The G2DAF Linear Amplifier

By G. R. B. Thornley (G2DAF)*

OVER the years the increasing use of single sideband transmission by amateurs has developed new interest in linear r.f. power amplifiers. Most sideband operators have a continuing desire to put new ideas to the test and have constructed and used a wide variety of linears employing many different valves and classes of operation. This experience has contributed greatly to the data available for practical designs.

The circuits to be discussed in the present article are the result of experiments made by the writer some time ago. The interest at that time was in some method of simplifing

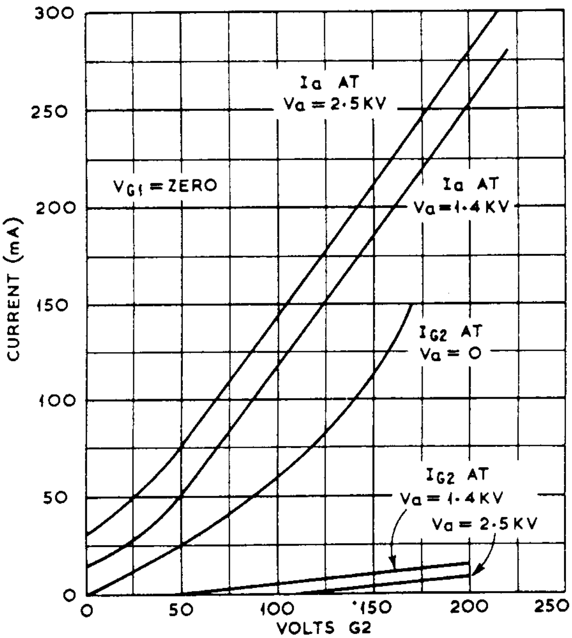

Fig. 1. Static anode and screen characteristics for two QY3-125 tetrode valves, with control grid held at zero potential.

existing circuits so that a tetrode or pentode valve could be used without any bias supply and in addition without the complication of screen dropper resistors, stabilizer valves, clamp valves or additional power packs. Initial experiments were undertaken using a pair of Mullard type QY3-125 (4-125A) valves in parallel with a passive grid input circuit.

Initial trials confirmed that the valves could be operated with zero bias and that very good control with a wide range of anode current was possible by variation of the screen voltage from zero to some positive value. A series of tests and measurements were made to determine the static anode-screen characteristics with differing anode voltages. Graphs were made from the readings obtained and these are shown in Fig. 1.

From the manufacturers' data for two QY3-125 valves operating in class AB1 with an h.t. supply of 2·5 kV, the maximum signal (single tone) anode current is 222 mA (111mA each valve). Inspection of the graph of Fig. 1 showed that the screen control was linear over a range of anode current from the resting value to a figure that was greater than the manufacturers' maximum-signal rating. Additionally the degree of control was little affected by wide variation in anode potential. It was further noted that the screen driving power was very small. At 2·5 kV anode potential with 625 watts input, the figure is 6 mA at 200 volts = 1·2 watts. At 1·4 kV anode potential, the figure is 10 mA at 190 volts = 1·9 watts.

Under *dynamic* operating conditions each positive r.f. cycle on the amplifier control grid causes the anode current to swing from its zero-signal resting point up to the maximum-signal value. At the same time the anode voltage swings down. This is the moment of time at which the screen current rises to its maximum value. If the instantaneous anode voltage drops *too* low, the screen current will rise to an excessive value. The graphs of Fig. 1 show this quite clearly. Under V_a = 1400 volt conditions $I_{g2\ max}$ =

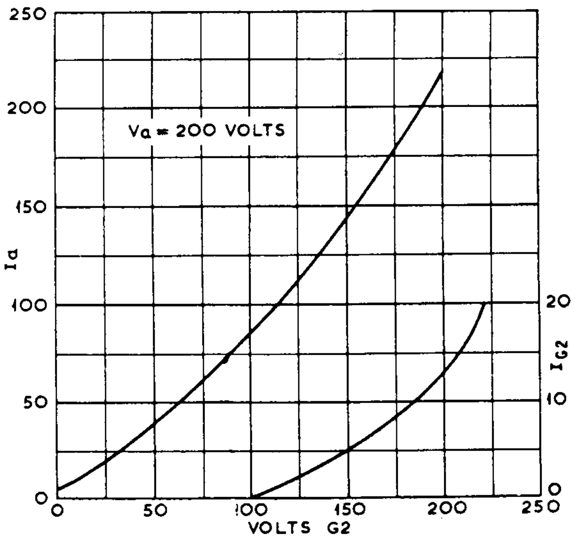

Fig. 2. Graph showing the anode and screen currents, under static conditions, for differing values of screen voltage with the anode held constant at 200 volts.

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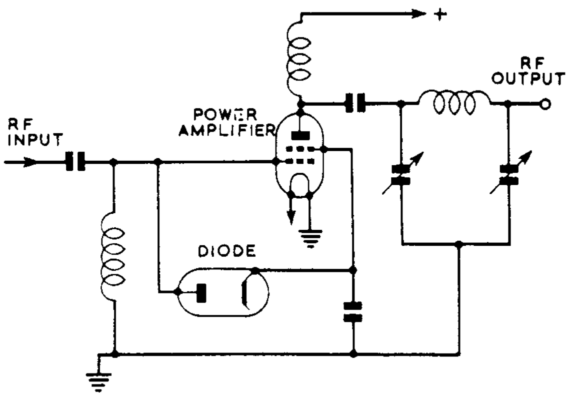

Fig. 3. Basic circuit diagram for the G2DAF Linear Amplifier.

10 mA. With $V_a = 0$ volts, $I_{g2\ max}$ has risen to 160 mA.

It was considered that in practice with the usual h.t. supply, the minimum value of $V_a\ min$ would be about 200 volts, and would represent normal operating conditions. Tests were then made and measurements taken of the screen characteristics with the amplifier anode potential held at a constant 200 volts. The screen potential was increased from zero in 20 volt increments and readings taken of both the screen and anode currents. These were plotted as a graph and are shown in Fig. 2.

As had been expected the ratio of screen current to voltage was quite low. At 200 volts on the screen the current value was 13 mA. Even at the point where the screen had risen above the anode potential by 20 volts and a quick rise in current would be expected, the value was only 20 mA. It was noted with increasing interest that the screen driving power at an anode current of 230 mA and a screen potential of 200 volts was the low value of 2·6 watts.

At this stage sufficient data was available from the plotted static characteristics to build up a mental picture of the valve operation under dynamic conditions. This was itemised as follows:

(i) The ratio of anode current to screen voltage should show excellent linearity.

(ii) The "zero signal" anode current and therefore the static anode dissipation should be a low value.

(iii) The zero-signal anode current should be at a suitable value for class AB operation over a wide range of anode h.t. supply voltage.

