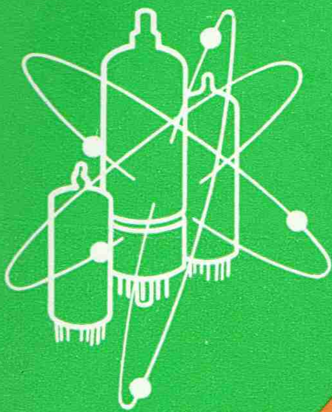

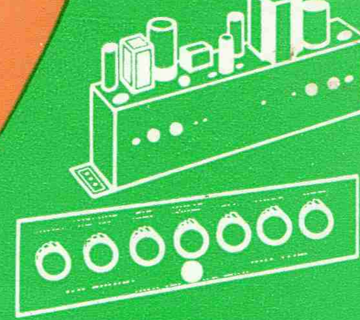

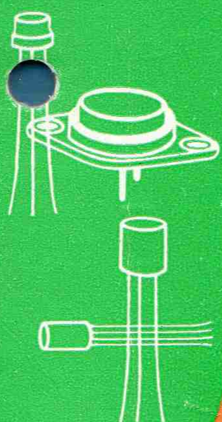

RADIOTRONICS

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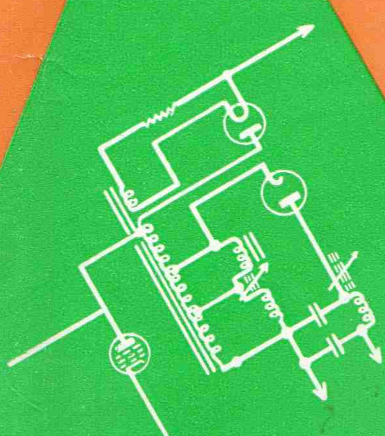
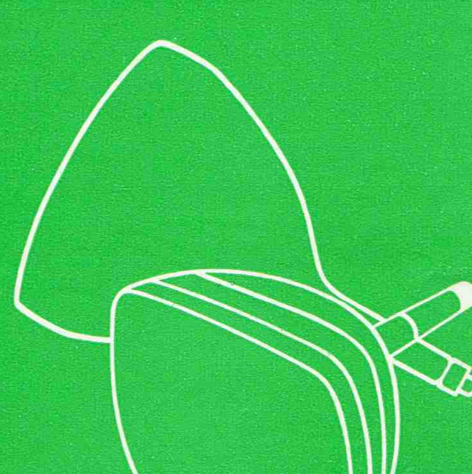
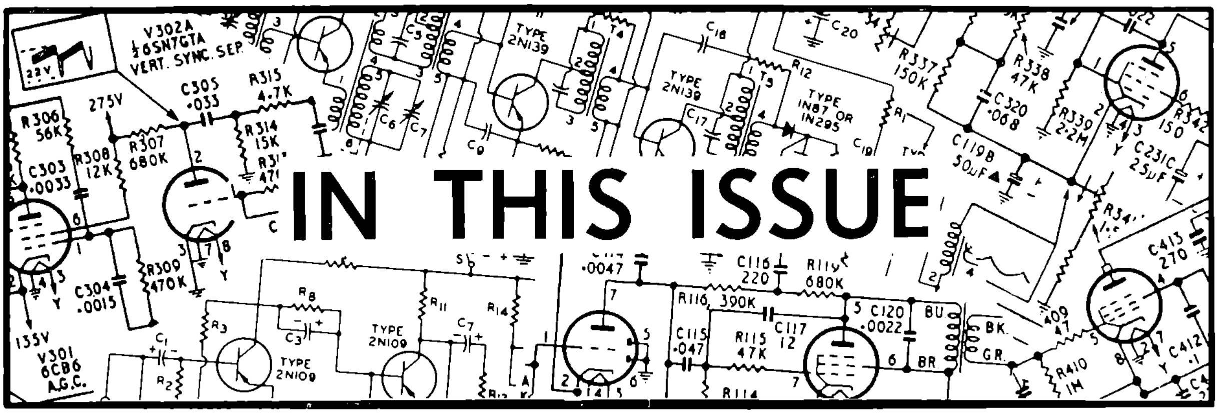
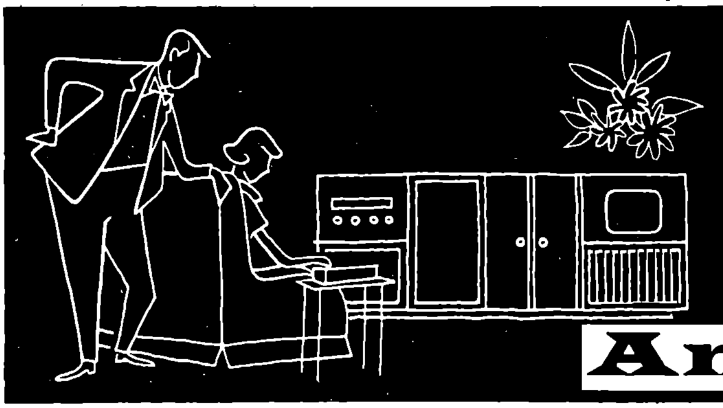






IN THIS ISSUE

For the second month in succession we are able to bring you an expanded issue of your magazine. Of particular interest is the short note on the miniaturization of electronic equipment by Dr. J. T. Wallmark. This note points the way to the manufacture of equipment of hitherto impossibly small size. It is anticipated that the principle of integrating many functions into a single small component will enable engineers to reverse the trend of increasing apparatus size occasioned by its increasing complexity, and lead to the development of more economical, versatile and reliable equipment.



KT88 in Audio Amplifiers

PART 1

INTRODUCTION

The KT88 may be used in pairs in either triode, pentode or ultra-linear push-pull circuits and this article gives circuits and operating data for amplifiers giving outputs of 30 to 400 watts in various conditions of operation. The ultra-linear circuit is recommended for outputs up to 100 watts when the maximum output coupled with the lowest distortion is required and it is also used in the multiple-pair 400w amplifier.

The curves in Fig. 1 show the output and approximate distortion obtained from push-pull KT88 valves with various positions of the screen grid taps on the output transformer. At the left-hand side (0% taps) the valves are, of course, pentode-connected and on the right-hand side (100% taps) they are triode-connected. Generally speaking, the taps should be spaced at 20% to 50% of the turns on each half-primary from the centre, with 30% to 40% as the optimum. The dotted curve shows the reduction in output impedance obtained.

Either cathode bias or fixed bias may be used and the circuits show both types of operation. The former has the advantage of simplicity whereas the latter provides the normal maximum output of 100 watts from a pair of valves and higher efficiency.

The output power of the ultra-linear circuit is not less than that of the pentode at any given supply voltage and has the advantage that a low impedance screen supply is not required. The ultra-linear circuit does, in fact, show a rather higher efficiency in that a lower current is required from the power supply. For example, with fixed bias and a supply voltage of 460 an output of 65 watts is obtained with an anode current of 240ma in both the pentode and the ultra-linear arrangements but the pentode requires, in addition, a screen current of 35ma. Furthermore, it is desirable that the pentode screen supply be stabilised. The output impedance and distortion are both more favourable in the ultra-linear circuit, the former being 6,500 ohms in this example, compared with 50k ohms, and the distortion is almost entirely independent of load impedance above the rated value. Information on the effect of the positions of the screen taps upon the degree of intermodulation is given in an Appendix.

The triode connection is sometimes preferred when a moderate power output with low distortion is required. An output of up to 27 watts is obtainable with a high tension supply of 485 volts and cathode bias and, even at lower voltages, the output is adequate for domestic amplifiers. The distortion will depend on the degree of matching between pairs but it is normal to obtain a distortion below 2% without negative feedback by the selection of two out of any three valves since this procedure, at worst, halves the normal variation in characteristics.

Due to the high mutual conductance of the KT88 some precautions have to be taken against parasitic oscillation whatever the circuit arrangement. Grid and screen series resistors of about 10,000 ohms and 270 ohms respectively, are recommended. In the ultra-linear circuit, resistance-capacitance networks may be found necessary between the anode and screen taps on the output transformer as shown in Fig. 4. With some transformers they will not be needed and they become less necessary as the taps include a greater part of the primary due to the consequent reduction in leakage inductance.

Details are also given for Class B operation of triode-connected KT88 valves for an output of up to 150 watts. This type of operation is useful when a distortion higher than that given by the ultra-linear circuit can be tolerated.

When an output exceeding 150 watts is needed, the KT88 valve may be used in multiple pairs in parallel push-pull instead of a single pair of larger valves. One of the advantages of this method is the low cost of the power supply which is required to give a high tension supply of only 550 volts. Another advantage is that a valve failure in the output stage merely reduces the available output power with a probable increase in distortion.

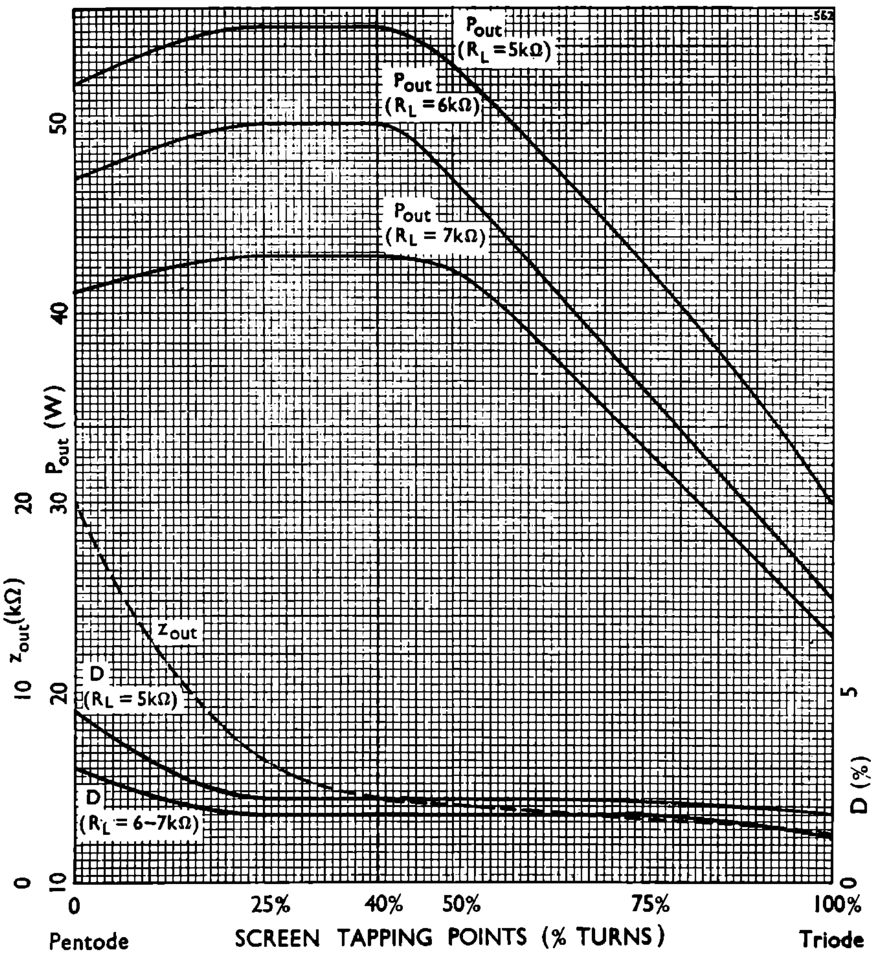


Fig. 1.—Output power, distortion and output impedance of KT88 ultra-linear output stage at various positions of the screen taps from 1% to 100% of each half-primary from the centre tap.

INSTALLATION

The bulb temperature rating of 250°C must not be exceeded and any cabinet used must provide adequate ventilation. Under free air conditions this temperature is not reached at maximum ratings but where a valve is enclosed the use of the temperature sensitive paint "Tempilaq"* is recommended to check that adequate air circulation exists, or other similar measure.

Although the KT88 may be mounted in any desired position it is recommended that, when valves are mounted horizontally, pins 4 and 8 should be in a vertical line as in Fig. 2.

When two valves are mounted vertically, in the more usual manner, on a horizontal chassis, there is a slight advantage obtained if pins 4 and 8 of both valves are mounted in line, as in Fig. 3. The coolest part of the bulbs then face each other whilst the hottest parts (nearest the anodes) are at right-angles.

* Obtainable from J. M. Steel & Co. Ltd., 36, Kingsway, London, W.C.2.

POWER SUPPLY

The type of power supply required will depend on the operating conditions but a capacitance input filter circuit is satisfactory for the ultra-linear cathode bias amplifiers.

With fixed bias, the large change in anode current requires a low impedance power supply and an inductance input filter is essential. It is desirable for the smoothing capacitor to be of high value to prevent an instantaneous fall in voltage upon the occurrence of a transient. Satisfactory performance will be obtained with a single inductor and a capacitance of 50-150μf. Two 160μf, 450 volt electrolytic capacitors in series have been found very successful with the ultra-linear higher voltage working conditions.

The choice of a rectifier will depend on the power output required. The U52 or U54 are suitable for powers up to 50 watts, but for amplifiers designed to cover the range 50-100 watts, two U19 or two GXU50 rectifiers should be used.

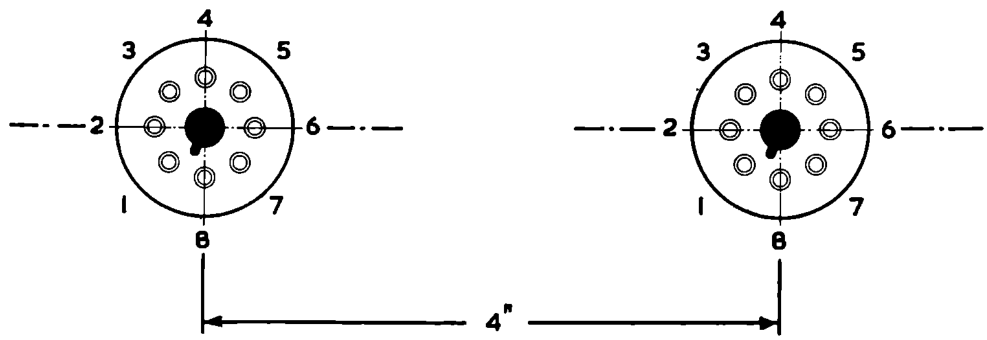


Fig. 2.—Correct orientation and spacing of the valve sockets when the valves are mounted horizontally.

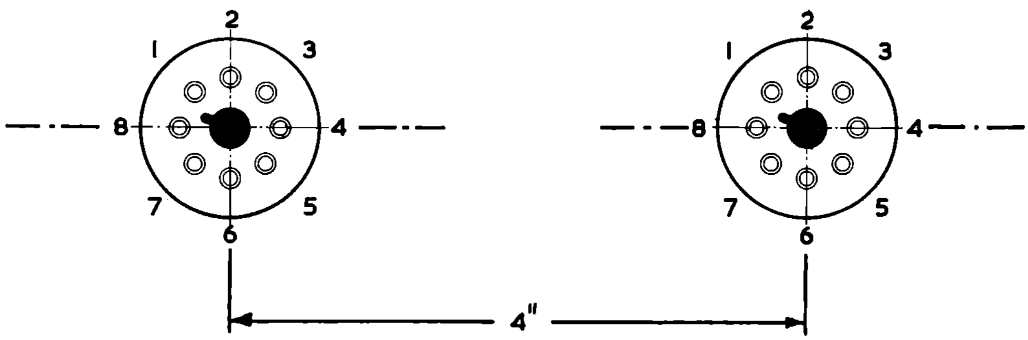


Fig. 3.—Correct orientation and spacing of the valve sockets when the valves are mounted vertically.

Alternatively the AWV 5AS4 may be used in lieu of the U52 or U54, whilst two 5AS4's may be used in the 50-100 watt range in lieu of the two U19's. Note that the U19 has a 4 volt filament.

A 30W ULTRA-LINEAR AMPLIFIER

With negative feedback, this amplifier (Fig. 4) will give 32 watts output with about 0.25% distortion at an anode potential of 335 volts. The input signal to the first stage of the amplifier for full output is 500 millivolts. Even when negative feedback is omitted the amplifier has only 1% distortion at 32 watts output and requires an input of 100 millivolts. As the KT88 valves are conservatively run in this circuit they will have a long life.

CIRCUIT DESCRIPTION

The output stage is preceded by a conventional double triode voltage amplifier V2 which is fed by a triode phase-splitter comprising one half of a further double triode V1. The first half of this valve is the input stage voltage amplifier, which is directly coupled to the phase-splitter. R8, in the cathode circuit of V2, reduces any signal unbalance in the phase-splitter and lowers intermodulation distortion.

Depending upon the particular characteristics of the output transformer, it may be necessary to prevent instability by connecting capacitors and resistors across part of each output transformer half-primary and "stopper" resistors in series with the grids and screens of the output stage. The curves of Fig. 5 illustrate the performance of the output stage of this amplifier without feedback.

The negative feedback network shown in Fig. 4 provides 14db of feedback and this is adequate for all normal purposes. This value reduces the output impedance, distortion and sensitivity of the basic amplifier by a factor of 5.

Feedback from the output transformer secondary is introduced into the cathode circuit of the first stage via R2. Since the sensitivity of the amplifier without feedback is approximately 100 millivolts, a feedback voltage of about 500 millivolts is required for 14db feedback. As the voltage across the output transformer secondary for 32 watts output is about 21.5 for a 15 ohm load and about 11 for a load of 4 ohms, the resistors R2 and R4 are chosen so that 500 millivolts will exist at their junction at full output. Assuming R4 to be 22 ohms, R2 is given by $225\sqrt{Z_0}$ (where Z_0 = the loudspeaker impedance) and the nearest standard value may be used. If $Z_0 = 15$ ohms, R2 should be 1000 ohms and if $Z_0 = 4$ ohms, R2 should be 470 ohms.

The operating conditions for the output stage of the amplifier of Fig. 4 are given in Table 1.

