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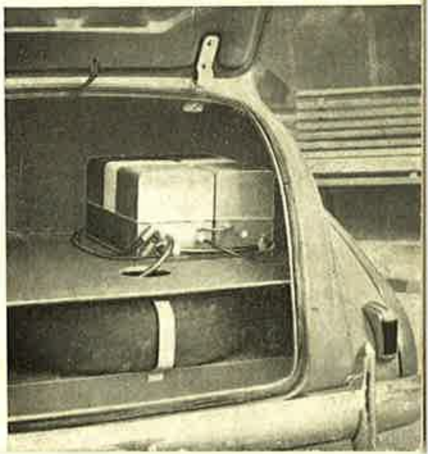

RESEARCH • DESIGN • MANUFACTURE

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CONTENTS

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Communication *Receiver Design*

By DENIS HEIGHTMAN

It is proposed in this paper to review the main requirements and factors governing the design of a modern general - purpose communications - type receiver from the point of view of (*a*) the potential buyer and operator and (*b*) the set designer and manufacturer. Consideration will be confined to A.M. (A3), M.C.W. (A2) and C.W. (A1) reception on frequencies between 150 Kc/s. and 35 Mc/s., with emphasis on those bands assigned to amateurs.

The average operator, when asked to indicate the most important features he requires in a receiver, will produce a list such as follows:—

(1) Good sensitivity and signal-to-noise ratio, permitting the reception of the weakest useable signals.

(2) High selectivity, preferably controllable to extreme limits for C.W. reception.

(3) Good stability, both electrical and mechanical.

(4) Freedom from spurious responses (images, whistles, etc.).

(5) Convenient layout of controls with pleasing appearance.

(6) Accurate calibration of tuning controls with adequate band spread, preferably with a means for checking calibration.

(7) Accurate-reading signal-strength meter.

(8) Effective noise-limiter.

(9) General reliability and robustness.

(10) Reasonable price.

Inevitably the last feature—that of price—will be one which has to be closely watched by the manufacturer, who, normally, within limits, controls the designer. Without having to consider cost a designer could undoubtedly produce the ideal receiver but it would be beyond the reach of all but the chosen few.

Signal-to-noise considerations

To achieve good sensitivity we should be quite clear on what decides the maximum gain that can be effectively used. Actually, to a marked degree, this has been set by nature! Cosmic and solar radiations set up in any receiving aerial small random voltages which produce noise in a sensitive receiver very similar to valve noise or hiss. The magnitude of this noise is dependent to some extent on the time of day, aerial gain and directivity, frequency and band-width. In the frequency band under consideration the noise generally tends to be greater the *higher* the frequency, due, it is thought, to the fact that on lower frequencies the ionosphere reflects or absorbs more effectively this incoming radiation. However on frequencies lower than about 15 Mc/s., the effects of static and electrical interference are more pronounced while the effects of extra-terrestrial radiation are negligible. In general too, on the lower frequencies the receiver will be working with higher signal strengths due to the shorter distances involved, and the gain requirement will be less than that on the higher frequencies. Thus the ultimate gain requirements will be set by the aerial noise received at the highest frequency range of the receiver, for, obviously, a signal which provides *less* voltage at the receiver input than the aerial noise will be unreadable.

To consider actual figures, very roughly it can be stated that at 30 Mc/s. (nearly the highest frequency normally used in this class of receiver) with a band-width of 10 Kc/s. the aerial noise will appear as a voltage of the order of 0.25 microvolt at the end of a 100-ohm feeder line. For a reasonably intelligible signal we should have at least a 6 db. signal-to-noise ratio. In other words our minimum usable signal becomes of the order of 0.5 microvolt, and neglecting other factors, the receiver has to provide a loudspeaker signal from this r-f input voltage, preferably ***not*** introducing additional noise of its own making. An overall gain of the order of 120 db., *i.e.* voltage $\times\ 10^6$ will bring the 0.5 μV. to 0.5 V., which is the order of audio signal required to drive the normal audio amplifier and output stages in the receiver. This is in terms of a C.W. signal or 100 per cent. modulated telephony. Thus we can fix an approximate gain from the aerial input to the second detector output (*i.e.* the r-f/i-f section) of 120 db., and further consider as to how this gain should be distributed over the various stages of the receiver.

High selectivity (2) and economy (10) dictate that as much gain as possible should be obtained at a low intermediate frequency, for it is much simpler to obtain optimum performance from stages operating at a single frequency, as with an i-f amplifier, than from continuously-tuned r-f stages. As the frequency is lowered so also will the band-widths of the tuned circuits be less for a given Q.

On the other hand good sensitivity and signal-to-noise ratio (1), as well as freedom from spurious responses (4), necessitate the use of a properly designed r-f stage (or stages) to overcome the effects of mixer noise and a sufficiently high i-f to give good image rejection.

Valve noise

To revert to the "front end" once more it will be of interest to examine what is required of the r-f stage. Valves produce noise voltages due to fluctua-

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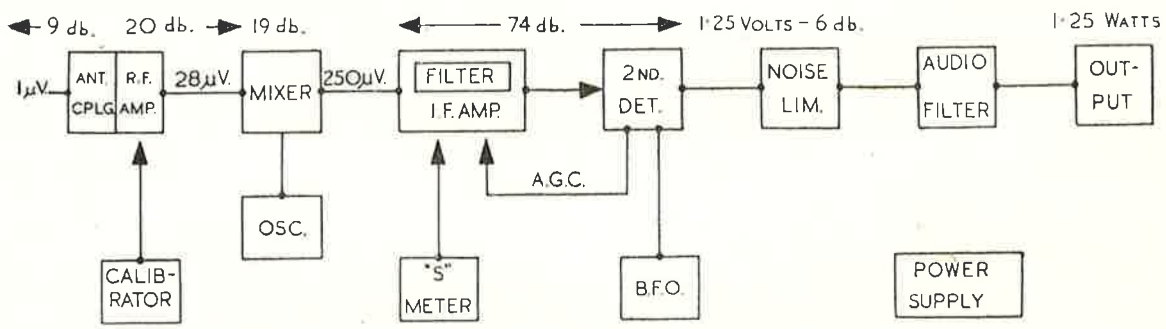


Fig. 1. Block diagram of the receiver discussed in the text, showing approximate gains (db.) per stage and indicating the voltages appearing at various stages when a 1µV signal is applied to the aerial terminals.

tions in the electron flow between cathode and plate. It is convenient to consider these valve noise voltages as if the valve were perfect and the noise source were in the grid circuit, as a resistor or generator. The noise voltage, referred to the grid of a typical r-f pentode, is of the order of 1.0 – 1.5 µV. at 10 Kc/s. band-width. Modern miniature high slope pentodes of the 6AK5 or the Z77 class produce a rather lower figure, of the order of 0.5 – 0.6 µV. Mixer valves unfortunately produce considerably more noise. On the same basis, the noise at the grid of a typical mixer is 5 – 6 µV.

To consider again the minimum signal-readable-above-aerial noise, *i.e.* 0.5 µV., the usual aerial coupling circuit will at 30 Mc/s. provide a step-up of the order of 3 – 4 times and thus the signal appears at the first grid at about 2 µV. Obviously, if this signal is applied to a mixer with 5 µV. of noise it will be unintelligible. Therefore our first requirement is to present the signal to the grid of the valve with the lowest noise figure and amplify it to a sufficient degree in the r-f stage to overcome the extra noise of the mixer. The miniature pentode with the figure of 0.5 µV. will add a negligible amount of noise to the aerial noise (which stepped up in the aerial coupling circuit will appear at the grid as 1 µV.).

