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# Controlled r.f. power generation using magnetically-beamed triodes

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SUMMARY

The magnetically-beamed triode has a structure radically different from a conventional industrial triode. The relatively small proportion of the current which is intercepted by the control electrode (gate), even when it is driven to a high positive potential, leads to useful design properties and novel applications circuits.
The cool gate needs no special coating to inhibit electron emission and oxide-coated cathodes are used to complete a robust structure and reduce ancillary heater power.
Low gate current results in low bias loss and low oscillator drive power. The latter permits the introduction of a new form of oscillator output power control by limiting the gate voltage swing by a catching-diode in conjunction with a solid-state regulator. Design parameters are considered for this unit which is less costly than alternative systems. Oscillator power may be controlled from maximum down to 10% by the swing of only a few volts at the regulator input. Provision may be made for external feedback loops in addition to a manual presetting potentiometer. Examples are quoted for the stabilization of input current or r.f. output voltage.

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## 1 Introduction

In the infancy of r.f. industrial heating, valves were used which had been specifically designed for use in telecommunication transmitters. Although some of these valves were, with care, used successfully in early equipments, it soon became apparent that their characteristics and mechanical construction resulted in quite severe limitations in their application in this new field. The primary reason for this is that transmitters are in general subjected to a constant electrical load which allows optimum matching to be obtained between the valve and the circuit to which the power is delivered, whereas for industrial oscillators the situation is quite different. Here there can be large differences in load impedance, for example, when a ferric load goes through the Curie point, and in consequence both the power delivered by the valve and the power dissipated on its electrodes are subject to considerable variation. This is particularly true for the control grid where the dissipation under low load conditions can be several times the dissipation under full load. This often resulted in the maximum allowable grid dissipation being exceeded when transmitting valves were used, leading to high grid emission, 'run-away' conditions, and consequent valve failure. In the mid-1950s, valves were specifically designed to operate with adequate safety factors under the onerous conditions met in industrial heating applications. A number of different approaches were made to solve this problem, such as optimizing the amplification factor[1] to minimize the variation in output power and grid dissipation with changing load impedance; alternatively lowering the amplification factor[2,3] to design in a large grid safety factor.

Though these improvements were significant, the construction of tubes remained conventional with the fundamental fragility of grids and thoriated tungsten filaments. The introduction of a magnetic beaming,[5] which dramatically improves the division of current between anode and grid (now called gate) represents the most significant change in the last quarter-century. Figure 1 compares the grid/cathode structures of a mesh industrial triode with that of a magnetically-beamed triode.

## 2 Principle of Operation of Magnetically-Beamed Valves

In the magnetically-beamed triode, the problems referred to earlier have been overcome by replacing the normal wire grid by a gate electrode. Electronic emission from the cathode is focused by a magnetic field into a rectangular beam of current, which is collected by the anode. Modulation of the beam is performed by the gate, which is equivalent to the grid of a conventional triode. The combined effect of the magnetic field and the perpendicular components of electrostatic field resulting from the anode and gate potentials causes emitted electrons to spiral towards the anode.

The radius of curvature $R$ of this helix is dependent on the magnitudes of the electrostatic and magnetic fields, in accordance with the relation

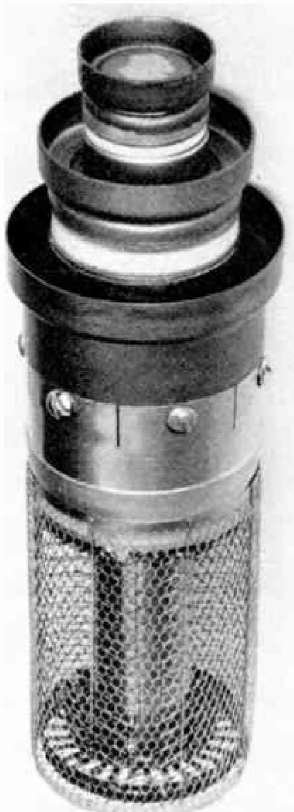
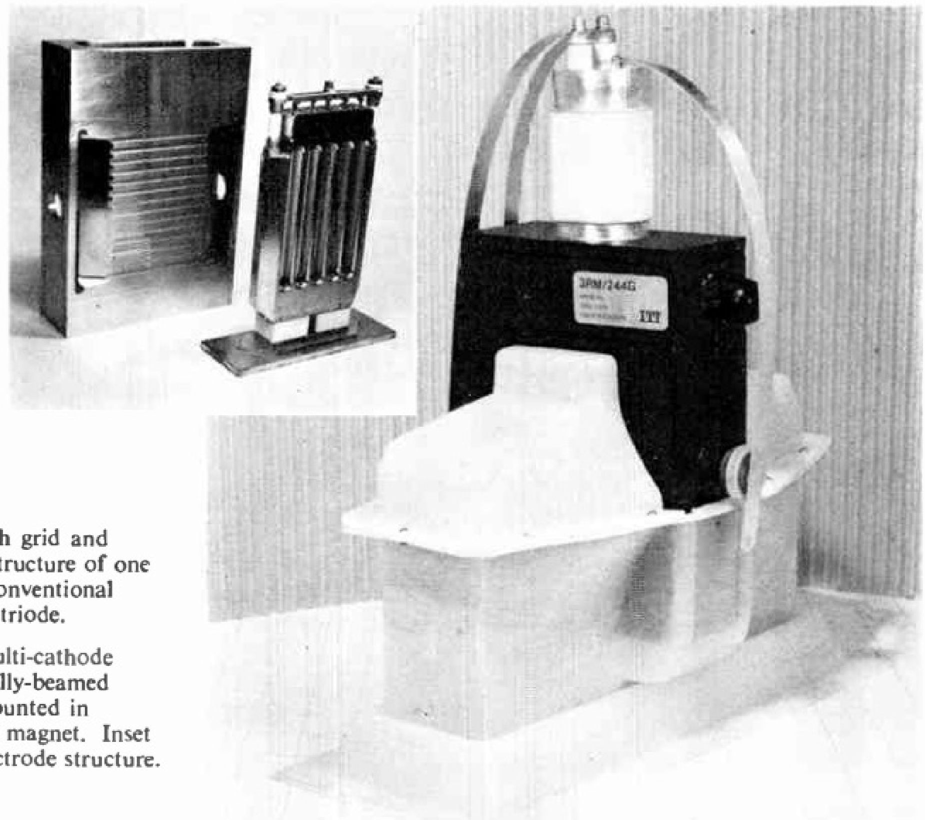


Fig. 1.
(left) Mesh grid and filament structure of one form of conventional industrial triode.