(iv) The maximum screen driving power should be less than 5 watts.

The following conclusion was reached: a screen driving power of 5 watts is a small proportion of the r.f. available from the exciter to drive the amplifier. It should therefore be perfectly feasible to obtain the required screen potential directly from the amplifier input signal.

The next step was to draw out the basic circuit diagram shown in Fig. 3 and convert the existing class AB1 amplifier to the new arrangement for " on the air " tests. During the autumn of 1961 this amplifier was used at G2DAF on the 80m band and the opportunity taken to obtain a considerable number of signal reports. The reporting stations were informed that a new experimental amplifier was in use and asked to comment particularly on the quality of the speech and the level of the intermodulation products. All stations reported favourably in regard to clean and smooth speech quality and low distortion product level.

Having obtained this verification in regard to the practicability and soundness of the theoretical approach, a further series of graphs was plotted to show the amplifier characteristics under dynamic conditions with both single-tone and two-tone input signals. From an analysis of these curves it was apparent that the required maximum-signal screen voltage was only being developed by r.f. drive voltages that produced a rather large peak grid current. This difficulty was overcome by a further development of the existing principle that gave a more favourable ratio between V_{g1} and V_{g2}.

Amplifier Operation

As may be seen from Fig. 4 the finalized amplifier makes use of two diode rectifiers in a Cockcroft-Walton voltage doubler circuit. This gives approximately twice the screen potential for the same r.f. drive voltage and results in an improved maximum-signal handling capability with a useful economy in the required driving power. The associated

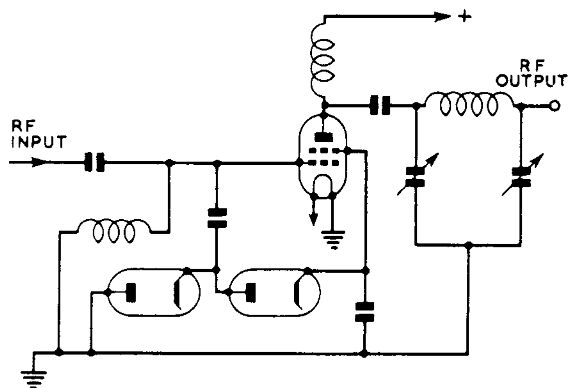

Fig. 4. Basic circuit diagram using Cockcroft-Walton voltage doubler circuit.

charging capacitor values are chosen to offer a low impedance to the signal input frequency, but a high impedance to all voice frequencies. The positive voltage on the screen is derived solely from the r.f. input signal and will at all times vary in sympathy with, and be directly proportional to, the amplitude of the modulation envelope. The screen voltage is therefore varying from zero in a positive direction at the frequency of the voice modulation.

It will be appreciated that as the amplifier valve is operating at zero bias, grid current is drawn throughout the 180° of each positive r.f. input cycle. Considering the input side of the amplifier only, the operation is class B. In regard to the output side of the amplifier the action of the valve is more complex than conventional operation because there is no reference point on the I_a/V_g characteristics against which to plot the anode current flow.

Consider first the dynamic characteristics for conventional class B operation shown in Fig. 5(a). This is a plot of the anode current in relation to the grid voltage for a particular valve at a rated fixed screen voltage. At zero bias the anode current would be very high as shown, and it would be necessary to bias the grid negative to reduce the standing anode current to a value that would maintain the static anode dissipation within the manufacturers' maximum rating. The r.f. driving signal swings either side of a reference point determined by the amount of the necessary negative bias voltage. Each half cycle drives the grid positive and the anode current increases in a linear manner. This increase can be plotted against the I_a/V_g curve to show the anode current pulse and the duration of the anode current flow.

In the G2DAF amplifier there is no bias and the grid potential is zero; at the same time the anode current is also at a low value because there is no screen voltage. With a small r.f. input signal the screen voltage will rise slightly above zero and the anode current will increase by a small amount—the I_a/V_g curve will appear as shown in Fig. 5(b). The important point to note is that the anode current flow is now much greater than the 180° of class B working. In fact anode current is flowing for 360° of the grid swing and the amplifier is operating in class A.

Consider next what happens when the input signal increases. The screen potential rises and the standing anode current rises. However, the grid is still at zero bias with the r.f. driving signal swinging equally positive and negative about the zero bias reference point. In effect the I_a/V_g curve has moved over towards the left of the graph as shown in Fig. 5(c). It will be noted that the positive half of the grid swing is now having more effect on the anode current flow than the negative half. The anode current flow is more than 180° but less than 360° and the amplifier is operating in class AB. At greater signal strengths the I_a/V_g curve moves further over, the zero bias operating point moves farther up the slope, anode current flows during a smaller part of the grid swing and the amplifier is approaching class B operation, as shown in Fig. 5(d).

The curve of Fig. 5(a) is representative of the usual tetrode or pentode operating in the conventional manner. Because of the curvature of the I_a/V_g line towards the cut-off point the valve is no longer linear with small grid swings. It is this curvature that produces non-linear operation and the increased generation of intermodulation distortion products. In the G2DAF amplifier, a small input signal produces a small screen voltage. The total length of the I_a/V_g curve becomes smaller and the sharpness of cut-off is much improved. In effect this is the curve of Fig. 5(d) reduced in size—the non-linear portion of this curve is reduced accordingly! All the graphs are drawn to the same scale and this point is shown clearly by comparison of Fig. 5(b) with Fig. 5(d). Because of this, and the class A operation at small signal levels, the linearity of the amplifier is improved with a corresponding reduction in intermodulation distortion product level.

In regard to the operation of the amplifier at maximum signal levels it will be noted that the grid swing of Fig. 5(a) and Fig. 5(d) are identical. However, the anode current pulse of the G2DAF amplifier is greater than the anode current pulse of the conventional amplifier. From this it follows that for the same maximum-signal input power, the G2DAF amplifier requires less driving voltage. Conversely, for the same driving voltage the amplifier will run to a greater power output.

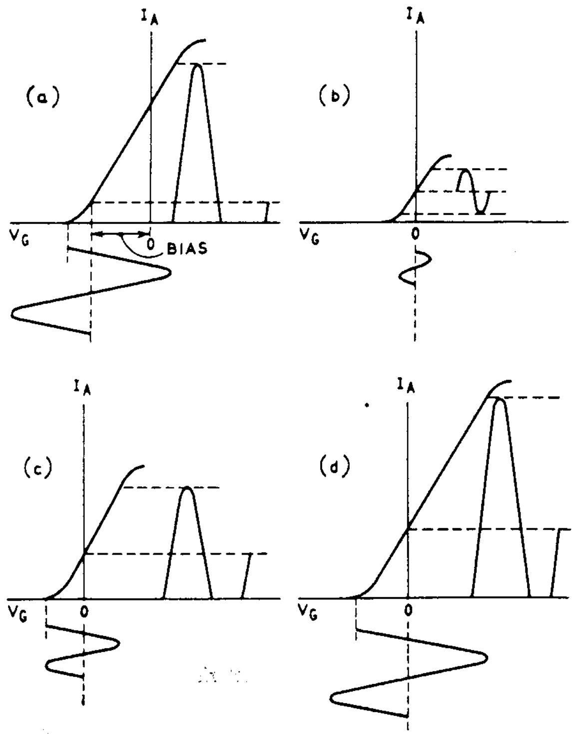

Fig. 5. Dynamic characteristics showing at (a) conventional ABI operation and (b), (c) and (d), the G2DAF amplifier with differing input signals.

