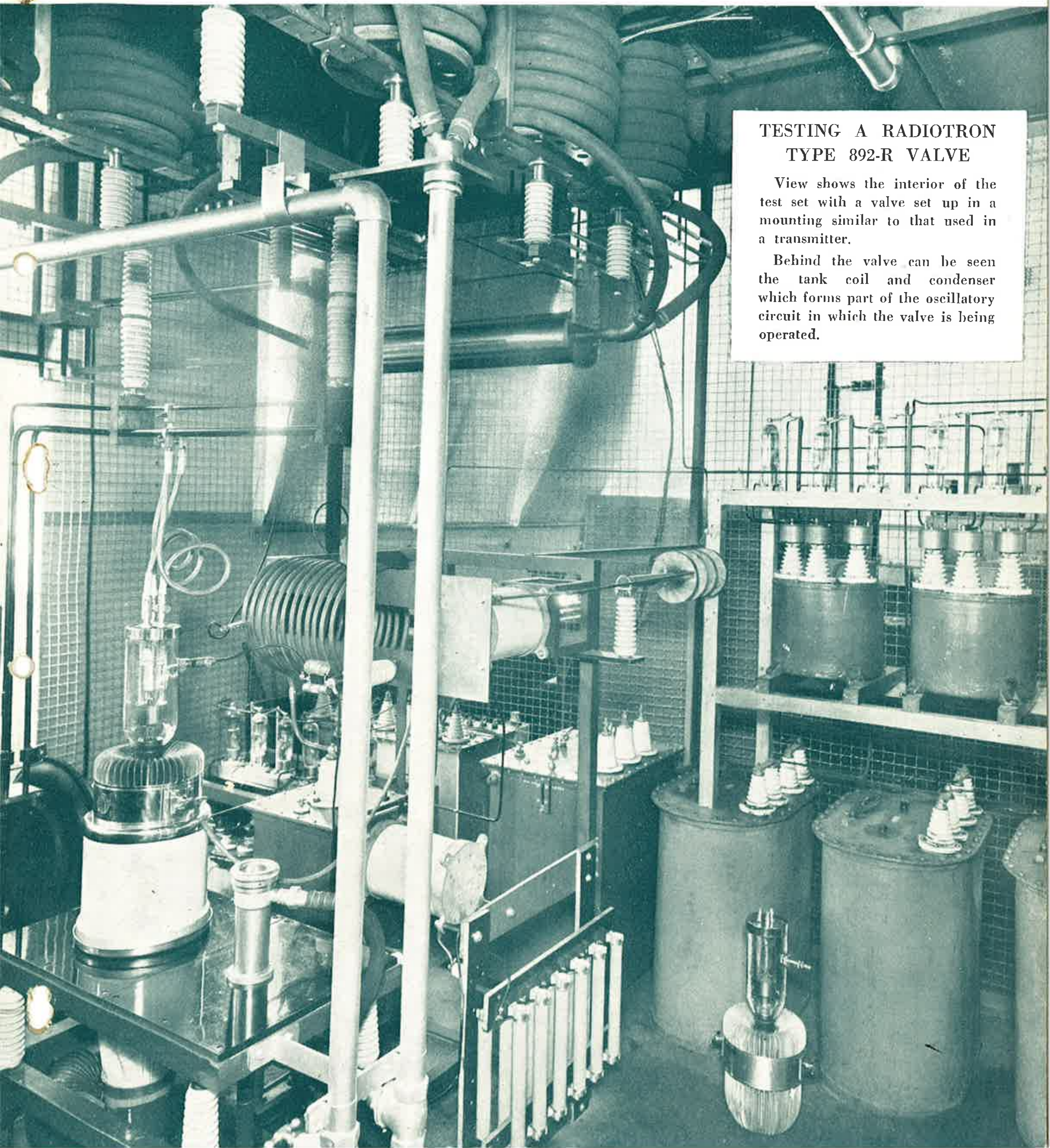
Radiotronics

Number 134 NOVEMBER — DECEMBER 1948



TESTING A RADIOTRON TYPE 892-R VALVE

View shows the interior of the test set with a valve set up in a mounting similar to that used in a transmitter.

Behind the valve can be seen the tank coil and condenser which forms part of the oscillatory circuit in which the valve is being operated.

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IN THIS ISSUE:

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Technical Editor

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All communications in Australia should be directed to Amalgamated Wireless Valve Co. Pty. Ltd. (Sales Promotion Department), 47 York Street, Sydney, N.S.W. — in New Zealand to Amalgamated Wireless (Australasia) Ltd., P.O. Box 830, Wellington, C.1, N.Z.

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New High Gain Converter Valve Radiotron Type X61M

The X61M is a frequency changer of the triode hexode type, having a high conversion conductance and a comparatively low cathode current giving a good signal to noise ratio. It is suitable for operation up to a frequency of 60 megacycles per second.

The triode hexode has been recognised as possessing advantages over other frequency changer valves, especially when operation is required at the high frequencies. The X61M has the following advantages over other triode-hexodes:

1. High conversion conductance for comparatively low cathode current.
2. Low wattage heater.
3. Large control ratio when operated with a fairly constant screen voltage.

Data sheets giving curves, circuits, and application notes are being circulated with this issue of Radiotronics.

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New AC/DC Series

A complete new series of valves for a.c./d.c. and d.c. receivers has recently been added to the Radiotron range. They all have a heater current of 0.16 ampere and are suitable for supply voltages from 200 to 260 volts when used with a barretter or dropping resistor, or from 117 volts when connected straight across the mains. The range includes

X76M	triode hexode converter	KT71	power output tetrode
W76	remote cut-off pentode	U76	half-wave rectifier
DH76	duplex-diode triode	161	barretter.

All valves (other than the barretter) fit standard octal sockets. For valve data, reference should be made to the loose leaf data sheets. We give below useful information regarding their application.

Valve Type X76M

The X76M is a frequency changer of the triode hexode type, having a high conversion conductance and a comparatively low cathode current, giving a good signal to noise ratio. It is suitable for operation up to a frequency of 60 megacycles per second.

The X76M is supplied with a metallised bulb only.

The triode hexode has been recognised as possessing advantages over other frequency changer valves, especially when operation is required at the high frequencies. The X76M has the following advantages:

1. High conversion conductance for comparatively low cathode current.
2. Large control ratio when operated with a fairly constant screen voltage.

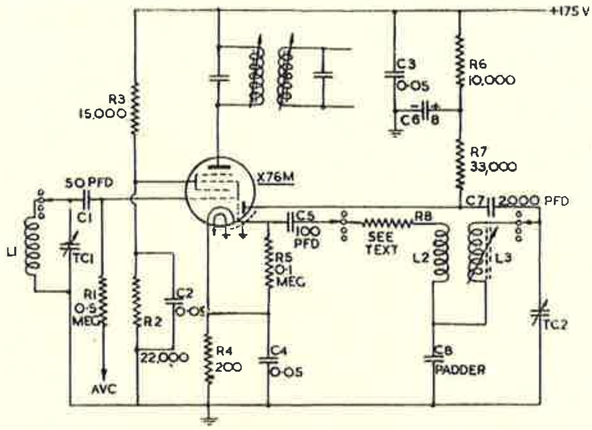


FIG. 1
NORMAL CIRCUIT

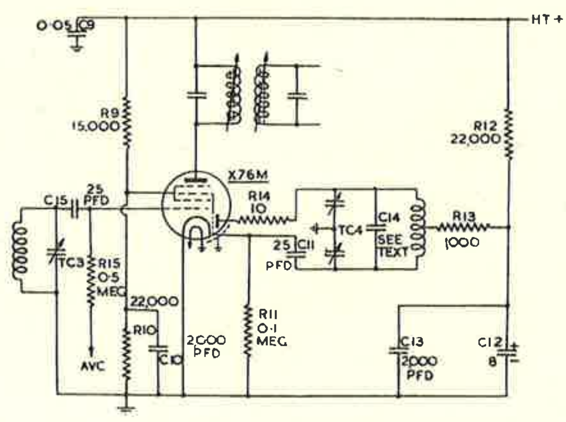
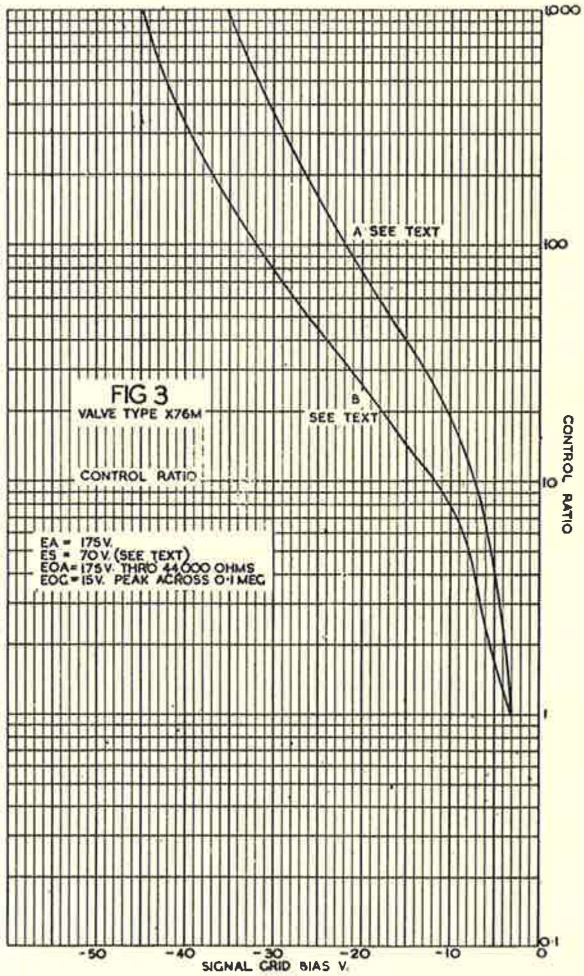
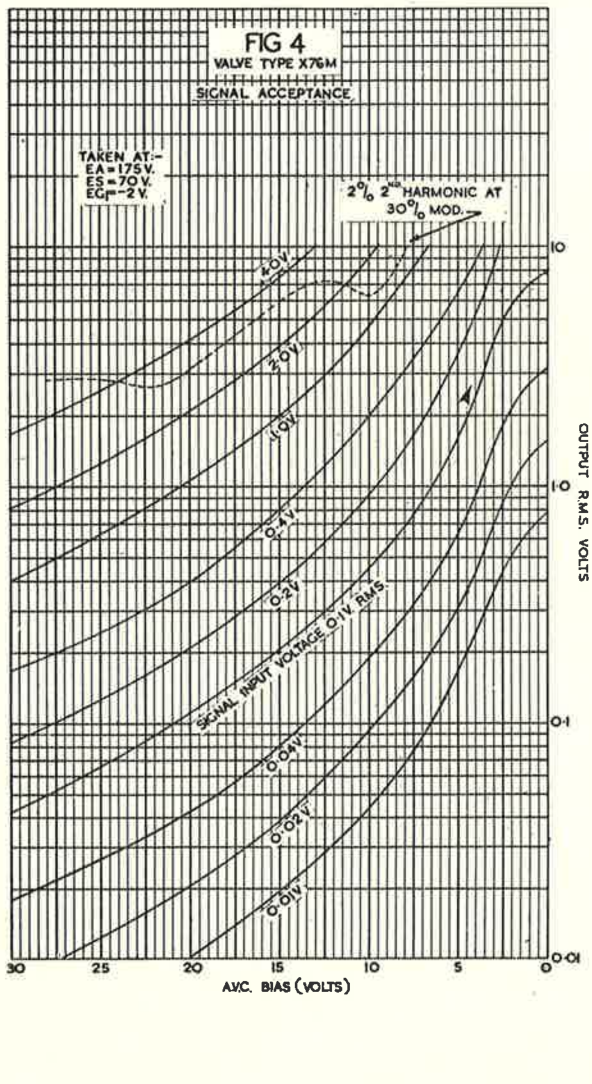


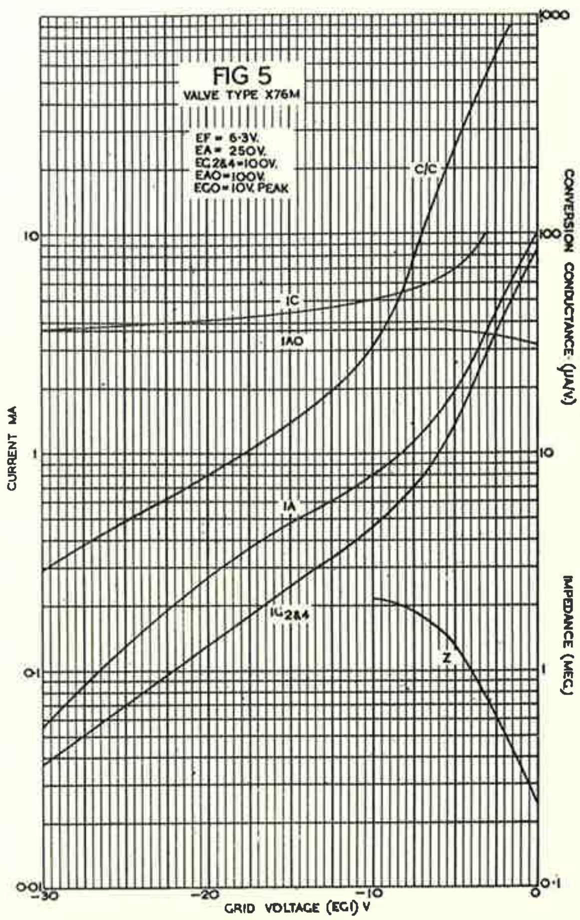
FIG. 2
COLPITTS OSCILLATOR FOR UHF RECEPTION



The recommended circuit is shown in Fig. 1; little explanation is necessary as it is conventional. The input signal is supplied to the control grid via the condenser C1 from the circuit formed by L1 and its tuning condenser. The control grid is returned to the AVC line by the resistance R1. This method of supplying the control voltage is preferred to the more usual method of injecting at the low potential end of L1, in that the bypass condenser, normally included in the tuned circuit, is no longer required: secondly the degree of coupling between the control grid and the tuned circuit may be varied by the condenser C1. For short wave operation a fairly small capacity is preferred so that the damping, due to the input impedance is reduced.