At 30 Mc/s. it is easy to obtain a gain of 20 db. in a single stage using one of these high-slope pentodes, *i.e.* a 2 µV. signal at the grid can be presented to the following stage as 20 µV. which will be at a sufficient level to make the contribution of the mixer to the general noise negligible. (The amplified noise from the r-f grid and aerial will appear as about 11.0 µV. at the following grid.) Thus considering requirements (1) and (10) it will be seen that a single r-f stage, provided it is well designed, will be sufficient.

Note that in the foregoing paragraphs a band-width of 10 Kc/s. has been assumed. By using considerably narrower band-widths, particularly for C.W. reception, the general noise level, which is proportional to the square root of the band-width, will be reduced and it will be possible to receive a weaker signal with the same signal-to-noise ratio. For example, if the band-width were dropped to 2.5 Kc/s.—*i.e.* $\frac{1}{4}$ the previous value the noise would be halved (*i.e.* $\times \sqrt{0.25} = 0.5$). Thus a signal of half the original voltage could be received (*i.e.* 0.25 µV. at the receiver input).

The picture can now be looked at from the point of view of selectivity (2) and freedom from spurious responses (4), particularly in regard to image rejection. Here the choice lies between using a few stages of low intermediate frequency with additional r-f stages, or rather more stages at a medium i-f with a single r-f stage.

Image rejection will become progressively worse as the receiver is tuned higher in frequency. Hence taking the worst possible case and considering reception at about 30 Mc/s., it will be found that an average single r-f stage with a 1,600 Kc/s. i-f (*i.e.* image 3.2 Mc/s. from the signal) will give a signal-to-image ratio of about 40 db., which is adequate for normal purposes. With one r-f stage and a 465 Kc/s. i-f however, the rejection will only be approximately 20 db., *i.e.* an S9 image signal would appear at S5/6. On the 28 Mc/s. amateur band this is particularly undesirable since the images 930 Kc/s. (i-f × 2) away from the desired signal may still fall within the band. For example, a signal on 29.0 Mc/s. may be interfered with by transmissions on 28.07 Mc/s. Thus two r-f stages are necessary with i-f's of the order of 465 Kc/s. in order to give only the *same* image rejection as with one r-f stage and a 1,600 Kc/s. i-f.

Two stages of r-f., with their attendant tuning, band-change, multiple coils, instability problems, etc., are costly in production and, as has been shown by the foregoing remarks, only confer the advantage of permitting the use of a lower i-f. If sufficient i-f selectivity can be obtained at 1,600 Kc/s., then it is obviously desirable to choose this frequency. In practice, by using three instead of the more normal two i-f stages, it will be found that ample selectivity is obtainable. In other words one of the two r-f stages needed with a 465 Kc/s. i-f can be replaced by an extra i-f stage if a 1,600 Kc/s. i-f is chosen. The latter arrangement is more economical. An i-f of 1,600 Kc/s., is about the most convenient value in practice, falling as it does between the h-f end of the medium wave-band (1,550 Kc/s.) and the l-f end of the 1.8 Mc/s. amateur band. With careful design and screening it will be found possible to work within 50 Kc/s. of the i-f before instability occurs, or considerably closer if the r-f gain is reduced.

Frequency range

Having sketched the broad outlines of a suitable receiver, it is now opportune to consider further, in rather more detail, other aspects of the design. Of these the first to be decided is the frequency range to be covered by the r-f tuned circuits. For the lower frequencies a 3 : 1 tuning ratio is satisfactory. If continuous coverage is required from, say, 175 Kc/s. (i.e. including the long wave broadcast band), this can be achieved by using three bands, namely, 175 – 525 Kc/s.; 515 – 1,545 Kc/s. (allowing 10 Kc/s. overlap); 1,650 – 4,950 Kc/s. (leaving a gap at the i-f of 1,600 Kc/s.). On the higher frequencies a 2 : 1 ratio is more desirable in a continuous coverage receiver, both in the interests of less critical tuning and also of circuit performance from h-f to l-f limits of any particular range. This can be achieved by adopting three more bands, namely, 4.8 – 9.6 Mc/s.; 9.4 – 18.8 Mc/s.; 18 – 36 Mc/s. The restricted ratio on these ranges can be easily obtained by the insertion of suitable fixed condensers in series with the main tuning condensers, without materially compressing the greater part of the frequency range to one end of the tuning scale.

Good performance necessitates the keeping of circuit capacities to a minimum, *i.e.* maintaining a high *L/C* ratio. The lower the minimum circuit capacities, the lower the maximum value of tuning condenser required (since higher value inductances will be used). In practice a tuning condenser with a maximum capacity of about 310 $\mu\mu$F. and a minimum capacity of about 11 $\mu\mu$F. will allow the required coverages to be obtained. A low value air trimmer can be used for adjusting circuit minimum capacities, within the tolerances required for production. Including valve input capacity, wiring and other stray capacities, a total circuit minimum of the order of 35 $\mu\mu$F. is permissible.

Low minimum capacities and questions of electrical stability are arguments in favour of mechanical methods of band-spread in preference to the addition of small variable condensers in parallel with the main tuning as is frequently done. The question of band-spread will be considered later, however.

Constant gain

Whilst not very obvious, there are distinct advantages in a circuit arrangement which gives substantially constant gain over the receiver tuning range. In particular, requirement (7)—accurate "S" meter readings—necessitates such an arrangement. With many receivers the gain in the r-f and mixer circuits is progressively greater the higher the set is tuned on any particular coil range. Often the amount of gain is decided by such questions as that of preventing instability at the h-f ends on the tuning ranges with a consequent falling off in performance at the l-f ends.

Obviously the gain in the i-f circuits is constant. Constant overall gain therefore will be dependent on the characteristics of the aerial input, r-f and mixer circuits. In the case of the first tuned circuit with its aerial coupling, the gain will increase with frequency on any particular range. This rising characteristic can be offset if an anode circuit which has a falling characteristic is used in the r-f stage. This falling characteristic can be obtained if a fixed high *L* circuit, tuned to a frequency about 0.6 of the lowest frequency on a particular range, is employed as the primary of the coupling transformer between the r-f anode and the mixer grid circuits. The gain on any particular range can also usefully be determined by correct choice of the *L/C* ratio in this circuit. A very small coupling capacity (about 2 $\mu\mu$F.) between the r-f anode and the mixer grid circuit will correct a tendency for excessive drop in gain towards the h-f end of the tuning ranges.