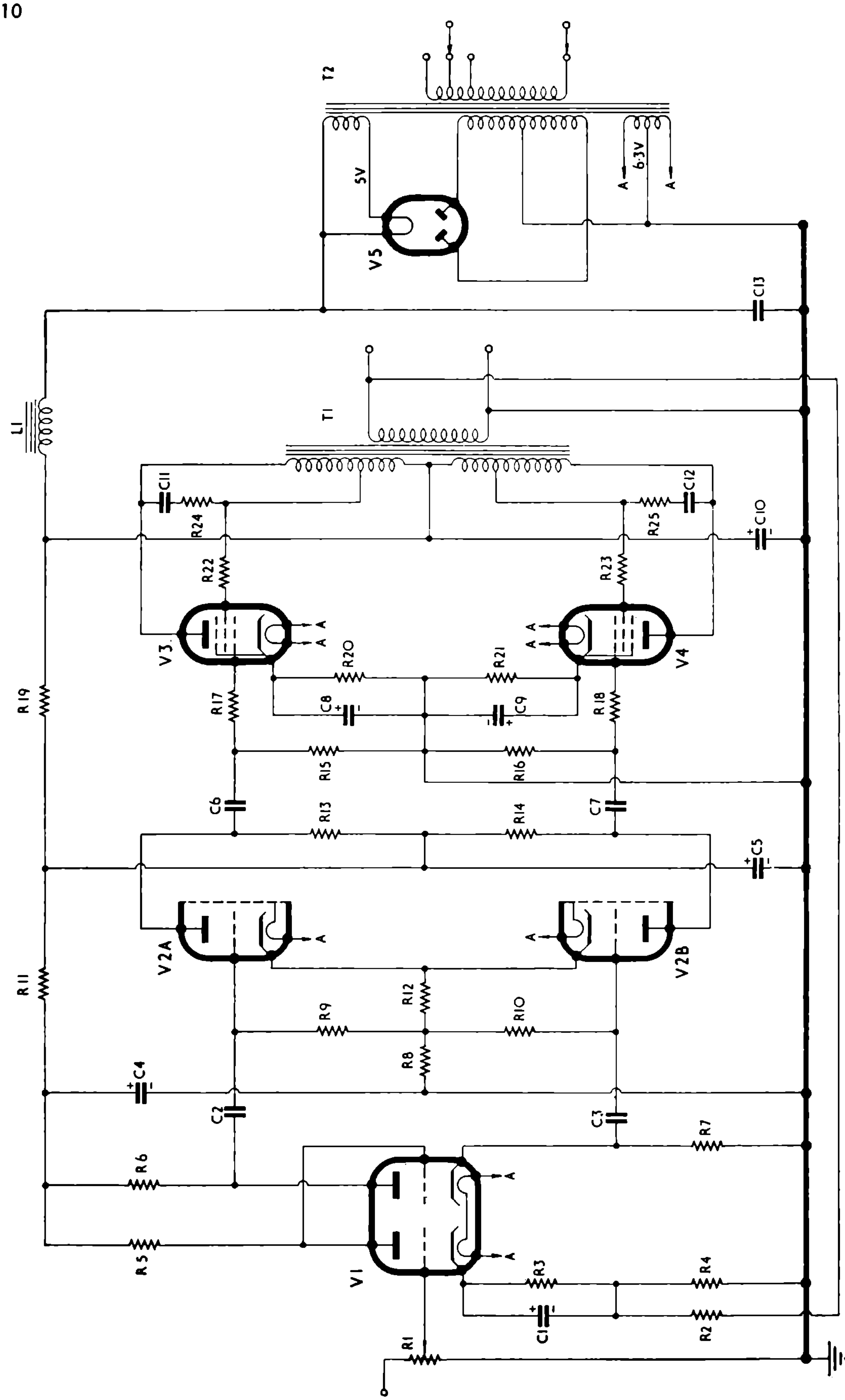


Fig. 4.—Circuit diagram of the KT88 30 watt ultra-linear amplifier.

TABLE I

30W ULTRA-LINEAR AMPLIFIER

OPERATING CONDITIONS

Plate Supply Voltage	375	v
Plate, Screen Voltage	335	v
Plate, Screen Current (zero sig.)	2 x 80	ma
Plate, Screen Current (max. sig.)	2 x 85	ma
Plate, Screen Dissipation (zero sig.)	2 x 27	w
Plate, Screen Dissipation (max. sig.)	2 x 12	w
Cathode Resistor	2 x 400	Ω
Grid Bias (approx.)	–32	v
Power Output	30	w
Plate Load (p - p)	5	KΩ
Output Impedance	1	KΩ
Distortion	0.25	%
Signal Input	500	mv
If negative feedback is omitted, the last three values become:—		
Output Impedance	4.5	KΩ
Distortion	1	%
Signal Input	100	mv

COMPONENT VALUES FOR THE 30W AMPLIFIER OF FIG. 4

VALVES

V1 B65 or 6SN7GT.
V2 B65 or 6SN7GT.
V3 KT88.
V4 KT88.
V5 U54 or 5AS4.

RESISTORS

All resistors 20%, 0.25 watt unless otherwise shown.

R1 1MΩ logarithmic.
R2 225 $\sqrt{\text{speech coil impedance}}$.
R3 1 KΩ.
R4 22Ω.
R5 100 KΩ.
R6 15 KΩ, 0.5 w } Matched 5%.
R7 15 KΩ, 0.5 w }
R8 4.7 KΩ.
R9 470 KΩ, 10%.
R10 470 KΩ, 10%.
R11 1 KΩ.
R12 1 KΩ.
R13 33 KΩ, 10%, 1 w.
R14 33 KΩ, 10%, 1 w.
R15 220 KΩ.
R16 220 KΩ.
R17 10 KΩ.
R18 10 KΩ.
R19 4.7 KΩ, 1 w.
R20 400 Ω, 5%, 5 w.
R21 400 Ω, 5%, 5 w.
R22 270 Ω, 0.5 w.
R23 270 Ω, 0.5 w.
R24 470 — 1500 Ω, 0.5 w.
R25 470 — 1500 Ω, 0.5 w.

CAPACITORS

C1 50 μf, 12 vw.
C2 0.05 μf.
C3 0.05 μf.
C4 8 μf, 350 vw.
C5 8 μf, 450 vw.
C6 0.05 μf.
C7 0.05 μf.
C8 50 μf, 50 vw.
C9 50 μf, 50 vw.
C10 8 μf, 500 vw.
C11 1000 μμf.
C12 1000 μμf.
C13 8 μf, 500 vw.

MISCELLANEOUS

L1 10 h, 200 ma.
T1 35 watt UL transformer,
6KΩ plate-plate, primary inductance not less than 50h.†
Leakage inductances:—
Prim-Sec:—not greater than 10mh.†
½ Prim-UL tap:—not greater than 10 mh.†
T2 Mains transformer,
375-0-375 volts, 200 ma.
6.3 volts, 5a centre tapped.
5 volts, 3a.

† With these values, R24, R25, C11 and C12 may be omitted.

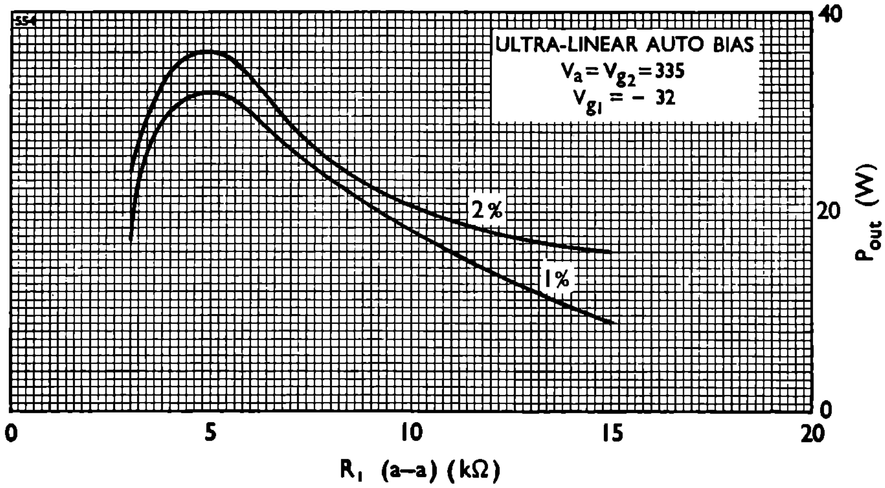


Fig. 5.—Performance of the KT88 30 watt ultra-linear amplifier of Fig. 4.

A 50W ULTRA-LINEAR AMPLIFIER

This section describes the use of the KT88 in a design for an ultra-linear amplifier with cathode bias giving 50 watts output at 0.2% distortion.

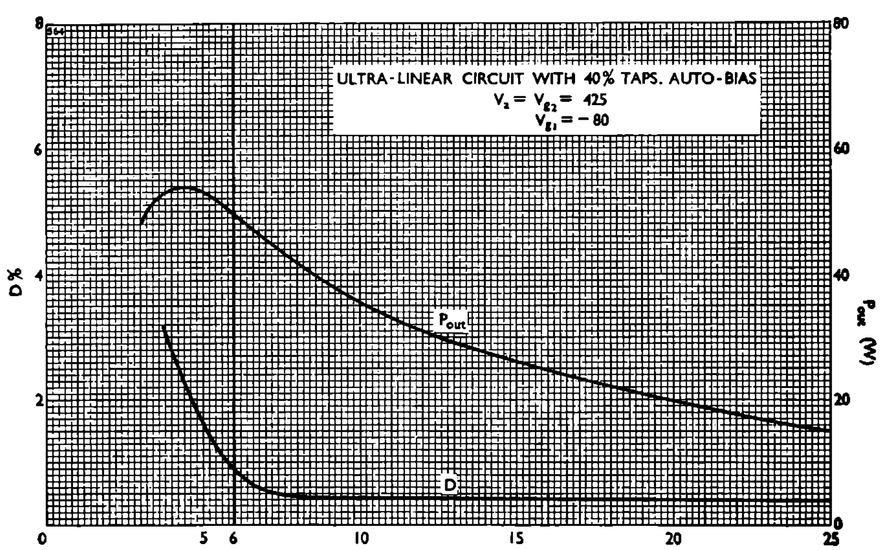
CIRCUIT DESCRIPTION

The circuit of the complete amplifier is given in Fig. 7. The design is similar to that of the 30 watt amplifier except for the type of input stage and the inclusion of the networks consisting of C6, R10 and C7, R11 between the first two stages. These networks reduce the amount of overshoot and consequent "ringing" in the output transformer.

The first double triode, V1, is arranged as a self-balancing floating paraphase phase inverter which feeds V2, the following push-pull voltage amplifier. Potentiometer R22 allows the signal input to the output stage to be adjusted for output stage dynamic balance.

The power supply incorporates a thermistor in the high tension output line in order to reduce the surge from the directly heated rectifier while the remaining valves are warming up. The performance of this amplifier is illustrated in Fig. 6 and the output stage operating data are given in Table II.

Fig. 6.—Performance of the KT88 50 watt ultra-linear amplifier of Fig. 7.



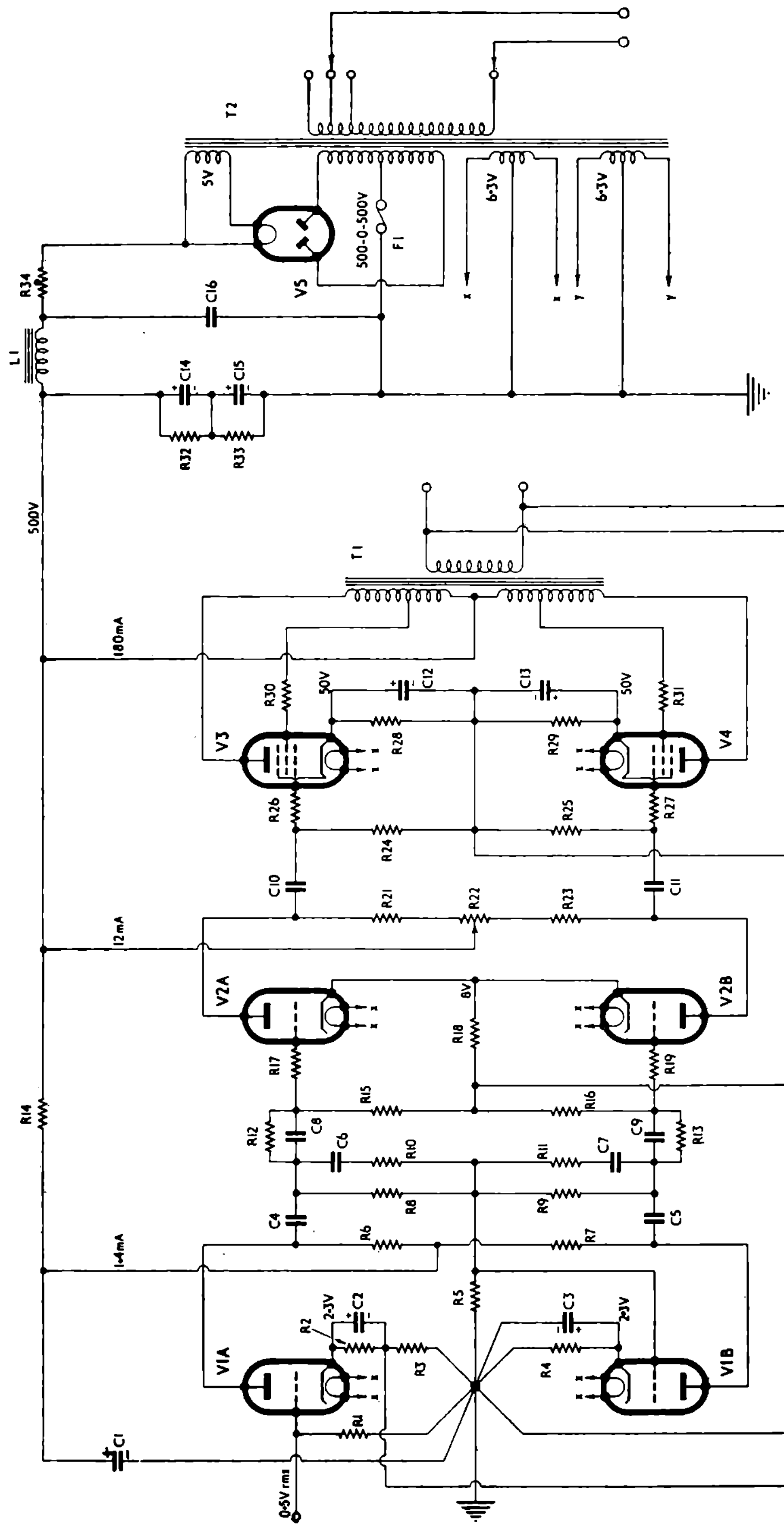


Fig. 7.—Circuit diagram of the KT88 50 watt ultra-linear amplifier.

TABLE II

50W ULTRA-LINEAR AMPLIFIER

OPERATING CONDITIONS

Plate Supply Voltage	500	v
Plate, Screen Voltage	425	v
Plate, Screen Current (zero sig.)	2 x 87	ma
Plate, Screen Current (max. sig.)	2 x 100	ma
Plate, Screen Dissipation (zero sig.)	2 x 40	w
Plate, Screen Dissipation (max. sig.)	2 x 18	w
Cathode Resistor	2 x 525	Ω
Grid Bias (approx.)	–50	v
Power Output †	50	w
Plate Load (p - p)	5	KΩ
Distortion †	0.2	%
Signal Input	500	mv

† These figures refer to an average pair.

COMPONENT VALUES FOR THE 50W AMPLIFIER OF FIG. 7

VALVES

V1 B339 or 12AX7.
V2 B329 or 12AU7.
V3 KT88.
V4 KT88.
V5 U52, U54 or 5AS4.

RESISTORS

All resistors 20%, 0.5 watt unless otherwise shown.

R1 1 MΩ.
R2 3.3 KΩ.
R3 100 Ω.
R4 3.3 KΩ.
R5 1 MΩ.
R6 220 KΩ, 10%.
R7 220 KΩ, 10%.
R8 1 MΩ.
R9 1 MΩ.
R10 10 KΩ.
R11 10 KΩ.
R12 4.7 MΩ.
R13 4.7 MΩ.
R14 100 KΩ.
R15 470 KΩ.
R16 470 KΩ.
R17 10 KΩ.
R18 680 Ω.
R19 10 KΩ.
R20 4.7 KΩ.
R21 33 KΩ, 2 w.
R22†25 KΩ, 4 w.
R23 33 KΩ, 2 w.
R24 220 KΩ.
R25 220 KΩ.
R26 10 KΩ.
R27 10 KΩ.
R28 525 Ω, 5%, 6 w.
R29 525 Ω, 5%, 6 w.
R30 100 Ω.
R31 100 Ω.

† Optional potentiometer for dynamically balancing the output stage. If omitted, R21 and R23 should be 47 KΩ, 2%, 2 w.

CAPACITORS

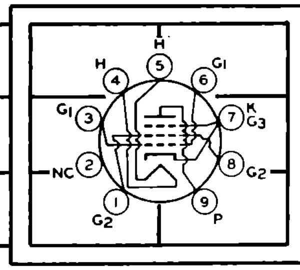
C1 8 μf, 500 vw.
C2 50 μf, 12 vw.
C3 50 μf, 12 vw.
C4 0.25 μf.
C5 0.25 μf.
C6 470 μμf.
C7 470 μμf.
C8 0.005 μf.
C9 0.005 μf.
C10 0.5 μf.
C11 0.5 μf.
C12 50 μf, 100 vw.
C13 50 μf, 100 vw.
C14 24 μf, 350 vw.
C15 24 μf, 350 vw.
C16 4 μf, 750 vw.
C17 8 μf, 500 vw.

MISCELLANEOUS

L1 5 h, 250 ma.
T1 50 watt UL transformer, 5 KΩ plate-plate, primary inductance not less than 30 h.
Leakage inductances:—
Prim-Sec:—not greater than 10 mh.
½ Prim-UL tap:—not greater than 10 mh.
T2 Mains transformer,
500-0-500 volts, 250 ma.
6.3 volts, 5 a centre tapped.
6.3 volts, 1-2 a centre tapped.
5 volts, 3 a.
F1 1 a delayed fuse.