The screen grid should be supplied from a low impedance potentiometer, and not via a series resistance from the plate supply; the latter method will extend the control grid base considerably and reduces the control provided by the AVC circuit; the effect of this is to overload the IF amplifier. Furthermore, comparatively small changes in screen current will cause wide changes in the anode impedance, which will affect the anode circuit considerably. The use of a potentiometer feed overcomes this. The oscillator circuit is quite conventional: in the interests





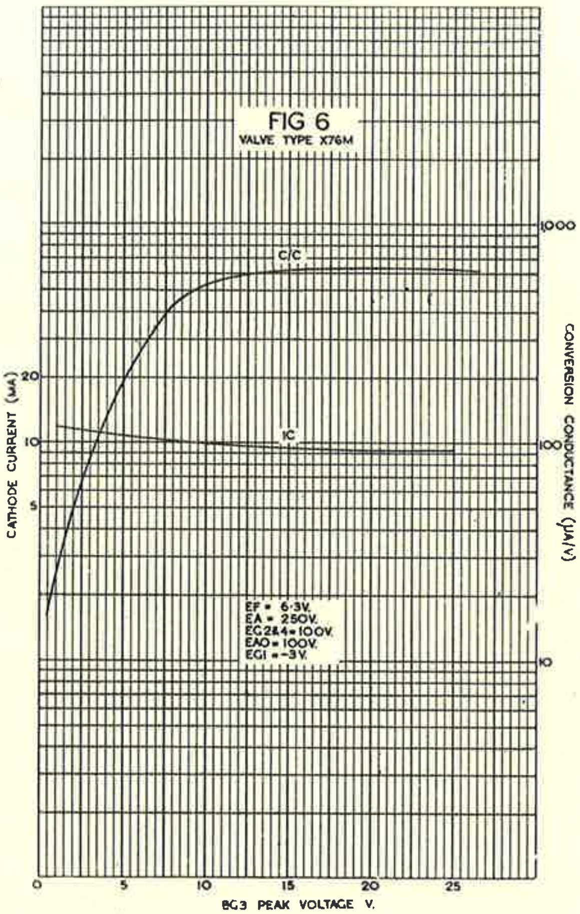
of frequency stability a tuned anode circuit is recommended and is shown in Fig. 1, although the usual tuned grid is satisfactory for many uses. The oscillator anode is supplied through the resistance R6, R7, the condenser C6 being chosen to give the necessary filtering action. It is convenient to return the earthy end of the grid coil L2 to the junction of the padding condenser and the coil L3 so that some capacitive regeneration is provided, particularly at the low frequency end of the waveband; this assists the normal inductive coupling at the point where oscillation is usually less easy to obtain. If excessive coupling is obtained, parasitic oscillation will occur at the high frequency end of the waveband; the incorporation of a resistance R8 will reduce the coupling and suppress the parasitic oscillation. The value of the resistance will vary between 25 ohms at 30 Mc/s to 5000 ohms at, say 300 kc/sec.

The curves of Fig. 3 show the control ratio with AVC voltage on the signal grid: curve A assumes a fixed screen voltage of 70V. This is not usually convenient and the screen is normally supplied from a voltage divider (R2 R3). This extends the grid base since the screen voltage increases somewhat with control grid voltage and the actual control curve obtained is shown by Curve B. The value of the resistances R2, R3, may be reduced if the maximum control is required.

Fig. 4 shows the acceptance and input-output voltage curves at various bias voltages. The acceptance curve is taken for 2% 2nd harmonic distortion of a carrier modulated 30%.

The data shown in Figs. 3 and 4 was obtained at Ea 175 Es 70 (at Egl = — 2v.): at lower anode and screen voltages the gain and acceptance will be reduced proportionately.

Full AVC must be applied to the X76M grid to provide a satisfactory control ratio and to allow the acceptance of large signals. There is a reduction

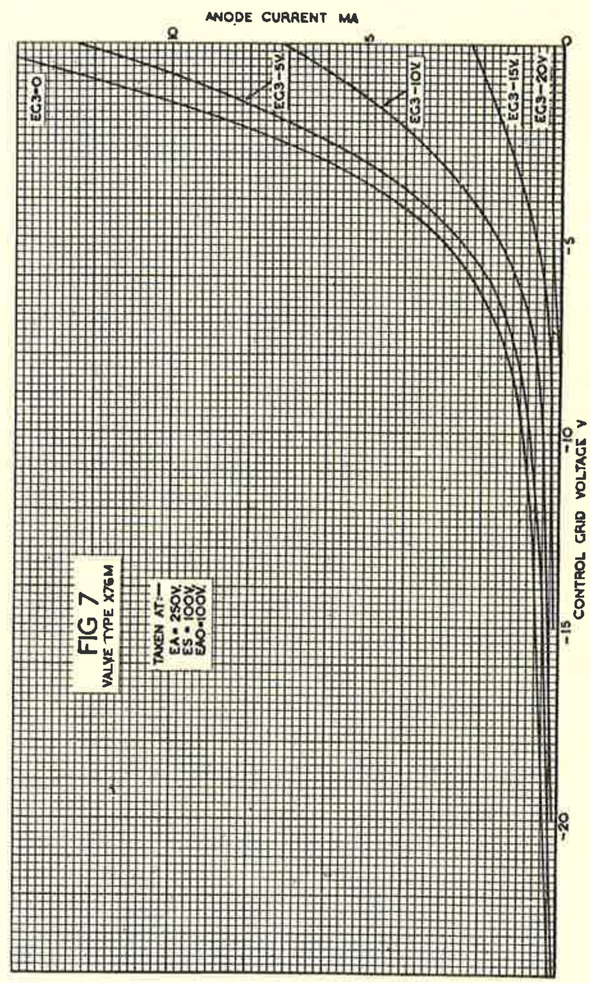


of input capacitance between — 2V and — 30V of approximately 1.5 pfd*: this variation is unimportant in the usual applications of this valve.

*1pfd = 1μμF

The X76M may be used with its heater either in parallel or in series with other valves of suitable voltage or current rating. In AC/DC receivers, where the heaters are series-connected, the X76M heater should be connected as near to the earthy ends as is convenient in order to minimise modulation hum; it is usually possible to place it second in the chain, the first place being reserved for the diode-triode.

At frequencies above 30 Mc/sec. the usual oscillator circuit is not as satisfactory as the Colpitts circuit shown in Fig. 2; a two gang condenser is employed and the circuit is completely symmetrical; oscillation is easy to obtain up to 60 Mc/sec., and operation above this frequency can be maintained. The degree of coupling to the grid is adjusted by the condenser C11. If a fairly wide frequency coverage is desired, a resistance R14 of 5 or 10 ohms may be included to suppress parasitic oscillation as the tuning capacity is reduced. It will be seen that no self-bias resistance is included; this enables the cathode lead to be kept short. The control grid is returned to a point three volts negative to the cathode. It is desirable to use a higher intermediate frequency than usual, and success has been obtained with an amplifier operating at 5 Mc/s. The response of the IF amplifier should be "flat topped" in order to minimise the effect of small shifts in frequency. Where an improved frequency stability is desired the L/C ratio of the oscillator circuit should be reduced by the addition of a further condenser C14.



It was found possible to operate the X65 valve with a low anode voltage; this was a convenience in receivers designed for 100-130 volt mains, in that the anode and screen could be operated at the same potential. The X76M is not as suitable for this class of service due to the design of the hexode, and it is necessary to employ an anode voltage at least twice that applied to the screen grid. When the X76M is used at an anode voltage lower than 175, the screen voltage should be reduced in order to maintain a high anode impedance; the use of an excessive screen voltage will produce a low impedance which will shunt the anode circuit. At lower voltages the conversion conductance is reduced, but the conversion gain will be highest when the correct Ea/Es ratio of 2.5: 1 is maintained.

The heater supply for the X76M must be within the limits of 0.16A± 7% under any operating condition; where wide variations of supply voltage occur, the use of a barretter is recommended. In other cases satisfactory results are obtainable if the heater current is adjusted by means of a tapped resistance: four taps should be provided, so arranged that the correct current is obtained with line voltages of 200, 215, 230 and 245.

Valve Type W76

Valve type W76 is an indirectly-heated variable-mu screened pentode primarily intended for use as an RF or IF amplifier in AC/DC receivers.

The heater of the W76 is designed for series operation at 0.16A, and under these conditions the comparatively low heater current, and hence low power dissipation, is a considerable advantage in smaller types of equipment where temperature rise is important. The total power dissipated in the heater circuit when operated from 230V supply is only 37 watts, of which an average of 20 watts will be dissipated in the barretter or series resistor. With 110V operation no series resistor is necessary for a number of common valve arrangements.

In order to minimise the possibility of modulation hum the W76 should be connected as near as possible to the low-potential end of the series heater chain. The order in which the heaters of the valves should be connected is normally, starting from the low-potential end of the chain: diode-triode, frequency changer, RF stage if any, IF stage, output stage, rectifier.

The heater of the W76 is also suitable for operation from a 12V accumulator in mobile equipment.

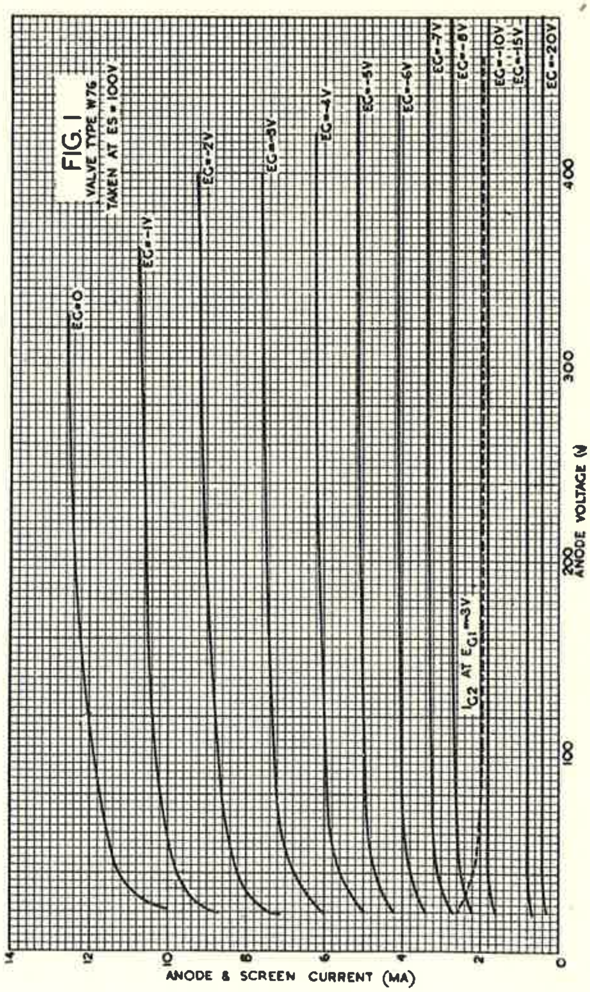
The W76 is fitted with an internal shield, and the use of an external screening can is unnecessary.

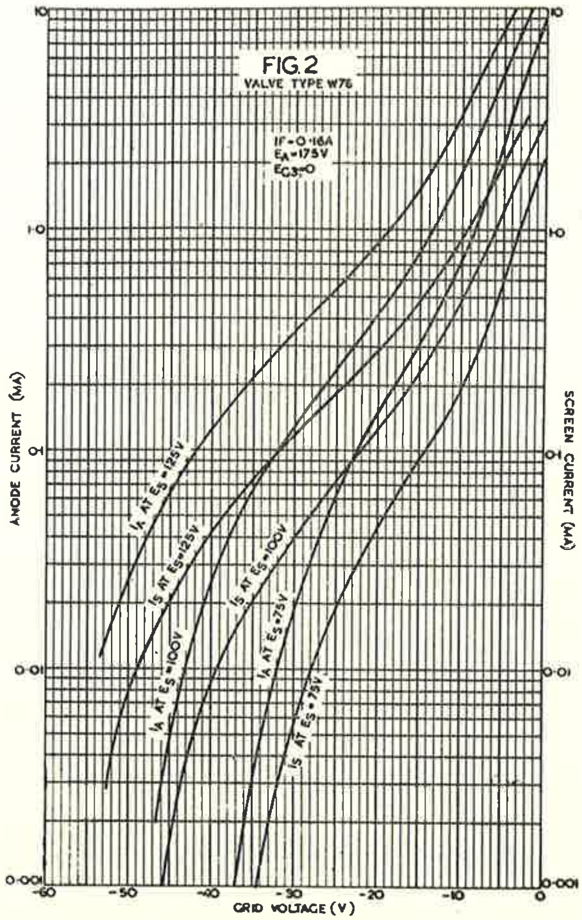
Care should be taken to decouple the anode, screen and cathode circuits, and a non-inductive bypass capacitor should be used. It may be desirable for high frequency operation to use a 100pF mica or ceramic capacitor in parallel with an 0.05 μF paper capacitor for decoupling.

The effective value of grid return resistance for the W76 should not exceed 4 megohms.

Application as an IF amplifier.

The W76 is particularly suitable for use as an IF amplifier in superheterodyne receivers. In this or any other class of operation, where the ability to handle large signal inputs is of importance the use of a series screen supply is recommended. By comparison with fixed screen voltage operation the series screen method gives a higher signal-handling capacity at the expense of a lower gain-control ratio for any given range of AGC voltage. Should a higher gain-control ratio than that given by series-screen operation be required, the AGC voltage may





be applied to the signal and suppressor grids simultaneously.

Fig. 8 shows a suitable circuit arrangement in which the W76 is used as an IF amplifier and is followed by a double-diode triode (Type DH76) as detector, AGC rectifier and AF amplifier. A series screen supply is shown, the screen resistor being chosen such that the screen voltage at minimum bias is approximately 100V. The AGC voltage is shown applied to signal and suppressor grids. Should it be desirable to control only the signal grid, the suppressor grid should be connected to cathode.

The detector is fed from the secondary of the IF transformer and with the component values shown it will accept modulation depths up to 95% with low distortion, and is suitable for operation at intermediate frequencies of the order of 450 kc/s.