Dynamic Characteristics

The most convenient method of expressing amplifier characteristics is in graphical form, and the mutual characteristics—where the anode current is plotted against the r.f. grid voltage—are particularly useful. From these it is possible to check (i) the amplifier linearity, (ii) the anode and screen current at different excitation levels, (iii) the maximum signal anode current, and (iv) the maximum signal power input. These characteristics are shown for both single-tone and two-tone input signals in Fig. 6. The r.f. grid input voltage was measured with a diode probe valve voltmeter, and the I_a and I_{g2} current values are the readings taken from the amplifier panel meters.

In regard to two-tone input power it is important to appreciate that the graph has been plotted using values of the anode current read on the panel milliameter. This is not the true maximum current value because the meter cannot follow at the two-tone (audio) rate. Fortunately the relationship between peak envelope input and indicated d.c. power input is accurately known. With a class B linear amplifier operating with low distortion and biased to cut-off, the peak envelope power input is 1·57 times the d.c. input as measured with a two-tone test signal. However this factor applies only in the case of a perfectly linear true class B amplifier.

In practice, amplifiers are never true class B and are run with less than cut-off bias. The relationship becomes: $I_{peak} = 1{\cdot}57\ (I_{dc} - 0{\cdot}363\ I_o)$ where I_{peak} = peak current; I_{dc} = anode meter reading; I_o = zero-signal (idling) current. At 2·5 kV anode potential the zero signal anode current is 30 mA as shown in Fig. 6 and the correct value of $I_{a\ peak}$ is $I_{a\ dc}$ multiplied by 1·5.

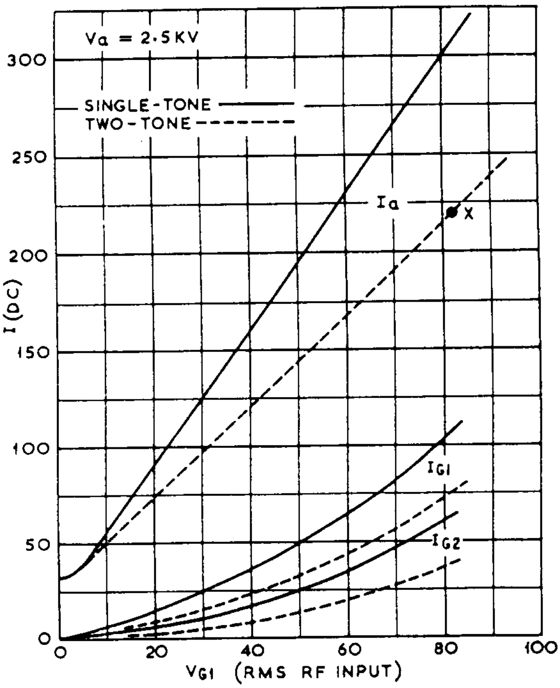

Fig. 6. Dynamic characteristics, showing anode, screen and control grid current for excitation voltages from zero to 80 volts for two QY3-125 valves.

Referring again to the graph of Fig. 6, the two-tone input current at point X is 220 × 1·5 = 330 mA and the input power is 825 watts. At this power level the amplifier was operating in a perfectly linear manner and the two-tone envelope monitored on the oscilloscope showed a clean cross-over, symmetrical lobe pattern and no evidence of flat-topping. The screen (I_{g2}) current is also plotted as a function of the excitation (V_{g1}) voltage on the same graph. It will be noted that the amplifier runs to a two-tone p.e.p. input of 660 watts for an r.f. grid swing of 63 volts, and at this level the screen current is 22 mA. By reference to the graph of Fig. 7 the screen voltage at this drive is seen to be 75 volts. The product of the two gives the screen driving power of approximately 1·65 watts.

Amplifier Efficiency

A considerable number of measurements have been made of the amplifier power output into a non-inductive load, at power input levels from zero up to the maximum possible. All figures for output have been based on r.m.s. voltage measurements across a 100 ohm dummy load. The amplifier tuning and loading was adjusted for maximum r.f. output, consistent with adequate loading to obtain a satisfactory two-tone envelope without flat-topping or other distortion. Measurements have been made under single-tone and two-tone input conditions.

At each power level the ratio (output watts)/(input watts) was calculated as a percentage. These percentage figures were then plotted against the corresponding value of input power. However, before considering the amplifier efficiency in detail the writer would like to make some general comments.

There is a considerable amount of confused thinking in regard to linear amplifier efficiency, and many amateurs do not realize that the often quoted figure of 66 per cent efficiency—for a "perfect" class B amplifier—only applies when the amplifier is operating at full output with maximum exciting voltage and under single-tone input conditions. At any other power level the efficiency is less than 66 per cent and is in fact proportional to the exciting voltage, i.e., if the drive voltage is halved the efficiency becomes 33 per cent; drive voltage one-third, efficiency 22 per cent—and so on. There is a simple reason for this:

Consider an amplifier with a grid drive of V_g, an anode current of I_a, an h.t. supply voltage of V_{ht}, an r.f. voltage across the tank coil of V_{tank}, and an r.m.s. voltage across the dummy load R of V_{load}.

Power input at full drive $= V_{ht} \times I_a = P_{in}$ watts.

Power output at full drive $= (V_{load})^2/R = P_{out}$ watts.

Consider now the conditions at half grid drive:

Excitation $= \frac{1}{2} V_g$. Anode current $= \frac{1}{2} I_a$. Tank voltage $= \frac{1}{2} V_{tank}$. Load voltage $= \frac{1}{2} V_{load}$.

Power input $= V_{ht} \times \frac{1}{2} I_a = \frac{1}{2} P_{in}$ watts.

Power output $= (\frac{1}{2} V_{load})^2/R = \frac{1}{4} P_{out}$ watts.

From this the statement follows: With a linear amplifier the d.c. anode current and the anode input power will be proportional to the exciting voltage, whereas the output power is necessarily proportional to the square of the exciting voltage.