(This article will be continued next month)

The Design-Maximum Rating System



NEED FOR A RATING SYSTEM

The conditions under which a valve may be operated are limited by fundamental capabilities of the valve itself. Physical limitations exist, for example, in the permissible temperatures at which the various electrodes may be operated, in the amount of current which can be emitted by the cathode, and in the voltage gradients which may be permitted between the various valve elements.

Maximum valve ratings have been established to define these various physical limitations of the valve in terms of readily measurable quantities. The numerical quantities presented as maximum ratings indicate the limiting operating values required to assure satisfactory valve life and performance.

Before the value of any rating can become meaningful, the rating system on which the rating is based must be specified. The system must define the interpretation required of the numerical values and indicate the procedure necessary to determine whether or not a valve is operating within its rating.

Until the present time, two rating systems have been commonly used in conjunction with receiving valves, the Design-Centre System and the Absolute-Maximum System. The problems created by their deficiencies have greatly intensified in recent years primarily as a result of the greatly increased scope of receiving valve applications. To overcome the deficiencies encountered with the two rating systems in current use, a new system for rating valves, designated the "Design-Maximum System" has been developed.

DEFINITIONS OF EXISTING SYSTEMS

The following definitions have been standardized by the industry to describe the two systems which have been in common use:

DESIGN-CENTRE RATING SYSTEM

"Design-Centre ratings are limiting values of operating and environmental conditions applicable to a bogey electron device of a specified type as defined by its published data, and should not be exceeded under normal conditions.

The device manufacturer chooses those values to provide acceptable serviceability of the device in average applications, taking responsibility for normal changes in operating conditions due to rated supply voltage variation, equipment component variation, equipment control adjustment, load variation, signal variation, environmental conditions, and variations in device characteristics.

The equipment manufacturer should design so that initially no design-centre value for the intended service is exceeded with a bogey device in equipment operating at the stated normal supply voltage.

ABSOLUTE-MAXIMUM RATING SYSTEM

Absolute-Maximum ratings are limiting values of operating and environmental conditions applicable to any electron device of a specified type as defined by its published data, and should not be exceeded under the worst probable conditions.

The device manufacturer chooses these values to provide acceptable serviceability of the device, taking no responsibility for equipment variations, environmental variations, and the effects of changes in operating conditions due to variations in device characteristics.

The equipment manufacturer should design so that initially and throughout life no absolute-maximum value for the intended service is exceeded with any device under the worst probable operating conditions with respect to supply-voltage variation, equipment component variation, equipment control adjustment, load variation, signal variation, environmental conditions, and variations in device characteristics.

To illustrate the meaning of these definitions, let us consider an electronic circuit in which it is desired to determine whether the valve employed is operated within the maximum ratings specified for it.

If these ratings are presented as design-centre ratings, the circuit must be arranged so that it is operating under normal conditions. That is, the supply voltage is adjusted to its normal value, all components employed are selected as average values, the equipment controls are adjusted for normal settings, and the valve employed is selected as a bogey valve. Under these average or most typical conditions, the circuit voltages, currents, and dissipations are measured in turn under the worst signal conditions for each particular rating and compared to the specified design - centre ratings. If each measured value is less than the corresponding rating and if the equipment is not subjected to supply voltages in excess of the stated variations, the operation of the valve satisfies the conditions of the design-centre system.

On the other hand, if the ratings are presented as absolute-maximum values, the worst probable operating conditions must be established in turn for each item rated. The measurements made with the combination of extreme supply voltage, limit components, extreme control settings, extreme signal, extreme environmental conditions, and any valve which produces the worst probable value for the particular rating under consideration, is compared to the specified absolute-maximum rating. Under these adverse conditions, if each measured value is less than the corresponding rating, the operation of the valve satisfies the conditions of the absolute-maximum system.

These two rating systems differ significantly in the four following aspects:

1. The operating conditions employed in in determining whether the valve is operated within maximum ratings.
2. The characteristics of the valve employed in determining whether it is operated within maximum ratings.
3. The permissible excess of the specified values of the maximum ratings. (Under the absolute-maximum system no excess of the rated value is permitted, while under the design-centre system the rated value may be exceeded with adverse operating conditions).
4. The assignment of basic responsibility for proper valve usage. (With the design-centre system, the valve manufacturer effectively assumes the responsibility for the effect that both variations in valve characteristics and circuit operating conditions will have on valve performance. On the other hand, under the conditions imposed by the absolute-maximum system, the complete responsibility for variations in valve characteristics and operating conditions is assigned to the circuit designer.)

DIFFICULTIES WITH EXISTING SYSTEMS

Although the design-centre system provides the circuit designer with an extremely convenient and usable system, it has the rather severe disadvantage that the degree of protection afforded the valve is variable and depends on the circuit and environmental operating conditions.

Historically this limitation was not particularly significant. The great majority of valves produced were used in radio receivers which involved essentially standard circuits. Hence the valve manufacturer could be reasonably certain of the effects that a specified line voltage variation would have on the currents, voltages, and dissipations associated with the various valves employed.

In recent years, however, electronic circuits have become more complex and more diversified. The circuits employed operate under widely different degrees of feedback and supply voltage regulation. As a result a fixed relationship no longer exists between the variations in supply voltage and variations to which the valve is subjected. Two extreme examples in this consideration would be the computer application and the television high-voltage rectifier application.

In the computer, particular attention is paid to the regulation of heater and plate voltages. Hence, normal fluctuations in the suply have little if any effect on valve operating conditions. In the case of the TV flyback high-voltage rectifier, the entire input power including the filament requirements is derived from the horizontal deflection amplifier. All variations in the horizontal-amplifier circuit which can result from variations in sweep valve characteristics, damper valve characteristics, output transformer and other components, and supply voltage, are imposed as variations in the filament power and inverse voltage of the high-voltage rectifier valve. Furthermore the output current required is established by settings of the brightness and contrast controls which in turn depend on the ambient light level of the room in which the receiver operates. All of these various sources contribute to the extreme variations associated with high-voltage rectifier operation. These two examples illustrate the fact that variations in operating conditions to which a valve can be subjected cannot necessarily be controlled by specifying a permissable variation in supply voltage.

Figure 1 illustrates qualitatively the variation of plate dissipation to which a bogey valve could be subjected in three different circuits, each of which is designed for operation at 100 per cent. of the maximum-design-centre dissipation rating. In each case, the circuit is subjected to ± 10 per cent. line voltage variations. The figure illustrates that the degree by which a design-centre rating can be exceeded is a function of the circuit and equipment in which the valve is operated.

These considerations indicate that there are at least two disadvantages to the design-centre system:

1. The valve manufacturer is not adequately protected. His valves can be operated

within the conditions of the design-centre system and yet be subjected to severe operating conditions as typified by Condition C of Figure 1. The design-centre system could adequately protect the valve only if standard circuits were used in which known operating variations are encountered.

2. The conservative circuit designer is penalized under the conditions of the design-centre system. The system requires only consideration of average operating conditions; no requirements are placed on permissable variation under adverse operating conditions. Consequently no incentive is provided for circuit design which tends to minimize variations applied to the valve. Condition C is as correct as Condition A of Figure 1 according to the rules of the design-centre system, yet certainly more satisfactory valve performance would be realized in the latter circuit.

Although the absolute-maximum system is not subject to the shortcomings of the design-centre system, it suffers from being extremely difficult to use. Effectively the circuit designer is given the responsibility to keep any valve of the type under consideration from exceeding the absolute-maximum values under worst probable circumstances.

To use the absolute-maximum system the circuit designer must have full knowledge of the tolerances associated with all the variables. He must know that complete range of values which can be encountered with valves, resistors, ambient temperatures, supply voltages, signal levels, and control settings. To evaluate his valve usage, he must then synthesize the variables in combinations which give rise to the worst probable conditions for each of the rated items.

The problem of selecting components which will combine to produce the worst probable operating conditions is extremely complex. Even in a simple triode circuit, for example, the high limit valve for one plate current is not necessarily the valve which produces the highest plate dissipation. When consideration of the plate dissipation rating is extended to a pentode circuit which incorporates a plate resistor, a screen-dropping resistor, and a cathode-bias resistor, the selection of the necessary valve must be made from valves which could exhibit maximum plate current/maximum screen current, maximum plate current/minimum screen current, minimum plate current/maximum screen current, minimum plate current/minimum screen current, or a combination in which the currents lie between the limits. The selection cannot be generalized and depends upon the specific application. Furthermore, the particular valve and other components required to evaluate the plate dissipation rating are generally not the same as those required to evaluate the other maximum ratings.

Thus the inherent complexity of selecting and synthesizing limit components and limit valves, coupled with the practical difficulty in obtaining the required limit valves, represents the major weakness of the absolute-maximum system, particularly for home or automobile applications without well-regulated power supplies. As might be expected, the time required to check the conformance of each valve to its specified absolute-maximum ratings in a relatively complicated piece of electronic equipment becomes excessive. Indeed, the practical result is that all too often the evaluation of the valve usage is not rigorously executed. Unfortunately inadequate evaluation can result in the misuse of valves and the consequent sacrifice in overall equipment performance and reliability.

PRINCIPLES OF THE DESIGN-MAXIMUM SYSTEM

Because of the serious difficulties encountered in the application of the two rating systems in current use, the need for an improved method to express the capabilities of receiving valves has become increasingly apparent.

Basically the design-centre system assigns the entire responsibility for valve usage to the valve manufacturer; he must accept full responsibility for the effects of variations in valve characteristics and variations in any possible circuit operating conditions. The absolute-maximum system assigns this entire responsibility to the circuit designer. A more logical division of these basic responsibilities would be to assign the effect of the variations in valve characteristics to the valve manufacturer and to assign the effects of variations in circuit operating conditions to the equipment manufacturer. This philosophy is incorporated in the Design-Maximum System.

The design-maximum system requires that the maximum ratings be checked with bogey valves rather than limit valves. In this way, the responsibility for variations in valve characteristics is effectively assigned to the valve manufacturer. The task of the circuit designer in determining conformance to the maximum ratings is greatly simplified, and the valve manufacturer is relieved of the almost impossible responsibility for supplying limit valves to the circuit designer for his maximum rating evaluation.

This system also requires that the specified maximum ratings not be exceeded when the circuit is so arranged that the bogey valve is operated under the worst probable conditions. In this way the responsibility for variations in operating conditions is effectively assigned to the circuit designer. He must realistically anticipate the worst probable operating conditions likely to be encountered for each valve in his equipment.

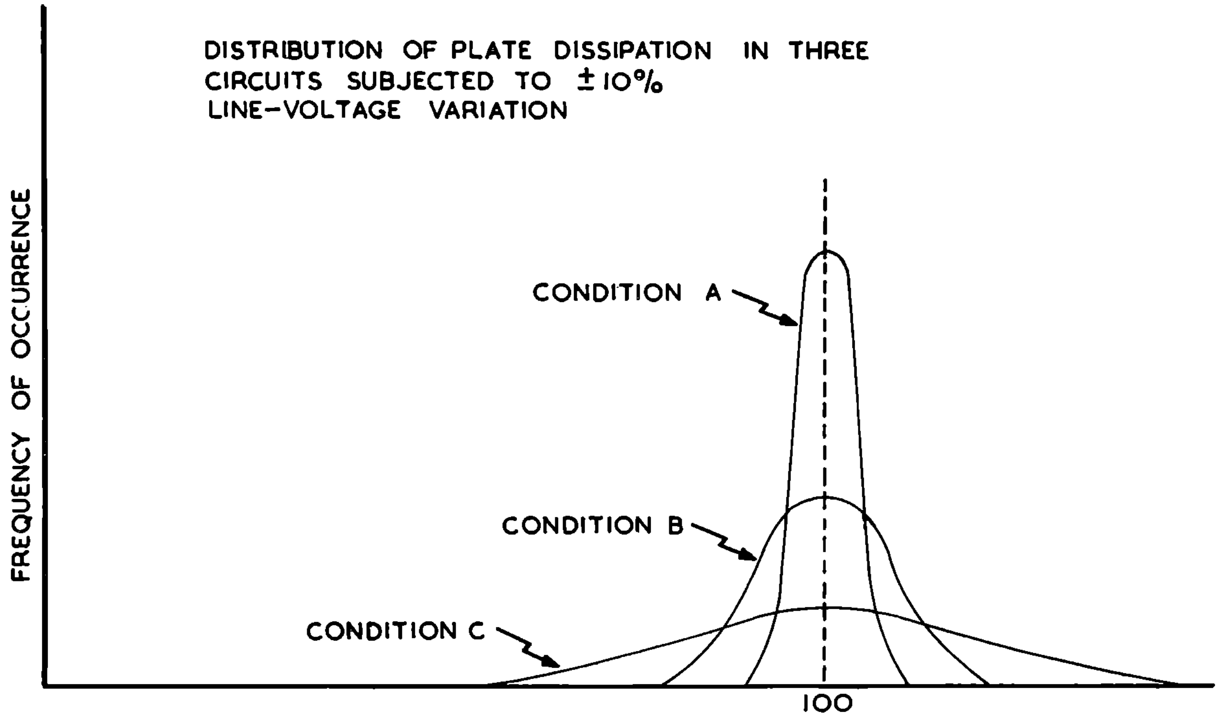


FIG. 1 — PLATE DISSIPATION AS PERCENTAGE OF DESIGN-CENTRE RATING.

In a television receiver, for example, it is possible—particularly at high line voltage—to exceed some of the ratings in certain stages by very great amounts when the controls are completely maladjusted. However, if the maladjustment of controls makes the picture unusable, the probability of this condition occurring for a prolonged period is remote as the receiver would normally be turned off. The worst probable operating conditions resulting from control setting would occur when the controls were maladjusted only to the extent that a minimum usable picture could still be obtained.

Another example of worst probable conditions occurs in the case of rectifier output current. In this case, particularly if the load consists of an appreciable number of valves, the difference in output current required between the probable high load and the theoretical maximum load is very large. Again the significant value from the standpoint of proper usage of the rectifier valve is the worst probable load current.

Therefore, under the design-maximum system, the valve manufacturer in effect supplies the circuit designer with maximum limiting values based on bogey valves. Because some variation in circuit operating conditions is present in any practical circuit, the equipment manufacturer must design his equipment below the design-maximum values so that no one of these ratings will be exceeded under the worst probable operating conditions which will be encountered. The degree to which the circuit must be designed below the design-maximum ratings depends on the extent to which the variables are controlled in the equipment.

Inherently, the proposed design-maximum system requires that the circuit designer anticipate the adverse conditions to which his equipment will be subjected. The circuit designer is the person to whom this responsibility should be delegated. Also the equipment manufacturer must assume the risk whenever bogey valves are operated above the design-maximum ratings.

It must be recognized that in some rare instances a valve which is operated at its design-maximum ratings under the worst probable operating conditions could exceed the ratings under unanticipated or freak conditions. For optimum design, therefore, the circuit designer must discriminate on a statistical basis between unlikely operation and adverse operation which can reasonably be expected to occur. The degree of risk which the equipment manufacturer could reasonably assume in this connection would necessarily depend on the reliabilty requirements of the final equipment. How competently the circuit designer evaluates his valve usage will be reflected in the final overall performance of the equipment.

(Continued on page 139)

Some Aspects of Synchronization in TV Receivers

By J. VAN DER GOOT, M.I.R.E. (Aust.)

PART 3 — CONCLUSION

SYNC PULSE INVERSION, AMPLIFICATION, LIMITING AND CLIPPING

As will be seen later in discussing the sync clipper, the voltage waveform across the plate load of the conventional sync clipper is of negative polarity. Sync pulses at the input of directly synchronized vertical oscillators must be positive. Sync pulses at the input of the synchro-guide control valve must also be positive. The horizontal phase discriminator requires positive as well as negative pulses. Often a valve is used for the purpose of phase reversal or phase splitting. At the same time this valve is often used for clamping and limiting. Fig. 26 shows a circuit diagram of a sync amplifier/phase splitter which also clamps and limits.

"Clamping" can be defined by the following description. Negative going sync pulses are applied to the grid of V via the coupling capacitor C. R1, which is connected to a positive voltage source, keeps the potential of the grid (with respect to the cathode) at approximately zero between pulses, due to the flow of grid current. During the pulses the grid is driven neagtive (with respect to the cathode) as soon as the amplitude rises above a certain level determined by the values of R1, R2 and the voltage of the positive voltage source with respect to the cathode. This clamping action reduces any irregularities of small amplitude at the base of the sync pulse waveform such as video information.

"Limiting" can be defined by the following description. The plate voltage of V is chosen so that the sync pulses drive the grid of V past plate current cut off. This limits the height of the sync pulses as they appear in the output of V. The limiting action is similar to that of the sync clipper discussed in the next section. R3 and R4 form the split plate load of V. The sync pulses will appear across R3 and R4 as indicated. Any noise pulse with an amplitude greater than that of the sync pulses appears in the output reduced to the same height as the sync pulses.

A pulse transformer as shown in the phase discriminator circuit diagram, Fig. 19, is another means by which phase reversal can be obtained. This has the additional advantage that the transformer can be designed to differentiate the sync information at the same time. Fig. 27 shows an oscillogram of some horizontal pulses obtained from such a transformer. As mentioned earlier in this section the vertical sync pulses at the input to the vertical oscillator must be positive going. If a pulse transformer is used for phase reversal of the horizontal sync pulses another phase reversing element is required for the vertical sync pulses (usually a triode valve). Naturally this will make this type of circuit more complicated.

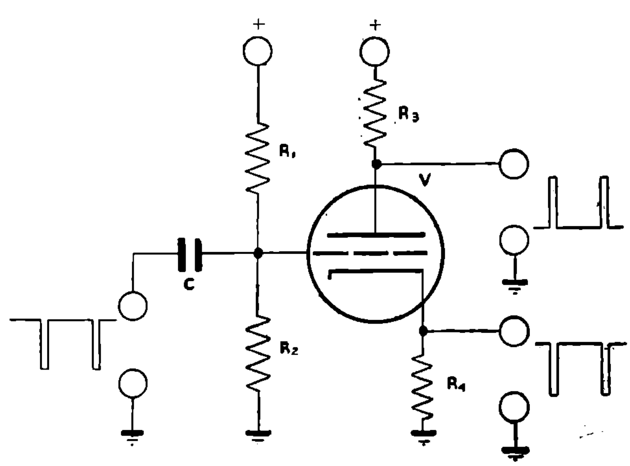


Figure 26.—Circuit diagram of a sync "splitter".