The maximum depth of modulation which a diode detector will handle with low distortion is determined by the ratio of the AC and DC load resistances, and for broadcast reception it is essential to keep this ratio as close to unity as practicable to avoid severe distortion with deeply modulated signals. In the circuit of Fig. 8 the maximum permissible modulation depth for low distortion is given by

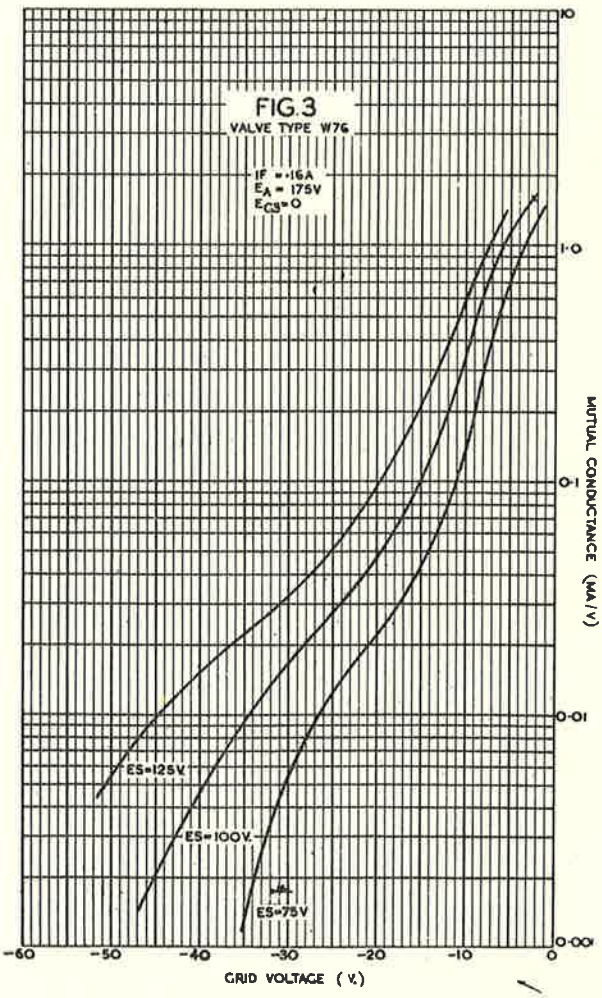
$$M \doteq \frac{\dfrac{R5 \times R7}{R5 + R7}}{R5} \times 100 \text{ per cent.}$$

neglecting the value of R6.

Since it is desirable to handle modulation depths up to at least 95% with low distortion it is evident that the value of R7 must be at least 20 times that of R5. A limit of about 2 megohms is placed on the value of R7 by grid current considerations, which gives R5 = 0.1 megohms.

For optimum performance the product of R5 in megohms and C6 in micro micro farads should be about 20.

An alternate arrangement which obviates the use of a high resistance volume control is shown in Fig. 9. In this circuit the maximum permissible depth of modulation is a function of the setting of the volume control R12. When this is at maximum, the critical modulation depth with the



component values shown will be:—

$$M \simeq \frac{\dfrac{R12 \times R14}{R12 + R14}}{R12} = 89\%$$

neglecting the value of R13.

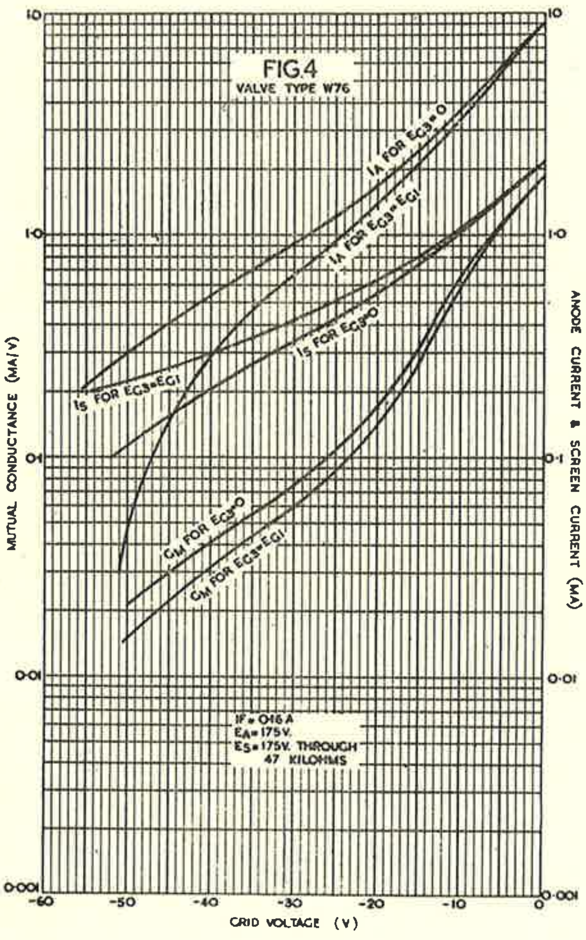
When the volume control setting is reduced by 20dB the AC load is applied only across the lower portion of R12 and the critical modulation depth becomes:

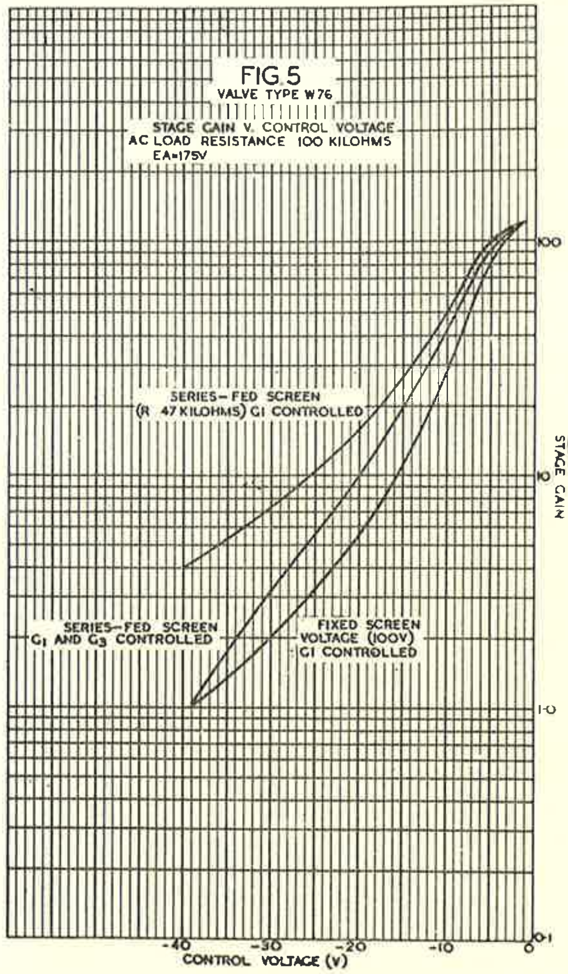
$$M \simeq \frac{0.9\ R12 + \dfrac{0.1\ R12 \times R14}{0.1\ R12 + R14}}{R12}$$

$$\simeq 99.5\%$$

The performance of this circuit is thus highest when the sensitivity is reduced, that is for local station reception. A further advantage of this arrangement is that a higher DC load resistance may be used than with the circuit of Fig. 8, permitting higher gain and selectivity to be obtained from the IF amplifier.

The AGC voltage is derived from the primary of the IF transformer, where the bandwidth and signal voltage are higher than on the secondary. This gives the advantages of increased available AGC voltage and reduced "sideband screech" when the signal is detuned. The AGC has a voltage bias or "delay" of about 1.2V due to the voltage across R8 (Fig. 8).

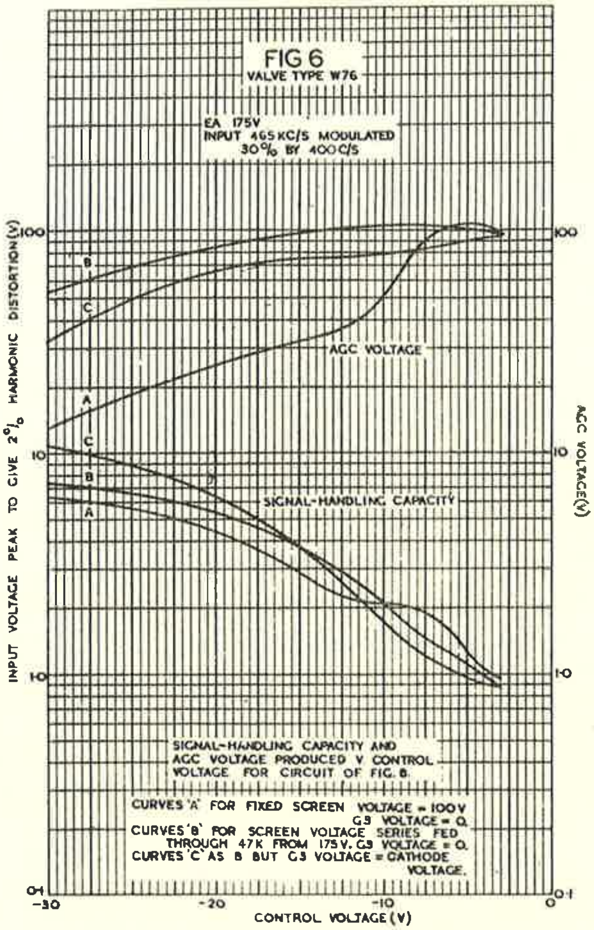


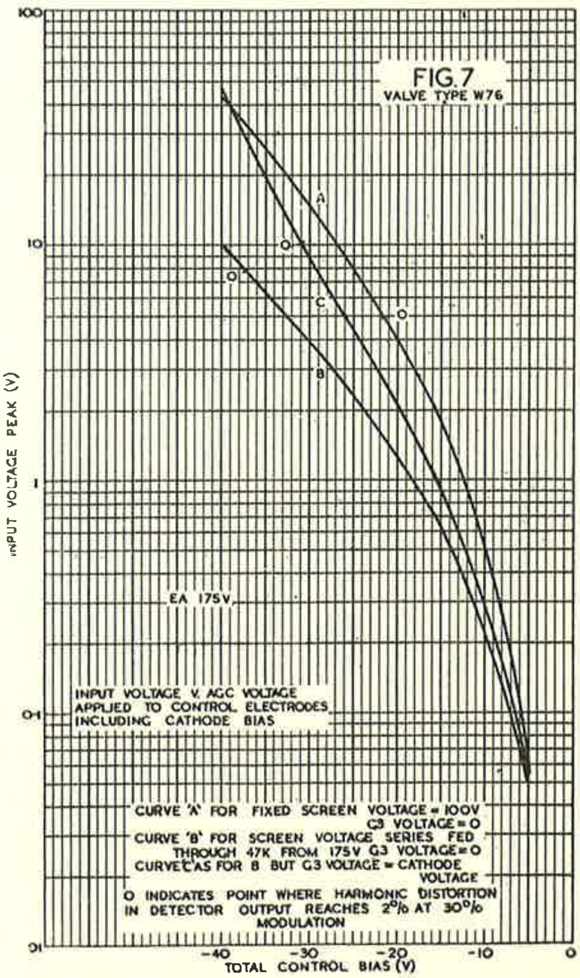


The performance of the W76 as an IF amplifier is shown in graphical form in Figs. 5, 6, and 7. The curves of Fig. 5 illustrate the variation of stage gain with control grid voltage. Curve A is for 100V fixed screen grid voltage and G3 connected to cathode. Curves B and C are for the conditions shown by Fig. 4 i.e., screen supply 175 volts through 47 kilohms with G3 connected to cathode and G1 respectively. The curves assume an effective anode load of 0.1 megohm which is typical of the primary load impedance of a commercial IF transformer when loaded by detector and AGC circuits.

Curves of signal-handling capacity and AGC voltage to a base of control voltage are given in Fig. 6 for each circuit condition. These curves were taken using a circuit similar to that of Fig. 8 and show the maximum signal (465 kc/s, modulated to a depth of 30% by 400 c/s) which may be applied to the W76, such that the output voltage from the detector shall contain not more than 2% harmonic distortion.

The signal-handling capacity of the W76 is illustrated in a different manner in Fig. 7. This figure





shows curves of input signal against control bias produced by the AGC rectifier (approximately equal to the peak carrier voltage). The point O

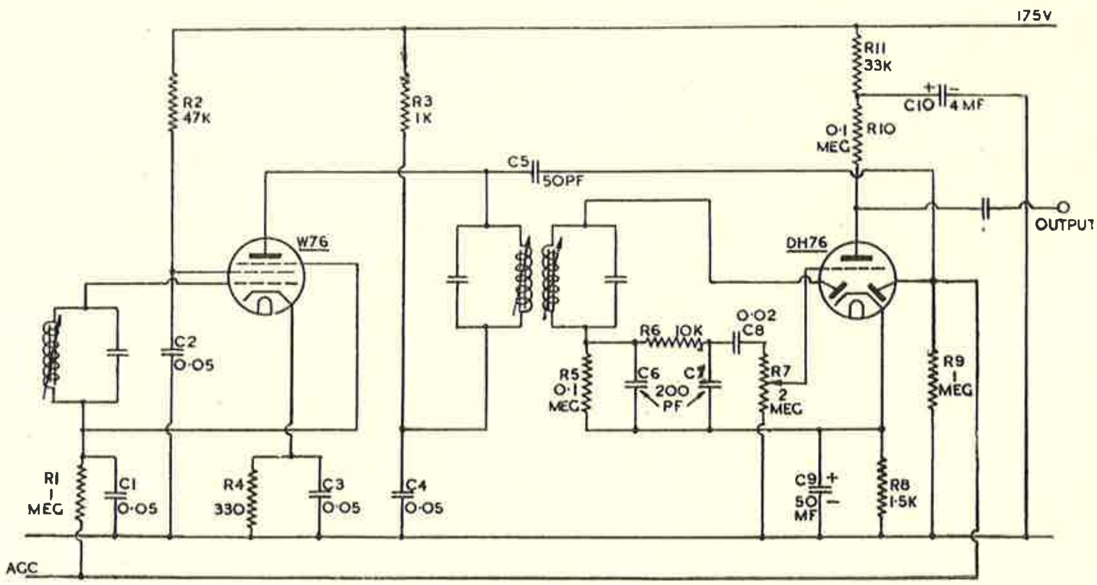


FIG.8
IF AMPLIFIER AND DETECTOR

indicates in each case the level at which the harmonic distortion in the detector output reaches 2% at 30% modulation.

The figures of signal-handling capacity at any other modulation depth for 2% harmonic distortion may be calculated from the expression.

$$e_1 = e_2\sqrt{\frac{M2}{M1}}$$

where e_1 = signal-handling capacity at modulation depth M1

e_2 = signal-handling capacity at modulation depth M2

The curves of Figs. 6 and 7 are for a single controlled stage. With the addition of preceding amplifier stages, the AGC performance will of course be modified according to the performance of each individual stage.