The actual figure of overall r-f stage gain which can be used over the receiver tuning range must be decided by the maximum that can be obtained on

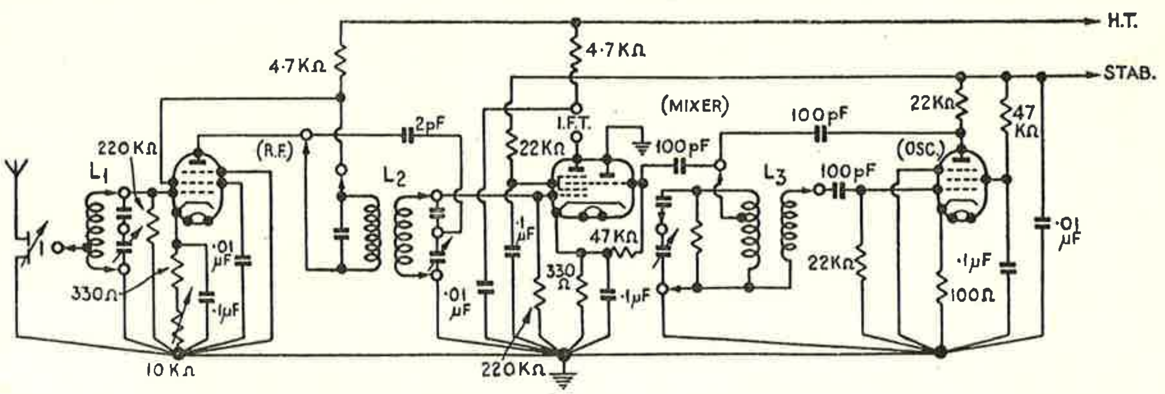


Fig. 2. R-F, mixer and oscillator circuit.

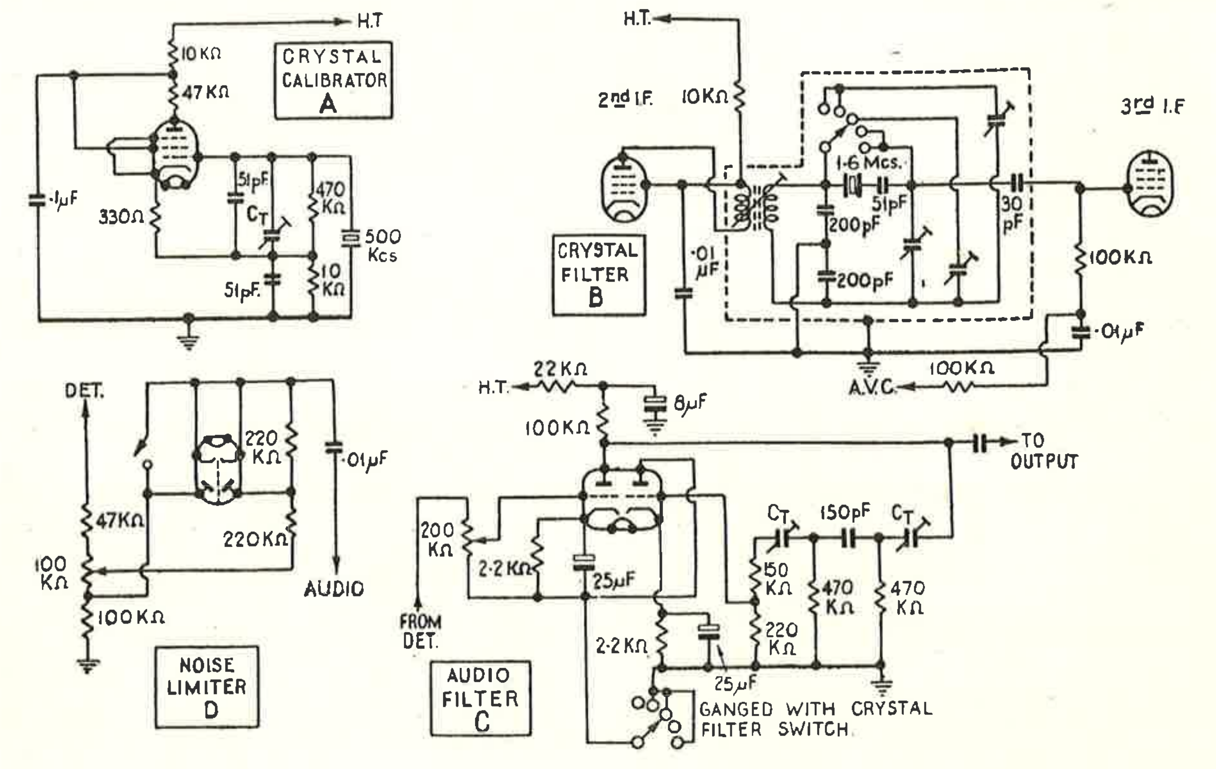


Fig. 3. Circuit Details of (A) crystal calibrator, (B) crystal filter, (C) audio filter, and (D) noise limiter.

the highest range, namely 18 – 36 Mc/s., since on these frequencies the valve input impedance and tuned circuit dynamic resistance will be lowest. With valves of the Z77 class it will be found that, if due precautions are taken, stable performance at a gain of 20 db. plus 9 – 10 db. in the aerial coupling can be obtained. Care should be taken to see that all unwanted or unapparent couplings from r-f grid to anode circuits are eliminated. For instance it has been found necessary to isolate the rotor spindle of the r-f section of the tuning gang condenser from that of the mixer section, as sufficient coupling in the common earth return through the spindle and wiper arm can take place to produce instability.

As has already been mentioned, the gain on the lower ranges can be made equal to that on the highest frequency range by careful adjustment of the *L/C* ratio in the r-f anode circuit and also the coupling factor to the mixer grid circuit. In this way stability is also assured since, obviously, excessive and uncontrollable gains would be obtained on the lower frequencies if a valve of this class were used without some such limiting arrangement.

Correct matching of the aerial or aerial feeder into the receiver is necessary for optimum performance. For a fairly wide range of impedance variations this can be effectively taken care of by a capacity potentiometer network shown which, in practice, takes the form of a 100 μμF. differential variable condenser.

Mixer stage

Mixer requirements for constant gain or conversion are somewhat rigid when considering the wide range covered by the receiver. With many forms of mixer valve there are stray internal couplings of which the effect, and consequently the conversion conductance, varies with frequency. On the higher frequencies small capacity couplings introduce grid injection which increases the conversion as the tuning capacity is reduced. If a high impedance d.c. grid circuit is used for a.v.c. application on frequencies over 25 Mc/s., a small amount of grid current flows which tends to bias the grid negatively and to reduce conversion conductance. In order to maintain constant conversion the injected oscillator voltage must not vary to any great extent over the tuning range. The restricted tuning ratio on the higher frequencies is advantageous in this respect, as is also the use of a separate high slope oscillator valve.

Of the present-day mixer valves, the X61M class has been found to be the most suitable, bearing in mind the above requirements, but it is preferable, in the interests of constant conversion and stability, to use a separate oscillator valve. On spot frequencies it is, of course, possible to obtain a much higher conversion conductance than the 0.7 mA./V. of the X61M but it is almost impossible to maintain the higher figure over a wide frequency range. With the suggested mixer an approximate gain from signal to i-f of 16 db. can be maintained.

The effect of variation of d.c. potentials on the grid and screen grid of the mixer is important when using high selectivity in the following i-f stages, even with a separate oscillator. Variation of the negative grid voltage as obtained by a.v.c. action is sufficient to produce a reflected variation, by capacity, of the order of 500 – 1,000 c/s. in the oscillator frequency, particularly on frequencies over 10 Mc/s. The effect of screen voltage variation is the opposite to that of the grid. If the B supply regulation is not good, such variation can occur with a.v.c. action. For these reasons it is considered preferable to exclude a.v.c. from the mixer grid and it has been found that the overall characteristic does not suffer materially thereby.

The screen voltage of the X61M mixer is fairly critical for optimum conversion. Owing to the fact that the screen current varies over wide limits with small variations in injected oscillator voltages, etc., and with the usual tolerances met in valves and components, the screen supply should preferably be obtained from a potentiometer network or from a stabilised source.