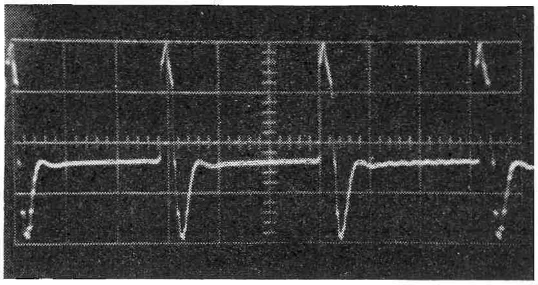


Figure 27.—Oscillogram of differentiated horizontal sync pulses obtained from a pulse transformer.
Sweep: 20 μ/sec/cm.

The Sync Clipper

There are several ways in which the video information can be removed from the composite video signal. The circuitry described here is of the grid clamping variety.

If a positive going composite video signal is applied to the input of the circuit of Fig. 28 (a) grid current will flow when the control grid of the valve V is driven positive, that is during the sync pulses. Grid current will charge C1. The charging time constant is mainly determined by the values of R2, C1 and the grid-cathode impedance of V. R2 is an isolating resistor which reduces high frequency shunting of the video amplifier load by the input capacitance of V. The higher the value of R2, the higher the high frequency shunting impedance will be. The maximum value of R2 is determined by the frequency response required at the input of V. The minimum frequency response required is discussed later in this section. The grid-cathode impedance during grid conduction is very low and small compared with R2. However, when the grid is not conducting the discharge time constant of C1 is determined by the values of R2, C1 and where R1 is large compared with R2. The result is that C1 is charged to a potential such that the sync tips drive the grid of V just positive, Fig. 28 (b). The energy contained in the hatched portions (grid conduction) is approximately equal to the energy dissipated in R1 between pulses.

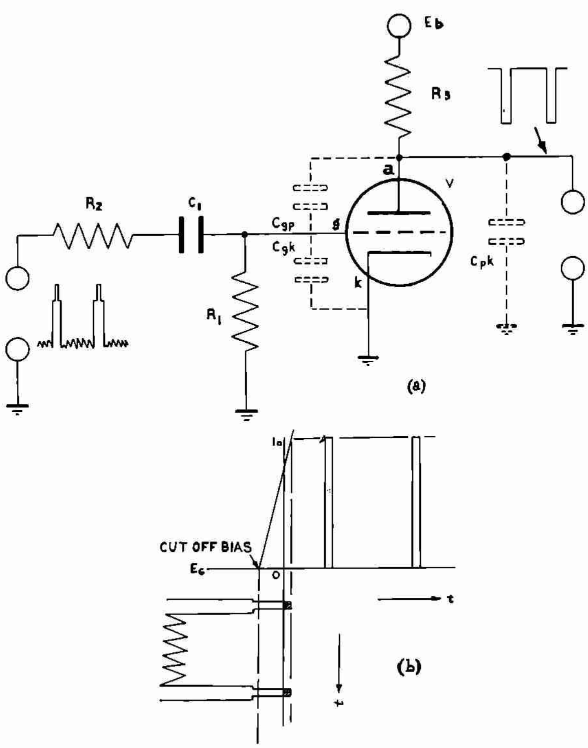


Figure 28.—(a) Circuit diagram of a sync clipper.
(b) Illustration of how plate current pulses are derived from the sync pulses in a positive-going composite video signal.

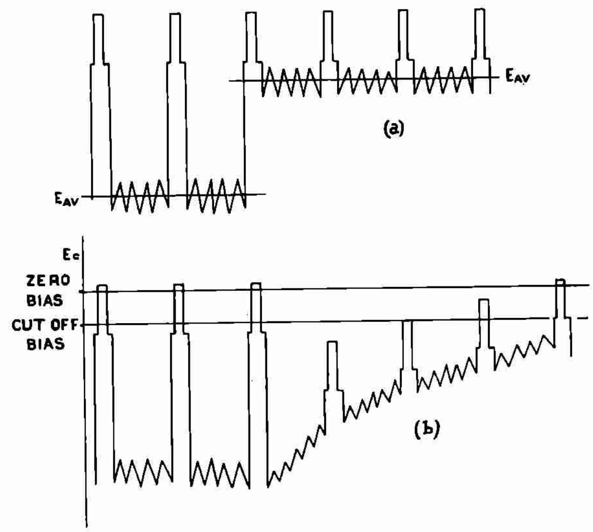


Figure 29.—Illustration of the effect of a sudden reduction in scene brightness when the grid discharges too slowly compared with the rate of change.
(a) Input voltage waveform.
(b) Resulting grid voltage waveform.

In Fig. 28 (b) an idealized transfer characteristic has been drawn. V is a sharp cut off valve. By applying a low plate voltage to V the plate current cut off grid potential is kept low so that the blanking level of the composite signal applied to the grid is always below cut off. The result is that the plate current is derived from the sync pulses only, Fig. 28 (b).

If the discharge time constant is comparable with the duration of several fields, say 50 msec, C1 will be charged with a voltage approximately equal to the average value of the composite voltage waveform taken over several fields. All the sync tips of the field will drive the grid of V

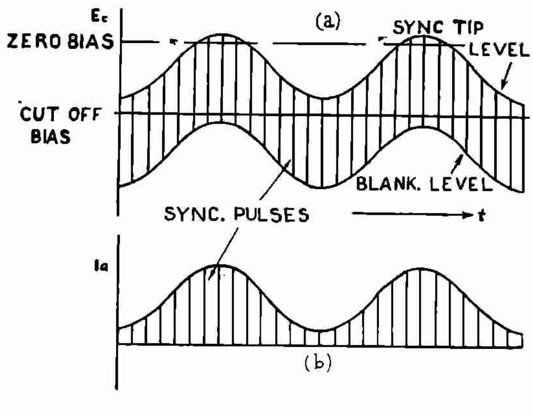


Figure 30.—Illustration of the effect of hum superimposed on the composite video waveform when the grid discharge time constant is long compared with the hum period. Only the sync pulses are shown.

(a) Waveform at the grid.

(b) Resulting plate current waveform.

just positive. When the average value of the composite signal changes, due to average scene brightness changes, the charge on C1 will change accordingly. If however the discharge time constant of C1 is made much longer, say 200 msec (1/5 second), a rapid reduction in average scene brightness will result in a temporary fall of the sync tip level at the grid of V due to the inability of the grid potential to follow the average voltage of the signal because of the long time constant, Fig. 29. The output of V will vary accordingly and

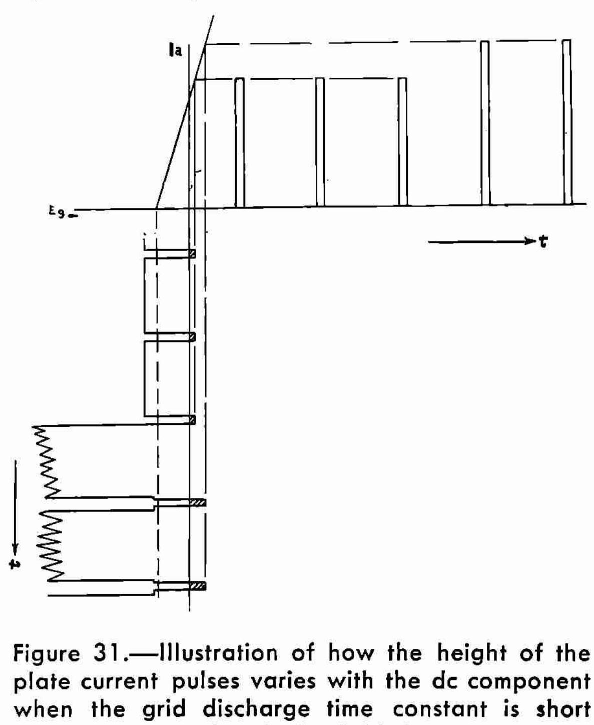


Figure 31.—Illustration of how the height of the plate current pulses varies with the dc component when the grid discharge time constant is short compared with the field duration.

temporary loss of sync may result. Another disadvantage of a long time constant is that any hum superimposed on the composite video waveform which is often encountered in practice will manifest itself as a reduction or even total loss of sync pulse output at the plate of V, Fig. 30. Accordingly, the frequency and phase of the horizontal oscillator will change and distortion of the picture will result.

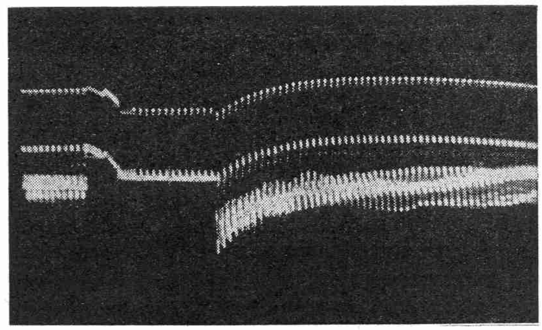


Figure 32.—Oscillogram of the sync clipper grid voltage waveform when the grid discharge time constant ≅ 100 μsec.
Sweep: 500 μsec/cm.

If the discharge time constant of C1 is made short, comparable with the duration of the field blanking, say 2 msec, the average grid potential will follow rapid scene brightness changes and respond to level changes such as those caused by hum. However, the average grid potential will also tend to follow the changes in energy content contained in the hatched portions of Fig. 28 (b). Fig. 31 illustrates this. The higher the average voltage of the composite video input signal the less the energy required to charge C1. Fig. 32 shows an actual oscillogram of the waveform at the grid of a sync clipper with very short discharge time constant of approximately 100 μsec.

During the field blanking period C1 is charged to a much lower potential than during the visible scan. The resultant plate current of V drawn in Fig. 31 shows how the height of the pulses varies. The variation in pulse height can be greatly reduced by the use of sync limiting as described in the previous section. However, the energy content of the pulses will also vary. The sync pulses as they appear in the composite signal at the grid of V in Fig. 28 are not rectangular. The frequency response at the grid of V is mainly determined by R2 and Cgk where Cgk consists of the grid-cathode capacitance of V and the wiring capacitance. Fig. 33 shows the shape of sync pulses at the grid of V if the frequency response is poor (time constant of the order of 2 μsec). Fig. 33 (a) shows the relative levels of the pulses during the field blanking and Fig. 33 (b) shows the relative level during the visible scan. Remembering how the pulse energy content affects the output voltage from the horizontal phase comparator, too short a discharge time constant in the grid circuit of the sync clipper will tend to cause a change in phase of the horizontal oscillator at the end of the field blanking period. Since it takes several line durations before the horizontal phase comparator and oscillator reach an equilibrium the change becomes visible in the picture. The effect on the picture is similar to that shown in Fig. 23. Fig. 34 is an oscillogram of the waveform at the grid of the synchro-guide showing the tips of the sync pulses only. This waveform was derived from the signal shown in Fig. 32.

The required frequency response or minimum time constant at the grid of V, Fig. 28 (a), is determined by the duration of the front porch, viz., 0.64 μsec. During the front porch the rise time should be such that the voltage can rise from white level to nearly blanking level. Assuming a rise from extreme white level to within 1% of blanking level in 0.64 μsec the time constant should be equal to or less than 0.14 μsec (see appendix). In Fig. 28 (a) the time constant R2 × Cgk should be less than 0.14 μsec.

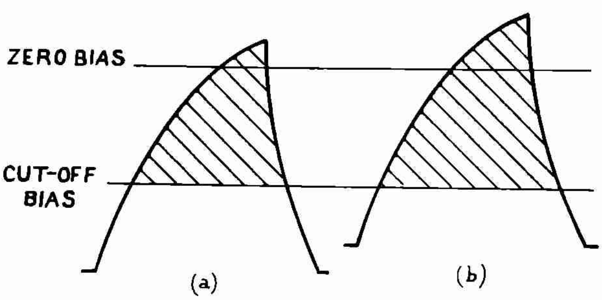


Figure 33.—Illustration of the change in energy content when the grid discharge time constant is short compared with the field duration and the frequency response at the grid is low.
(RC ≅ 2 μsec.)
(a) Pulse during field blanking.
(b) Pulse during active picture scan.

As was mentioned before the sync clipper valve should have a sharp cut off characteristic so as to enable clipping above blanking level. Then no video will appear at the plate of the clipper. There is, however, another path via which video information can appear at the plate of the sync clipper, viz., the grid-to-plate capacitance of the valve and the associated circuit capacitance. Sharp rises and falls in the video signal, black to white and white to black, will result in pulses at the plate of the sync clipper. In Fig. 28 (a) R2, Cgp, Cgk, Cpk and R3 form a differentiating network. C1 is much larger than Cgk, Cgp or Cpk. Increasing Cpk results in a smaller amplitude of video in the output. At the same time however it reduces the frequency response at the plate of V.

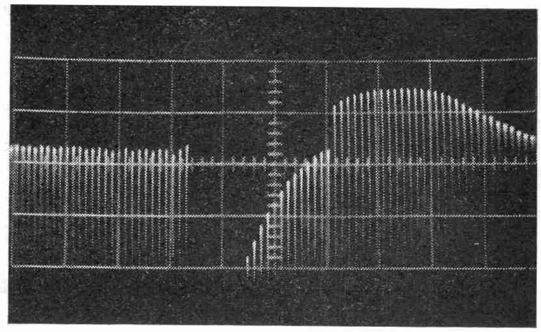


Figure 34.—Oscillogram, showing the tips of sync pulses at the grid of the synchro-guide control valve, derived from the waveform of Fig. 32.
Sweep: 500 μsec/cm.

Reducing R3 reduces both the sync pulse and video amplitude. It is therefore of utmost importance to keep the value of Cgp down. A pentode clipper will give some reduction of video content in the output. However, with a sharp cut off triode and a sensible lay-out the video content at the plate can be kept down low enough to permit a peak-to-peak composite video input at the grid of the sync clipper of a high as 120 volts without ill effects on the quality of the picture.

The effect of too high a video content is illustrated in Fig. 35 which shows video at the grid of the synchro-guide control valve. The result is that the output voltage from the synchro-guide is affected where the video pulses extend past the plate current cut off grid potential of the synchro-guide valve. Fig. 36 is an oscillogram of the tips of the sync pulses at the grid of the synchro-guide control valve showing the effect of video. Fig. 37 shows the resultant distortion of the picture. Generally an amplitude of video of 10% of the sync amplitude at the grid of the control valve can be tolerated.

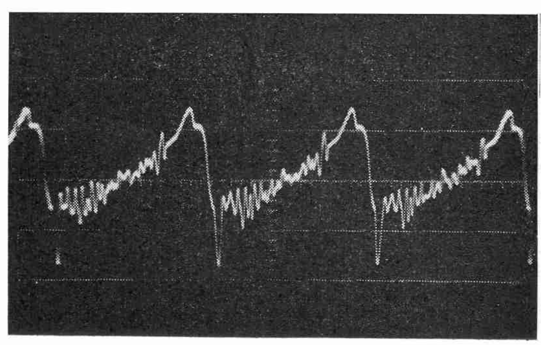


Figure 35.—Oscillogram showing video at the grid of the synchro-guide control valve.
Sweep: 20 μsec/cm.

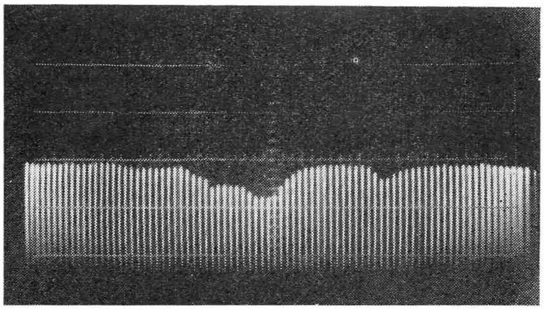


Figure 36.—Oscillogram of the tips of sync pulses at the grid of the synchro-guide control valve showing the effect of video at the grid. Sweep: 500 μsec/cm.

Sync Circuit

Fig. 38 shows the circuit diagram of a simple sync circuit which meets the requirements previously discussed. R1 is the plate load resistor of the video amplifier. The sync is taken off via an isolating resistor R2. The frequency response at the load side of the peaking coil is adequate for the sync circuitry, and sync take-off at this point has the advantage that the effective capacitive plate load of the video amplifier is less than if the sync were taken direct from the plate. C1 and R3 provide coupling to the clipper, one half of a 6CG7. R4 in parallel with C2 and the grid impedance of the clipper form a high-pass filter during grid conduction. This filter attenuates noise of comparatively long duration.

R5 and R6 provide a plate voltage source for the clipper of approximately one-eighth of the B+ voltage (35V for a B+ voltage of 275V). The equivalent plate load resistance is approximately 30,000 ohms. With a plate voltage of 35 volts the 6CG7 has a plate current cut off grid voltage of approximately —2.5 volts. A composite video signal of 12 volts peak-to-peak at the plate of the video amplifier will have a sync amplitude of approximately 3 volts. At the grid of the clipper the blanking level of such a signal will be slightly more negative than cut off and no video will be clipped.

C5 and R9 with the effective plate load resistance of the clipper form a differentiating network with a time constant of approximately 1.0 μsec. The differentiated pulses appear across R9. With C3 as a dc blocking capacitor, R7, C4 and the clipper plate load form an integrating network with a time constant of approximately 50μsec. The integrated pulses appear across C4. Capacitor C6 and R8 with R10 couple both differentiated and integrated pulses to the grid of the sync amplifier/phase inverter, the second half of the 6CG7.

The plate load of the amplifier is in the form of a high-pass filter, R12 and C7 in parallel and R11. The differentiated sync pulses which appear across R11 are coupled to the synchro-guide control valve via C8. The signal at the plate of the amplifier, where the amplitude of the differentiated pulses are substantially smaller than the amplitude of the vertical pulses, is passed through a low pass filter, consisting of R14 and C10 in parallel and R13 via a dc blocking capacitor, C9. The vertical sync pulses appear across R14 and C10 which are in series with the grid resistor of the vertical blocking oscillator.