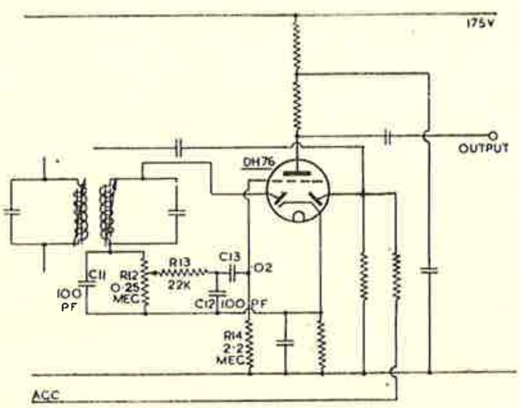


FIG. 9
ALTERNATIVE DETECTOR CIRCUIT

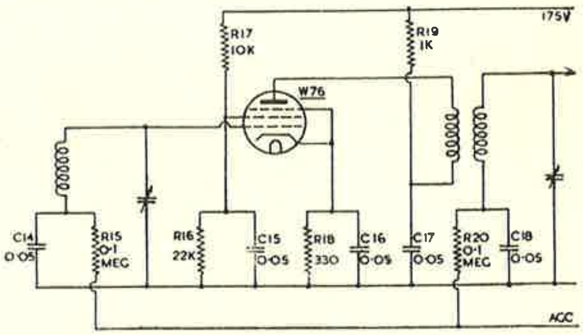


FIG. 10
RF AMPLIFIER

Application as an RF amplifier

The use of a fixed-voltage screen supply is recommended when the W76 is used as an RF amplifier. The full AGC voltage should be applied to the control grid. A typical circuit arrangement is shown in Fig 10 and values for a suitable potential divider network for screen supply are given.

The data on signal-handling capacity given in the preceding section is applicable to the W76, when used as an RF amplifier, for similar bias conditions. More detailed information for the design of AGC circuits may be found in the KTW61 "Application Report".

The gain as an RF amplifier will depend on the Q of the tuned circuits and is likely to vary considerably over the tuning range.

Valve Type DH76

The DH76 is a double-diode triode valve of the indirectly heated type with high impedance and amplification factor. It is principally for use in AC/DC receivers with series-connected heaters.

The DH76 is a double-diode triode having a fairly high anode impedance: it is suitable for use in RC coupled amplifiers. Due to its low current consumption, economical operation of receivers with series connected heaters is effected.

The circuit recommended is that shown in Fig. 1 in which one diode is used for providing a DC

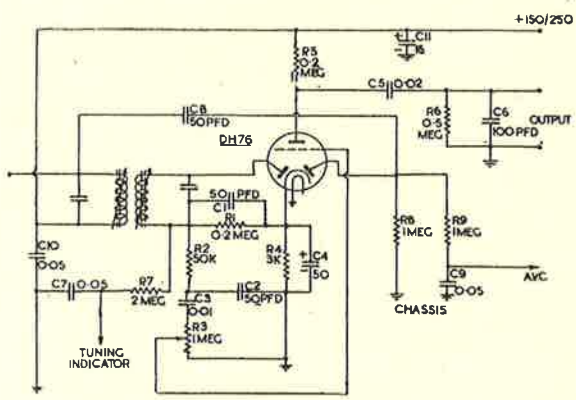


FIG. 1

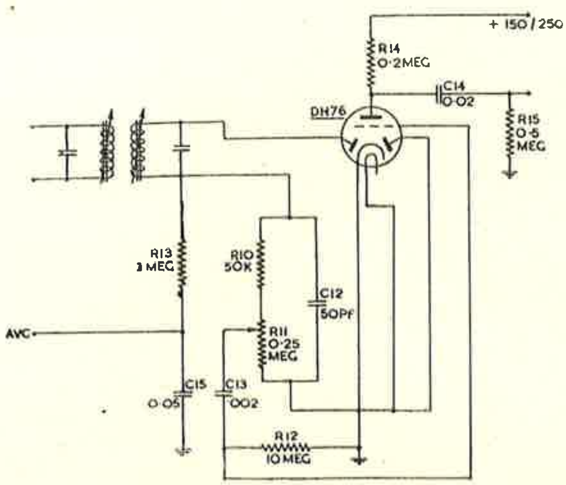
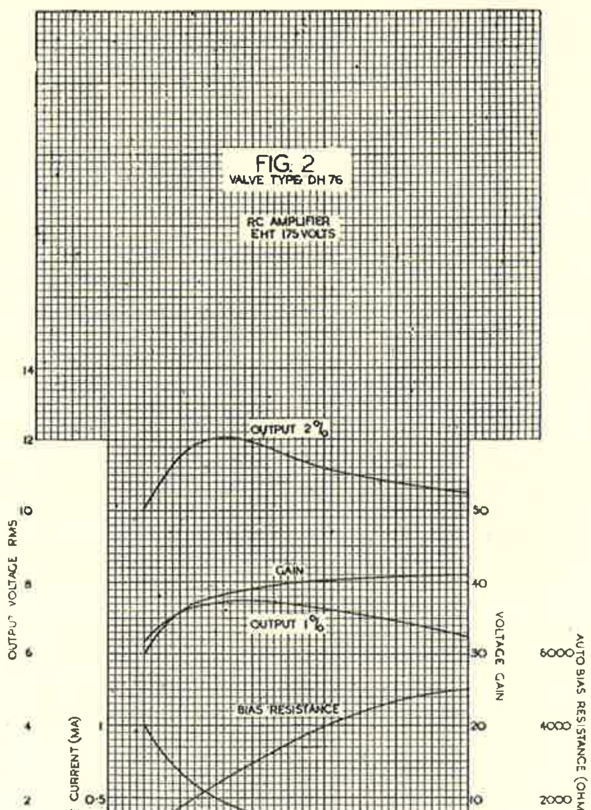


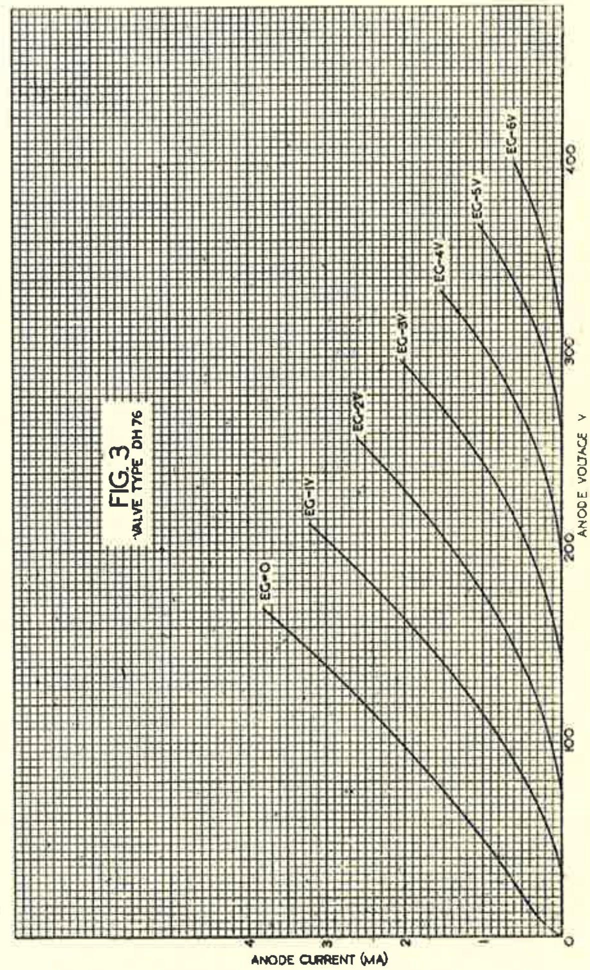
FIG 1A

voltage for the AVC circuit and the other for producing an audio signal from the RF input signal. The AVC diode is fed from the primary of the IF transformer so that a large AVC voltage is produced, and the signal diode is supplied from a point of greater selectivity which obviates, to a large extent, "side band screech." A small amount of delay is provided for the AVC diode by the voltage across



the auto bias resistance R4. The tuning indicator, if used, should be supplied from the signal diode so that the indicator will have a better discrimination than if it were supplied from the AVC circuit.

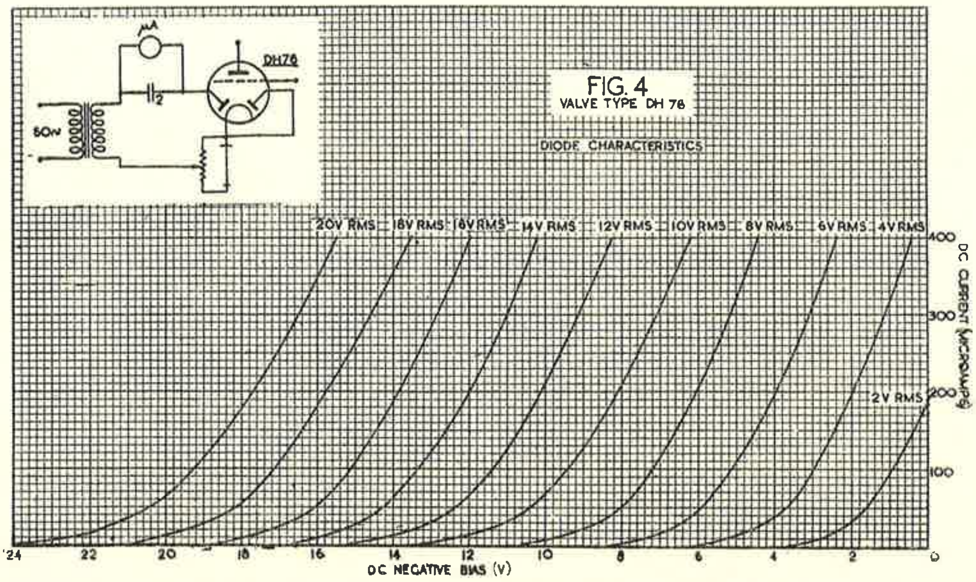
The value of the anode load resistance is not critical and the performance of the DH76 with various values may be found from the curves, Fig. 2, together with the correct value of bias resistance. It will be seen that, assuming a following grid resistance of 0.5 meg., adequate output for most purposes



is given with a distortion lower than 1%. The available output voltage is proportional to the anode supply voltage.

In circuits where the diodes are not required, they should be earthed. For minimum hum the heater should be connected at the earthy end of the heater chain in series-connected circuits, so that the AC heater-cathode voltage is reduced to a minimum.

An alternative circuit is shown in Fig. 1A, where the control grid is connected to cathode via a high resistance R12 and the usual cathode bias resistance is omitted. This arrangement is satisfactory when



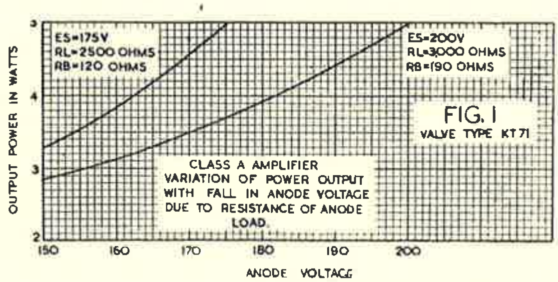
the input signal is small, as it is when the subsequent valve is either a KT71 or KT76.

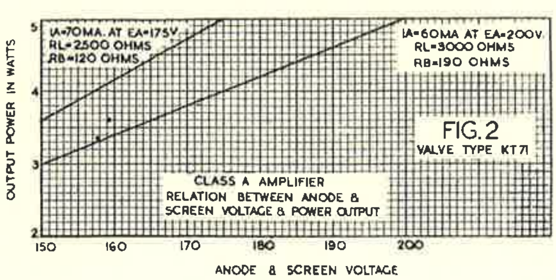
The AVC circuit is shown connected to the secondary of the IF transformer, but it may be connected as in Fig. 1 if the maximum AVC voltage is required.

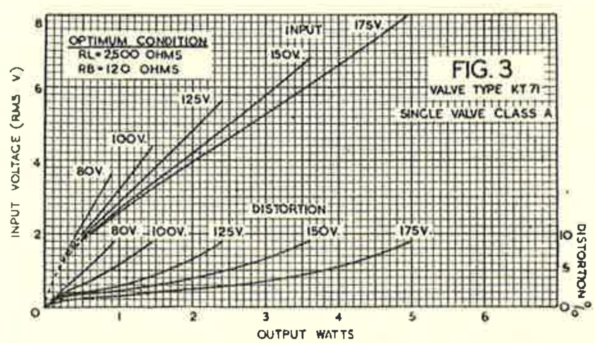
It will be noted that the circuit of Fig. 1A shows some savings in components.

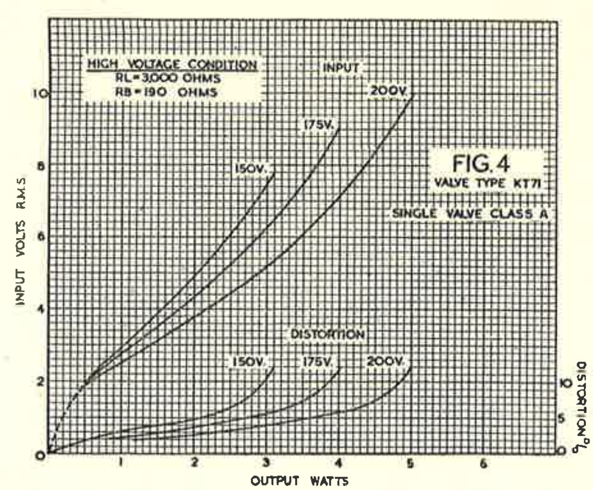
Valve Type KT71

The KT71 is an output tetrode with a heater rating of 48 volts at 0.16 ampere, designed for use in AC/DC receivers; satisfactory operation is obtained over a wide range of anode and screen voltages, five watts output being obtained at 175 volts and one watt at 90 volts.