(right) Multi-cathode magnetically-beamed triode mounted in shrouded magnet. Inset shows electrode structure.




$$R \propto \sqrt{V_r}/B$$

where $V_r$ = voltage component perpendicular to the magnetic field,

$B$ = flux density.

The pitch $P$ is given by

$$P \propto \sqrt{V_p}/B$$

where $V_p$ = voltage component parallel to the magnetic field.

The intercepted gate current is an inverse function of magnetic field but is relatively independent for field strengths well above the 'knee' region (Fig. 2).

As with conventional triodes, tube characteristics may be described in terms of mutual conductance and amplification factor, and are determined by geometric relationships, particularly gate dimensions and gate-cathode spacing: dimensions $m$, $n$, $d$ in Fig. 3. The present range of m.b. triodes has an amplification factor of about 25 which is near optimum[1] for industrial oscillators.

## 3 Construction

The very low gate dissipation of a magnetically-beamed triode when operated as an induction heating oscillator, typically one-sixth of that on the grid of an equivalent conventional triode, allows a wide choice of materials for the gate construction. In the range of valves described here, plain copper replaces the molybdenum and multi-layer anti-emissive coatings used for conventional grids. The gates are conduction cooled, through beryllia† blocks, to the water-cooled anodes. In conjunction with the low dissipation, this allows gate operating temperatures as low as 250°C to be attained,

† The latest design has eliminated the need for this material.

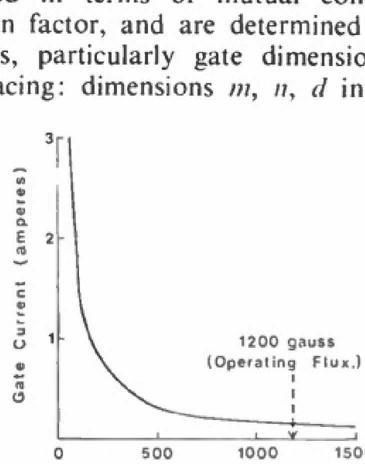



Fig. 2. Intercepted gate current vs. magnetic field (3RM/245G).

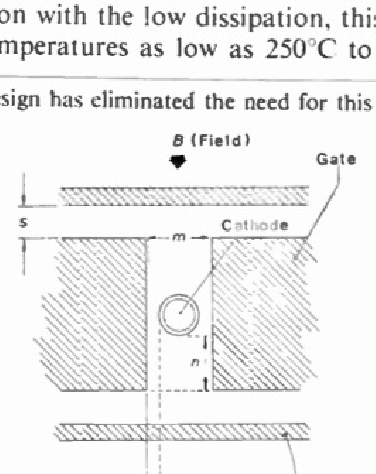



Fig. 3. Electrode and magnetic field configuration.

**Table 1**

| | | M | C | M | C | M | C |
|---|---|---|---|---|---|---|---|
| Valve Type | | 3RM/189G | 3R/199E | 3RM/244G | 3R/252E | 3RM/245G | 3R/262E |
| Anode voltage | kV | 6·5 | 6·5 | 9·0 | 9·0 | 7·0 | 7·0 |
| Anode current | A | 1·8 | 1·8 | 5·7 | 5·7 | 6·8 | 6·8 |
| Gate/grid current | mA | 25 | 500 | 70 | 475 | 70 | 500 |
| Gate/grid resistor | Ω | 30 000 | 900 | 13 500 | 1650 | 10 000 | 1700 |
| Gate/grid dissipation | W | 20 | 150 | 45 | 130 | 40 | 130 |
| Gate/grid resistor dissipation | W | 20 | 200 | 65 | 220 | 49 | 430 |
| Cathode heating power | W | 79 | 403 | 176 | 1000 | 209 | 2400 |
| Output power (anode) | kW | 8·1 | 8·6 | 35 | 35 | 35 | 33 |
| Drive power (feedback) | W | 37 | 350 | 110 | 350 | 90 | 580 |
| Net output | kW | 8 | 8·2 | 34·9 | 34·6 | 34·9 | 32·4 |
| Efficiency (anode) | % | 68·0 | 70·0 | 68·0 | 68·0 | 73·0 | 68·0 |
| (overall | | 67·9 | 67·7 | 67·8 | 66·0 | 72·9 | 64·8 |

M—magnetically beamed triode   C—conventional triode

typically about 800°C below the temperature of grids in conventional valves. This means that anti-emissive coatings are no longer required.

A further important advantage results from the low operating temperature of the gate in that it allows the use of oxide-coated cathodes in place of the thoriated tungsten filaments used in conventional triodes, with a consequent saving in heater power of about 80%. Oxide cathodes cannot be used with grids operating at high temperature because any barium reaching the grid by evaporation from the cathode would cause excessive electron emission from the grid. The direction of this current would be in opposition to the normal grid current flow and would effectively reduce the negative bias. This is because the bias is derived entirely by the use of a grid bias resistor which, apart from being relatively cheap, effects some compensation by adjusting the bias against varying grid drive voltage. The loss of bias would lead, in turn, to increased anode and grid currents, higher grid temperature and electron emission and hence run-away conditions.