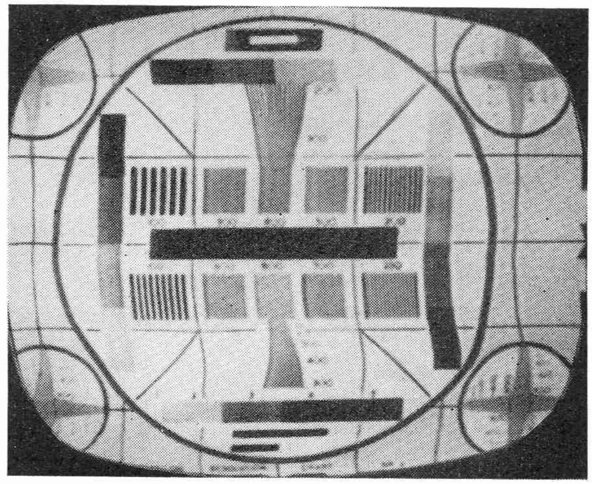


Figure 37.—Photograph of picture tube showing the effect on the picture of excessive video at the grid of the synchro-guide control valve.

Appendix

All calculations made for this article refer to the 1st issue of the ABCB standards which were in force until 4th November, 1957. On the 4th November, 1957 revised standards were issued[3]. However, the changes were relatively small. The numerical examples presented in the article are not very much affected by the revision of the standards.

In the section on "Integration", Figs. 4, and 5 the voltage levels were calculated assuming a simple series resistance-capacitance network:

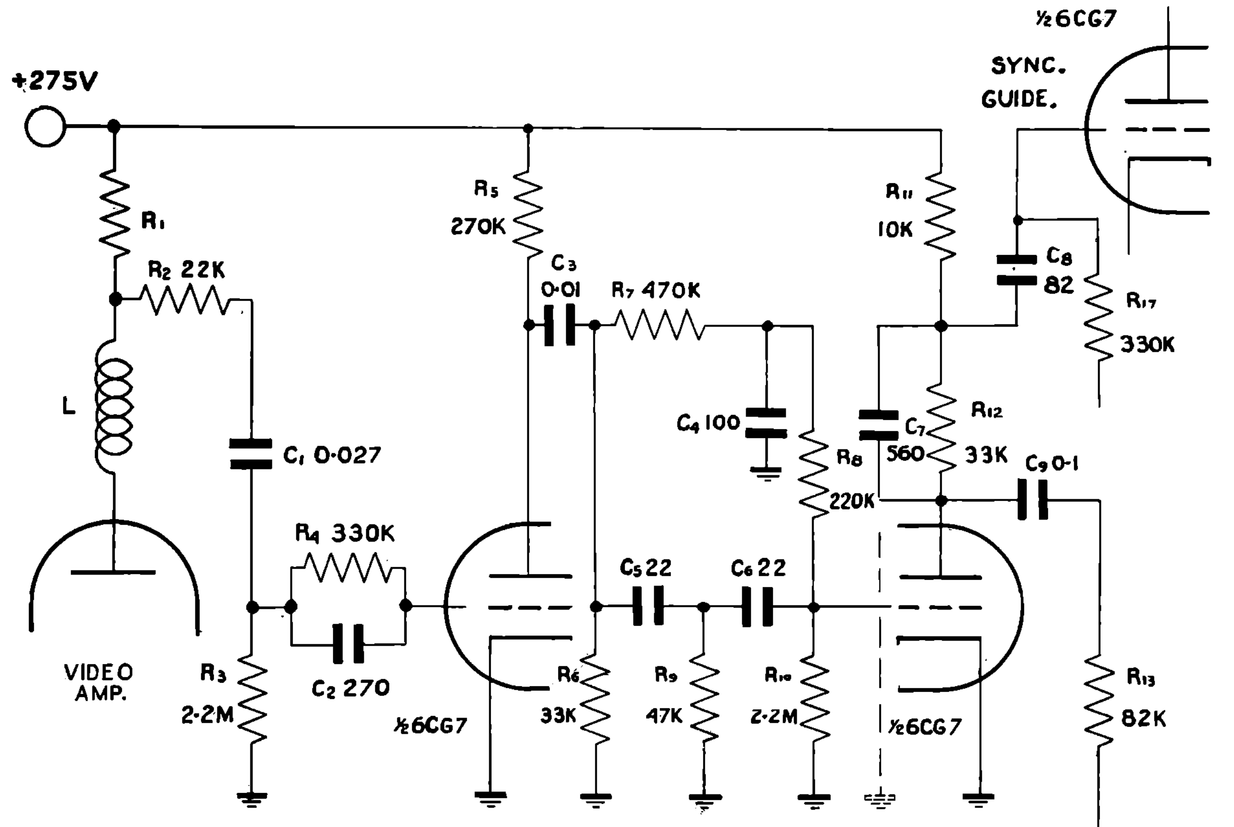
Time constant:	50μsec.
Source impedance:	zero.
Load impedance:	infinite.

If the voltage level at the start of a pulse is aE, where E is the pulse amplitude, then the voltage across the capacitor rises exponentially during the pulse. The voltage at the end of the pulse is then:

$$Ee = aE\ (1 + \exp\ (-tp/RC)) \quad \text{.............. (1)}$$

where tp = pulse duration. If the voltage at the end of a pulse is bE, the voltage across the capacitor falls exponentially during the pulse interval.

3. "Australian television, revision of technical standards" Proc. I.R.E. (Aust.), 19, March, 1958, 122-4.



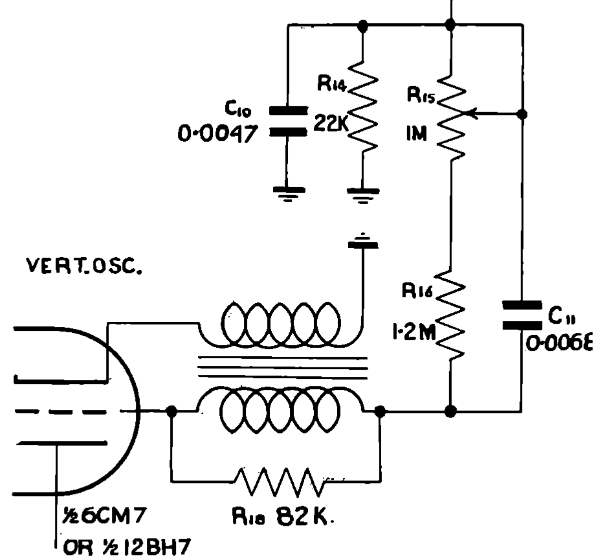


Figure 38.—Circuit diagram of a double triode sync circuit embodying the principles discussed.

The voltage at the start of the next pulse is then:

$$Es = bE \times \exp(-ti/RC) \quad (2)$$

where ti = time between the trailing edge of one pulse and the leading edge of the next pulse.

Referring to the section dealing with "Necessity for Differentiation" and Fig. 21 (b), and assuming that at the trailing edge of a serration (leading edge of following vertical pulse) the voltage level should be 1% from the zero voltage level, then:

$$e/E = \exp(-t/RC) \quad (3)$$

where e/E = 0.01 and t = 5.12 μsec. Solving the equation, RC = 1.11 μsec. With the new standards t has been given a minimum value of 4.8 μsec. and then RC = 1.04 μsec.

Referring to the section on "The Sync Clipper", the rise time from extreme white to 1% from blanking level should not exceed the duration of the front porch which in the old standards is 0.64 μsec. Assuming a simple series network of a resistance and a capacitance (the isolating resistor, R2, and the input capacitance of V in Fig. 38) the voltage rises exponentially:

$$e = E(1 - \exp(-t/RC)) \quad (4)$$

when e/E = 0.99 and t = 0.64 μsec, then RC = 0.14 μsec. With the new standards t has been given a minimum value of 1.0 μsec and then RC = 0.22 μsec.

ACKNOWLEDGMENTS

Acknowledgment is hereby given to the management of the Amalgamated Wireless Valve Company Pty. Ltd., for permission to publish the paper, to Messrs. R. G. Howells and J. McDonald for assistance in preparation of the photographs, to Mr. A. Dombrovski for his work in preparing the drawings and to members of the Staff of the Application Laboratory for helpful discussions.

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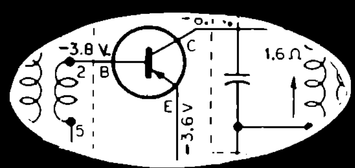
TRANSISTOR APPLICATIONS



CHAPTER TWO

POWER AMPLIFIERS

Transistor power amplifiers can be divided into two basic types: single ended and push-pull. Also, as in valve circuits, transistor power amplifiers can be classified by their modes of operation, such as, Class A amplifiers, Class B amplifiers, etc.

Class A Amplifiers

A typical class A, single ended, power amplifier is shown in Figure 19. The base-emitter circuit is biased in the direction of greater current flow by the bleeder arrangement of R_1 and R_2. This causes the emitter to be positive with respect to the base. The emitter current is stabilized by R_3 and the ac component is bypassed by capacitor C_1. Since this amplifier is used to drive a loudspeaker, a matching transformer, T_1, must be used. Capacitor C_2 is often used to limit the bandwidth to prevent high frequency distortion.

This type of circuit is limited as to power output. The circuit is arranged so that the collector current with no input signal is 11 ma and the collector to emitter voltage is 6 volts. This means the transistor is dissipating 66 milliwatts. If this exceeds the maximum dissipation of a transistor a "heat sink" must be used to prevent transistor destruction. The case of the transistor is usually connected to a large metal object such as a speaker frame by means of a heavy braided metal strap.

A load line can be constructed as shown in Figure 20. The supply voltage is 9 volts, therefore, the collector voltage (V_{cc}) when the collector current (I_c) is zero is also 9 volts. The load line is drawn from the 9 volt point through the dc operating point of the I_c-E_c curve. As can be seen the output will be linear from zero base current (I_b) to 170 microamps of I_b. Any greater swing in base current will cause distortion of the output signal. Therefore, the maximum undistorted output can be calculated as follows:

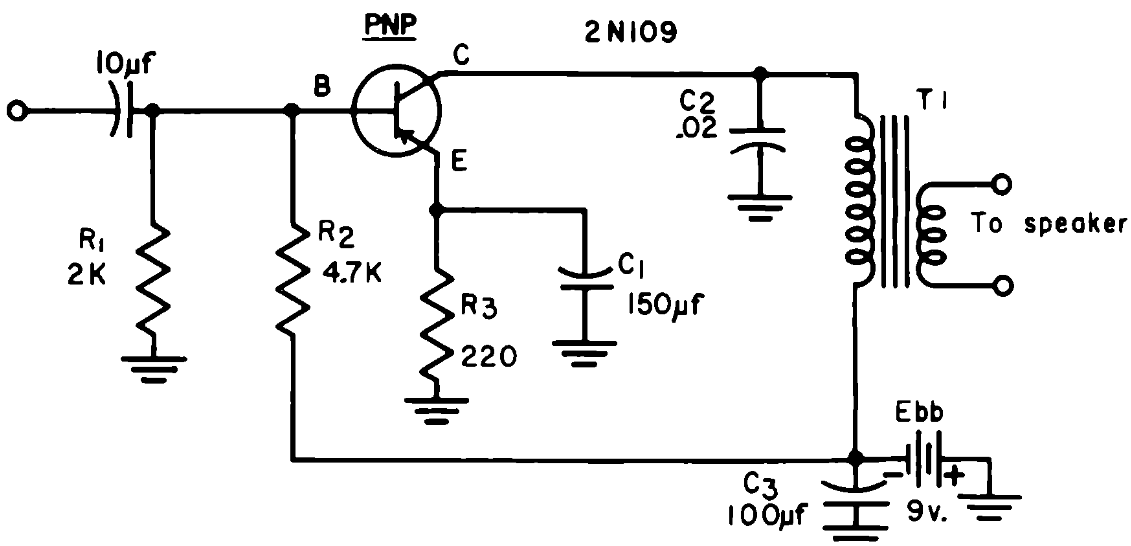


FIGURE 19

$$\text{Power out} = \frac{V_{ce}\ (p\text{-}p) \times I_c\ (p\text{-}p)}{8}$$

$$= \frac{9-3 \text{ volts} \times 22 \times 10^{-3} \text{ ma}}{8}$$

$$= \frac{6 \times 22 \times 10^{-3}}{8}$$

$$= 16.5 \text{ milliwatts.}$$

A greater output can be obtained, but, it will be distorted due to clipping of the input signal.

In order to provide a higher power output with less distortion, a push-pull circuit arrangement should be used. A Class A push-pull audio amplifier is illustrated in Figure 21. A similarity can be noted between the single ended and push-pull circuits. Actually identical components are used, the difference being that the two transistors in the push-pull stage are driven 180 degrees out of phase.

The chief advantage of this circuit is one of less distortion at greater power outputs.

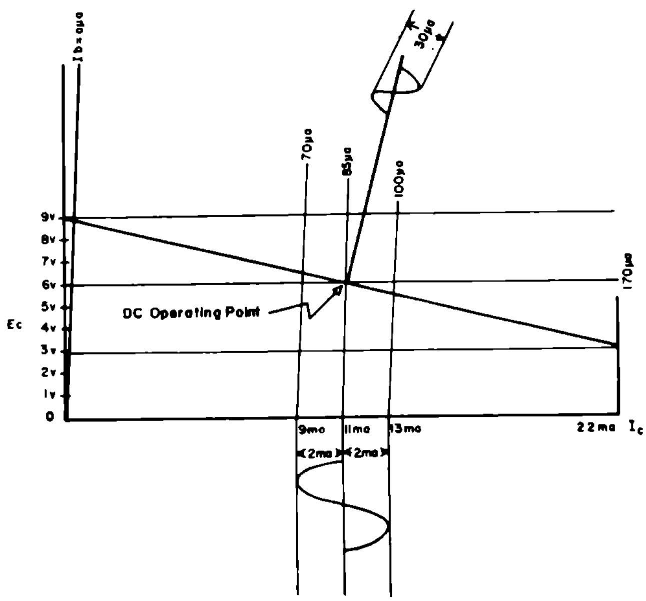


FIGURE 20

Class B Amplifiers

One of the disadvantages of a transistor Class A amplifier is that collector current flows at all times. The transistor dissipation, therefore, is high even when no ac signal is present. The dissipation can be greatly reduced by the use of an emitter-base bias such as that very little collector current flows when no input signal is present. This type of operation is called Class B. When

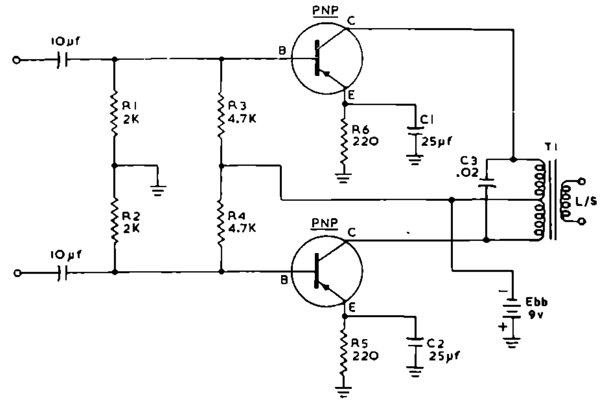


FIGURE 21

p-n-p transistors are operated under Class B conditions collector current flows only during negative signal excursions; when n-p-n transistors are used, collector current flows only during positive signal excursions. The resulting distortion is minimized by the use of two transistors connected in push-pull.

A Class B push-pull audio amplifier is illustrated in Figure 22. The base-emitter circuit is biased near collector cutoff so that very little collector power is dissipated under no signal conditions. Ideally, the transistors would be biased to cutoff and no power would be dissipated under no signal conditions. However, at low signal inputs the resulting signal would be distorted as shown in Figure 23 (a). This is known as **cross-over** distortion. By biasing the transistors so that a small collector current flows at all times the greater portion of this distortion can be eliminated. Figure 23 (b) shows how the conincidence of the projected cutoff points corrects this distortion. Any residual distortion can be minimized by the use of negative feedback. In Figure 22 this feedback is provided by resistor R_4.

Resistors R_1, R_2 and R_3 form a bleeder network which provides proper bias for the transistors. To minimize distortion at low signal levels and prevent thermal destruction of the transistors the characteristics of this network must be very carefully chosen. It was pointed out earlier that

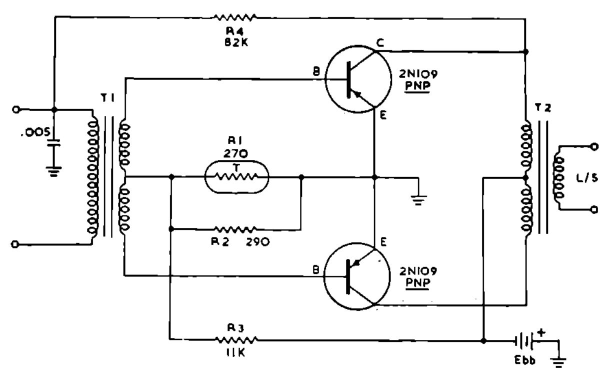


FIGURE 22

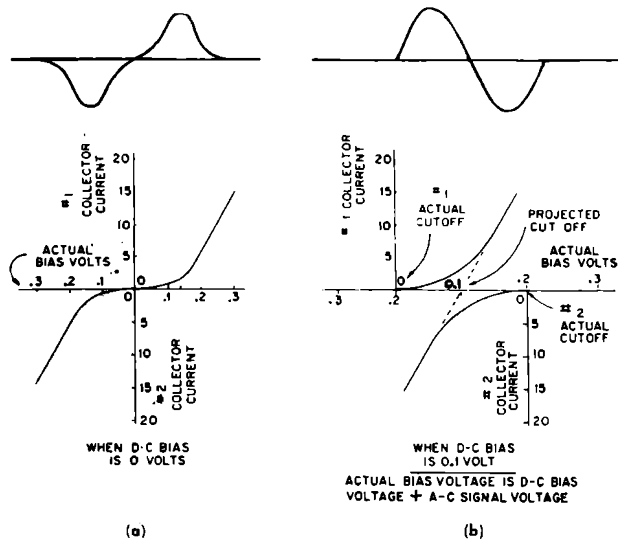


FIGURE 23

the collector current, collector dissipation, and dc operating point of a transistor vary with the ambient temperature. To minimize the effects of these variations a thermistor (R_1) is used in the biasing network. When the ambient temperature increases the resistance of the thermistor decreases, and vice versa, so as to maintain constant voltage across the biasing network. Since the bias voltage controls the emitter and collector currents, the thermistor stabilizes the dc operating level over a wide range of ambient temperature.