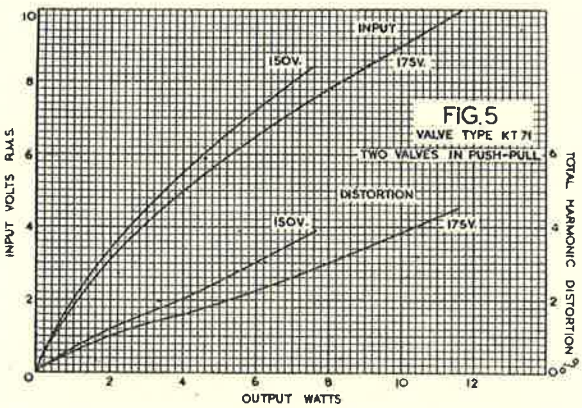


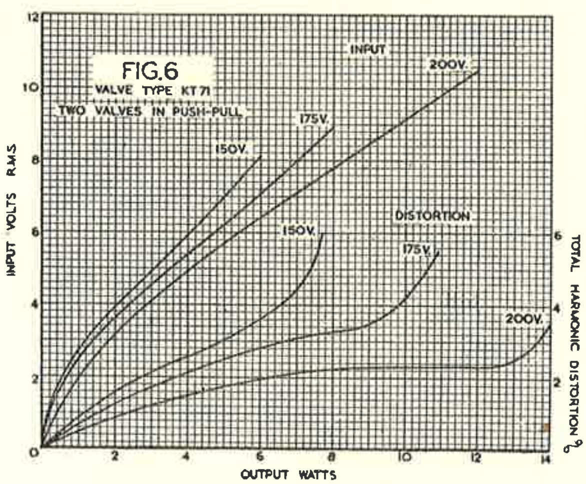
OPERATING CONDITIONS Single Valve Class A amplifier. Valve type KT71.

	Optimum Condition					High Voltage Condition		
Anode Voltage	80	100	125	150	175	150	175	200
Screen Voltage	80	100	125	150	175	150	175	200
Grid Bias Voltage approx. (No signal)	4.5	5.5	7.5	8.4	9.8	9.9	11.6	13.3
Anode Current (mA) (Max. signal)	32	38	49	60	68	42	49	53
Anode Current (mA) (No signal)	33	40	52	60	70	44	52	60
Screen Current (mA) (Max. signal)	9	12	14	19	22.5	16	19.5	22
Screen Current (mA) No signal	5.5	6.5	8.5	10	12	8	9	10
Auto-bias Resistance (ohms)	120	120	120	120	120	190	190	190
Output (5% D) watts	0.55	0.9	1.7	2.5	3.8	1.9	2.7	3.6
Power Output (watts) max.	0.9	1.45	2.4	3.6	5.0	3.1	4.0	5.0
Signal Input Volts (pk) max.	5.0	6.0	8.0	9.5	11.0	11.0	13.0	14.0
Total Distortion (%)	9	9	9	9	9	12	12	12
Load Resistance (ohms)	2500	2500	2500	2500	2500	3000	3000	3000

The anode dissipation must not exceed 13 watts; to get the best performance over a range of supply voltages, the operating conditions should be arranged so that this figure is reached at the maximum supply voltage. In the single valve case there is no gain in power output through the use of an anode voltage of 200v. instead of 175v., but the remainder of the receiver may show some improvement. The conditions given above are for the two cases where the maximum anode and screen voltage is 1) 200v. and 2) 175v.; the bias resistance is set for the maximum voltage and the conditions at lower voltages are those that automatically obtain.

It will be seen that a reasonable output is given at Ea = Es = 80/100V. so that the valve may be used in 110 volt AC—DC receivers.

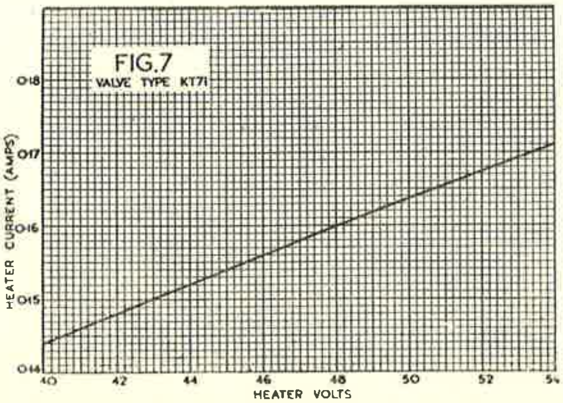




Two valves in push-pull, Class AB1.

Anode Voltage	150	175	200	150	175
Screen Voltage	150	175	200	150	175
Grid Bias Voltage (v) No Signal	10.6	12.5	14.2	8.8	10.2
Combined Anode Current (mA) Max. Signal	89	105	120	120	145
Combined Anode Current (mA) No Signal	76	89	102	107	128
Combined Screen Current (mA) Max. Signal	21	26	39	24	30
Combined Screen Current (mA) No Signal	12	15	17	18	22
Auto Bias Resistance (per valve) ohms	240	240	240	140	140
Power Output (watts)					
Normal	6	8	12 }	7.5	11.5
Maximum	7.5	11.0	15.5 }		
Distortion %					
Normal	3.5	3.5	2.5 }	4	4.5
Maximum	6	5.5	7.5 }		
Maximum Input Signal (Peak Volts)	31	36	44	24	28
Anode to Anode Load Resistance ohms	4000	4000	4000	2500	2500

The anode and screen dissipation figures of 13.0w and 3.0w respectively must not be exceeded: to obtain the best performance over a range of HT supply voltages the operating conditions should be arranged so that these figures are reached at the maximum supply voltage and with full drive. The increase of screen wattage with signal is permissible, being an intermittent overload only.



The conditions given above are for the two cases where the range of HT voltage gives an anode and screen voltage variation of 1) 150 to 200v. and 2) 150 to 175v. The bias resistance is set for the maximum anode voltage of the range, and the conditions at lower voltages are those that automatically obtain.

Two sets of figures for power output and distortion are shown, the first giving the undistorted output, and the second the full output with some amplitude distortion.

APPLICATION

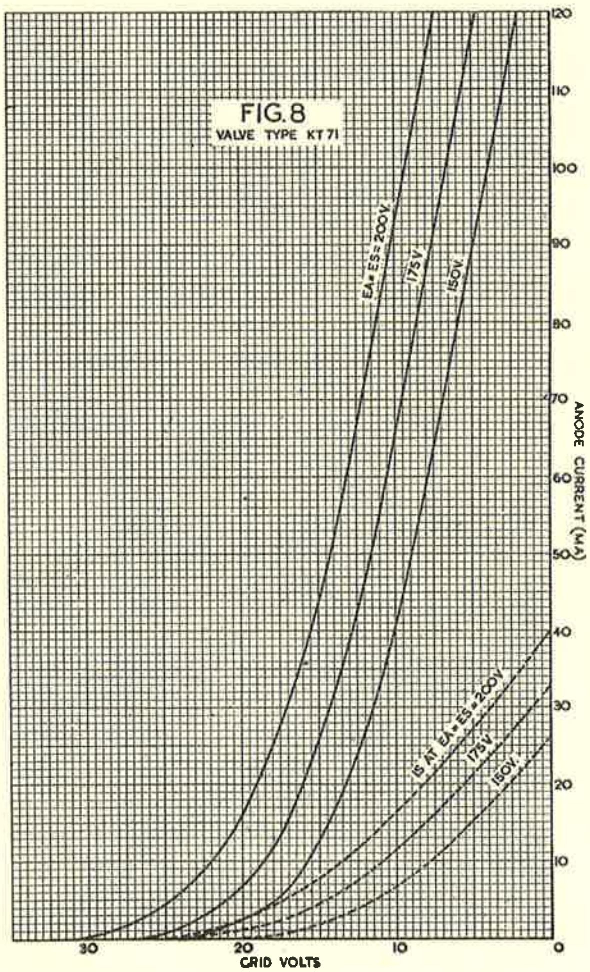
Single valve Class A amplifier.

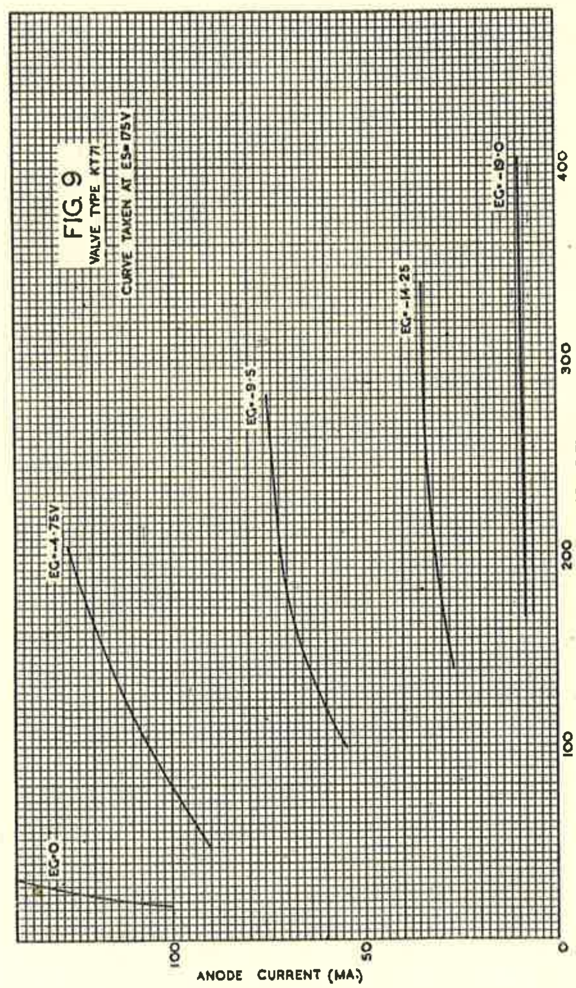
The circuit of Fig. 10 shows the arrangement for operating a KT71 in a single valve Class A amplifier.

A signal from the previous stage is fed to the grid, and where a grid resistance is used, its value should not exceed 0.5 megohm. The screen is supplied direct from the HT. line, so that the actual anode voltage will be lower than the screen voltage by the DC drop in the output transformer primary winding. A consequent loss of power must be expected, as shown in the curve of Fig. 1, which shows variation of power with anode voltage, the screen voltage being kept constant. No series screen resistance should be used because the increase of screen current with load results in a loss of output power.

Any tendency to instability is prevented by the use of a grid stopper (10,000—100,000 ohms) and/or screen stopper (100 ohms) mounted close to the valve holder.

The condenser C2 prevents over-accentuation of the higher frequencies and in itself may prevent oscillation in this stage; it should be adjusted in





value to give the best quality and will be found normally to be about 0.005 μfd. There is a possibility of the circuit becoming unstable if the external grid and anode circuits are too close together.

The smoothing choke L1 should be of low DC resistance so that the receiver HT voltage may be kept as high as possible on low mains voltages. If it is found that the maximum anode and screen voltages are exceeded on a high supply voltage, the value of the resistance R5 may be increased.

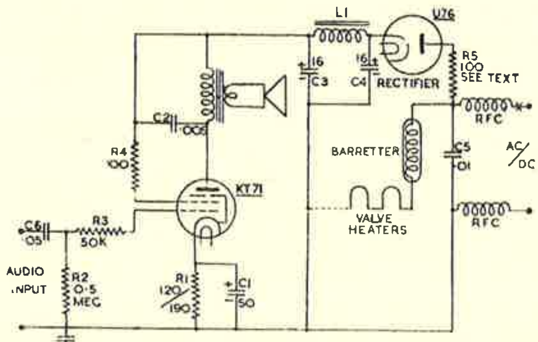
Power output varies directly as the anode and screen voltage; this is illustrated in the curves of Fig. 2. Input voltage plotted against output power and distortion is shown in Figs. 3 and 4.

Two valves in push-pull Class AB1 amplifier.

Fig. 11 shows a circuit arrangement for two KT71 in Class AB1 push-pull, preceded by a transformer coupled DH76: alternatively an RC coupled circuit can be used with another DH76.

Instability is prevented by the use of a grid stopper (10,000—100,000 ohms) and/or screen stopper (100 ohms) mounted close to the valve holder. It is also important that the anode and grid circuits should be spaced well apart. The condensers C12/13 prevent over-accentuation of the high frequencies and in themselves may obviate oscillation in this stage: their value will probably be of the order of 0.005 μfd, but this will be determined by the characteristics of the loud speaker.

In general, less smoothing of supply voltages is necessary when using push-pull systems. In some cases it is possible to take the anode supply direct



OPERATION OF KT71 AS CLASS A AMPLIFIER

FIG. 10

from the rectifier, but the screen supply should be filtered in every case. Where a smoothing choke is used, this should be of low resistance so that the receiver HT voltage is kept as high as possible.

The resistances R15, R16 are included to limit the peak current through the rectifier and must not be less than 100 ohms each.

It will be observed from the operating data on page 105 that in certain cases it will be necessary to use 2 U76 rectifier valves in parallel in order to obtain the required output current.

Figs. 5 and 6 give curves showing the output power plotted against input voltages and distortion, for various anode voltages and loads.

Precautions.

Under the maximum steady conditions likely to be encountered, the anode and screen dissipation figures given must *not* be exceeded.

It is not advisable to omit the anti-squegger resistances in the grid or screen leads because of the risk of instability troubles. Separation between anode and grid circuits should be as wide as possible.

Adequate ventilation should be provided in a receiver as considerable heat must be dissipated by an output stage consisting of 2—KT71 valves.

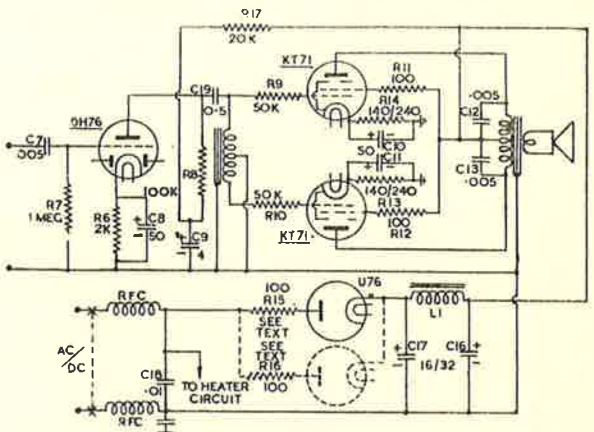


FIG. 11

COMPONENT VALUES

R1	120 or 190 ohms See text. .	C1	50 μfd
R2	0.5 megohm	C2	0.005 μfd
R3	50,000 ohms	C3	16 μfd
R4	100 ohms	C4	16 μfd
R5	100 ohms minimum	C5	0.01 μfd
R6	2000 ohms	C6	0.05 μfd
R7	1 megohm	C7	0.005 μfd
R8	100,000 ohms	C8	5.0 μfd
R9	50,000 ohms	C9	4 μfd
R10	50,000 ohms	C10	50 μfd
R11	100 ohms	C11	50 μfd
R12	100 ohms	C12	0.005 μfd
R13	140 or 240 see text	C13	0.005 μfd
R14	140 or 240 see text	C16	16/32 μfd
R15	100 ohms minimum	C17	16/32 μfd
R16	100 ohms minimum	C18	0.01 μfd
R17	20,000 ohms	C19	0.5 μfd

Valve Type U76

The U76 valve is a half-wave, vacuum rectifier, fitted with an indirectly heated cathode. Its low heater current is of advantage where economy of power is essential.