A graph of amplifier efficiency plotted against power is shown in Fig. 8. In order to give a direct basis of comparison, the efficiency of a perfect class B amplifier operating with a single-tone input and zero no-signal anode current is also plotted to the same scale with the 66 per cent reference

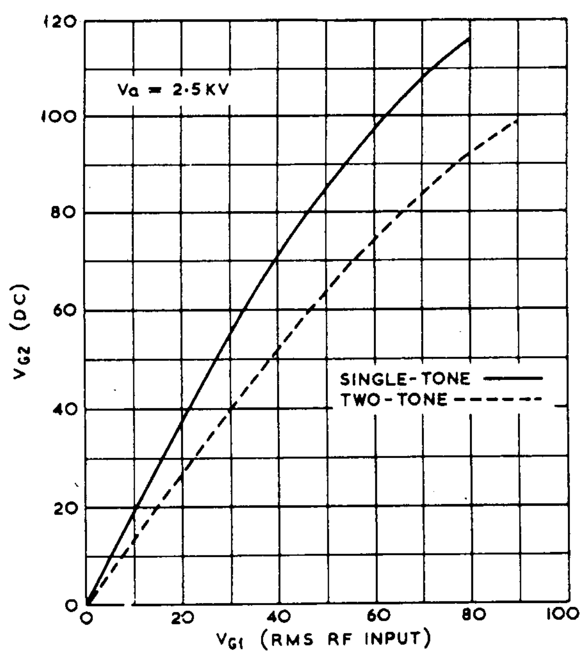

Fig. 7. Dynamic screen characteristics, showing the screen potential as a function of r.f. excitation voltage.

point taken as 750 watts. If the amplifier is operating in a linear manner, by definition—with a linear amplifier the d.c. anode current and the anode input power will be proportional to the exciting voltage, whereas the output power is necessarily proportional to the the square of the exciting voltage—the plot of either single-tone or two-tone efficiency should be a straight line. The graphs shown in Fig. 8 are curved because at each of the input measuring points, the amplifier tuning and loading was adjusted to give the maximum r.f. output consistent with satisfactory linearity. This has the effect of increasing the value of R_L and gives improved operating conditions at the lower power levels and therefore a greater output for the same input power. This procedure would be adopted in practical operation and is therefore perfectly legitimate.

It is normal procedure for valve manufacturers to quote power output and efficiency as measured at the anode. In practice the power output is calculated from voltage readings across a known value of non-inductive dummy load. Obviously the value obtained—and any efficiency figure based on this value—would be less than the *anode* efficiency because it includes transfer (tank circuit) loss. In order to be able to give a true basis of comparison for the G2DAF amplifier the power output has been corrected to allow for tank circuit loss, and the efficiency figures quoted in Table 1 and shown in Fig. 8 include this correction.

This gives an anode efficiency at maximum signal conditions of 65 per cent, almost equal to a conventional "perfect" class B amplifier. Finally, before concluding this section the writer would like to make it quite clear that he does not claim an efficiency for the G2DAF amplifier of 65 per cent at maximum drive level, based on an estimated transfer loss correction factor and power output measurement alone. It is generally possible to find an alternative measuring technique and this has been used as a check against the first.

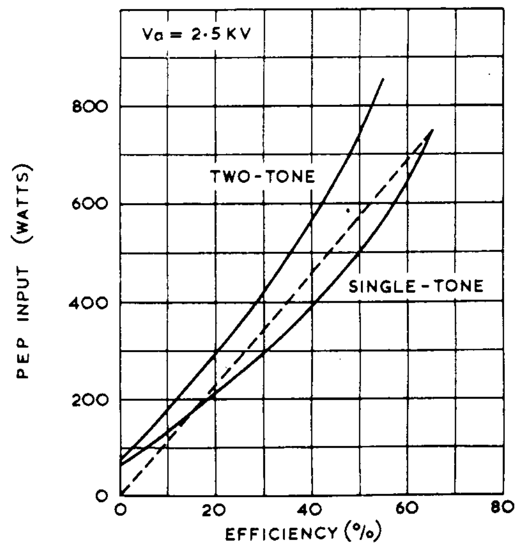

Fig. 8. Graph showing input power and efficiency for single-tone and two-tone signals. The dotted line shows for comparison the efficiency of an imaginary "perfect" class B amplifier operating with zero bias and 66 per cent efficiency at 750 watts input power.

TABLE 1

Maximum signal operating conditions for an h.t. supply of 2·5 kV using two QY3-125 valves.

	Single-tone	*Two-tone*
Anode current (d.c.)	300 mA	220 mA
Power input (d.c.)	750 watts	550 watts
P.E.P. input	750 watts	825 watts
Anode dissipation	260 watts	250 watts
P.E.P. output	490 watts	600 watts
Power output (mean)	490 watts	300 watts
Anode efficiency	65 per cent	54 per cent

NOTE These ratings are in excess of permissible power output within the conditions of the G.P.O. Amateur (Sound) Licence.

Amateur Band operating conditions (400 watts p.e.p. output) for two QY3-125 valves.

V_a = 2·5 kV	*Single-tone*	*Two-tone*
Anode current	250 mA	175 mA
Power input (d.c.)	650 watts	440 watts
P.E.P. input	650 watts	660 watts
I_{g2}	38 mA	22 mA
I_{g1}	70 mA	45 mA
V_{g2}	105 volts	75 volts
V_{g1} (r.m.s.)	65 volts	64 volts
Driver Load (p.e.p. out)	30 watts	35 watts
Anode Dissipation	250 watts	240 watts
Power Output (mean)	400 watts	200 watts
P.E.P. Output	400 watts	400 watts
Anode Efficiency	61·5 per cent	46 per cent

Typical operation with 300 ohm grid swamping resistor.

With an anode supply of 3 kV the efficiency figures will show an improvement on those quoted. It is estimated that they will be as follows:

Single-tone efficiency = 72 per cent. Two-tone efficiency = 58 per cent.

Amplifier efficiency can be accurately checked by measuring the temperature of the valve envelope by a thermometer or thermocouple in close contact with the glass. With a single-tone input driving the amplifier to an indicated anode meter reading of 260 mA and an h.t. supply of 2·5 kV, the input power is 650 watts. At this input the envelope temperature indicated a dissipation of 250 watts (125 watts per valve) and this is an anode efficiency at 650 watts input of 61·5 per cent.* This experiment has been repeated with a two-tone input driving the amplifier to an indicated anode meter reading of 194 mA at 2·5 kV. This is a d.c. input of 480 watts for 250 watts anode dissipation giving a *mean* output of 230 watts and an anode efficiency at 480 watts d.c. input (720 watts p.e.p.) of 48 per cent.