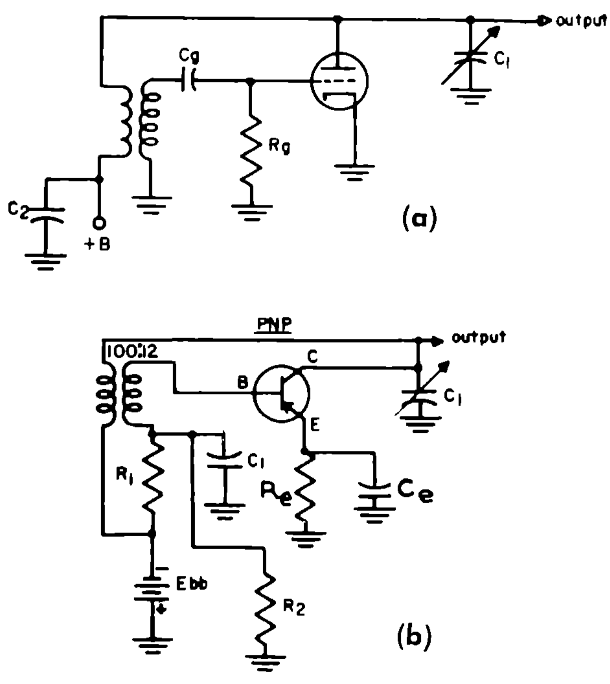


FIGURE 24

OSCILLATOR CIRCUITS

In general, transistor oscillators function in the same manner as valve oscillators. To sustain oscillation an amplifying device having a gain greater than 1 is required. The output of the amplifier is fed back in phase with the input signal with sufficient amplitude to overcome circuit losses. Two basic types of feedback will be discussed. One type uses tuned reactive circuits and the other type uses R/C networks.

Tuned L/C Oscillators

A transistor oscillator and its valve equivalent are illustrated in Figure 24. The valve circuit in Figure 24 (a) is a typical Meissner oscillator. The plate signal is fed back in phase with the grid signal in order to sustain oscillations. The frequency can be varied by the variable capacitor C_1. This type of oscillator can be adjusted to produce continuous sine waves or be self-quenching by choice of the proper time constant for C_g and R_g.

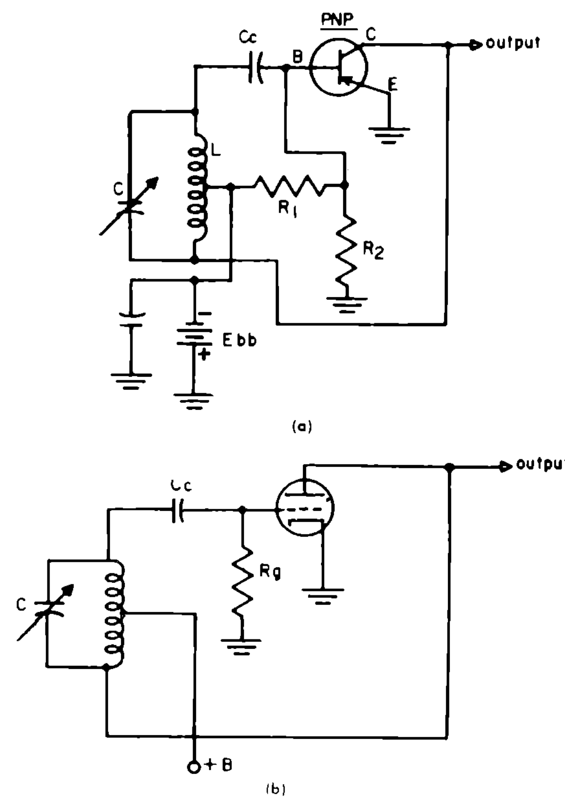


FIGURE 25

The transistor circuit shown in Figure 24 (b) is just as versatile. For sine wave output resistor R_1 (base resistor) is adjusted for enough bias to prevent the transistor from cutting off on positive voltage swings. The circuit can be made self-quenching by readjustment of R_1.

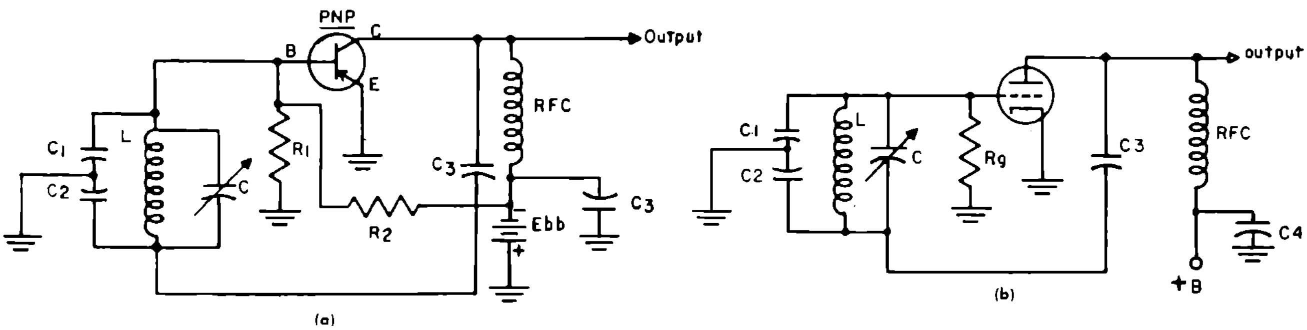


FIGURE 26

Since the transistor is a current operated device, a voltage step down is used between the collector and base. Thus, a small current change in the collector causes a large current change in the base. This action sustains oscillations by overcoming circuit losses. A stabilizing resistor R_e is normally included in the emitter circuit to compensate for variations between production transistors. Resistors R_1 and R_2 provide a low impedance bias source. To prevent self-quenching the time constant of R_1, R_2 and C_1 is made long compared to that of R_e and C_e. R_e and C_e have a function similar to that of C_g and R_g in the valve circuit. Therefore, the emitter time constant is an important consideration.

Another common type of oscillator known as the **Hartley** is illustrated in Figure 25 along with its valve counterpart. The L/C tuned circuit is common to both the input and output circuits. Voltage from the collector circuit is developed across a portion of L, inducing a current of the proper phase into the base circuit in order to maintain oscillation. Again since the transistor is a current operated device, there is a voltage step down from collector to base. However, there is a **current** step up to meet the requirement for the oscillator mode of operation.

The **Colpitts** oscillator is another widely used circuit. The transistor circuit and its valve counterpart are illustrated in Figure 26. As in the Hartley circuit the tuned circuit is common to both the input and output. Capacitors C_1 and C_2 split the signal to provide the proper feedback in order to maintain oscillation. Resistors R_1 and R_2 of the transistor circuit provide the proper bias between the base and emitter.

In servicing these types of transistor oscillators the base voltage waveform can be checked with an oscilloscope to indicate the amplitude of the ac voltage. This waveform is not always a pure sine wave. For rf conversion in superheterodyne receivers a pure sine wave is not absolutely necessary. Therefore, for stability reasons in some transistor circuits the amplitude of the sine wave will appear somewhat compressed. Actual VTVM base to emitter voltage measurements are impractical since the voltage difference is very small. Care must be taken not to load the circuit. A medium to high impedance probe is generally required.

R/C Oscillator

A transistor R/C (multivibrator) oscillator circuit is shown in Figure 27. As can be seen from the circuit it is very similar to a valve circuit. Components C_{b1}, C_{b2}, R_{b1} and R_{b2} form a time constant which roughly fixes the operating frequency. However, the collector current cutoff point (I_{co}) and the emitter cutoff point (I_{eo}) also must be considered. The output waveform is essentially a square wave.

POWER SUPPLIES

The power requirements of most transistor circuits, such as found in transistor radios, are very small as compared to valve circuits, and often allow economical and practical operation with batteries. The power requirements of some transistor circuits are so small that the operating battery life will nearly equal its normal shelf life. The familiar zinc-carbon batteries as used in valve portable radio receivers and flashlights may be used as a power source.

Another type of battery that can be used as a power source is the mercury cell. These cells offer several advantages over the zinc-carbon type. They are more stable and rugged, both mechanically and electrically. They have a much longer shelf life and a greater current capacity than similar sized zinc-carbon batteries. Thus, they are well suited to the design of sub-miniature equipment.

A 9 volt mercury cell battery is composed of seven individual mercury cells stacked together. Each mercury cell has a voltage output of approximately 1.4 volts. Electrical energy is produced by an electro-chemical reaction between amalgamated zinc powder and mercuric oxide, with potassium hydroxide solution used as an electrolyte. A layer of cotton saturated with the electrolyte separates the zinc and the mercuric oxide.

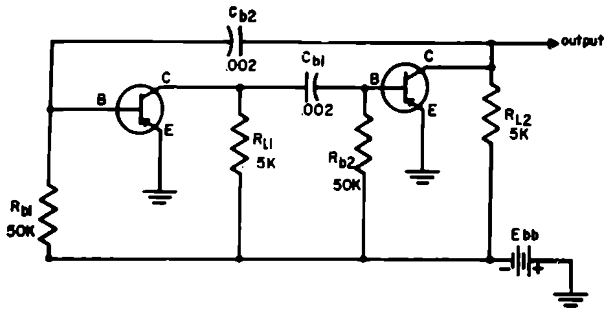


FIGURE 27

The case of the mercury cell is made of steel. Due to the non-linearities of Q_1, an if signal the inactive case material, there is no internal cell reaction until electrical energy is drawn from the cell. The shelf-life, therefore, is extremely long.

In more complex transistor circuits, intended for continuous operation, it is impractical to use battery power supplies. AC power supplies are used in these applications. A transistor power supply is illustrated in Figure 28. The circuit is a full-wave rectifier with an R/C filter. This particular circuit supplies 30 volts at 10 ma of current drain.

PRACTICAL TRANSISTOR CIRCUITS

Portable radios were the first to use transistors for entertainment type equipment. Transistors are also employed in automobile radios, computers, television receivers, etc. A discussion of several circuits which have been used in practical applications or have been developed experimentally will follow.

Pocket Radio

The circuit diagram of an extremely compact pocket radio is presented in Figure 29. This circuit covers the AM broadcast band from 550 to 1600 Kc. The if frequency is 455 Kc, and the undistorted audio output is 25 milliwatts. A ferrite loop antenna is used to capture the signal energy which is transformer coupled to a low impedance secondary winding. The current developed in this secondary winding flows in the base-emitter circuit of Q_1 (2N140). Q_1 serves a dual purpose as an oscillator and converter. Transformer T_1 combined with C1A, tuning capacitor, forms the oscillator tank circuit. An ac regenerative voltage is fed back from the collector circuit to the primary of T_1. The secondary is a low impedance winding causing the fedback current to flow in the base-emitter circuit of Q_1. Therefore, both the signal and oscillator currents flow in the base circuit. Due to the non-linearities of Q_1, an if signal current exists in the collector circuit. The if transformer T_2 transfers the 455 Kc if signal to the base of Q_2, the if amplifier. The collector of Q_2 is terminated into another 455 Kc if transformer, T_3. The secondary is a low impedance winding which drives the crystal detector.

An agc circuit is used to automatically control the gain of Q_1 and Q_2. The agc voltage developed by the detector opposes the bias current of transistors Q_1 and Q_2, thereby operating them closer to cutoff. To prevent if overloading on large signal strengths, an overload diode is incorporated in the collector circuit of Q_2. The overload diode is normally biased to cutoff by R_{16}. However, large signal peaks will cause conduction, loading the primary of T_3. This in turn lowers the stage gain improving the agc action of the circuit.

The detected audio is R/C coupled to the base of the first audio amplifier, Q_3. The collector terminates into the audio interstage transformer, T_4. Capacitor C_8 rolls off the high frequency response preventing high frequency distortion. The volume control is connected across the low impedance secondary winding of T_4. It can be seen from this arrangement that the emitter-base bias is unaffected by the setting of the volume control.

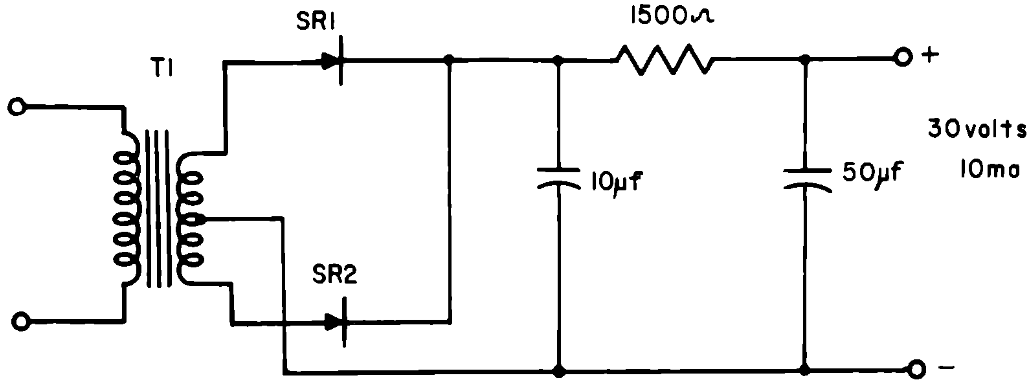


FIGURE 28

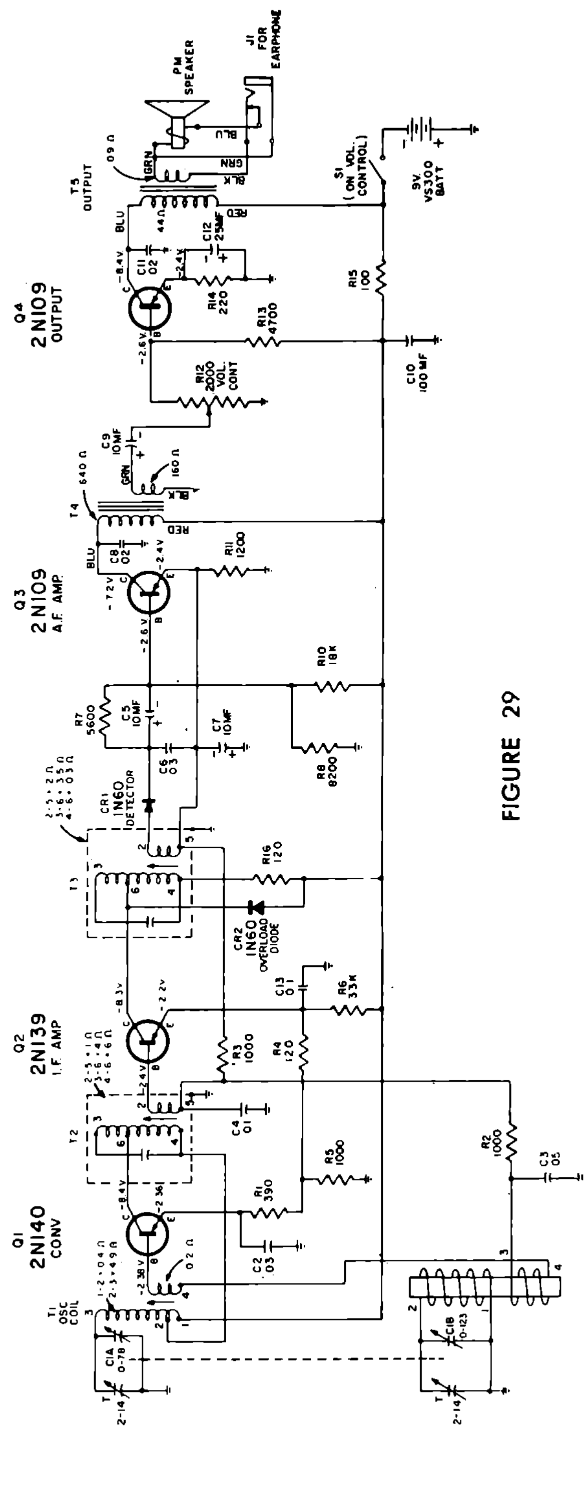


FIGURE 29

To prevent the base current from flowing in the secondary of T_4, capacitor C_9 is used as a dc blocking capacitor. The value of the capacitor must be made large for adequate low frequency coupling.

The audio output stage, Q_4, is a grounded emitter amplifier stage. The collector is terminated into the output transformer, T_5, whose secondary matches the impedance of the loudspeaker. An earphone jack is provided for private listening. The earphone should have an impedance of approximately 200 ohms.

The power supply consists of a single nine volt mercury cell battery. Used in this circuit the battery has a useful life (intermittent service) of approximately seventy five hours. The battery life is comparatively short since the Class A output stage consumes approximately 11 milliamperes with no signal input.

Portable Radio

The schematic diagram of a transistor portable is shown in Figure 30. This receiver utilizes seven transistors and has an undistorted output of 250 milliwatts. The circuit is of the superheterodyne type consisting of: an oscillator-converter, two stages of if amplification, a crystal diode detector, an af amplifier, an audio driver and a push-pull Class B output stage.

The antenna and input circuits are very similar to the circuit shown in Figure 29. Two if's are used in this receiver for better selectivity and gain. The output of the crystal detector is fed directly to base-emitter of Q_4, the 1st af amplifier. A volume control is located in the emitter circuit of Q_4, and the collector is at ac ground. This circuit should not be confused with an emitter follower circuit. The common electrode is the emitter and the characteristics are the same as a grounded emitter amplifier.

The driver stage, Q_5, provides the peak to peak voltages required to drive the push-pull output stage to rated output. The output transistors Q_6 and Q_7 are operated Class B, and a thermistor is used in the base return of both transistors. As the ambient temperature increases, the thermistor decreases in value resulting in a lower bias applied to the base.