The drive power requirement for m.b. triodes is far below that required for a conventional valve. In the same proportion so are the gate dissipation and gate leak resistor dissipation. To achieve the required bias voltages the ohmic value of the leak resistor must be proportionally higher, but its physical size and cost are reduced roughly in proportion to dissipation.

Table 1 makes quantitative comparison between three m.b. triodes and their nearest equivalents (in terms of supply voltage and output power) among conventional triodes.

## 4 Oscillator Circuits

Magnetically beamed triodes are used in industrial oscillator circuits of conventional design such as Colpitts, or tuned anode/coupled grid (which is essentially a split-coil Hartley) and others. Since exactly the same voltage and power handling conditions prevail in the anode circuit, component values here are traditional. The differences occur in the control electrode circuit; as already mentioned, the leak resistor has a substantially higher value (typically 10 times greater). To avoid 'squegging' the gate coupling capacitor must be reduced by a corresponding factor. This presents no problem in increased reactive impedance since the drive current which it must pass is commensurately lower.

Figure 4 illustrates a typical tuned anode/coupled gate oscillator together with output power control by gate limiting circuit described in Section 3.3.

## 5 Output Power Control

The control in output power from an oscillator may be effected in several ways which divide into three main groups:

1. Control of the main supply voltage.
2. Change of load coupling away from that for optimum efficiency.
3. Control of the oscillator behaviour in the grid or gate circuit.

The first group sub-divides again into three classes:

1.1 The use of a variable transformer regulator. This is an expensive and bulky item limited to mechanical adjustment only.

1.2 Control of the e.h.t. power supply by delayed firing of thyratrons or, in modern technology, thyristors. These are high power (and relatively high cost) elements and are prone to the generation of spiky ripple on the e.h.t. voltage, and resultant troublesome feed-back into the electricity supply authority's distribution network.

1.3 Control of a series element between the e.h.t. supply and the oscillator circuit. This system is satisfactory at the expense of an extra power tube. Normally this is a saturable, tungsten filament diode but the m.b. triode itself offers easier electronic control, fail-safe features and economy. The m.b. tube has the advantages that the relatively high amplification factor at low anode currents makes the anode current at given bias level relatively independent of voltage drop: furthermore, the low gate currents permit the control voltage to remain steady in the positive region and within dissipation limits even at low anode voltages. As a 'fail-safe' feature the bias may be tripped out in the event of a detectable malfunction of the equipment to reduce the oscillator output power to a low level. By contrast, a low impedance conventional triode as a series regulator would normally operate in the negative bias region and loss of such bias would result in a dangerous, higher than normal current flow.

Considering the second group:

2.1 Power control by effectively mismatching the load to the oscillator is usually only achievable by mechanical means and leads to lower efficiency. Without reducing the e.h.t. voltage the only direction in which the load matching can be safely offset is by uncoupling, i.e. decreased effective load admittance. If this is done to limit the heating rate of a ferrous load below Curie point, the mismatch becomes even worse above that temperature just when tighter coupling is desirable. Overcoupling to a load results in higher anode current, lower efficiency and the risk of excessive dissipation in the valve anode.

The third grouping also sub-divides:

3.1 Interrupted cut-off technique

This was originally conceived for use with conventional tubes and is even easier to achieve with m.b. triodes because it is so much easier to develop a cut-off bias across their relatively high resistance grid (gate) leak component. The power control is effected by varying the mark space ratio in which oscillations are interrupted and the method does have the merit of running the oscillator at full efficiency during the ON periods. However, there is a serious disadvantage in that convenient interruption frequencies are in the audio frequency range, e.g. 50 (or 60) Hz and a corresponding low-frequency component becomes superimposed upon the radio frequency output of the oscillator. This can present an operator with a severe shock hazard if the work coil is inadvertently touched. Hence, the use of this system is inadvisable except in remote process applications.

3.2 Reduced drive coupling

This is limited in range because the oscillator efficiency is reduced and there is risk that oscillation will cease. Furthermore, such an arrangement would be mechanical and not readily adaptable to an electronic feedback system.

3.3 Variable gate loading techniques[6]

In early experiments with m.b. triodes, it was discovered that variation of the magnetic field (towards the knee region) produced a significant change in output power and that power control by a mechanical system of adjustable magnetic shunts was feasible. It was deduced that the fundamental effect of this alteration of magnetic field was change in gate current and that this could be effected electronically by a shunt element, either thermionic valve or semiconductor.

Initially shunt triodes and tetrodes were tried, but difficulty was experienced with screen grid dissipation or finding a triode of sufficiently low impedance for wide range power control. For low impedance a power transistor was more attractive as well as more fashionable. The use of a series connected (thermionic) diode to block the destructive reverse voltage led directly to the development of the gate-limiter system of power control which is the principal topic of the final Section of this paper.

## 6 Gate Limiting Power Control

The basic circuit is shown in the lower part of Fig. 4. A low-power, high-voltage diode D1 connects the gate of the m.b. power triode to a potential divider formed between resistor chain R1 (comprising a chain of about 100 kΩ per kilovolt e.h.t. voltage) and the power transistor shunt regulator. If the regulator is set in a turned-off condition, such that the potential at the diode cathode exceeds the peak of the m.b. triode gate drive signal, the diode is backed-biased throughout the whole r.f. cycle and the oscillator yields full power—the same as if no power controller were attached.