It will be appreciated that the numerical difference between the input power and the dissipation power is the power output from the amplifier—at the anode. If simultaneous output measurements are made across the external dummy load, the discrepancy between the "load" figure and the

* The anode temperature for 250 watts dissipation is determined by running the amplifier up under static condition—no r.f. drive. Connect a 20 K or 25 K ohms 3 watt, wire wound, potentiometer across the exciter h.t. supply and the slider to one of the terminals of the amplifier screen current meter. Increase the potentiometer from zero until the required anode power is being drawn. At 2·5 kV supply voltage, this would be 100 mA. An alternative but less accurate method is to observe the colour of the anode at the rated dissipation for the valve in use. The amplifier is then run under single-tone input to the same anode colour and the p.e.p. input calculated from the product of the h.t. supply voltage and the steady anode current meter reading.

"anode" figure will indicate the transfer power loss in the tank circuit. This method was used to determine the correction value included in the output and efficiency figures that have been given. Really accurate r.f. power measurement is not easy and there must always be some possibility of error. However, the writer has attempted by repeated checking and by using two dissimilar methods to arrive at the right answer. The efficiency figures quoted and those plotted on the graph of Fig. 8 are given in good faith that they are reasonably correct.

Due to the method of operation, the screen voltage is always proportional to the exciting voltage, and is very much lower than the normal value for class AB operation. Because of this the anode can swing down to a much lower value without excessive screen current. The numerical value of $V_a - V_{a\ min}$ is the term in the formula:

$$\textit{Anode Efficiency (single-tone)} = \frac{3 \cdot 14}{4}\left(\frac{V_a - V_{a\ min}}{V_a}\right)$$

$$\textit{Anode Efficiency (two-tone)} = \left(\frac{3 \cdot 14}{4}\right)^2 \times \left(\frac{V_a - V_{a\ min}}{V_a}\right)$$

This governs the efficiency figure that is obtainable.

It is usual with power tetrodes to find that the two-tone efficiency figure is 15 to 20 per cent below the single-tone figure. In the G2DAF amplifier the maximum two-tone efficiency is particularly good and is in fact only 11 per cent below the peak efficiency obtainable with a single-tone input. This is attributed to the self-compensating action of the circuit under two-tone input conditions, where the mean screen potential is actually lower for the same p.e.p. input, thus allowing a further improvement in the ratio $V_a - V_{a\ min} / V_a$, with increased efficiency and greater power handling capability without exceeding the rated anode dissipation.

Driving Power

The amplifier valves are operating with zero bias and are drawing grid current throughout the positive half cycle of the r.f. signal, and they require driving power. The grid drive requirements are similar to those of any other class B amplifier—from 2 to 5 watts or so—depending on the valve type. Additionally there is the screen driving requirement, and allowing for a small loss in the rectifiers, this is likely to be 2 to 4 watts—again depending on the operating conditions for the valve or valves in use.

The final requirement is the loss in the passive grid resistor and this will depend on the value used. This resistor not only stabilizes the linear amplifier and makes neutralizing unnecessary, but additionally provides a constant load for the driver valve—in this respect the lower the value, the better the performance. Taking as a specific example two QY3-125 valves, the maximum signal requirements are as follows (approximate values):

Driving Power, Grid	= 5 watts.
Driving Power, Screen	= 3 watts.
Dissipation, Grid Resistor of 300 ohms	= 17 watts
Total	= 25 watts.

(There is additional loss in the driver tank circuit and this transfer loss must be added to the total.)

A golden rule is to provide a driver stage with a p.e.p. output rating of double the calculated driving requirements. Therefore:

Driver p.e.p. output = 50 watts

This ensures that the driver is never overrun and never produces a distorted signal before it goes into the power amplifier. There is no point whatsoever in building a low distortion amplifier and then driving it with a signal that is already distorted by non-linear operation of the driver stage.

Two 6146 valves with 500 volts anode potential are rated at p.e.p. output of 57 watts, or with 600 volts anode potential 76 watts, and would in practice be very suitable for the requirement. If the exciter output was less than this—for instance a single valve with half the above ratings, the passive grid resistor would have to be increased in value to say 400 ohms. The figures would then be (approximate values):

Driving Power, Grid	= 5 watts.
Driving Power, Screen	= 3 watts.
Dissipation, Grid Resistor of 400 ohms.	= 10 watts.
Total	= 18 watts.

Allowing a 2 to 1 safety margin:
Driver p.e.p. output = 36 watts.

The correct value of passive grid resistor can be made up using standard 1 watt carbon resistors in a series parallel arrangement. Under two-tone input conditions the average output power is one half the p.e.p. rating. Therefore the wattage dissipation of the passive grid network need only be one half of the maximum signal figure shown, i.e., the 400 ohm resistor could be made up with five 1 watt resistors.

Because of the original method of operating the valve, manufacturers' figures for conventional AB1 or AB2 tetrode or pentode operation no longer apply. The grid driving voltage is smaller and the grid current larger than data sheet figures. An approximation for grid driving power that is near enough for all practical use is:

$$\textit{Maximum Signal Driving Power} = V_g(r.m.s.) \times I_g(d.c.)$$

For initial calculation $V_g(r.m.s.)$ can be taken as approximately half the peak voltage normally quoted for class AB1 or AB2. $I_g(d.c.)$ is the grid current meter reading. When the amplifier is completed and driven to the required maximum signal (p.e.p.) output with a single-tone signal the actual grid voltage can be measured with a diode probe valve voltmeter. (These instruments are normally peak reading but are calibrated in r.m.s. values.)*

The d.c. screen voltage can also be checked under the same driving conditions using a standard AVO or similar test meter, and this voltage multiplied by the screen current meter reading is the approximate screen driving power.

The exciter at G2DAF uses two 6146 valves in the driver stage with an h.t. supply of 650 volts. As a check against the calculated driving requirements of 25 watts (300 ohm grid loading resistor) one of these valves was removed. Operating the exciter with the remaining 6146 valve, it was found just possible to fully drive the linear amplifier to the maximum allowable 400 watts p.e.p. output, either under two-tone or single-tone conditions. Allowing for the inevitable loss of

* An accurate figure for grid driving power is obtained from the formula:

$$\frac{V_{g\ peak} \times I_{g\ peak}}{4}$$

where $V_{g\ peak}$ is 1·4 times the r.m.s. value as measured with a diode probe valve voltmeter, and $I_{g\ peak} = I_{g\ d.c.} \times 3 \cdot 14$. This power is being dissipated at the grid in the form of heat—if the energy dissipated exceeds the manufacturers' rating, the grid wires may become white hot and melt. The most punishing form of input is single-tone (continuous carrier). Under voice input conditions it is permissible to allow the grid current meter to swing on peaks to a slightly higher value.

r.f. due to normal transfer losses, this is consistent with the estimated requirements.

Practical tests indicate that the same driving power—to either a G2DAF amplifier or a conventional class AB1 amplifier—can produce approximately the same maximum signal anode current. This occurs because zero-bias operation requires a smaller r.f. excitation voltage than normal AB1 operation and there is therefore a smaller power loss in the passive grid resistor. With low values of grid resistor, this can almost compensate the screen driving requirement.