APPLICATION

The recommended circuit for the U76 is shown in Fig. 1, and is the conventional arrangement used with AC-DC mains. The resistance R1 is essential to limit the surge on switching; the anode-cathode impedance of the U76 is very low and an abnormally high current would be passed by switching off and on the supply voltage with the cathode in an

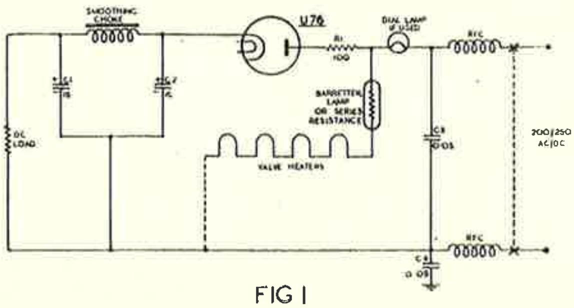


FIG I

emissive stage. This resistance also tends to equalise the DC output voltage obtainable on AC and DC mains and the receiver performance is thereby rendered more constant. When the mains voltage is below 130 volts AC the resistance may be omitted.

The heater of the U76 valve is designed to operate at a mean current of 0.16 ampere, and should be kept within ± 7% under all conditions of operation. In order to cater for variations in supply voltage either a barretter or a tapped resistance may be used: a barretter is preferable since the correct operating current is obtained automatically. An adequate number of taps must be available if a

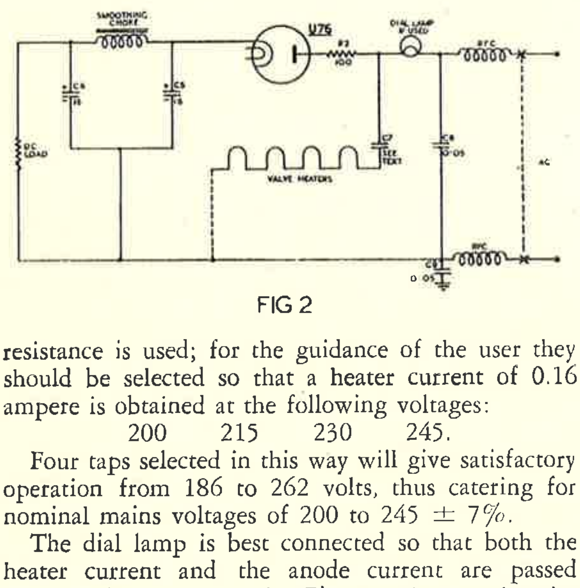


FIG 2

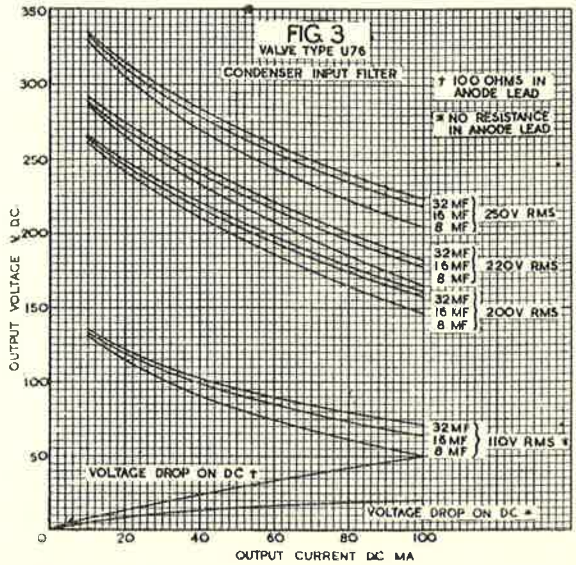
resistance is used; for the guidance of the user they should be selected so that a heater current of 0.16 ampere is obtained at the following voltages:

200 215 230 245.

Four taps selected in this way will give satisfactory operation from 186 to 262 volts, thus catering for nominal mains voltages of 200 to 245 ± 7%.

The dial lamp is best connected so that both the heater current and the anode current are passed through it as shown in Fig. 1, since a heavier current rated lamp may be used which is better able to withstand the switching surge. The anode current of 50-100 mA does not flow until the valves are at operating temperature and then the heater current has dropped to a normal value. A suitable lamp is that rated at 6.5 volts 0.3 ampere (Osram type "S") for cases where the current is high.

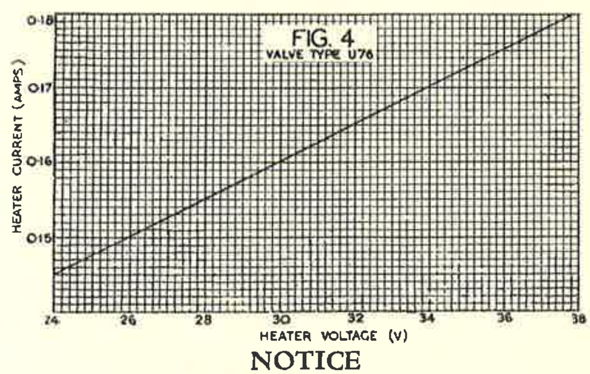
A circuit suitable for AC operation is shown in Fig. 2, where a condenser C7 is used instead of



a resistance; this method shows an economy in power consumption since the loss in the condenser is negligible, though DC operation is not of course possible. The value of the condenser will depend on the valve combination and the supply voltage: the approximate value is 2.5 μfd., additional smaller condensers being connected in *parallel* to accommodate lower mains voltages. This circuit is of use where it is impossible to provide adequate ventilation for a voltage dropping resistance.

The performance of the U76, with various values of smoothing condensers, is shown in Fig. 3: the output voltages are obtainable only when a low impedance source is used, as is normal in AC-DC receivers. Should a transformer of poor regulation be interposed, lower DC voltages will be obtained. A curve is also given showing the voltage dropped in the U76 and the 100 ohm resistance on a DC supply.

A curve connecting heater voltage and current is given in Fig. 4.



NOTICE

It should be understood that the information given in the foregoing memoranda is of a technical nature only and does not imply any licence under any patents which may be involved.

Use of Miniature Tubes in Stagger-Tuned Video Intermediate-Frequency Systems

R.C.A. application note AN-126. Reprinted by courtesy of the Radio Corporation of America

Application Note 126

[Although type 6AG5 is not available in Australia owing to the dollar shortage, this article is included as being of general interest. Types 6AU6 and 6BA6 will be available as part of the new season's programme of equipment types. It will be some considerable time before broadcast television commences in Australia.]

This Note describes the design and construction of stagger-tuned video intermediate-frequency systems utilizing miniature tubes and compares the performance of miniature types 6AG5, 6AU6, 6BA6, 6AK5, 6BJ6, and 6BH6 in such wide-band systems.

A comparison of the important tube characteristics involved in wide-band amplifier design given in Table I (last page) indicates that type 6AK5, because it has slightly lower grid-to-plate and input capacitances than type 6AG5, provides slightly higher gain. The 6AK5, however, has a more complicated construction and, consequently, is more expensive to manufacture. The higher gain of the 6AG5, compared to that of the 6AU6 or the 6BA6, and the lower cost compared to that of the 6AK5 justify a preference for the use of the 6AG5 in most receivers. Type 6AU6, however, may find preference in receivers in which high average stage gain at reduced bandwidth is desired. Types 6BJ6 and 6BH6, designed to operate with a heater current of 150 milliamperes, are recent additions to the miniature line for use where low heater current is a requisite. The gain of the 6BH6 is approximately equal to that of the 6AU6 and the gain of the 6BJ6 is approximately equal to that of the 6BA6.

A circuit diagram for a video i-f system using a 6AG5 complement is shown in Fig. 1; a circuit diagram for a system using a 6AU6 complement is shown in Fig. 2. Each system includes the plate circuit of a converter stage and a 6AL5 video detector. The i-f tubes are placed in the stages labelled V1, V2, V3, and V4. Each stage is tuned by adjusting its inductance for resonance with the tube and circuit capacitances. Because tube capacitances vary from type to type, the system requires retuning to obtain the same band-pass characteristic for all tube complements.

Formulas for the calculation of the resonant frequencies and the values of grid damping resistors for stagger-tuned systems are given in the Appendix. Slightly higher values of damping resistors than the calculated values are necessary because of the loading due to the tube and circuit components. Introduction of rejection traps also causes a slight departure from theoretical alignment frequencies. Adjustments of the values of damping resistors and frequencies calculated for the circuits of Figs. 1 and 2 were made empirically.

The values of cathode resistors needed for complete compensation of input-capacitance change with change in bias can be calculated from the information given in Table I.

Gain and Band-Pass Data.

In this Note, over-all video i-f gain is considered as the amount of dc voltage developed by the video detector per ac (RMS) volt applied to the grid of the first i-f stage. Over-all gain data were obtained by applying a 23.5-megacycle signal to the control grid of the first i-f stage. Mixer gain, therefore, is not included as part of the video i-f gain. The band-pass characteristics of the system were obtained by placing a 6BA6 in the mixer position and applying the i-f signal to its control grid.

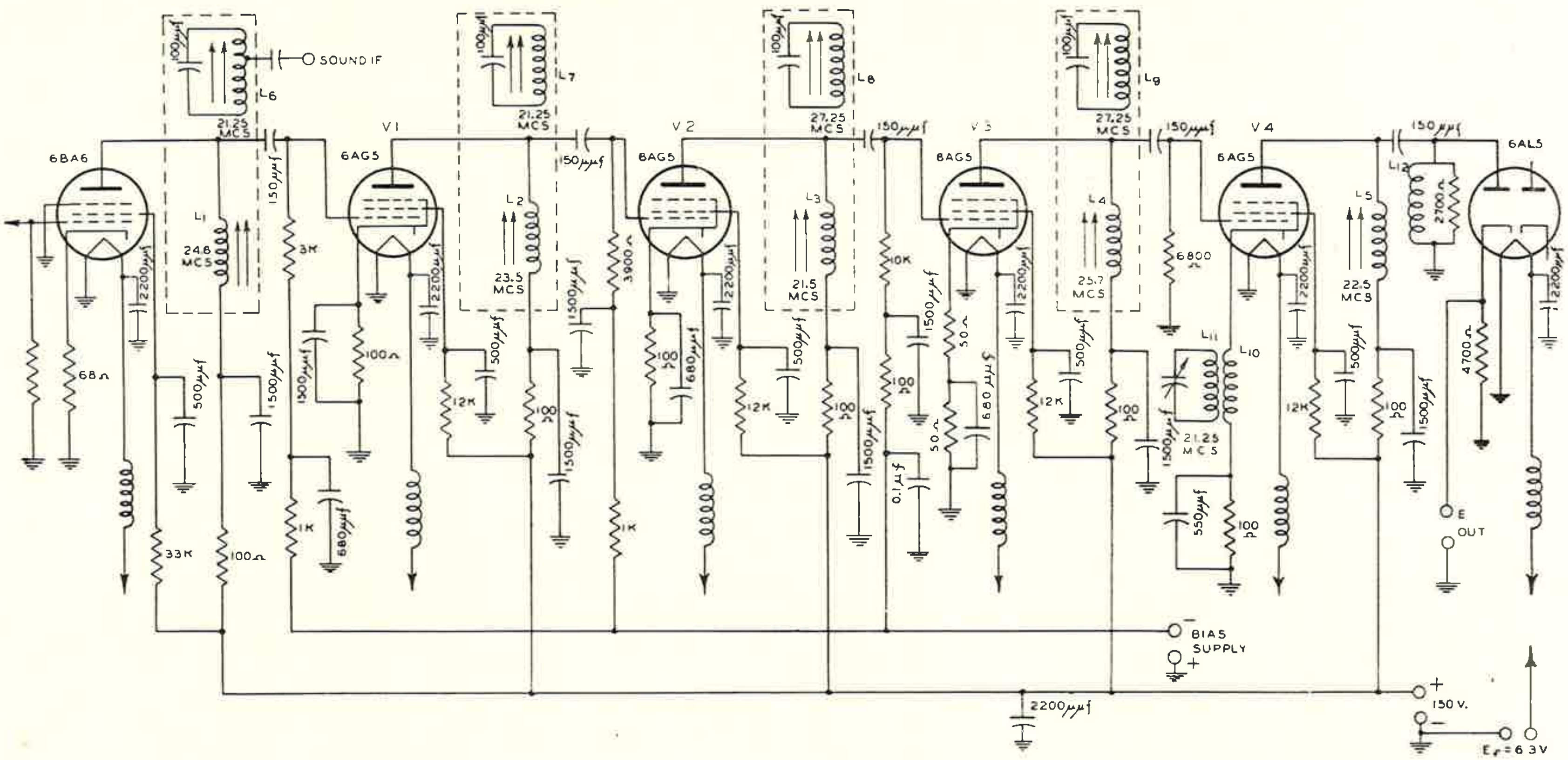
The following tabulation shows average gain-per-stage values for each four-tube complement.

Type Employed in Four-tube Complement	Over-all Gain	Effective Average Stage Gain
6AK5	10,000	10.0
6AG5	6,500	9.0
6AU6	3,000	7.5
6BA6	1,280	6.0
6BJ6	not measured	—
6BH6	not measured	—

Although the gain of a single stage at its resonant frequency in a four-stage stagger-tuned system is not necessarily the fourth root of the over-all gain, the fourth root is used for purposes of comparison as the effective average value of stage gain. Gain and band-pass data for the 6AG5 and 6AU6 complements are shown in Figs. 3 and 4 for several values of control-grid bias. The curves are labelled according to the value of over-all gain and the value of the fixed grid bias. The actual values of grid bias are somewhat greater than those tabulated because self-bias is also used and contributes to the actual values.