Oscillator and i-f stages

The requirements of the oscillator have been largely considered under the mixer heading. Questions of tracking will not be dealt with here for the subject has already received considerable attention in various publications. In the interests of stability, maintenance of calibration, etc., great care, both electrical and mechanical, must, of course, be paid to this circuit, probably more than to any other section of the receiver. Stabilisation of the B supply is most desirable. In the interests of standardisation it has been found that the same class of valve as that used in the r-f stage performs satisfactorily as an oscillator. The high slope permits resistance damping of the tuned circuit so as to maintain constant output.

Thus far gains have been obtained of say 10 db. in the aerial circuit, 20 db. in the r-f stage and 16 db. in the mixer—a total of 46 db., leaving an overall i-f amplifier gain requirement of 74 db. to meet our previous estimate of 120 db. necessary up to the second detector. A gain of 74 db. *could* be obtained in two stages but the use of an additional stage allows better selectivity. Three stages also provides a more stable arrangement as it will be no longer necessary to squeeze the utmost from each stage. In fact, other considerations may favour the *loss* of some of the possible gain of three stages; first, to improve selectivity by using under-coupled i-f transformers (about 0.3 coupling factor); second, to avoid detuning effects of a.v.c. action and give wider tolerances in production by using fairly high fixed tuning capacities in the i-f's, and lastly to provide a gain-compensating arrangement in order to equalise gains when a crystal filter is brought in and out of operation.

Suitable valves for the i-f stages to conform with the B7G series* used in the r-f stage, etc., are the W77 and 6BA6, variable-*mu* type. When used together with modern pot-cored small type i-f transformers a compact form of construction is possible. The a.v.c. decoupling and feed resistors should not exceed a total of 1 megohm in order to avoid a positive voltage building up on the a.v.c. line due to residual gas effects, etc., in these valves.

The use of three i-f stages also permits a good a.v.c. characteristic to be obtained without the necessity of applying a.v.c. to the r-f or mixer stages. For certain purposes, manual control of i-f gain is desirable. This can easily be provided by the inclusion of a potentiometer in the negative B supply line.

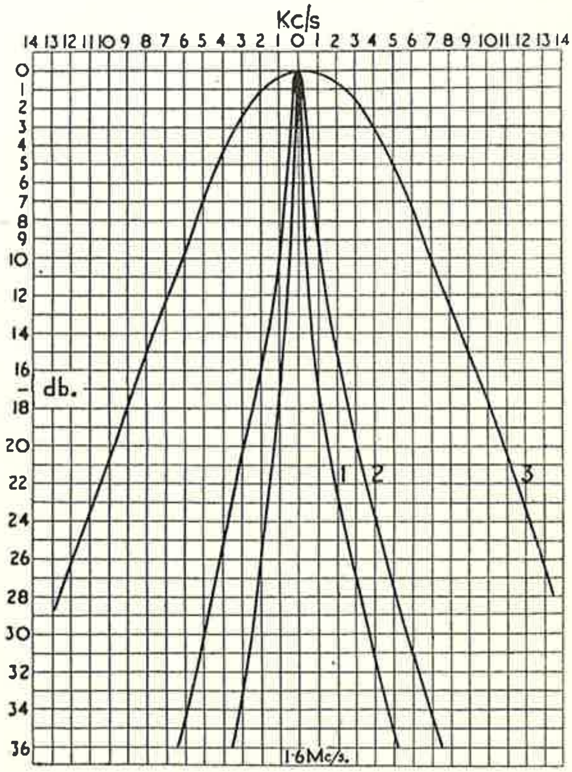


Fig. 4.

I-F response curves obtainable with the circuit of Fig. 3. (1) 0.5 Kc/s. crystal (2) 1.5 Kc/s. crystal (3) 8 Kc/s. crystal out.

Selectivity

Considering feature (2)—high controllable selectivity—it has to be decided what arrangement will meet the average practical requirements. Generally speaking for reasonably good quality A-M 'phone reception, a band-width of the order of 8 – 10 Kc/s. for 6 db. down is required if excessive cutting of the higher audio frequencies is not to take place. For 'phone reception under difficult adjacent channel interference conditions, an equivalent figure of 3 Kc/s. or less would be desirable. Unless, however, a rather expensive and complicated band-pass crystal filter is used the value of 3 Kc/s. is too wide to achieve in a normal crystal filter. A value of 1.5 Kc/s. bandwidth, however, can be obtained with a single crystal

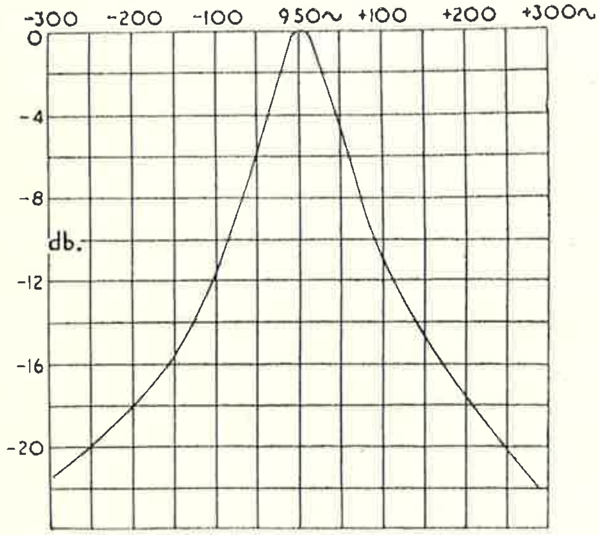
* Standard 7-pin miniatures.

filter at 1,600 Kc/s. when it is tuned for maximum band-width and in practice this value has been found to be about the narrowest useful band-width for intelligible 'phone reception under bad conditions of interference.

C.W. reception can be accommodated in much narrower band-widths provided both the transmitter and receiver frequencies are sufficiently stable. A value of 500 c/s. can be obtained with the simple crystal filter circuit shown and this has been found adequate, used in conjunction with a tuned audio filter, for reception under the worst conditions of C.W. interference.

Thus is arrived at a minimum selection of degrees of band-width. Modern trend, which is probably desirable, is to pre-align the crystal filter to the required band-widths and use a switch to control the selectivity rather than bring various filter controls, which are tedious to adjust, to the front panel. Provided good quality air trimmers are used in the crystal filter it will "stay-put" almost indefinitely.

The crystal filter is most conveniently placed between the second and third i-f stages where it is unaffected by other variable factors. If inserted in the mixer anode circuit difficulties arise at certain frequencies due to the untuned primary of the filter transformer causing regeneration.



Audio filter response curve.

In the complete receiver a ganged switch can be arranged for selectivity control which also adjusts the value of cathode resistor in the final i-f stage in order to compensate for any loss of gain when the filter is switched in. This switch can also control on a further position, the switching in or out of the audio filter, of which more later.

The second detector/a.v.c. rectifier should preferably be of the isolated double-diode type in order to avoid unwanted stray capacities which occur in double-diode-triodes, etc. The 6AL5 or D77 are most suitable.

The i-f transformers may be designed so that the a.v.c. diode, fed from the i-f amplifier, receives approximately three times the voltage applied to the detector diode from the secondary of the last i-f transformer. This allows for a reasonable delay voltage to be used to improve the a.v.c. action. The delay is arranged so that a.v.c. action will just commence with 1 μV. input to the receiver.