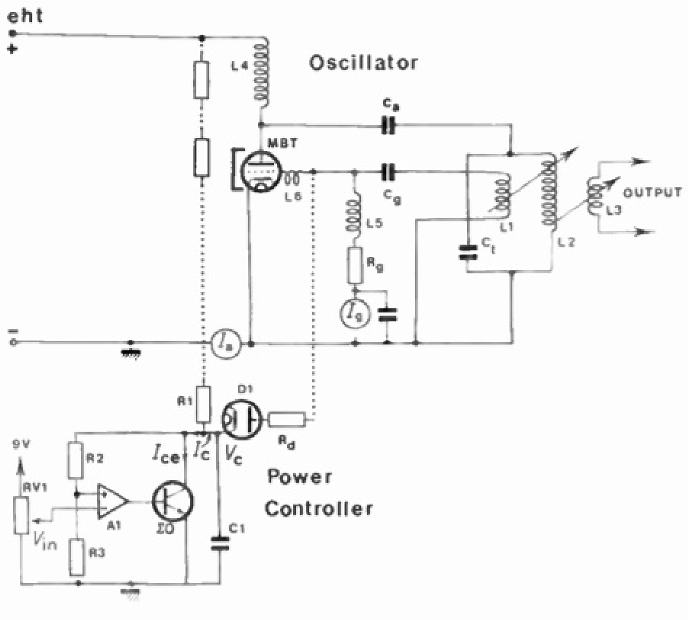



Fig. 4. Tuned anode/coupled gate oscillator with gate limiting circuit for output power control.

If the shunt regulator voltage is reduced until the diode begins to conduct at the crest of the drive waveform, the drive is reduced, the m.b. triode anode voltage swing is reduced, the oscillator output power is reduced but so, also, is the gate drive signal. Hence the oscillator settles to a new lower output power level. Thus, whilst the diode and regulator must have a dynamic impedance low compared with the drive source impedance of the oscillator (substantially the reactive impedance of capacitor $C_g$), they need clip only the crest of the new drive waveform rather than to load it at its original amplitude and this requires the controller to dissipate only a small power relative to that which it controls.

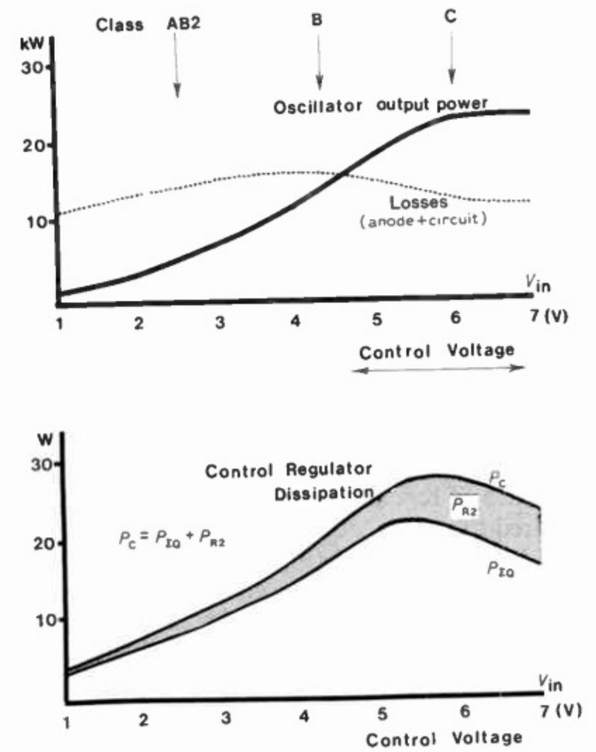



Fig. 5. Oscillator power and dissipation related to control setting.

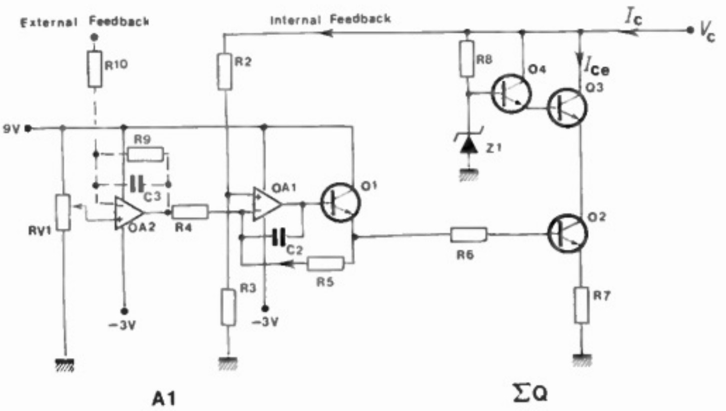



Fig. 6. Power control regulator.

The smoothness and range of such a power-controlled oscillator is typified by the output power curve in Fig. 5 which is plotted against control potentiometer voltage. It will be seen that the control regulator dissipation ($P_c$) is 3 orders smaller than the power controlled and is shared between the regulator transistor(s) $P_{\Sigma Q}$ (lower curve) and the feedback resistor $P_{R2}$ (shaded area).†

In its simplest form, the shunt regulators may comprise a single power transistor with its base connected to the level-set potentiometer. However, in most applications it is desirable to insert an amplifier, with differential input to enable a negative feedback signal to be applied via divider R2/R3 (Fig. 4).

R2 is typically 100 kΩ and R3 pre-set to about 750 Ω; thus at a maximum regulator voltage of 900 V the feedback is about 7 V, which is within the range that can be balanced by the input from RV1 (either direct or via OA2 as in Fig. 6). Lower inputs from RV1 cause the feedback via R2/R3 to stabilize the regulator voltage at the lower levels of $V_c$ for which RV1 has been set.

Capacitor C1 has a low reactive impedance at the oscillator frequency, removes ripple and thus makes the control regulator essentially a steady direct-current device.

