


Fig. 2 (*left*). Increase of impedance and loss resistance, caused by single beads 56 590 65/3B and 56 590 65/4B threaded on straight wires.

Fig. 3 (*right*). Impedance and loss resistance of a Ferroxcube choke as a function of frequency.



comings of self-resonant air-cored chokes.

A number of beads (total length small compared with the wavelength) simply strung on the supply leads, or a single wide-band choke may be used in a given application. However, for a given space, somewhat better performance can be obtained using the chokes. These possess six axial holes through which are threaded $1\frac{1}{2}$, $2\frac{1}{2}$ or $2 \times 1\frac{1}{2}$ turns of wire.

The forms in which beads and chokes are readily available, together with some performance details of the chokes, are shown in Fig. 1.

Performance curves for single beads and chokes are given in Figs. 2 and 3.

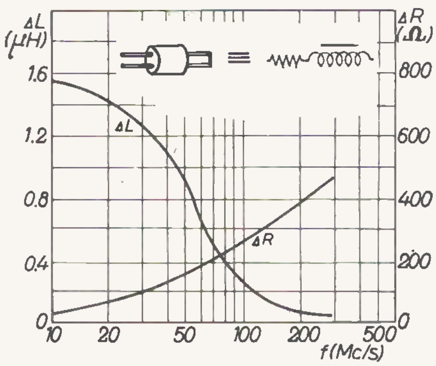


Fig. 4. Increase of inductance and loss resistance caused by one double bead 56 390 31/4B threaded on two straight wires.

It will be noted that above about 60 or 80 Mc/s the impedance is substantially resistive and tends to be constant. This is especially so for the chokes.

Insulated or bare wire may be strung through the beads, but if grade 3B is used with bare wire a maximum fall off in resistance of 8% could be expected, due to its lower resistivity.

Double beads and double chokes are available for twin leads, in which case the advantages of mutual inductance can be obtained by using them in place of single units. Fig. 4 gives performance curves for one type of double bead. Grade 4B should be used for double units as it will provide ample insulation between the two windings even if bare wires are used.

Either beads or chokes may be used in conjunction with small ceramic capacitors in "damping circuits," to provide additional rejection at the self-resonant frequency of the capacitor with its leads. Figs. 5 and 6 compare typical performance obtainable.

Damping is defined as

$$20 \log_{10} V_1/V_2 = 20 \log_{10} \omega C.Z_\omega$$

where

$$Z_\omega >> \frac{1}{\omega C} \text{ and } R_L >> \frac{1}{\omega C}.$$

In selecting an RF choke for a feed-line, a self-resonant single-layer air-cored winding on a former of small diameter could be considered. This has several disadvantages:—

(a) A large number of turns may be required with consequent high effective shunt capacitance. The high L/C ratio results in a sharp fall off in impedance if the choke is slightly off tune.

(b) On the other hand, if the shunt capacitance is too low, the choke is easily detuned by a variation in stray circuit capacitance.

(c) Unless damped by a parallel resistor, spurious resonances may occur.

If a Ferroxcube core is now considered, (a) will no longer apply, as the turns will be reduced because of high core permeability. Also, the substantial losses in grades 3B and 4B Ferroxcube at elevated frequencies cause heavy damping of the inductance. This will widen the bandwidth and hence (b) no longer applies. As a direct consequence of the damping caused by the Ferroxcube core, no parallel resistor is required as in (c). Furthermore, as the core losses are extremely low at supply frequencies (e.g. 50 c/s), there is no interruption of the supply.

The extremely flat impedance response of the chokes (Fig. 3) arises from the fact that for coils with closed Ferroxcube cores, both resistance and reactance tend to be constant with frequency.

Winding (and testing) of special coils is expensive and full advantage should be taken of these small inexpensive components which are known quantities often capable of superior performance.

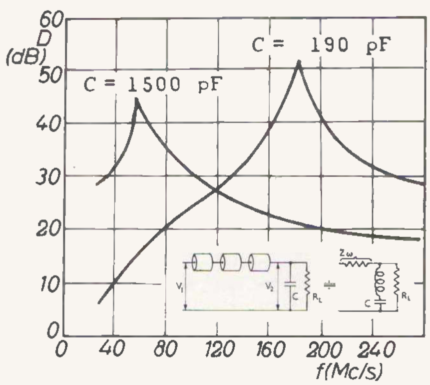
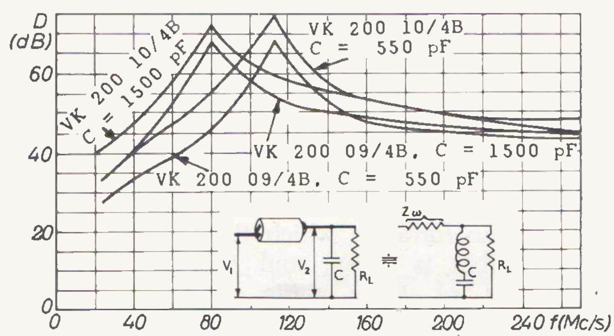


Fig. 5 (*left*). Damping in an LC circuit consisting of a string of three Ferroxcube beads 56 590 65/3B and a ceramic capacitor.