It is convenient to measure signal strength by a meter which indicates changes in voltage, due to a.v.c. action, at an i-f valve cathode. Other methods require a meter amplifier stage or a very sensitive meter both of which are uneconomical. Obviously, the meter will not register until the signal has overcome the delay voltage of the a.v.c. diode. Thus 1 μV. will produce the first meter reading. From experience a 1 μV. signal under average conditions is S2. It is therefore preferable to omit the S1 meter reading (which will be about a 0.5 μV. signal) and commence at S2 with 6 db. steps to S9, which will be equivalent to an input of 125 μV.

For C.W. reception the B.F.O. is an important consideration. Stability of a high order can be obtained by the use of high C/L circuits, using an electron coupled or similar oscillator circuit. It is desirable to use a.v.c. even for C.W. but normally this cannot be done, due to the output from the B.F.O. leaking into the i-f stages and operating the a.v.c., etc. If considerable care is taken in screening the entire B.F.O. circuit (including the injection feed line and with r-f chokes in the heater circuits) this detrimental effect can be overcome. Using a longer time constant in the a.v.c. circuit enables signal strengths to be read and effective a.v.c. action to be obtained with C.W. reception.

A built-in crystal calibrator, consisting of a very simple 500 Kc/s. oscillator circuit, rich in harmonics, provides the only certain manner of checking accurately the calibration of the receiver and should be considered almost essential.

Noise limiters have received considerable attention in the technical press, and for that reason will not be dealt with in detail here. The circuit shown is automatically adjusted to the signal level by bias derived from the detector load resistor and has an adjustment for modulation depth—both desirable features. It should be noted that, due to a.v.c. action with strong noise or interference, it may be preferable for improved limiting to use the manual i-f gain control.

An audio filter circuit is a decided advantage in providing still greater selectivity for C.W. reception. The circuit shown, avoids the use of an iron-cored choke or transformer, both of which tend to give hum pick-up troubles and which are liable to more service troubles than a resistor capacitor network. The filter arrangement can most conveniently be switched-in (as already mentioned) by means of the i-f selectivity switch. In the "off" position the first half of the double triode acts as a straightforward audio amplifier. In the filter position the second triode is switched in series with the cathode of the first triode, and provides considerable negative feed-back, attenuating all frequencies other than a small band at 1,000 c/s. or as determined by the constants of the *R/C* network between the anode of the first triode and the grid of the second. With this arrangement an audio band-width of the order of 100 c/s. is provided. Even when used without a crystal filter (for example, with the 8 Kc/s. i-f band-width), the audio filter is found to be an extremely useful feature for C.W. reception.

The audio stage needs no comment as the usual modern output pentode or tetrode provides 3 – 5 watts of audio power. Power supplies are also more or less standard practice. Frequently a mains transformer of insufficient wattage rating is used, tending to overheat with continuous running, particularly in the tropics. A stabilised supply should be provided for the B.F.O. and oscillator stages.

Mechanical considerations

Having dealt fairly comprehensively with the electrical design of the receiver, it is now possible to consider mechanical matters. The final receiver layout will be largely decided by the factors (5) and (9) in the original list of operator's requirements, with some compromise in view of electrical requirements. These factors are convenience of layout of the controls, general reliability and robustness.

Since most operators are *right* handed and require to be able to *write* with the right hand whilst operating or tuning with the left, it is obviously preferable that the most frequently used controls should be towards the *left*-hand side of the front panel.

Probably the two controls used most in operation will be band-spread tuning and audio gain. The control knob operating the band-spread arrangement should therefore be placed at a height which allows the arm to lie comfortably on the operating desk so that the wrist and hand do not tire. The audio gain and other controls will preferably be fairly closely associated with this tuning control thus allowing for a mere movement of the wrist in order to change rapidly from one control to the other. Less frequently used controls can be distributed on the panel as decided by convenience of electrical connection. Normally the band-set and band-change controls will be comparatively seldom operated and thus their positions are not so critical.

Band-spread and band-set dials should be large and clearly marked. Long-term reliability and accuracy make it desirable that they should be directly connected (mechanically) to the tuning condenser or band-spread arrangement, rather than be dependent on cord or some other form of drive. The *drive* to the dials may, for economic reasons, take the form of a reliable spring tensioned cord and drum drive device, provided cord replacements do not entail a major operation.

It was indicated earlier in this paper that there are arguments in favour of mechanical as compared to "electrical" methods of band-spread. These may be summarised as:—

(1) Lower and less "distributed" circuit capacities resulting in better performance at the highest frequencies with less possibility of feed back.

(2) A more even band-spread ratio at all positions of the band set dial, especially in the case of general coverage receivers.

(3) A more compact arrangement is usually possible.

The arguments against mechanical band-spread lose their point if the design of the device is good. They are (1) backlash (2) insufficient band-spread.

There is much to be said for the provision of a complete angular framework as a basis for the receiver chassis. Not only does this provide a very strong and stable arrangement but also in either assembly or service readily allows the receiver to be placed upside down or in any other position without damage to internal components. The front panel, sides, top, back and bottom (all of which should be capable of being easily unscrewed for service purposes) may then be affixed directly to this framework in order to form a cabinet.

A point worthy of attention is that the alignment of the receiver should not be affected when placed in its cabinet unless provision is made for final adjustment by means of holes suitably positioned to give access to trimmers, etc. Careful internal layout and screening, however, should avoid this necessity.

Strong handles suitably fixed on the panel are not only useful for handling a receiver but provide important protection in transit.

Considering individual components and chassis layout, attention will be focussed largely on the tuning condenser, band-change arrangement and coil assemblies. Nowadays the typical gang condenser is a well engineered job and is likely to give very little trouble, although in order to avoid microphony the vanes should be well braced. Ceramic insulation of the stators is much preferable to phenol-impregnated-laminated-paper material not only because of the better insulation provided, but also because of the lower minimum capacity possible. The earlier reference to the need for the isolation of the r-f gang-section should also be noted. Trimmers should be of the air-spaced type.

Coils have a better long-term stability if wound on formers of a non-hygroscopic, efficient insulator such as polystyrene. Careful positioning and screening of the coils to avoid stray couplings, instability, etc., is essential. For band changing, a turret arrangement, which rotates the coils to their respective contacts, has decided advantages over the more frequently used wafer switch. The turret arrangement allows for (1) shorter circuit connections, (2) lower capacities between switch contacts and wiring, (3) longer leakage paths, and (4) easier service.

Adequate ventilation and the placing of the tuning

(Continued on page 121)

Circuit Laboratory Report. **Number 5.**

The Reduction of Playthrough

Playthrough, that is the output which remains when the volume control of a receiver is turned to its minimum setting, is experienced in many receivers. In Radiotronics 123 (January-February, 1947) most of the causes were described, and for a general discussion of the subject reference should be made to that article.

However, one further cause may be mentioned: in this case the effect occurs due to a capacitive voltage divider network between diode plate and control grid, and control grid and ground, as shown in Figs. 1(a) and (b). In some valve types the capacitance between diode plate and control grid is comparatively high, e.g. in the 6SQ7-GT it is of the order of 0.5 $\mu\mu$F for diode No. 1 and 1.0 $\mu\mu$F for diode No. 2. The other capacitance in the voltage divider, that between control grid and ground when the volume control is turned to its minimum setting, is seen from Fig. 1 to be the audio coupling capacitor. When grid leak bias is used with a 5 or 10 megohm grid leak this capacitor is often made quite small, perhaps 0.005 μF or less.