In Class B output circuits, the battery current increases with increased signal input. Therefore, prolonged operation at high volume levels will reduce the battery life. The current consumption of this receiver (with no signal) is approximately 8 milliamperes. Therefore, battery life under these conditions is greater than for the previously described radio. However, the current consumption at 50 milliwatts output jumps to 29 milliamperes.

Phonograph Pre-Amplifier

Transistors are ideally suited as phonograph pre-amplifiers for the following reasons: low noise, low hum due to the elimination of heater cathode leakage and no microphonics. A phonograph transistor pre-amplifier is illustrated in Figure 31.

A low impedance **moving coil** type pickup is used to drive the base of the p-n-p type transistor. Capacitor, C_1, is required to block the dc current from the pickup and also to maintain the proper base biasing. The B+ voltage from the amplifier's power supply is used to bias the transistor. The emitter terminal voltage averages 6 volts positive in this particular pre-amplifier. The combination of resistors R_1 and R_2 provides a base terminal voltage of 5.85 volts positive. The base is biased slightly negative with respect to the emitter, as is normally required for p-n-p transistor operation. The collector terminal in this case must be grounded so that it is biased negative with respect to the emitter. This necessitates placing the load resistor R_4 in the emitter circuit. This type of circuit arrangement is not to be confused with the emitter follower. It can be seen in Figure 31 that the magnetic pickup is connected between the base and emitter. Since the collector is grounded, and the output is taken between the emitter and ground, the circuit is a common emitter type. Capacitor C_3 is a decoupling capacitor used to prevent interaction from other circuits using the same B+ supply.

Capacitor, C_2, couples the amplified audio signal to the tone control amplifier.

Video Amplifiers

Video output amplifiers differ from audio power amplifiers in that they operate into voltage operated devices such as picture tubes or oscillograph tubes rather than into current operated devices such as loudspeakers. Since transistors are current operated devices they present no problems in audio amplifier applications. However, the fact that the peak to peak driving voltages required by picture tubes are greater than the maximum voltages that can safely be applied to the collectors of most transistors has limited the application of transistors as video output amplifiers in television receivers and similar equipment.

A two stage transistor video amplifier capable of delivering an output of approximately 30 volts peak to peak is shown in Figure 32. The transistors used in this amplifier should have a high frequency cutoff of approximately 30 Mc, and the one used in the output stage must have a collector voltage rating of at least 40 volts.

The **drift** type transistor which was discussed previously is well suited since it meets both requirements. The base region is thick enough to prevent voltage breakdown and the drift design assures the high frequency response.

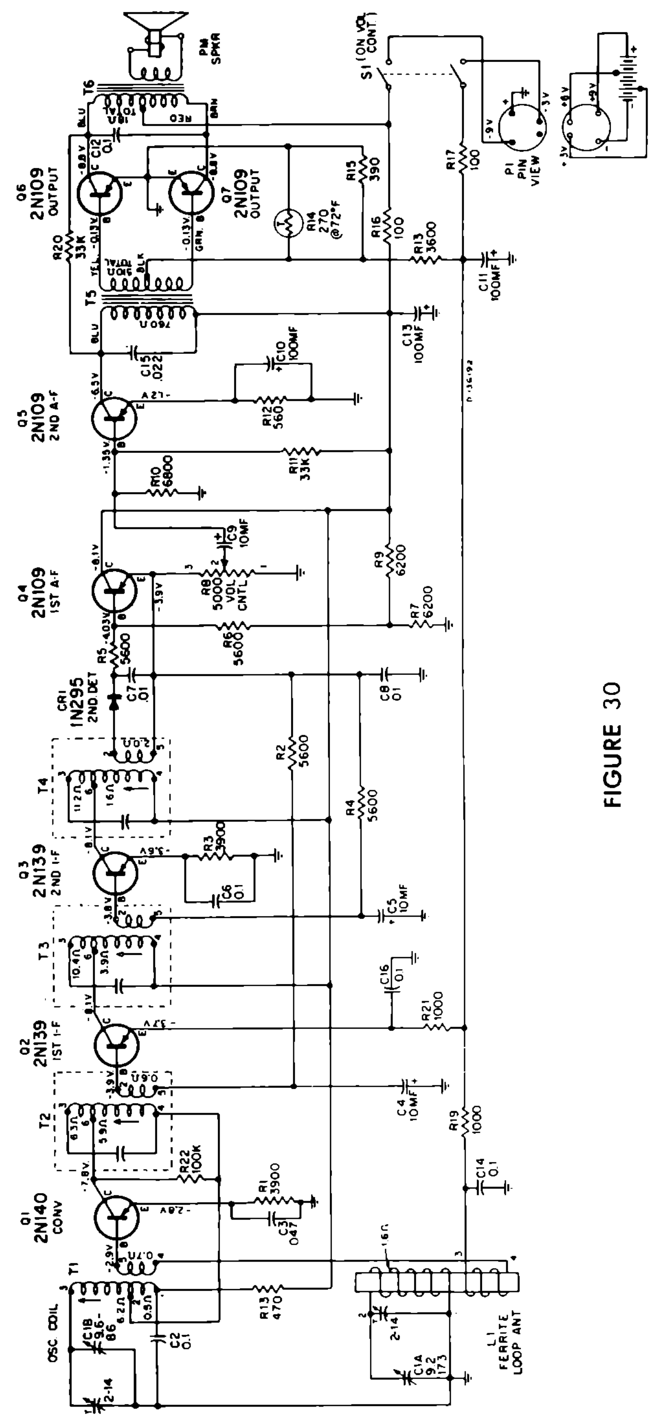


FIGURE 30

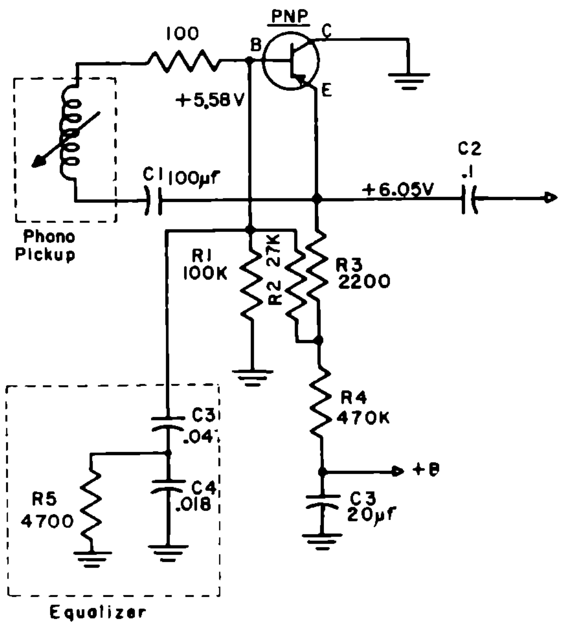


FIGURE 31

As can be seen from Figure 32, both transistors are operated as grounded emitter amplifiers. Series peaking is used at the input of both stages and in the output of the second stage. Also, the emitter resistors are partially bypassed providing additional peaking. This peaking compensates for high frequency losses due to internal and external capacity associated with the transistors. A gain (contrast) control is used in the base circuit of the first stage. The gain of this stage is varied by regulating the base-emitter bias. As the base is made more positive, greater current flows between the emitter and base, thus increasing the stage gain.

A greater output can be obtained from a similar arrangement using a transistor which can operate with higher collector voltages. Another method is to use push-pull output transistors. In this case one transistor drives the cathode of the picture tube while the other transistor simultaneously drives the grid 180° out of phase with the cathode. In this case a Class A push-pull amplifier is used.

Television Sync Separator

A transistor sync separator is shown in Figure 33. In this example a positive voltage supply is available, thus an n-p-n transistor is used. The transistor is biased such that little current flows without the presence of signal. When a sync positive video signal is applied to the stage, the developed bias across the resistors in the base circuit produces a clamping action which cuts off the transistor during picture intervals and permits it to conduct only during sync pulse intervals. This dynamic biasing action produces clean, amplified sync pulses at the output of the stage. The gain of the transistor is sufficient to assure that, even on very weak input signals, the collector current saturates and clips off the tips of the sync pulses below the noise level. Horizontal sync pulses are prevented from reaching the vertical sweep circuit by the R/C network in the vertical sync output lead.

Other Transistor Circuits

There are other transistor applications too numerous to cover in this lesson. However, if the technician understands the basic operation of transistor amplifiers and oscillators, little additional reading is necessary to basically understand the operation of a completely transistorized television receiver.

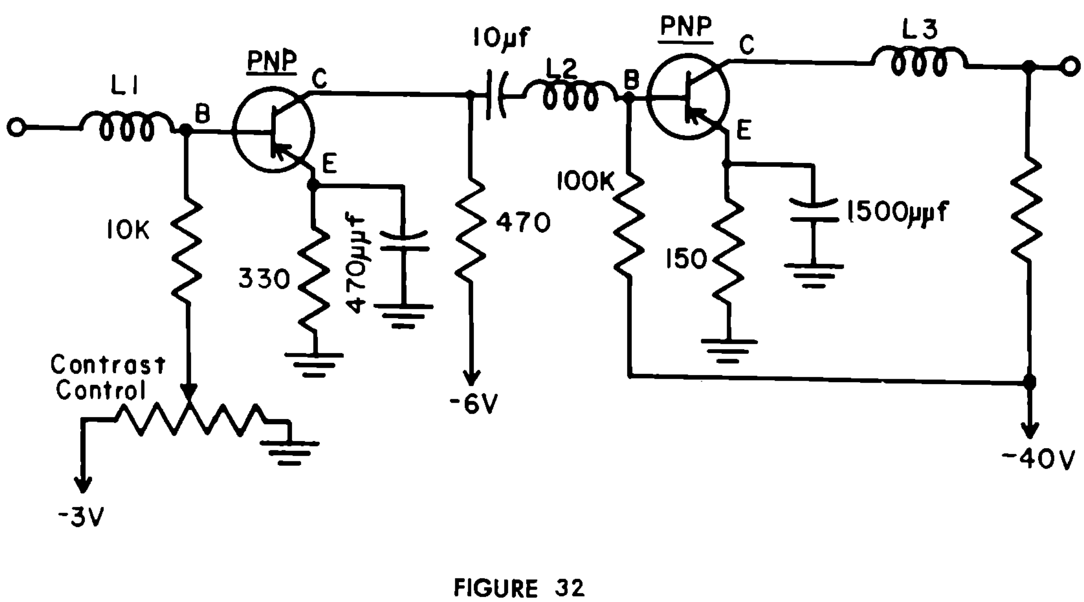


FIGURE 32

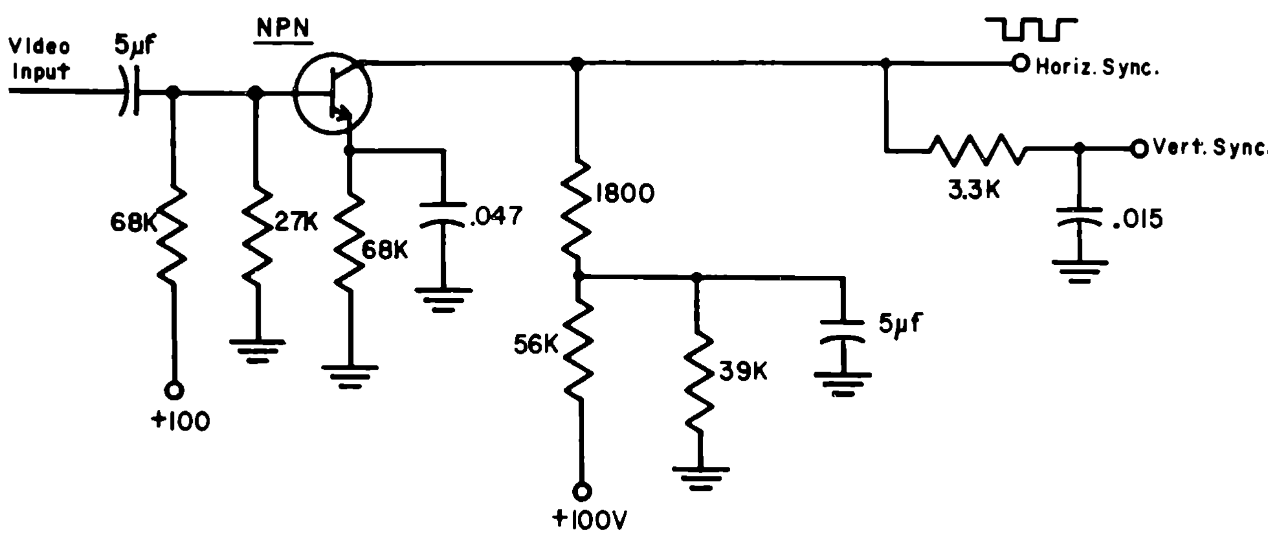


FIGURE 33

TRANSISTOR COMPONENTS

Due to the small size of the transistor it is desirable that the associated circuit components be small to allow compact transistorized electronic equipment design. Since the voltage and current requirements are usually low in this type of equipment, resistors, coils, controls, transformers and sockets can be made small in size without danger of voltage arc-over and overheating.

Electrolytic Capacitors

The capacitor has presented a different design problem. Due to the low circuit impedances involved, the value of capacitance must be large. Even though the working voltages are low the actual value of capacitance may range from 1 to 100μf or higher. If ordinary electrolytic capacitors of this value were used in transistor circuits they would be so large that compact design and styling would be impossible.

However, a metal known as **tantalum** overcomes the capacitor size problem. Tantalum is oxidized and this oxide serves as a dielectric. Since tantalum is also used as the capacitor base material, miniature electrolytic capacitors can be made with large capacity values and low working voltages. A typical capacitor, case diameter $\frac{7}{16}$" and length $\frac{3}{4}$", is rated 100 μf at 10 volts. Units of lower capacity can be encapsuled in even smaller cases.

Transformers

The size of a transformer is generally determined by two factors: the amount of unbalanced direct current flowing in the windings and the amount of power to be handled at a specified frequency. If the transformer must handle large dc currents the core material must be large to prevent magnetic saturation which in turn will cause distortion of ac signal currents. Since the dc currents in transistor circuits are usually relatively small, the core material of the transformers can also be made small. However, in some applications, such as audio power amplifiers, the transformer may be as large or larger than those required by valve circuits. This is due to the fact that primary currents in the order of 10 amps may be encountered. Generally, rf, if and oscillator coils can be made quite small.

SERVICING TRANSISTOR CIRCUITS

Servicing transistorized electronic equipment presents no major problems when compared with valve type equipment. With printed wiring, component parts are neatly arranged with little or no stacking. However, care must be taken to prevent damage of the printed wiring.

One of the outstanding benefits of transistor equipment is the reliability of the transistor itself. Transistors have long life and this factor tends to decrease the amount of service required.

Precautions

Although transistors have the advantages of long life and reliability, certain precautions must be observed. Mechanically they are rugged but not indestructable so reasonable care should be employed. The leads are the most fragile part and whether they are the long flexible type or the short rigid type they should be treated in the same manner as the leads of a crystal diode. The transistors with long flexible leads are usually soldered directly into the circuit. When unsoldering or re-soldering, caution should be taken not to overheat the transistor. As in the case of crystal diodes, long nose pliers should be used to grip the lead between the transistor and point of soldering in order to control the heat. Transistors with short rigid leads are usually plugged into a socket. In some cases, however, these transistors are plugged directly into the printed board and then dip soldered. In this case long nose pliers cannot be used, therefore, the soldering iron should be hot before touching the solder joint and the transistor should be soldered or unsoldered as fast as possible to prevent damage. When un-soldering a component from a transistor socket, the transistor should first be removed to prevent damage from heat.

Due to the low operating voltages required, relatively small voltage changes can greatly upset the biasing of the transistor. Depending on the circuit, small bias changes can result in destruction of the transistor due to excessive transistor dissipation. Therefore, it is important that circuit components are not inadvertently shorted when servicing transistorized equipment.

Servicing Techniques

The actual servicing techniques used when servicing transistor equipment are similar to the techniques used when servicing normal valve circuits. Basically these techniques are:

1. Check Battery.
2. Visual Inspection.
3. Check Transistor.
4. Voltage Check.
5. Resistance Check.
6. Capacitance Bridging.
7. Probing.
8. Signal Injection.
9. Alignment.
10. Component Substitution.

Before performing any transistor service, the battery should be checked and direct substitution is generally the best method. However, it is usually best to first check the voltage of the original battery with the receiver on and maximum volume setting. If possible also tune in a strong local station. Generally, where a 9 volt battery is used, the receiver should operate with a battery voltage as low as 6 volts. This, of course, will depend on the design characteristics of a particular receiver. However, circuit trouble should be suspected if the receiver will not operate if the battery voltage is 7.5 to 8 volts. In some cases the customer will complain that batteries only last for a short duration. In all cases an ammeter should be inserted in series with the battery and the total current drain of the receiver measured. These current figures are usually stated in the service notes. If the current reading is high the leakage should be traced and corrected. If not, explain to the customer the operation and limitations of battery operated receivers. It is often desirable to check the operation of the receiver with a reduced battery voltage. This condition can be duplicated by inserting a small resistor in series with the battery. This is shown in Figure 34. The actual value of the resistor used depends on the condition of the battery and the voltage desired. Using a rheostat in this application allows a selection of voltages and allows the technician to check the sensitivity of the receiver under varying voltage conditions. In cases where an ac power supply is used, the supply voltages should be checked before attempting any service work. Improper supply voltages can cause odd effects and many headaches can be eliminated by first checking the power supply.

Visual inspection is also a good service technique. Occasionally a loose wire or faulty connection can be found before extensive voltage checks are made. Faulty components such as burned resistors are seldom encountered since the power of the supply voltage is usually very low.