Construction

The complete circuit diagram of the G2DAF linear is shown in Fig. 9. The diode rectifiers may be either semi-conductors of the point contact type suitable for r.f. use, or alternatively thermionic valves. Suitable rectifier valves are already available at low initial cost and were used in the original amplifier. The main requirement is a good heater-cathode insulation and a heater rating suitable for the additional 6 volt winding generally provided on standard p.a. heater transformers. Suitable valves are the Brimar 6U4G or the Mullard EY81. The Brimar valve is preferred because the anode connection is brought out to a base pin and, as this is underneath the chassis, it is screened from the p.a. output circuits.

It will be noted that the circuit is inherently simple and straightforward. The only panel controls necessary are the anode tuning, band change, aerial loading and a small switch to bring in the second section of C1 as required. The simplicity of the r.f. side enables a compact layout to be used without any risk of instability. This is shown in detail in the chassis and panel diagrams in Fig. 10.

The pi tank coil is wound on an Eddystone 2½ in. diameter ceramic former grooved eight turns per inch, and this is attached to the switch (from a TU5B unit) before it is fitted to the panel. A gap of one groove is left between the 15 and 20, and 20 and 40m sections. The total winding length of 16 s.w.g. tinned copper wire is 3¼ in. For the 10m band the coil of six turns of 12 s.w.g. wire is spaced to approximately 2 in. long and is self-supporting—with the axis at right angles—between the tuning capacitor and the end of the main tank coil.

The pi tank values depend on the required value of anode load (R_L) and as with any amplifier it is important that the valves are operating into the correct load. If they are not, both the power handling capability and the efficiency will suffer. Assuming that V_a is the h.t. supply voltage; $V_{a(min)}$ the instantaneous anode voltage at its lowest point; $I_{a(dc)}$ the maximum signal anode current meter reading; then $I_{a(peak)} = I_{a(dc)} \times K$ (where K is a constant dependent on the angle of current flow—in this case approximately 3) and $R_L = 2(V_a - V_{a\ min})/I_{a\ peak}$. For the amplifier under consideration R_L = 2(2500 0—250/300 × 3 = 5000 ohms.

The pi constants are then R_L = 5000 ohms; $R_{L(out)}$ = 75 ohms. The ratio $R_L/R_{L(out)}$ = 5000/75 = 66. The square root of 66 is approximately 8 and this is the reactance ratio (XC1 : XC2). For a Q of 12:

$$XC1 = R1/Q = 5000/12 = 416 \text{ ohms}$$
$$XC2 = XC1/8 = 416/8 = 52 \text{ ohms}$$
$$XL = XC1 + XC2 = 416 + 52 = 468 \text{ ohms}$$

These values are a simple approximation but are quite near enough for amateur purposes. From the reactance chart the values for 80m are C1 = 116 pF; L = 20 μH; C2 = 900

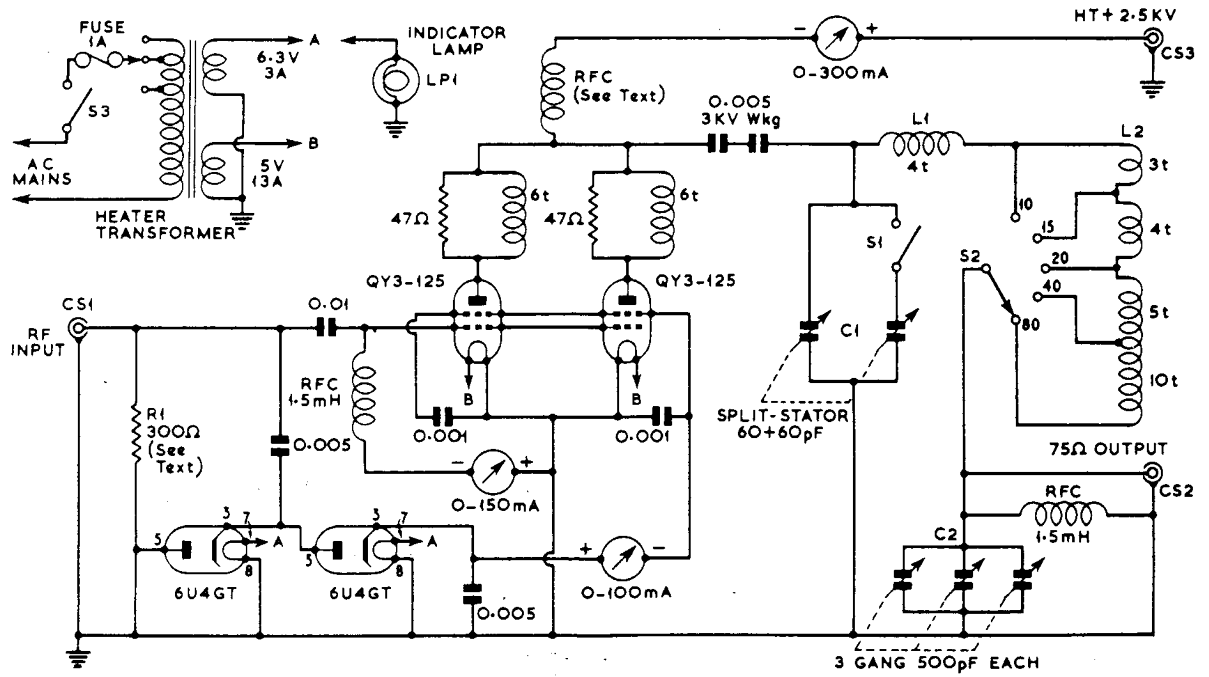

Fig. 9. Circuit diagram of the G2DAF Linear Amplifier. S1 open for 20, 15 and 10 metres, closed for 80 and 40 metres. All fixed capacitors mica except 0·005 μF,3K volt anode blocking which are disc ceramic. For CS3 Belling Lee chassis sockets and coaxial plugs are available coloured red. S1 and S2 ceramic switches. (All connections to panel meters via coaxial cable.) For winding details of L1 and L2 see text.

pF. Values for other bands scale down in the same ratio as the band wavelength as follows:

Band	C1	*L*	C2
80m	116	20	900
40m	58	10	450
20m	29	5	225
15m	20	3·5	160
10m	15	2·5	113

The type of transmitter tuning capacitor suitable for C1 generally has semi-circular rotor plates and therefore a large minimum capacity value—usually about 15 to 20 pF. This, together with circuit and valve anode capacity, will make up a total that is greater than required for the 10 and 15m bands. It is possible to overcome this in two ways, (i) redesigning the tank circuit for a higher *Q* value of, say, 15 or 20; (ii) reducing the minimum capacity of C1. A high value of *Q* in the tank coil will increase the circulating r.f. currents and therefore the losses. Accordingly the second expedient has been adopted and the tank capacitor was, in fact, made into a two-gang unit of 60 pF each section by sawing through the bars holding the stator plates. One section only is connected to the 10m coil and the anode r.f. blocking capacitor and this tunes the three higher frequency bands. The other section is switched in parallel for the 40 and 80m bands. In addition to reducing the minimum capacity value, this method also doubles the dial bandspread and makes tuning less critical on the 10, 15 and 20m bands.