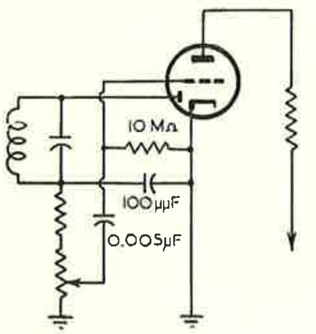


Fig. 1 (a).—Typical detector and a-f amplifier circuit using duo-diode-triode valve.

In a particular case with a capacitance of 0.5 $\mu\mu$F from diode to control grid and 5,000 $\mu\mu$F (0.005 μF) from control grid to ground, one ten-thousandth of the voltage on the diode will appear on the control grid.

In a receiver tuned to a strong station it is quite usual to find 30 volts developed across the diode load, and with heavy modulation this represents an a-f voltage of about 20 volts r.m.s. One ten-thousandth of this voltage, i.e. 2 millivolts will appear at the triode control grid and as most receivers using say a 6SQ7-GT and a 6V6-GT have an a-f sensitivity of about 14 millivolts input for 50 milliwatts output, this 2 millivolts will produce approximately 1 milliwatt with the volume control at zero.

An output of 1 milliwatt may represent excessive minimum volume from a sensitive speaker in a quiet room, and as assumptions of developed voltage and capacitances have been conservative throughout, this output may often be exceeded.

A simple method of improving playthrough from this cause is to increase the size of the grid coupling capacitor. With an 0.05 μF capacitor, the voltage on the triode grid will be reduced to one tenth and the power output of the receiver to one hundredth of its previous value.

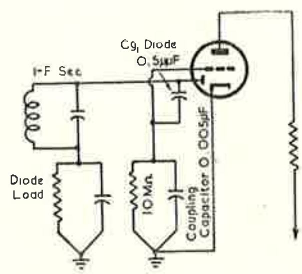


Fig. 1 (b).—Equivalent circuit when volume control is at minimum setting, showing capacitive voltage divider.

There is, however, a limit to the improvement which can be obtained in this way. If the triode is grid leak biased the grid leak is usually not smaller than 5 megohms (although 2 megohms is sometimes used) so that with an 0.05 μF coupling capacitor the time constant of the grid circuit is $5 \times 10^6 \times 5 \times 10^{-8}$, i.e. 0.25 seconds. This is about the maximum value that can be tolerated without time delay being noticeable when the volume control setting is changed rapidly with a strong station tuned in, or when the receiver is tuned rapidly over a strong station, with the volume control turned up. Generally, however, the playthrough is reduced to an acceptable level before this trouble is experienced. A lower value of grid leak together with a larger value of coupling capacitor can be used with still better results if the valve is separately biased.

If playthrough cannot be reduced sufficiently by this means, as for example in a reflex receiver, the circuit of Fig. 2 can be used. In this circuit, negative feedback is used to decrease the a-f gain of the receiver by an amount which depends on the setting of the volume control. The feedback is applied to the bottom of the control and, as there is a conducting path to ground through the control and the diode, the closer the arm of the volume control is

Contributed by the Circuit Design Laboratory,
Valve Works, Ashfield.

to its maximum setting, the smaller the amount of feedback. Thus, it is possible to decrease playthrough say five or more times for a gain reduction, with the volume control at maximum, of perhaps two times.

Other effects, for example hum, which are evident at the minimum volume control setting are similarly reduced. It may even be possible in some cases to decrease power supply filtering and minimize the resulting increase in hum by applying feedback to the bottom of the control if the hum can only be heard at the minimum volume setting. In this way the circuit of Fig. 2 may result in reduced receiver costs despite the addition of two resistors.

If only a very small loss of gain can be tolerated, the use of a volume control tapped at 50,000 ohms with say 5,000 ohms to ground from the tap will almost completely remove the effects of feedback above the tap, while allowing greater gain reduction with the volume control at minimum.

There is, however, a limit to the amount of gain reduction that can be obtained, because excessive feedback will lead to instability which usually becomes evident as motor-boating. This is the reason for the two resistors *R*1 and *R*2 in Fig. 2. With an a-f amplifier using 6SQ7-GT and 6V6-GT or similar valves and conventional coupling circuits, output transformer and loud-speaker, a receiver will be unstable if the full voice-coil voltage is returned to the bottom of the volume control. From one third

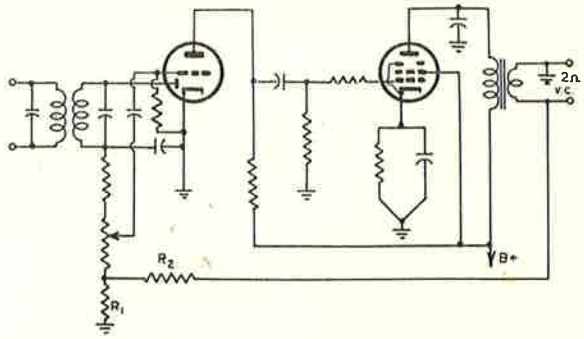


Fig. 2.—Circuit using negative feedback to minimize playthrough without seriously affecting maximum gain.

to one fifth of this voltage usually provides adequate feedback with a 2 ohm voice coil so that *R*1 can be 25 ohms (¼ watt carbon) and *R*2, 50 or 100 ohms. The resistance of *R*1 in series with the cold end of the volume control will not of itself noticeably affect playthrough and many commercial controls have a higher minimum resistance.

It will be seen that this type of playthrough reduction has been used in circuit RD34 published in issue 144. It has been possible to use the full voice-coil voltage for negative feedback in this case because the a-f gain of the reflex stage is much lower than that of a 6SQ7-GT. Nevertheless care was needed to obtain complete stability.

New RCA Releases

Radiotron type 1V2 — is a high voltage, half wave rectifier, utilizing a 9-pin miniature base. It is designed for use in compact pulse-operated rectifying systems. In voltage-doubler circuits, it is particularly suitable as a rectifier of higher voltage pulses produced by the scanning system for magnetically deflected 10-and 12-inch kinescopes. The 1V2 has a filament operating at only 0.625 volt and requiring only about 0.18 watt.

Radiotron type 6AU5-GT — is a high-perveance, beam power amplifier of the single-ended type. Because of its features including low mu-factor, high plate current at low plate, and a high operating ratio of plate current to grid-No. 2 current, the 6AU5-GT makes possible the design of an efficient horizontal-deflection circuit in which the plate voltage for the valve is supplied in part by the circuit and in part by the low-voltage, d.c. power supply of the receiver. A power supply of 250 volts, or less, is all that is required for a receiver utilizing such a deflection circuit and the 6AU5-GT.

Only one 6AU5-GT in a suitable circuit is required to deflect fully a 10BP4, a 12LP4, or any other similar kinescope having a deflection angle up to about 60° and operating at an anode voltage up to 12 kilovolts.

Radiotron type 16GP4 — is a short, directly viewed, 16-inch kinescope of the metal-cone type for use in television receivers designed for it. A rounded-end picture 11″ x 14⅝″ is obtained by utilizing the full-screen diameter. The 16GP4 has a maximum overall length essentially 5 inches shorter than the 16AP4.