Transistors, like valves, can be checked by direct substitution. Transistors, however, have a characteristic known as **leakage current** which may affect the results obtained when using the direct substitution method. The leakage current

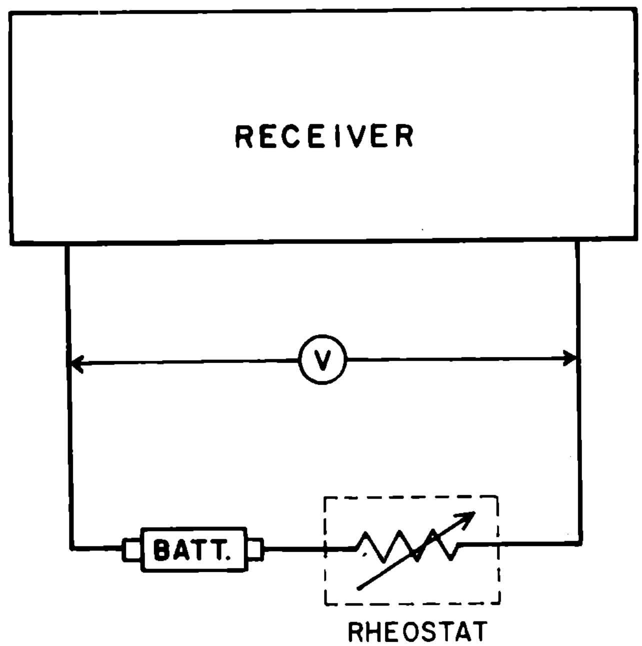


FIGURE 34

can affect the amplification factor or gain of the transistor and is more critical in certain applications than others. Thus it is possible that a particular transistor will operate satisfactorily in one circuit and not in another. It also has been found that the amount of leakage current will increase slightly as the transistor ages.

Voltage measurements provide a means of checking receiver conditions similar to valve circuits. The major difference is the magnitude of the voltages encountered. For example, the bias voltage between the base and emitter is in the order of 0.05 to 0.2 volts. Therefore, a sensitive VTVM is usually required. When making voltage checks the polarity should be observed. In the case of valve circuits, if a positive voltage is measured on the grid of a valve a leaky coupling capacitor is indicated. However, in the case of transistors the base to emitter voltage may be positive or negative depending on the type of transistor used. For example, the p-n-p type normally operates with the base negative with respect to the emitter, whereas, the opposite is true in the case of n-p-n transistors. The schematic should be checked for the proper polarity as well as the magnitude of the voltage.

In some circuits it may also be desirable to make a current check, although with printed wiring this is not always possible. The amount of current can be calculated by measuring the voltage across a resistor in the circuit and also checking the value of the resistor with an ohmmeter. The current can then be determined with the aid of Ohm's Law. For example, if the collector current is to be measured, measure the voltage drop across the emitter resistor and check the resistor with an ohmmeter. Using Ohm's Law the collector current can be calculated.

Resistance measurements are not generally made in transistor service other than checking for open circuits in transformers and coils. Due to the low voltage power supplies used in transistor circuits the resistors have little tendency to burn up or change values. It is important that the transistor or component be removed from the circuit before attempting resistance measurements. Since the ohmmeter contains a battery, the wrong polarity voltage may be applied to a critical stage and cause permanent damage to the transistor. Always disconnect supply voltages before removing a transistor from its socket to prevent damage to the transistor by current transients.

Capacitance bridging, probing, and component substitution are all good techniques which are commonly used when servicing valve circuits and are equally good when servicing transistor circuits. They are generally resorted to when servicing the so-called "dog" troubles. Care must be taken to observe polarity when substituting or bridging electrolytic capacitors.

Alignment is another excellent service technique when checking rf and if stages for sensitivity. A shorted turn in an rf or if coil is quickly identified by poor response when tuning the coil. For example, turning the slug in an if coil one-half turn normally has a great effect on the sensitivity of a transistor radio. If this adjustment produces little or no effect the transformer is probably defective or the associated stage inoperative for other reasons.

Signal injection is also an excellent service technique. A defective stage in a radio can be rapidly isolated by injecting a signal at the input to each stage starting at the loud-speaker and moving back to the antenna coil. The defective stage can be identified by a decrease in gain from the previous stage, or by a very slight increase where a large increase should occur.

All of the service techniques mentioned are suitable for all types of transistorized equipment. Keeping in mind the aforementioned precautions, servicing transistor receivers should present no greater problem than their electron tube counterparts.

NEXT MONTH . . .

A series of review questions will be presented, based on the material printed on transistor fundamentals and applications. The questions will be accompanied by a set of multi-choice answers, and of course, a key to the correct answers.

Integrated Electronic Devices:

A New Approach to Microminiaturization

By J. T. WALLMARK, Research Engineer, RCA Laboratories

While performance and reliability requirements for future electronic systems are mounting, weight and size allowables are shrinking. To meet such conflicting specs, radically new design approaches are needed. The "integrated electronic device" may fill this bill.

Miniaturization of electronic circuits and equipment has been brought about mainly by two techniques:

- component miniaturization through improved design and the use of transistors and other solid state devices;
- modular packaging, in which parts are grouped in subassemblies that, in turn, are built into compact blocks from which larger systems can be easily put together.

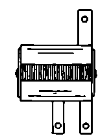
We haven't yet squeezed the limit out of these "conventional" techniques, but we're fast reaching the point of diminishing returns. From here on out, these approaches undoubtedly will require a high engineering effort compared with the gains in size and weight.

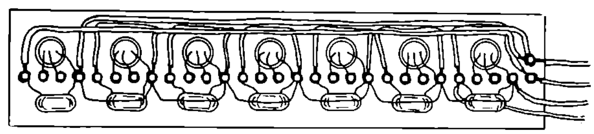
Yet in tomorrow's vehicles equipment size and weight will be even more critical than they are today. In addition, electronic gear will have to operate reliably over long periods with extreme economy of power.

To meet these mounting requirements and at the same time subminiaturize systems, radical design approaches will be needed. One promising technique uses the "integrated electronic device" concept.

If we consider the functional roles of the various circuit components, we see that amplifying and switching are the "active" job of devices such as valves and transistors. "Passive" components—resistors, capacitors, and inductors—perform such functions as determining frequency, delaying signals, and transferring charges and storing them for short periods.

The true "integrated device" can do the jobs of both active and passive components. Moreover, it goes beyond a mere combination of active and passive elements in the same enclosure—it performs the entire circuit function by control of electron motion within a vacuum or a solid.





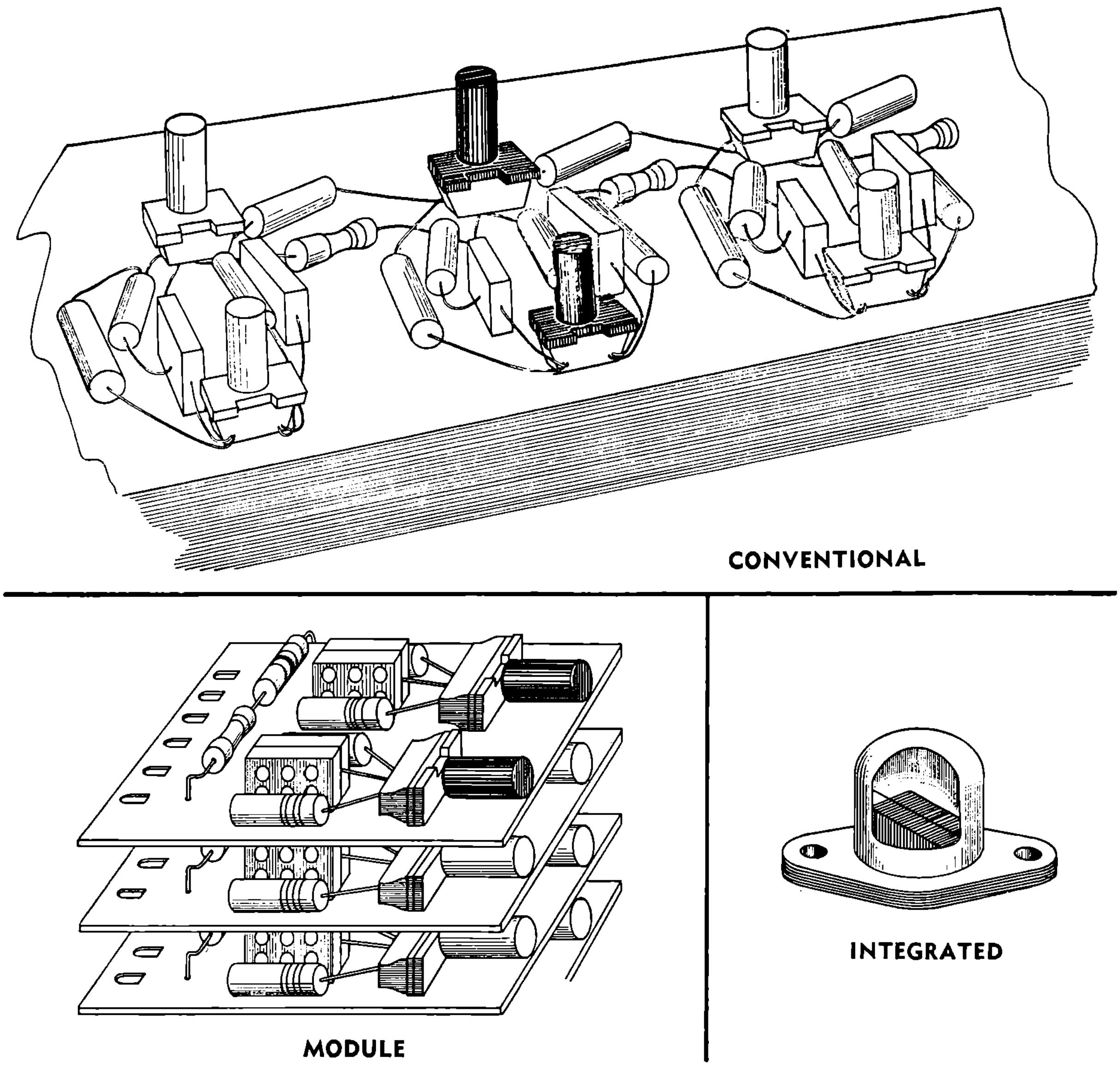
Experimental shift register transistor (top) is 0.5 in. long and 0.04 in. thick, has 20 positions. Seven-position transistorized shift register (bottom) uses transistors, ferrite cores, resistors, and capacitors, is shown for size comparison of new concept with present miniaturized construction. With further refinements, experimental unit is expected to perform functions of 20 transistors, 40 resistors, and 20 capacitors.

Much of modern electronic circuitry is iterated—identical stages follow one another in cascade, "trees", or matrices. If one stage can be replaced by an integrated device — a semiconductor, for example — it should be unnecessary to carry on the function through an array of separate but identical devices. The entire array may be built into a single, tiny semiconductor.

Ten transistors made into single piece

Lab studies are trying to find out how close practical devices can come to this ideal. Circuits in which the passive elements play a relatively simple role will be considered first. As the concept is explored, there will undoubtedly be many expeditious compromises that stop short of complete integration.

An example of the integrated electronic device is the semiconductor shift register and counter now under development (with USAF support) at the RCA Laboratories, Princeton, N. J. This unit



Size comparison of three circuit configurations that would perform precisely the same function. The three drawings are made to the same scale.

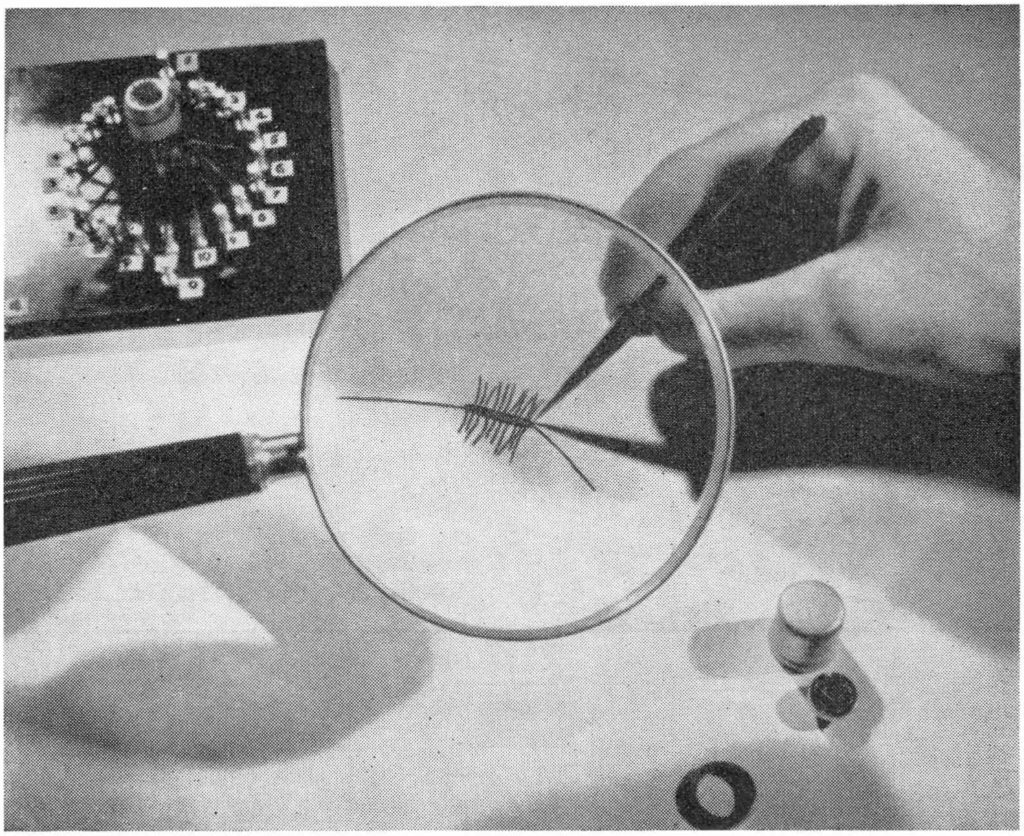
consists of 10 transistors made into a single piece of semiconductor material. Also part of the same piece are the interconnecting circuit functions—the jobs that in conventional shift registers are performed by separate RC couplers between the transistors. The result is a single microminiature part into which are integrated the functions formerly performed by 10 transistors, 20 resistors, and 10 capacitors.

The function of the shift register is the short term storage of a multiple-digit number or similar information for subsequent transfer (intact) at the right moment into the computing circuits. In this function, the developmental semiconductor shift register would replace a complex circuit with its active and passive components (see photograph).

The developmental register is a bar of germanium measuring only 0.5 x 0.02 x 0.004 in. Within the bar are integrated 10 "thyristor"-type transistors that perform high speed switching over a certain voltage range. Between the stages, the germanium bar itself performs the functions of the interconnecting circuits. It serves as a short term memory, or delay, retaining the information stored in the register while it is shifted from one stage to the next.

Integrated device is extreme module

The integrated device, being an extreme form of a module, offers all the modular advantages. Furthermore, bulk and weight have been cut to a bare minimum, since many components can be made to share a single enclosure and a single structural support without interconnecting metallic conductors.



Research breakthrough in ultra-miniature circuitry is how RCA describes its "integrated electronic devices". This experimental shift register does the job of 10 transistors and 30 passive elements through molecular arrangements within a bar of germanium. At upper left is test unit for the register, at lower right are elements of capsule in which unit is enclosed for testing.

Although it has not yet been obtained in experimental results, increased reliability may also be expected. This is a possibility because (1) use of a single element will reduce the number of manufacturing and assembling procedures and (2) a large number of units may be manufactured simultaneously in one piece of crystal under identical conditions. In a sense, the integrated device concept shifts part of the responsibility for and cost of producing a reliable system from the system assembler to the component manufacturer.

It is still a little too early to make firm predictions about the future of integrated devices, but interesting speculations are possible. At first there appears most to gain in the fields of electronic computing and control systems, since these use subassemblies made up of many identical elements or stages. This includes shift register, matrix switches, transfer trees, adders, and similar widely used circuits. Another application appears in equipment that commonly uses individually designed stages, such as radio receivers, audio amplifiers, and control circuits.

Although the integrated device concept greatly reduces the number of external interconnections when many units are integrated, the great reduction in physical size makes external leads a tricky problem in terms of present techniques. The answer must be the use of printed circuit techniques in the device itself.

Our concept can lead to the integration of ever larger numbers of units into subsystems, reducing both number and size of building blocks. Ultimately, complete systems might be "custom-built" from conventional logic charts without any need for extensive knowledge of electronics or circuit theory.

Looking well beyond the present research state we can foresee the application of this technique to complete microminiature electronic control and guidance systems of extreme versatility. Such systems, made up of a variety of integrated building blocks, might enable us to pack into a missile or satellite many of the complex computations and analysis functions that are left out today because of lack of room. In manned aircraft, the integrated device could cut navigation, communication, and flight control electronics gear to a mere fraction of today's bulk.

(With acknowledgements to RCA)

THE DESIGN-MAXIMUM RATING SYSTEM

(Continued from page 118)

The foregoing concepts are summarized in the following definition:

DESIGN-MAXIMUM RATING SYSTEM

Design - Maximum ratings are limiting values of operating and environmental conditions applicable to a bogey electron device of a specified type as defined by its published data, and should not be exceeded under the worst probable conditions.

The device manufacturer chooses these values to provide acceptable serviceability of the device, taking responsibility for the effects of changes in operating conditions due to variations in device characteristics.

The equipment manufacturer should design so that initially and throughout life no design-maximum value for the intended service is exceeded with a bogey device under the worst probable operating conditions with respect to supply-voltage variation, equipment component variation, equipment control adjustment, load variation, signal variation, and environmental conditions.

Not included in the definition is any reference to the permissible heater or filament voltage variation. Actually, the conditions of the design-maximum system in no way restrict the heater or filament voltage tolerance. For this reason, no fixed tolerance is included in the fundamental definition; the permissible variation in heater or filament voltage will be specified for each individual valve type.

It should also be noted that the design-maximum rating system overcomes the fundamental difficulties associated with the two existing systems. As opposed to the design-centre system, incentive is provided to the circuit designer to study and control the variations in operating conditions to which the valves are subjected. The valve manufacturer is fully responsible for control of quality in areas in which he operates, but is not responsible for those areas which he can neither evaluate nor control. As compared to the absolute-maximum ratings, design-maximum ratings are significantly simpler to use. The principles of the design-maximum rating system can be applied to such diverse classes of service as television receivers, computers, military equipment, and industrial equipment.

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