The required air gap for C1 is approximately one-tenth of an inch. A standard three gang broadcast tuning capacitor of 500 pF each section is suitable for C2 and provided the amplifier is working into a load (as it should be) the plate spacing is ample to prevent flashover.

The r.f. choke comprises 300 turns of 32 s.w.g. enamelled wire wound in unequal sections—165, 65, 35, 20 and 15—on a ceramic former 1 in. in diameter and 5½ in. long with a ⅛ in. spacing between each section. Standard multi-section pie wound r.f. chokes are unsuitable for pi tank circuits and should not be used. A standard 1·5 mH r.f. choke rated for at least 300 mA is connected across the output co-axial socket as a safety precaution that should never be omitted when high voltages are in use. Should there be failure of the r.f. blocking capacitor the h.t. current through the choke will blow the main h.t. fuse and prevent h.t. voltage reaching the aerial circuits.

Operation

Tuning and loading is exactly the same as a conventional class AB amplifier. Initially the drive level is increased until the anode current meter reads 150 or 200 mA. With the loading capacitor fully meshed, the anode tuning is adjusted for a dip in anode current. With the tank circuit at resonance, the screen current will be a high value. As the loading is increased by reducing the capacity of C2, the anode current will rise and the screen current will fall in the usual manner. The drive can now be increased until the grid, screen and anode currents are the required values.

Should the amplifier have been built using some other type of valve, the manufacturers' figures for class AB1 or AB2 working can be used initially. If an oscilloscope is available the amplifier should be driven with a two-tone input and the modulation envelope monitored on the c.r.t. It is then a simple matter to adjust excitation, tuning and loading for maximum r.f. output consistent with adequate loading to prevent flat topping or other distortion of the modulation

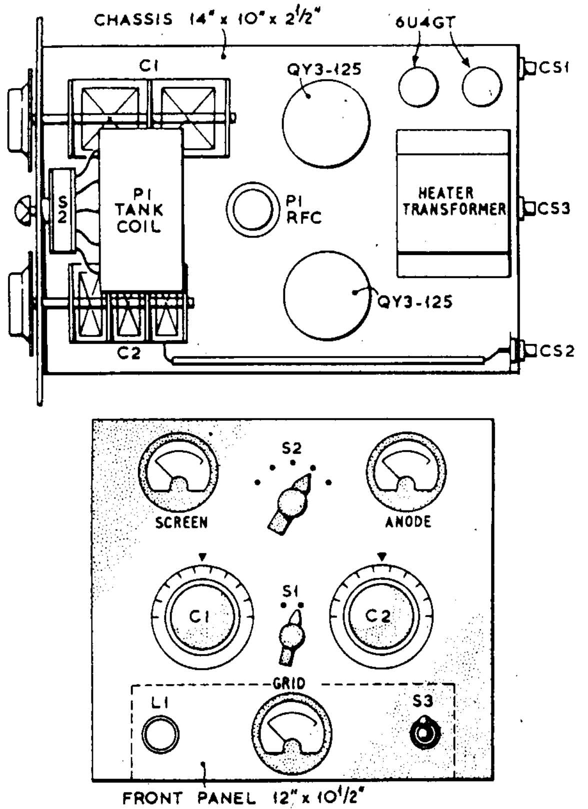

Fig. 10. Chassis and panel layout for the G2DAF Linear Amplifier.

envelope. The grid, screen and anode currents are then noted and in all subsequent operation the amplifier is adjusted to obtain these values. If an oscilloscope is not available, the loading should be adjusted so that the dip in anode current is not more than about 20 per cent of the off-resonance value, i.e. 250 mA off-resonance—load to 200 mA at resonance. A golden rule to observe is, "If in doubt, load heavily!" Under speech conditions adjust the exciter audio gain or r.f. drive so that the anode meter does not swing beyond half the steady signal value. Remember that the meter movement cannot follow at syllabic-rate—if it swings up to 150 mA, the true maximum signal current is at least twice this value.

The question of harmonic generation and TVI as a result of the method of operation is of importance to all amateurs. This can only be answered by stating that careful measurement of harmonic output using the identical amplifier under (1) conventional class AB1 conditions, and (2) the G2DAF method of operation, indicates clearly that there is not in fact any appreciable difference between the two methods. Additionally, at the writer's home station during 18 months continuous operation—much of it during peak viewing hours—there have been no complaints.

Conclusion

There are many different methods of operating a linear power amplifier, and it is of course possible to dispense with bias and screen supplies by using zero bias triodes. It is also possible to dispense with neutralization by using grounded grid operation. However, the grid driven tetrode or pentode linear power amplifier has become increasingly popular among amateurs and there are many good reasons for this. A method of operation that can provide tetrode advantages and at the same time give the simplicity of zero bias triode operation is sure to be of interest to a large number of single sideband workers.

The advantages of the writer's method of operation can be summarized as follows:

(i) Greater power handling capability.
(ii) Increased efficiency.
(iii) Low static anode dissipation and exceptionally cool running.
(iv) No screen power supply, dropper resistors, voltage regulators or clamp valves.
(v) No bias power supply.
(vi) No wasted power.
(vii) No adjustment or setting up procedure necessary.
(viii) An inherently high safety factor, i.e. the p.a. valves cannot be damaged by failure of either grid, screen or anode voltages.
(ix) Simple and stable operation.
(x) Heavy grid swamping resulting in a more constant load to the driver valve.
(xi) Very low intermodulation distortion product level.
(xii) High power gain. (Power gains of 10 are quite practicable.)

The method of operation is suitable for any of the commonly used tetrode or pentode amplifier valves including the 4-65, 4-125, 4-250, 4-400, 4X150A, 4X250B and 813. With regard to h.t. supply requirements, the amplifier will operate satisfactorily over a wide range. For instance the writer has used the QY3-125 valves with a supply of 1400 volts, and received favourable signal reports. As a matter of interest the zero signal anode current for the two valves at this voltage is 10 mA, representing a resting anode dissipation of 14 watts. The only proviso in regard to anode voltage is that there is sufficient to draw a small amount of static anode current, i.e., the valve must not be operated at cut-off or beyond.