The comparatively flat face of the 16GP4 is made of "Filter-glass" to provide increased picture contrast particularly in a lighted room. The special glass in the face plate contains a neutral light-absorbing material which reduces ambient-light reflections from the phosphor and reflections within the face plate itself by a much greater ratio than it reduces the directly viewed light of the picture.

The 16GP4 has a new design of cone-to-neck section which makes possible the design of a longer and more efficient yoke than would otherwise be practical. This design also facilitates centring of the yoke on the neck and, in combination with better centring of the beam inside the neck, contributes to improved uniformity of focus over the entire picture area.

Other outstanding features incorporated in the 16GP4 include an ion-trap gun which requires only a single-field magnet, and a duodecal 5-pin base which permits the use of a lower-cost segment socket.

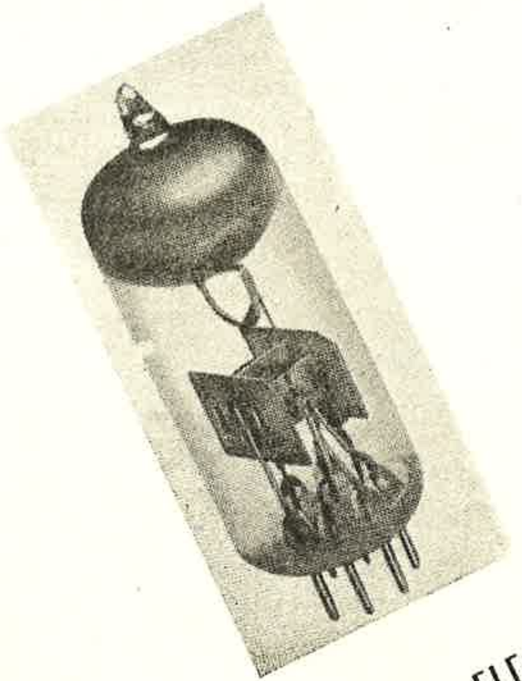


Fig. 1.—Parallel electrode triode (A1714) on pressed glass base.

PLANAR ELECTRODE VALVES FOR V-H-F

During the past ten or fifteen years considerable progress has been made in improving the high-frequency performance of triodes and pentodes by reducing the inductance of the leads to the electrodes. One of the first attempts in this direction was the "acorn" valve, which was designed with a very small electrode system, the leads from which projected as radial pins passing through the all-glass envelope. It is interesting to note that the earliest forms of this type of valve employed planar electrodes[1] similar in some respects to those which will be mentioned later. However, this construction was abandoned in favour of a very small cylindrical electrode system when "acorns" were eventually produced and marketed. The "acorn" type of valve, while enabling a considerable improvement to be obtained in the effective amplification at very high frequencies, has proved to be a difficult manufacturing proposition and has been superseded by valves with conventional electrode systems, mounted on flat glass bases through which pass the lead-out wires, which themselves form the valve pins. Two forms of such designs are represented in present-day commercial products in the button seal pressed-base valves, commonly known as the miniature, and the ring seal moulded-base type. In all these valves the electrode lead-out wires themselves form the connecting pins and the necessity for an external base with separate pins has been obviated.

These glass-based valves represent a big step forward in valve design, and there seems little doubt that the majority of receiving valves in the future will be mounted on this form of base. Quite apart from the advantages of this construction for high-frequency operation, it has led to a reduction in size and freedom from loose base troubles, which, under some conditions, occur with the cemented plastic base. Furthermore, with large-scale production the cost of manufacture of some forms of pressed glass base valves may be less than with earlier designs.

In a wide-band amplifier it is normal for the dynamic resistance of the circuits to be of a comparatively low order and several considerations arise in the design of a suitable valve for high gain combined with low noise in such amplifiers.

The gain of a single stage of a wide-band amplifier is proportional to the ratio of the mutual conductance (g_m) to the sum of the input capacitance (C_i), the output capacitance (C_o) and the stray

Contributed by the Research Staff, M.O. Valve Company. Reprinted from Wireless World, May, 1949, by courtesy of the publishers, Iliffe & Sons Ltd., London.

[1] "Vacuum Tubes of Small Dimensions for Use at Extremely High Frequencies," B. J. Thompson and G. M. Rose. Proc. I.R.E., Vol. 21, p. 1707, 1933.

capacitances (C_s). It is important therefore to make this ratio as high as possible. In addition, for successful high-frequency operation the interelectrode capacitances should be kept small, in order to keep as much as possible of the circuit external to the valve, and the electron transit time should be reduced to a minimum.

Now it can readily be shown that the requirements of high ratio of mutual conductance to capacitance and of low electron transit time require a high ratio of electron current density to grid-cathode spacing. The further requirement of low interelectrode capacitance necessitates a small cathode area. Thus the best performance is likely to be obtained with a valve having a small cathode area, small grid-cathode spacing and operating at a high current density.

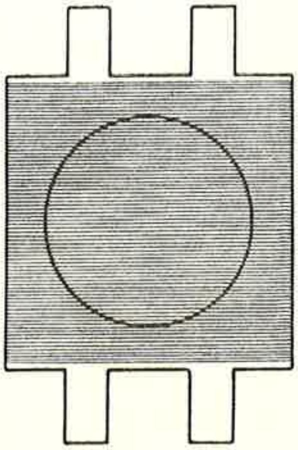


Fig. 2.—Grid assembly of planar-electrode valve.

The ultimate sensitivity of a high-gain amplifier depends on its signal-to-noise performance. If the gain of a receiver is more than about 5 db then most of the noise output is contributed by the first stage. The amount of noise contributed by a valve is usually regarded as being equivalent to that generated in an imaginary resistance, R_s in the grid circuit of the valve. R_s is known as the "equivalent noise resistance" of the valve and is approximately inversely proportional to the mutual conductance. If R_1 is the dynamic resistance of the input circuit, then it can be shown that the signal-to-noise ratio is a function only of the ratio R_1/R_s and will increase as this ratio increases. Now R_1 cannot be increased indefinitely owing to the inherent losses in circuit components so that the only way to improve the signal-to-noise performance is by reducing R_s and this means increasing the mutual conductance of the valve.

For frequencies above a few hundred megacycles per second a greater decrease in lead inductance proves necessary than has been achieved in the conventional concentric cylindrical arrangement of electrodes, and this improvement has been achieved by making the electrodes integral with metal discs which pass through the envelope and which may be directly connected to cavity resonators if desired. Such valves have been described elsewhere.[2]

These valves are known as the disc-seal type and such are capable of operation at frequencies up to about 4,000 Mc/s. The valves employ planar electrodes which allow very small interelectrode spacings to be achieved, permitting a high mutual conductance from a small cathode area and a high ratio $g_m/C_{g\text{-}k}$.

An example is the Marconi disc-seal triode type DET 23 in which the mutual conductance is 7.0 mA/volt at an anode current of 10 mA, and the total input and output capacitances including the discs which pass through the envelope are 2.4 $\mu\mu$F and 1.1 $\mu\mu$F respectively, of which the discs themselves account for about 0.7 $\mu\mu$F in each case. Thus: $C_{g\text{-}k}$ is 1.7 $\mu\mu$F and $C_{a\text{-}g}$ is 0.4 $\mu\mu$F. This high ratio of mutual conductance to input capacitance is better than has hitherto been achieved with concentric electrode arrangements, and is due to the fact that the spacings are small only at the operating surfaces of the electrodes.