The differential drive amplifier may comprise an operational amplifier OA1 (Fig. 6) plus a low-power transistor Q1 in emitter-followed connection to provide adequate current to the base of power transistor Q2 to turn it hard on for minimum oscillator power. Feedback loops (C2, R5) within this drive amplifier circuit limit overall gain to a suitable level (typically 10 dB set by ratio R5/R4) and provide stability. The reactance of C2 effectively shunts R5 and its low value (relative to R4 = 15 kΩ), at high frequencies restricts the frequency range to avoid response to any spurious signals.

The broken lines in Fig. 6 (A1) show how a second differential-input operational amplifier OA2 further splits the input options to feedback from an external sensor in addition to the manual level-set (RV1). The gain of OA2 is likewise set by the resistor ratio R9/R10 and the frequency response limited by the capacitor C3. The gain of OA2 coarsens the response to RV1 and its adjustment relative to the feedback signal is mentioned later.

† $P_{\Sigma Q}$ is the total power dissipated in transistors ΣQ in Fig. 4 and $P_{R2}$ is dissipated in R2.

## 7 Power Dissipation Ratings

Reference to Fig. 5 shows that oscillator power loss (mostly anode dissipation) and controller power loss both peak at intermediate power levels. The anode dissipation increases because the anode voltage swing is reduced and the anode current peak, though lower, occurs at a higher anode voltage. The relationships of voltages and currents at anode and grid (gate) of a power oscillator tube are usually analysed by reference to the tube's constant-current characteristics. Both anode and grid voltages are sine-wave and, for a resistively-loaded oscillator, their relationship is a straight line with its centre on the ordinate through $V_a = V_s$, the anode supply voltage. In Fig. 7, only one half of each operating line is shown. (In class B or C no current flows during the other half-cycle and it does not enter into the calculation of mean or peak-fundamental currents. Such calculations are discussed more fully in Appendix 1.)

At full output it is conventional and sensible to set the conditions to achieve the best efficiency in class C. As the gate voltage swing is limited, the gate current and, for fixed leak resistor, the gate bias $V_g = I_g R_g$ is reduced; hence the (half) operating line moves as shown through class B to AB2 with reduced input current but also reduced efficiency.

At low power levels the loss in efficiency is not serious because the input power is low. Many applications will use the oscillator near full output and considerations for high reliability in an industrial environment demands that adequate anode cooling be provided. Excellent margin is available in actual tube ratings.

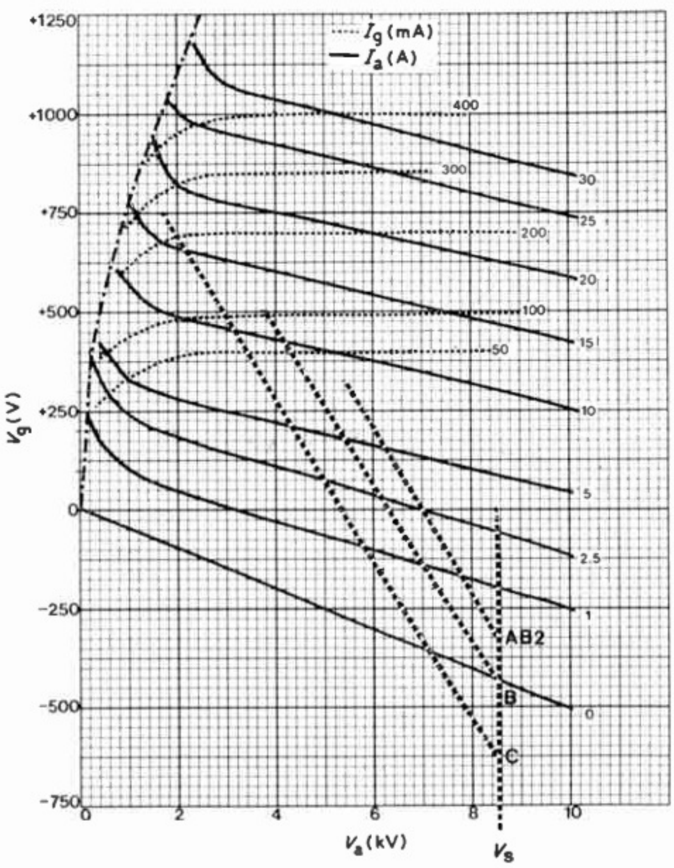



Fig. 7. Typical constant current characteristics (3RM/244G).

By the same token, the control regulator must also be reliable. Care must be taken with both dissipation and voltage ratings of the power transistor(s) and voltage/current plots of this part of the circuit need to be made over the full range of control of the oscillator with the various degrees of drive coupling and work load impedance likely to be experienced. If this $V_c/I_c$ plot is made on log-log graph paper the constant power curves are straight lines and may be readily added as an additional graticule; likewise the effective total impedance, though this is mostly of academic interest.

Figure 8 shows such a set of curves (thick lines) for a range of oscillator loading. (Tube type 3RM/244G operated at 8·5 kV anode voltage.) These dynamic curves intersect the (broken) static curves at the corresponding input control voltage, $V_{in}$, applied to resistor R4 (Fig. 6). Each dynamic curve corresponds to a particular load coupling. The left-hand curve corresponds to optimum coupling for maximum performance.

The static curves may be plotted by connecting the regulator (separated from a power oscillator) across a d.c. power supply plus a series load resistor. The independence of regulator voltage relative to the current it is required to pass demonstrates the effectiveness of the feedback loop via R2, R3 (Figs. 4 and 6). The locus of the 'knee' at the low current end of the static curves corresponds to an impedance marginally smaller than R2. It would be exactly equal to this value if the transistor chain ΣQ were to be cut off completely and R8 were infinite rather than the same order as R2.

Reduced gate drive impedance moves a given dynamic curve in the same direction as reduced load admittance and if the gate coupling capacitor ($C_g$ in Fig. 4) is larger than necessary it requires the transistor regulator to have an unnecessarily high dissipation rating. In this matter special design care needs to be exercised with respect to feedback coupling in a Colpitts oscillator circuit.