It is seldom possible in this world to get something for nothing; the G2DAF amplifier is no exception to this rule and the many advantages have to be paid for. The price is a little more driving power than is usual for conventional class AB1 passive grid operation. However, in recent years it has become fashionable to build exciters with larger driver valves, and many amateurs today are operating small transmitters—both home constructed and commercially made—rated for 50 or 100 watts p.e.p. output and have driving power to spare.

The circuits and methods of linear power amplifier operation described in this article are protected by British Patent No. 926081. Amateurs who wish may nevertheless construct and operate the G2DAF linear amplifier for their own personal use.

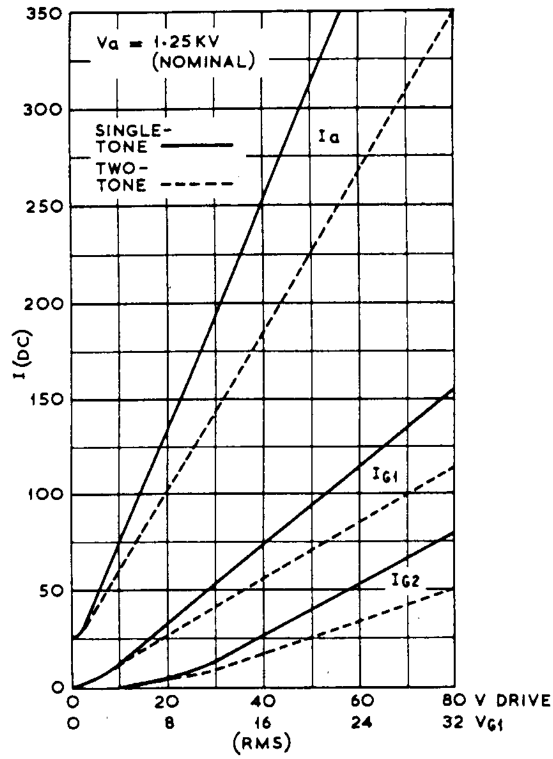

Fig. 11. Dynamic characteristics showing the anode, screen and grid currents. For excitation voltages from zero to 80 volts for two 4X150A valves.

LINEAR AMPLIFIER USING TWO 4X150A VALVES

Maximum Signal Conditions
V_a = 1·25 kV (nominal)

	Single-tone		*Two-tone*
V_a	1200 volts	1000 volts	1100 volts
I_a	350 mA	400 mA	350 mA
$P_{in\ (dc)}$	420 watts	400 watts	385 watts
P.E.P. $_{in}$	420 watts	400 watts	578 watts
I_{g2}	50 mA	60 mA	50 mA
I_{g1}	105 mA	125 mA	115 mA
V_{g2}	94 volts	110 volts	85 volts
V_{g1}	22 volts	26 volts	32 volts
$V_{drive\ (r.m.s.)}$	56 volts	64 volts	80 volts
P_{diss}	180 watts	144 watts	195 watts
P.E.P. $_{out}$	240 watts	256 watts	380 watts
$P_{out\ (mean)}$	240 watts	256 watts	190 watts
Efficiency	57 per cent	64 per cent	50 per cent

(The variation in anode voltage is due to poor regulation of the power supply).

Appendix

The small metal anode, high slope valves of the 4X150A and 4X250B class are becoming popular for linear amplifier use. These valves have a close electrode spacing and a low maximum grid dissipation rating. In the case of the 4X150A this is 2 watts maximum for a single valve.

Because of the high slope the required grid drive voltage is much smaller than usual, and under zero bias conditions, two 4X150A valves can be driven to 130 mA d.c. grid current with an excitation of only 36 volts peak. Using the formula: $V_{g\ peak} \times I_{g\ d.c.} \times 3.14/4$, this is equivalent to a grid dissipation of 3·6 watts—very close to the maximum allowable. However, this small value of excitation would only produce a screen potential of approximately 50 volts and this is insufficient to allow the valves to draw the required maximum signal anode current. It follows therefore that, (i) a greater r.f. drive voltage is necessary to produce the required screen potential, (ii) because of the low dissipation rating, the whole of this voltage cannot be applied to the control grid. The practical solution is the very simple one of tapping the grid feed point down the passive grid resistor. This then becomes a potential divider and any desired ratio of V_{drive} : V_{g1} can be selected to give the optimum operating conditions for the particular high slope amplifier valve in use.

The writer has had the opportunity of undertaking some measurements on an experimental G2DAF amplifier using a pair of 4X150A valves. At the time the measurements were made the only 1250 volt power supply available was one with a rather poor voltage regulation, and at full load current the anode potential dropped to 1100 volts. This is not the optimum operating conditions to get the best performance. With a full 1250 volts on the anode the figures for both power output and efficiency would have been improved.

The dynamic characteristics are given in Fig. 11 showing the anode, screen and grid currents, for different excitation levels from zero up to 80 volts r.m.s.

It will be seen from Fig. 12 that the grid is tapped down the passive grid resistor (total value 150 ohms) so that it receives 40 per cent of the total drive voltage. This is not necessarily the optimum ratio for the best possible performance, and there is in fact ample scope for the experimentally minded amateur to find the right ratio to suit his own particular valves and operating parameters.

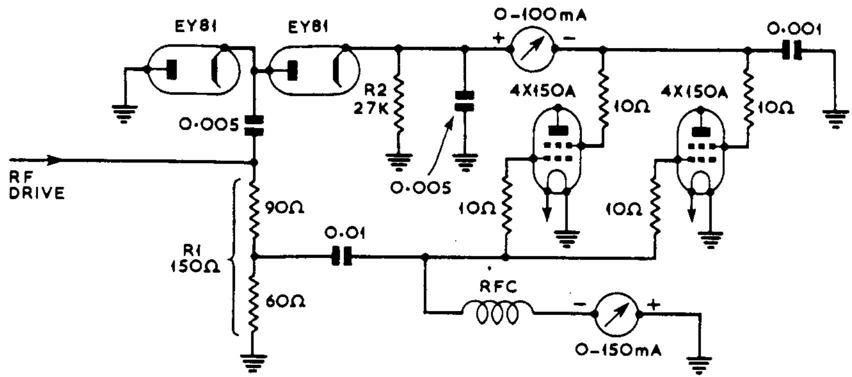

Fig. 12. G2DAF Linear Amplifier using two 4X150A valves, showing modified input circuit. R1 passive grid resistor.

Most transmitting tetrodes employing oxide-coated cathodes exhibit negative screen current under certain conditions of loading. This occurs because secondary electrons are emitted by the screen grid. Small values of negative screen current are not detrimental to valve operation and are quite normal for some tetrodes. Under conditions of negative screen current there would be no return path through the two diode valves for the reverse electron flow. Accordingly an additional resistor R2 of 27 K ohms is included in the screen circuit as shown. It is most important that under all operating conditions there is a return path from the screen-grid to earth, and this resistor is an essential part of the circuit and must not be omitted.