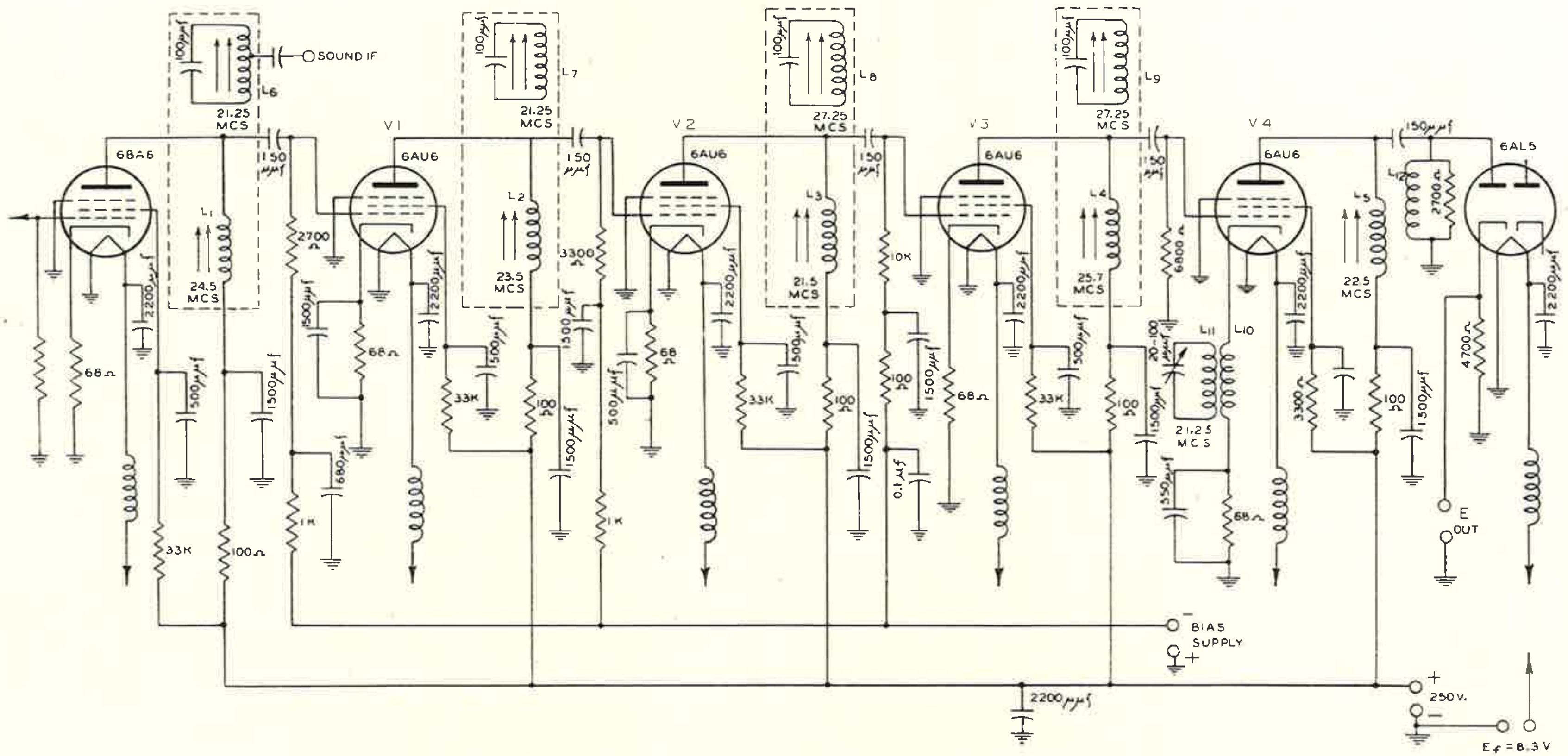
The curves show that increasing the grid bias causes a change in the shape of the band-pass. The change in shape, however, is not enough to result in any readily discernible effect on picture quality. In addition, the position of the carrier remains at the point of approximately 50 per cent. response even though the grid bias is increased enough to cause large decreases in over-all gain. Similar results can be obtained with 6AK5 and 6BA6 complements.

The curves in Figs. 3 and 4 are not intended to show adjacent channel information. The attenuation at the sound carrier frequency, however, is greater



K = 1000 ohms.
Ω = Value in ohms.
Filament Chokes: 14 turns #20 enam. 1/4" dia. - 1/2" lg.
L1: 21 turns #28 enam. on 1/4" form.
L2: 21 turns #28 enam. on 1/4" form.
L3: 25 turns #28 enam. on 1/4" form.
L4: 21 turns #28 enam. on 1/4" form.
L5: 25 turns #28 enam. on 1/4" form.
L6: 8 turns #18 enam., on same form as L1 spaced 1/4" from plate end of L1.
L7: 8 turns #18 enam., spaced 1/8" from L2 on same form.
L8: 6 turns #18 enam., spaced 1/8" from L3 on same form.
L9: 6 turns #18 enam., spaced 1/8" from L4 on same form.
L10: 4 turns #24 enam., 3/4" dia. on same form as L11 spaced 1/4" from cold end of L11.
L11: 6 turns #18 enam., spaced to - 1/2" length 3/4" dia.
L12: R.F.C. 60 turns #32 enam., over 1/4" dia. 1/2" length 2700 ohm resistor.

Fig. 1 - Schematic Diagram of Four-Stage Stagger-Tuned Video IF Amplifier System Using 6AG5 Complement



K = 1000 ohms.
Ω = Value in ohms.
Filament Chokes: 14 turns #20 enam. 1/4" dia. - 1/2" lg.
L1: 21 turns #28 enam. on 1/4" form.
L2: 21 turns #28 enam. on 1/4" form.
L3: 25 turns #28 enam. on 1/4" form.
L4: 21 turns #28 enam. on 1/4" form.
L5: 25 turns #28 enam. on 1/4" form.
L6: 8 turns #18 enam. on same form as L1 spaced 1/4" from plate end of L1.
L7: 8 turns #18 enam., spaced 1/8" from L2 on same form.
L8: 6 turns #18 enam., spaced 1/8" from L3 on same form.
L9: 6 turns #18 enam., spaced 1/8" from L4 on same form.
L10: 4 turns #24 enam., 3/4" dia. on same form as L11 spaced 1/4" from cold end of L11.
L11: 6 turns #18 enam., spaced to 1/2" length 3/4" dia.
L12: R.F.C. 60 turns #32 enam., over 1/4" dia., 1/2" length 2700 ohm resistor.

Fig. 2 - Schematic Diagram of Four-Stage Stagger-Tuned Video IF Amplifier System Using 6AU6 Complement

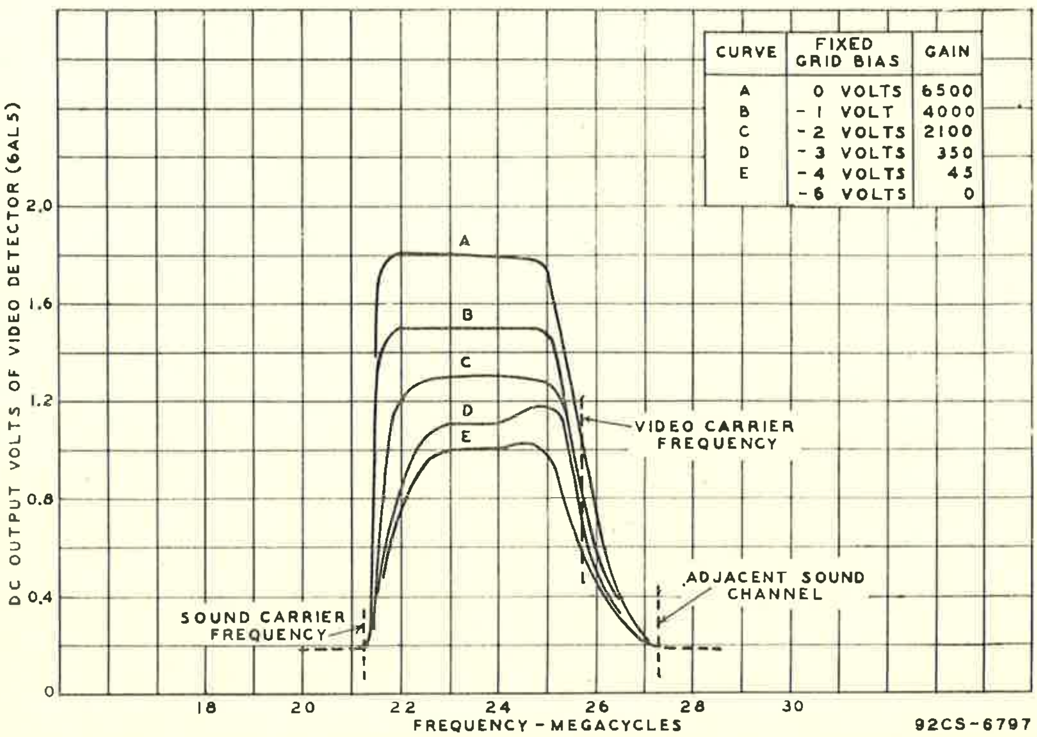


Fig.3 - Frequency Response of Video IF Amplifier Using 6AG5 Complement

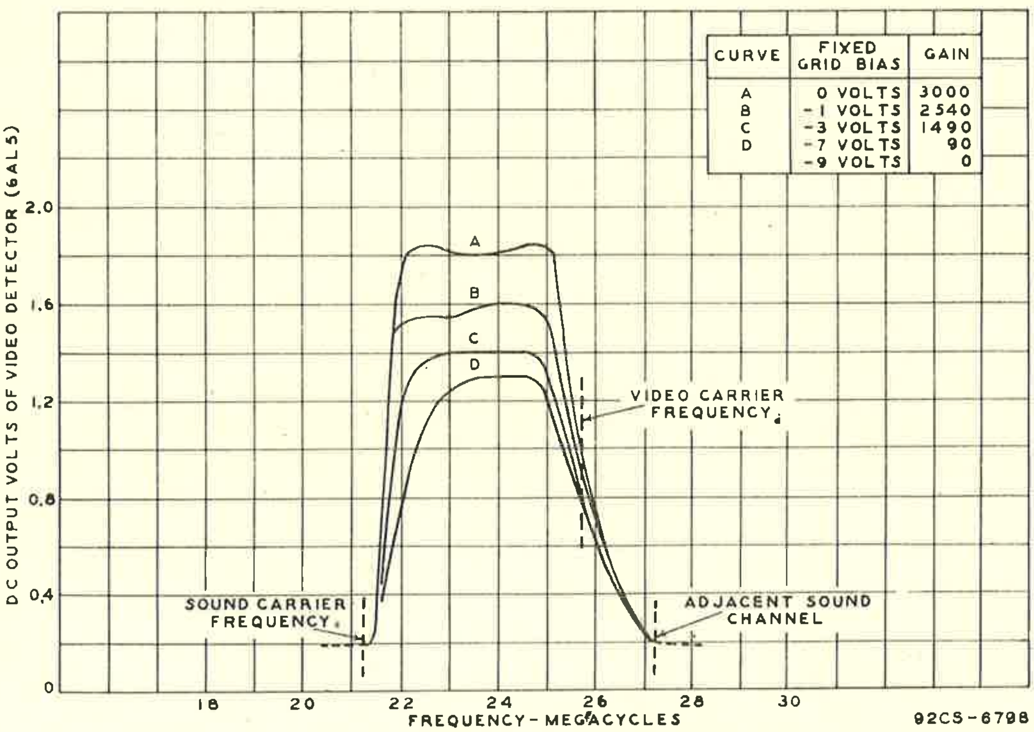


Fig.4 - Frequency Response of Video IF Amplifier Using 6AU6 Complement

than 40 db. In the adjacent channels slight responses occur.

When the i-f response is allowed to be appreciable at 21.25 megacycles, the insertion of the rejection traps results in the development of transients which occur when a signal is applied having frequency components extending into this region. A circuit having a very steep-sided response curve with a sharp corner at the 21.25-megacycle sound-carrier frequency will cause a loss in resolution evidenced by blurring of the centre portion of the vertical wedge of the television test pattern. It is good practice to avoid this defect by rounding off the i-f response curve slightly between 21.25 and 22 megacycles.

Amplitude Characteristics.

When a steady carrier is impressed on the video i-f system, a dc voltage is, of course, developed by the video detector. In Figs. 5 and 6, dc voltage obtained with varying amounts of fixed grid bias is plotted against impressed signal strength to show the amplitude characteristics. These characteristics are nearly straight lines which represent a desired condition for essentially undistorted video signals derived from modulation of the carrier. The dc output voltage indicated in Figs. 5 and 6 is adequate for receivers incorporating one stage of video amplification, assuming the video gain to be in the order of 30 to 50.