If the dissipation margin is insufficient with a single power transistor, a second transistor must be provided. Shunt connection is readily achievable with low value emitter resistors† to balance the current. Alternatively, if the voltage rating is insufficient, series connection must

† Positioned as R7 in Fig. 6.

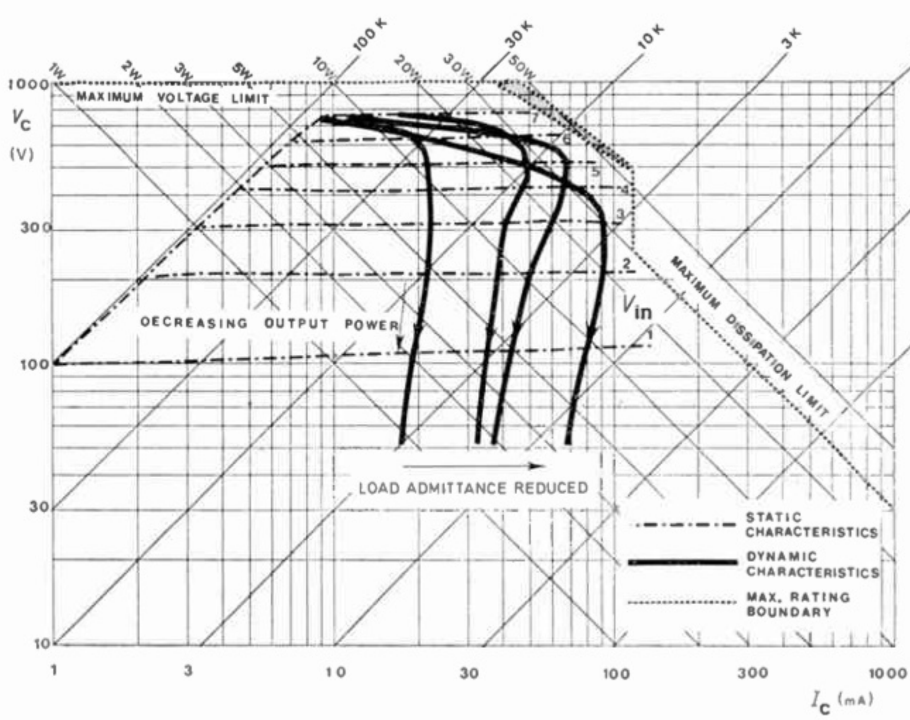



Fig. 8. $V_c/I_c$ plot for range of oscillator loadings for 3RM/244G.

be used. Equal sharing of voltage by a ladder of resistors linking the transistor bases presents problems in achieving the extremes of turn-on and turn-off. A more effective arrangement is shown in Fig. 6. When Q2 is turned on hard, the upper transistor Q3 is also turned on hard by current through R8 amplified by Q4. (Q4 must also have a high voltage rating but its dissipation is considerably lower than Q3.) The voltage across Q2 is low and as Q3 and Q4 are in emitter-follower connection, the voltage at the base of Q4 is not much higher and well below the conduction level of Zener diode chain Z1.

As Q2 is turned off by the level set (or external feedback), it takes most of the voltage $V_c$ until the Zener diode Z1 voltage is reached. Then the $V_{CE}$ of Q2 is held at that level and further voltage rise is taken up by Q3 (and Q4): much of the current through R8 is now directed from the base of Q4 through the Zener chain Z1.

By careful choice of Z1 Zener chain voltage and series/parallel combinations of Q2 and Q3, the voltage/dissipation rating boundary curve of the total unit can be tailored to provide a good safety margin relative to the most onerous $V/I$ curve plotted empirically. The rating boundary shown in Fig. 8 assumes a Zener chain (Z1) voltage of 250 V and that the $V_{CE}$ rating of Q2 exceeds this value: also that the $V_{CE}$ rating of Q3 and Q4 is 750 V and that the maximum dissipation of Q2 and Q3 is 30 W each at the maximum heat sink temperature to be encountered.

The shaded area of the boundary curve is the power dissipated by feedback resistor R2. Formulae defining the ratings boundary are given in Appendix 2.

## 8 Power Controller Applications

Some industrial heating applications merely need a manual control or presetting by the operator: other processes benefit, or indeed may only be successful if the change of power level is rapid, thus needing automatic feedback or programming. The low voltage and power level at the input end of the regulator driver circuits make them ideal for coupling to solid-state logic circuits for programming oscillator output power.

If it is desirable to maintain constant input power to the oscillator, the anode current (or, more readily, the cathode current which is nearly the same) may be monitored as a voltage by insertion of a low-ohmic resistor plus integrating circuit to obtain a mean level for feedback to the auto-control input (Fig. 9(a)).

Perhaps, more usefully, the r.f. output voltage can be kept constant by monitoring its level converted to low voltage d.c. via a capacitive potential-divider, detector diode and smoothing circuit (Fig. 9(b)).

If component values, including those in the regulator amplifier, are chosen for reasonable frequency response up to 300 Hz, then the ripple at that frequency from the usual 3-phase, full-wave rectified, e.h.t. power supply can be suppressed significantly.

Feedback may be from a point even nearer the load in the form of an electronic temperature sensor with an output voltage which may be suitably amplified for presentation to programmed logic control, or, for simple

(a) Automatic current regulation.

(b) Automatic r.f. voltage regulation.