Fig. 6 (*right*). Damping in an LC circuit consisting of a Ferroxcube choke and a ceramic capacitor.



Commercial Comment

THE CREDIT SQUEEZE AND TV

There is a feeling among some of those connected with the TV industry that the credit squeeze is entirely to blame for the recent fall in sales of TV receivers. We do not wish to underestimate the effects of the credit squeeze. However, it should be borne in mind that the growth of the TV market has surely indicated that a turn-down in sales was bound to occur as the development of various areas moved to higher degrees of saturation. This growth of TV homes is clearly indicated in the graphs shown.

The first graph, showing the saturation of TV homes in the principal viewing areas, highlights the rapid approach towards saturation and the speed with which the relatively "new" viewing areas are catching up with the "older" ones.

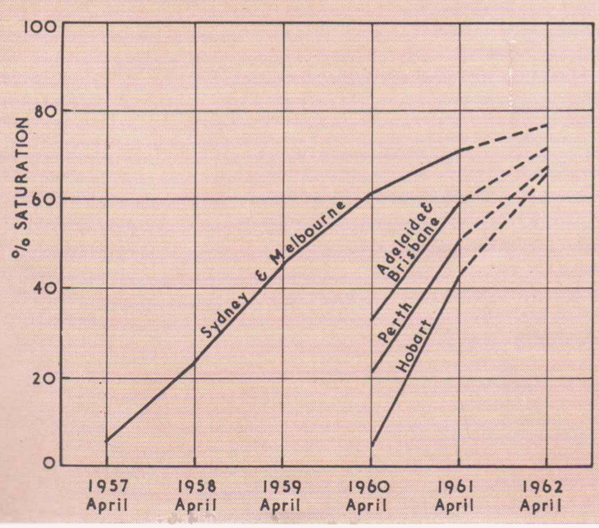
The second graph depicts the annual sales of TV sets from inception, with an estimate of next year's sales. Obviously, the "older" States, N.S.W. and Victoria, have already experienced a fall in annual sales; but because the "new" States have moved more rapidly towards saturation, the fall in sales in these areas will be proportionately greater.

The fact emerges that the TV industry has naturally entered a new era, quite apart from the temporary effects of the credit squeeze. Of course, a means of establishing additional turnover now becomes essential, and preferably one that may be expected to yield long-term benefits.

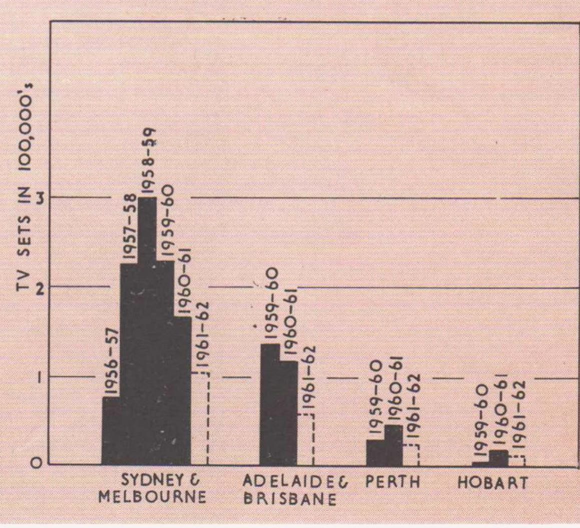
To independent service organisations and retailers possessing their own service facilities, we recommend that careful consideration be given to planning an expansion of their activities in the service field. It has been found in America and, in fact, in all countries of high TV home saturation, that the annual value of TV service by far exceeds the total annual value of all TV receiver sales. Admittedly, there is no current shortage here of TV service, but neither is there a surplus of organisations providing the most efficient type of service and at the same time making themselves known by sales promotional activities. Naturally, efficient service demands competent technical personnel, suitable equipment and adequate space. It should not be overlooked, however, that your service organisation now needs to be very effectively advertised to secure more of the market in this profitable field.

In this regard we mention that the Miniwatt Division has a variety of sales promotional material available. A particularly effective item is the adhesive label for attachment to the back of a TV receiver after service. This ensures that your company's name, address and 'phone number will be always readily available in the right place at the right time, in the customer's home.

TV Saturation in Six Capital Cities



Annual Sales of TV in Six Capital Cities



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