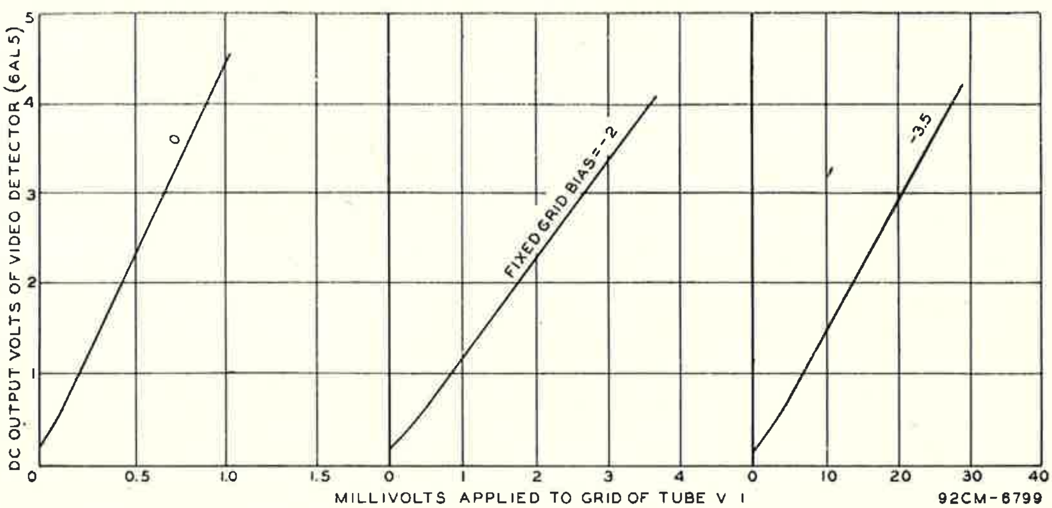


Fig.5 - Amplitude Characteristic of Video IF Amplifier Using 6AG5 Complement

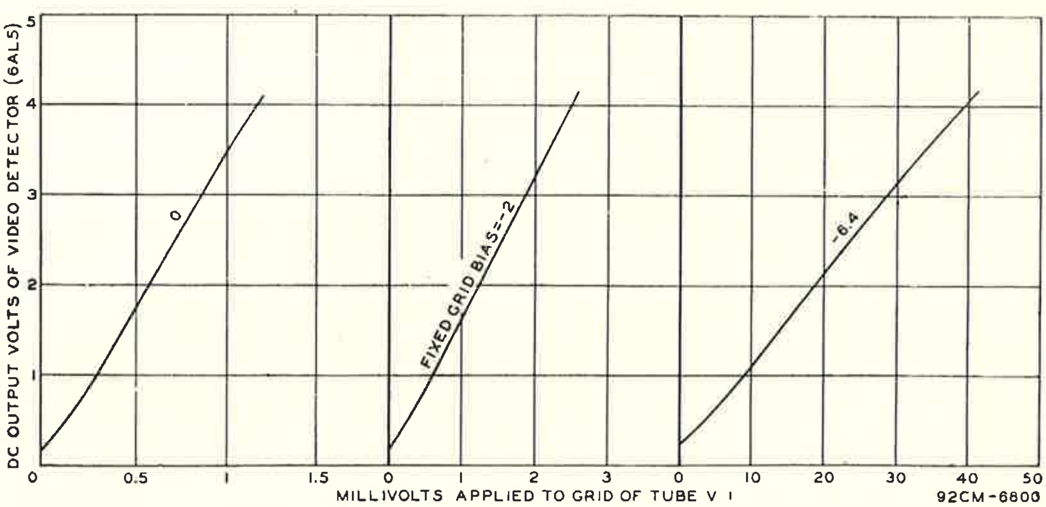


Fig.6 - Amplitude Characteristic of Video IF Amplifier Using 6AU6 Complement

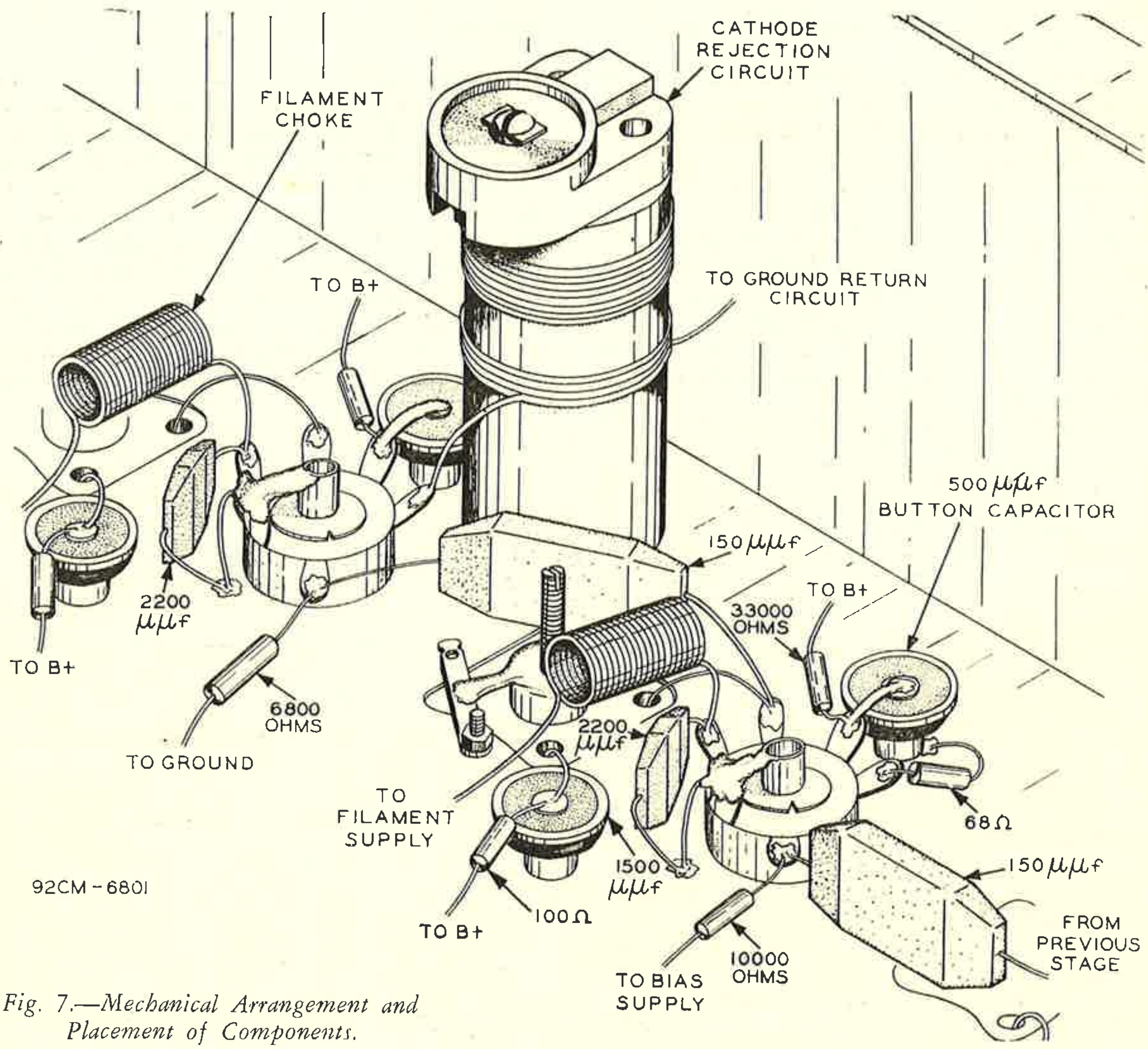


Fig. 7.—Mechanical Arrangement and Placement of Components.

Miscellaneous Constructional Details.

Construction of multistage video i-f amplifiers requires careful technique in design and wiring in order to minimize regeneration and inter-stage coupling. The use of miniature tubes is helpful because they permit a very compact construction. If the coils used are exactly as described in this Note, it is necessary to duplicate closely the mechanical arrangement (Fig. 7) used in this i-f amplifier in order to obtain similar results. Appreciable departure from this layout will undoubtedly result in the need for redesigning some of the tuned circuits because stray wiring capacitances will not be duplicated.

The grid No. 2 (screen) bypass capacitors used in this amplifier are of the ceramic button type and are mounted as close as possible to the tube base pin (Fig. 7). Heater chokes are employed to prevent undesirable coupling between stages. The untuned type of choke which serves to decouple each stage from the common supply impedance is generally satisfactory. In extreme cases, however, it may be necessary to use tuned parallel circuits resonating at the i-f frequency in place of the chokes. Resistance-capacitance filters are included in each plate circuit to prevent coupling through the common B+ line. Although the amplifier will work satisfactorily with certain of these decoupling circuits omitted, such practice is not recommended because the arrangement of the leads would become more critical and production difficulties, therefore, would probably result.

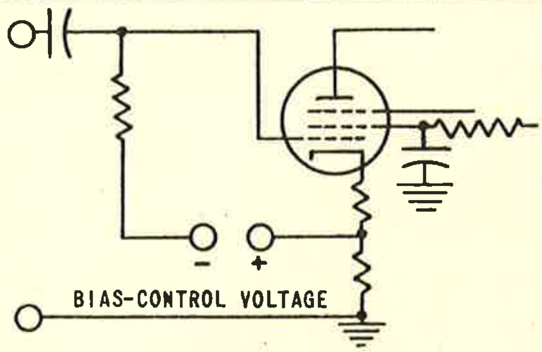
It is also important to use a mechanical arrangement which will afford short leads so that wiring capacitances may be kept to a minimum value. The control-grid and plate circuit leads should not be dressed against the chassis but should be spaced away from it so as to minimize capacitance to

(Continued on page 116)

TABLE I

Tube Type	Interelectrode Capacitances			Increase in short-circuit input capacitance from value at cutoff to value at g_m indicated.		Value* of unbypassed cathode resistor needed for complete compensation of input capacitance change with bias change.	Gain Factor. Degeneration resulting from use of unbypassed cathode resistor.
	C_{in} μμf	C_{out} μμf	C_{gp} μμf	g_m μmhos	ΔC μμf	R_k ohms	
6BA6	5.5	5.5	0.0035	4400	2.6	100	0.62
6AU6	5.5	5.0	0.0035	5200	2.6	85	0.61
6AG5	6.5	1.8	0.025	5100	1.4	50	0.75
6AK5	4.0	2.8	0.02	5100	1.2	50	0.75
6BJ6	4.5	5.0	0.0035	3800	1.8	135	0.59
6BH6	5.4	4.4	0.0035	4600	1.8	110	0.59

* The values of the unbypassed cathode resistor given for types 6BA6 and 6AU6 exceed that required for biasing. To realize full transconductance, either the control-grid circuit should be returned to a positive compensating potential or else the grid resistor should be tapped on the cathode bias resistor as illustrated. In Fig. 2, note that the cathode resistor is 68 ohms as only partial compensation is required.



$$\text{Gain Factor} = \frac{\text{Gain with cathode unbypassed}}{\text{Gain with cathode bypassed}} = \frac{1}{1 + R_k \left[g_m \left(\frac{I_b + I_{c2}}{I_b} \right) \right]}$$

Where: R_k, the cathode resistor, for complete compensation of input capacitance change with bias is approximately

$$R_k = \frac{\Delta C}{C_{gk} \left[g_m \left(\frac{I_b + I_{c2}}{I_b} \right) \right]}$$

ΔC = change in input capacitance in farads from normal operating condition to cutoff.

C_{gk} = grid-to-cathode capacitance in farads measured with tube cold.

g_m = grid-to-plate transconductance in mhos at normal operating condition.

I_b = plate current (dc) in amperes.

I_{c2} = grid-No. 2 (screen) current (dc) in amperes.

ground. The i-f amplifier transformers (Fig. 8) are shielded in this design, although satisfactory results may be obtained without the use of individual shields because no two circuits are tuned to the same frequency. Removing the shields, however, may seriously affect the ease of circuit adjustment because regeneration may occur before the circuits are properly aligned.

Rejection traps of the absorption type are somewhat critical with respect to coupling and, therefore, adjustment of the spacing between the coils may be required. The traps reflect a low impedance at their parallel-resonant frequency, thereby providing attenuation. At a slightly lower frequency, the traps introduce a high impedance effecting a slight rise in the response curve. The cathode rejection trap in stage V4 is tuned to attenuate this response. The coil in the cathode circuit (L10) should be in series resonance with the stray cathode-to-ground capacitance at a frequency approximately equal to the resonant frequency of L5. If the coil and capacitance are not in resonance, a loss in stage gain will result because of the degenerate effect of the inductive reactance in the cathode circuit.

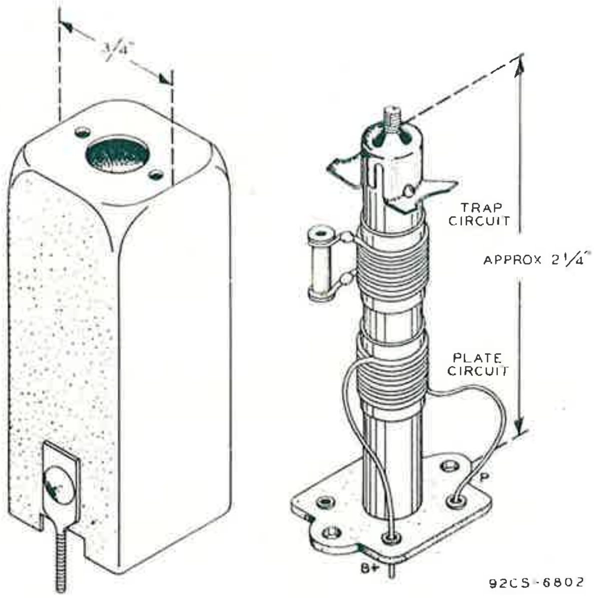


Fig. 8 - Mechanical Details of IF Transformer

Alignment Procedure.

It is suggested that the following procedure be used for aligning the i-f system. From a fixed-frequency signal generator, apply the theoretical alignment frequency to the grid of each stage starting at the last i-f stage and working back to the converter grid. Connect a low-capacitance high-frequency voltmeter to the plate circuit of the stage being aligned and adjust the tuned circuit for maximum output. After each circuit has been tuned to the theoretical alignment frequency, a sweep generator may be used to check the over-all response. Only slight additional adjustment of the i-f coils should be needed to provide proper response.

APPENDIX

The following formulas* apply to the calculation of the resonant frequencies and the values of the grid damping resistor for stagger-tuned circuits when the system dissipation factor (δ), which is the ratio of bandwidth (Δf) to geometric mean frequency (f_o) as defined below, is less than 0.3.

$$\delta = \frac{\Delta f}{f_o} = \text{ratio bandwith to geometric mean frequency.}$$

$$\Delta f = f_2 - f_1 \qquad f_o = \sqrt{f_1 f_2}$$

where f_1 = lower frequency edge of pass-band in cycles per second (0.707 response);
and, f_2 = upper frequency edge of pass-band in cycles per second (0.707 response).

With the values f_o and δ of the amplifier determined and with the desired number of stagger-tuned stages assumed (either four or five), substitution of these values in the formulas below furnish the required resonant frequency and the dissipation factor for each stage.

*Based on M.I.T. Radiation Laboratory Report No. 524 by H. Walman.

Formulas for Four-Stage Stagger-Tuned System.

†Stage	Resonant Frequency	Dissipation Factor
A	$f_o(1+0.46\delta)$	0.38δ
B	$f_o/(1+0.46\delta)$	0.38δ
C	$f_o(1+0.19\delta)$	0.92δ
D	$f_o/(1+0.19\delta)$	0.92δ

Formulas for Five-Stage Stagger-Tuned System.

†Stage	Resonant Frequency	Dissipation Factor
A	f_o	
B	$f_o(1+0.29\delta)$	0.81δ
C	$f_o/(1+0.29\delta)$	0.81δ
D	$f_o(1+0.48\delta)$	0.31δ
E	$f_o/(1+0.48\delta)$	0.31δ

The grid damping resistor (theoretical value) for each stage is determined from the dissipation factor and the resonant frequency of the stage by means of the following equation:

$$R = \frac{1}{2\pi f_r C d}$$

where R = theoretical value of damping resistance in ohms.
f_r = resonant frequency of the stage in cycles per second.
d = dissipation factor of the stage.
C = total capacitance across circuit in farads (including tube and wiring capacitance).

The average stage gain (m) at resonance is

$$m = \frac{g_m}{2\pi C \Delta f}$$

where g_m = grid-plate transconductance in mhos.
The over-all gain (M) of the stagger-tuned system is

$$M = m^n$$

where n = number of stagger-tuned stages
m = gain per stage.

†Sequence of stages is determined by circuit considerations such as loading effect of diode, location of sound i-f selection coil, and location of rejection traps.