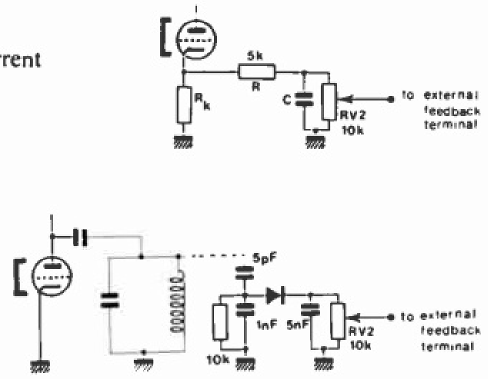



Fig. 9.

stabilization of load temperature, directly to the external feedback terminal of the regulator.

For practical realizations, it is convenient to develop the net external feedback voltage across a 10 kΩ potentiometer RV2, the sliding contact of which is connected to the regulator input. Since this input level acts in conjunction with, and in the reverse sense to, the manual preset (RV1 in Fig. 6), it is a useful setting-up technique to start with RV2 output at zero, set the oscillator output level with RV1 and finally adjust RV2 and RV1 alternately to keep the required output level whilst introducing sufficient feedback for automatic control of the process or stabilization of work-coil voltage or work-load temperature, as the case may be.

## 9 Conclusions

A new concept of industrial oscillator valve design and output power control has now been substantially proved in laboratory test and workshop floor experience in a number of industrial heating equipments of various designs and powers ranging from 6 to 30 kW.

These new methods of power control open the way for more refined industrial heat treatment techniques in addition to substantial cost saving.

## 10 Acknowledgments

The authors are indebted to a number of helpers, especially to Mr. G. H. Phipps who was involved in the detail of the valve engineering, to Mr. B. G. Kennett who pioneered the gate-limiting system of power-control, and to Mr. A. E. Miller who contributed significantly to the detail of the circuit design and carried through the extensive proving tests. Thanks are also due to the Management of ITT Components Group, Europe for permission to publish.

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## 12 Appendix 1 : Oscillator performance analysis[4]

The general-case locus of the $V_a/V_g$ relationship is an ellipse which virtually collapses to a straight line AGA′ for a resistively-loaded oscillator (Fig. 10). (Significantly reactive loads can cause serious tube dissipation problems.) In Fig. 10, G is at the centre of the line and has co-ordinates

$x$ = anode supply voltage $V_s$

$y$ = grid bias voltage $V_{gb} = -I_g R_g$

where $I_g$ is the mean grid (gate) current and $R_g$ is the leak resistor shown in Fig. 4.

As the grid voltage excursion is reduced by the gate limiting system of power control, the gate current falls and the reduction of $I_g R_g$ causes the movement of the point G to a less negative level. Point $G_1$ corresponds to the condition B in Fig. 7 and $G_2$ corresponds to condition AB2.

The locus of A, $A_1$, $A_2$ is not necessarily a straight line but it is approximately so in the present example. Line length AA′ reduces and finally vanishes at $V_a = V_s$, $V_g = 0$. In practice oscillation would cease before this point was reached. This situation does not occur with tubes of present design if the gate catching potential $V_c$ is not taken below zero.

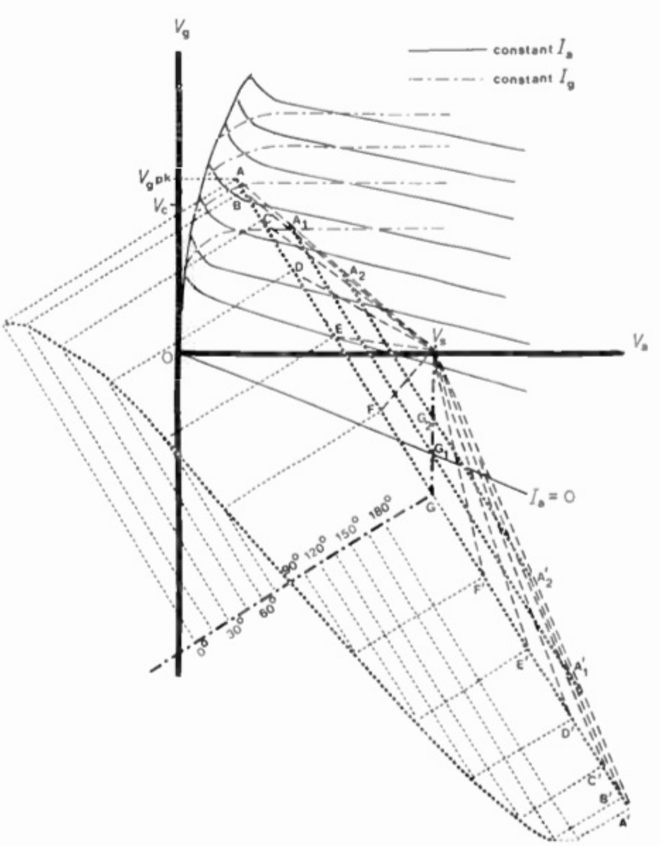



Fig. 10. Locus of $V_a/V_g$ relationship for analysis of oscillator performance.

For an oscillator under power control but with unchanged drive coupling, it may be expected that the slope of the operating line AA′ remains constant and the lines at lower powers are parallel to the original.

Practical measurements of mean anode or gate current may be compared with theoretical assessments obtained by summing the incremental values read off the characteristic curves at points A, B, C, D, E, F, G, F′, E′, D′, C′, B′ and A′ which are at cos 15° intervals. (More frequent intervals increase accuracy but not significantly so.) Thus

$$I_{mean} = \tfrac{15}{360}[(A+A')+2(B+B')+2(C+C')+\ldots+ +2(F+F')+2G].$$

For class AB2 this complete formula must be used, but for classes B and C the formula simplifies to[4]

$$I_{mean} = \tfrac{1}{12}[\tfrac{1}{2}A+B+C+D+E+F+G].$$

The same incremental readings of anode and gate current are used to calculate the peak fundamental components of these currents from the formula

$$I_1 = \tfrac{1}{12}[(A-A')+2\cos 15^\circ(B-B')+2\cos 30^\circ(C-C')+ +2\cos 45^\circ(D-D')+2\cos 60^\circ(E-E')+ +2\cos 75^\circ(F-F')].$$

Normally, a transparent film is used on which is printed a graticule comprising a series of parallel lines ranging in length but all divided in the same cosine ratios. This graticule is placed over the characteristic curves and eliminates the need to draw on the graph itself.

If a study begins with a paper design, $R_g$ is chosen to obtain the bias level selected at G from the product $I_g R_g$, where $I_g$ is the mean gate current calculated.

Where analysis of a working oscillator is required, point G is fixed by the d.c. anode voltage and the product of $R_g$ and the $I_g$ reading obtained from the meter. Point A is not as readily fixed. The $V_g$ peak level may be obtained using a suitable peak-reading voltmeter or, if a power controller is effectively clipping the drive voltage, $V_g$ peak $= V_c + \Delta V$ where $\Delta V$ is the voltage drop across the clamping diode and its series resistor and is typically about 50 V. The precise value of $V_{a(min)}$ is not readily obtained by measurement but some trial at positioning A will usually yield a condition which gives mean $I_a$ and $I_g$ readings in close agreement with meter readings.

If output power can be measured, for example with a water-cooled calorimeter as load, a further check on the analysis of working conditions can be made since the output power

$$P_0 = \tfrac{1}{2}(V_s - V_{min}) \,.\, I_{a1}$$

where $V_s$ = anode supply voltage

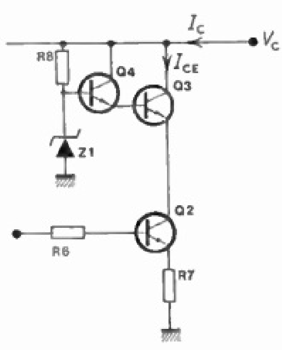
$V_{min}$ = lowest anode voltage reached

$I_{a1}$ = peak fundamental anode current.

In practice about 85% of this power reaches the load. The other 15% is absorbed as heat losses in the oscillator circuit.

Such theoretical assessments combined with practical measurements have shown good correlation and confirmed the action of the gate limiter on the operating conditions.

## 13 Appendix 2: Control regulator ratings boundary (Figures 6, 8 and 11 refer)



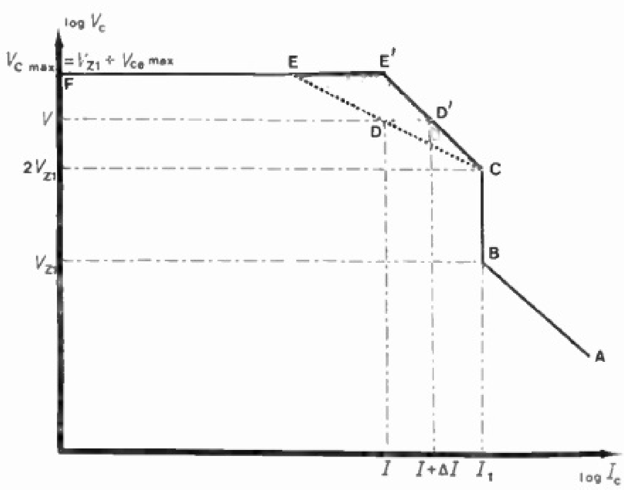




Fig. 11. $V_c/I_c$ relationship for control regulator.

BOUNDARY SECTION AB

High-power transistors usually have a maximum current rating in excess of 1A and the pertinent limit for $V_{CE}$ greater than 30 V is the maximum dissipation. This will be related to heat-sink area and ambient temperature range. At voltage $V_c$ below the Zener diode chain voltage $V_{z1}$, transistor Q2 takes virtually all of the voltage and AB is its dissipation alone. $V_{z1}$ is less than the $V_{CE}$ maximum rating of Q2.

$$I_1 = P_{Q2}/V_{z1}.$$

BOUNDARY SECTION BC

At voltages $V_c$ above $V_{z1}$, transistor Q3 takes a share of the voltage and its dissipation limit is reached also at C when it drops an equal voltage $V_{z1}$, making the total $V_c = 2V_{z1}$ at C where

$$P_c = P_{Q2} + P_{Q3} = 2V_{z1}I_1.$$

To be precise, Q3 handles about 10% less current at its collector than Q2 because Q2 also passes the Q4 collector current. For simplicity in the formulae, Q3 current is assumed to be the same as Q2 current and errs on the side of additional safety.

BOUNDARY SECTION CDE

At voltages $V_c > 2V_{z1}$, transistor Q3 takes the larger share of the voltage and the common current level at the limit must be reduced to keep the dissipation of Q3 within its limit.

At an arbitrary point D, the dissipation of Q3 will be at its maximum rating, $P_Q = I(V - V_{z1})$ and that of transistor Q2 will be $IV_{z1}$ making a total limit of $P = P_Q + IV_{z1}$ (which is also equal to $IV$ as must be the case).

BOUNDARY SECTION CD'E'

At voltages above $2V_{z1}$, the additional current $\Delta I$ via the resistor feedback chain R2, R3, becomes more significant and extends the regulator total dissipation boundary to CD'E.

At D:

$$\Delta I = \frac{V}{R2+R3} \simeq \frac{V}{R2} \text{ since } R3 \ll R2$$

and the tota. dissipation $= V(I+\Delta I) = P_Q + V_{z1}I + \dfrac{V^2}{R2}.$

BOUNDARY SECTION E'EF

Eventually transistor Q3 will reach its maximum $V_{CE}$ limit when

$$V_c = V_{CEmax} + V_{z1}.$$

The ratings boundary is then independent of current.

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