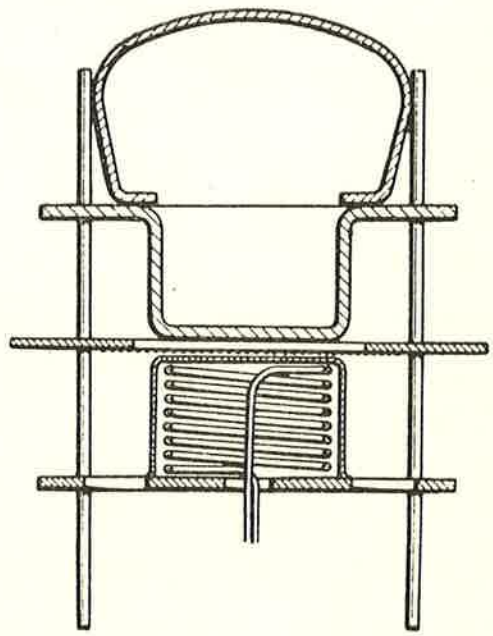


Fig. 3.—Electrode assembly in the type A1714 triode.

These disc-seal valves which were designed primarily for ultra-high frequencies will be seen to satisfy the wide-band amplification requirements set out above. It therefore seemed desirable to employ a similar electrode arrangement in valves designed for more general use in the u-h-f range, such as valves mounted on pressed glass bases with the pins forming the lead-in wires. Valves of this type are easier to use and less costly than the disc-seal valve. A typical triode of this class is the Radiotron type A1714 and is illustrated in Fig. 1.

The very small grid-cathode spacing employed

[2] "Triodes for Very Short Waves," Bell, Gavin, James, Warren, Journal I.E.E., Vol. 93, Part IIIA, p. 833, 1946.

(0.003 in.) necessitates the use of extremely fine and closely spaced wires for the grid, and the design of the grid (Fig. 2) is one of the principal features of valves of this type. In the conventional type of electrode system in which the grid wires are located on two separating rods the wires themselves must be sufficiently strong to carry the separate rods so that the whole structure is rigid enough for handling during the assembly of the valve electrodes without risk of distortion, and this sets a lower limit to the diameter of wire which can be employed. In planar electrode valves a departure from convention has been made, which enables rugged grids to be manufactured with wires as small as 0.0006 in.

The grid is in the form of a metal plate pierced by a circular aperture across which the grid wires are stretched, while the cathode and anode are the end surfaces of two short cylindrical members, supported from or integral with a relatively thick and therefore rigid plate. These plates and the grid frame are located in slotted mica bridges which serve to hold the electrodes in the correct relative positions. Stray capacitances between the electrodes are in this way reduced to a minimum, only the operating surfaces of the electrodes being in close proximity.

The leads connecting the electrodes to the pins in the valve base are also well spaced and contribute little to the total capacitances. The electrode assembly for this type of valve is shown in Fig. 3.

The very small diameters of grid wire possible with this construction allow adequate grid dissipation for amplifiers and for low-power oscillators. Furthermore, the grid frame serves to radiate heat and thus minimizes the risk of primary grid emission.

The characteristics of the Radiotron A1714 are as follows:—

Rating

Heater voltage	6.3 Volts
Heater current	0.55 Amps
Plate voltage	300 max. Volts
Plate Dissipation	2.5 max. Watts
Amplification factor	35*
Mutual conductance	8* mA/V
Plate resistance	4380* Ohms

* Measured at $Va = 150V$ $Ia = 10mA$

Capacitances (Micro-microfarads)

Taken on a valve with external shield.

(Ia = 10 mA)	Cathode cold.	Cathode hot.
Grid — plate	0.9	—
Grid — cathode	1.6	3.0
Grid — all except plate	2.6	4.0
Plate — all except grid	1.1	—

Typical operation

Plate voltage	150 Volts
Plate current	10 mA
Grid Voltage	–2.0 Volts
Input Resistance	40,000 Ohms at 45 Mc/s
Equiv. Noise Resistance	550 Ohms

These characteristics undoubtedly represent the best performance which has been obtained with a triode operating at frequencies of the order of 45 Mc/s, covering a bandwidth of 10/15 Mc/sec.

"SHORT - WAVE RADIO AND THE IONOSPHERE", by T. W. Bennington, of the Engineering Division, British Broadcasting Corporation. 2nd Edition. Published by Iliffe & Sons Limited. Size 8¾" x 5½" (D8vo). 138 pages and 61 illustrations.

All who carry on radio communication over long distances by short waves — whether professionals or amateurs — must be interested in the role of the ionosphere, which is one of paramount importance.

This book presents all the available information in simple form so that it is of use to those with only a limited technical knowledge. The author is a member of the Engineering Division of the British Broadcasting Corporation and has been able to draw freely on the Corporation's experiences in the development of short-wave overseas services.

The use of mathematics has been avoided and the physical processes involved are explained in simple descriptive language. The author has kept the practical side of the subject in mind throughout and shows how existing ionospheric data can be applied to everyday problems of short-wave transmission and reception.

This new edition of the work first published six years ago (under the title "Radio Waves and the Ionosphere") is to all intents and purposes an entirely new book. Mr. Bennington has completely rewritten it; it has been re-set in new type; and 56 illustrations have been added.

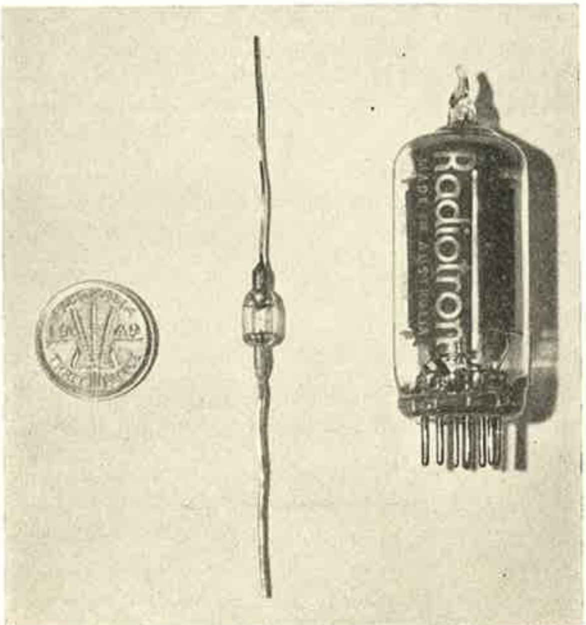
Contents include:—

Preface. Fundamentals of Long-distance Communication. Formation and Structure of the Ionosphere. Radio Waves and the Ionosphere. Measurement of the Ionospheric Characteristics. Ionospheric variations — Short-wave Transmission. Multiple-hop transmission and Ionospheric Forecasting. Amateur Transmission of High Frequencies. Radio Noise, Ionospheric Absorption, and the Low Limiting Frequency. Ionospheric Storms and Other Phenomena. Conclusion. Index.

Our copy received with the compliments of the publishers.

Germanium

Crystal Rectifiers



IN this type of Rectifier, as distinct from Silicon Rectifiers, the outstanding advantage lies in an ability to handle large voltages in the reverse direction, while in the forward direction, the slope (measured at + 1 volt) may be of the order of 100 ohms (10 mA/volt).

The slope in the reverse direction reaches a maximum resistance value at the order of – 2 volts, beyond which the resistance falls slowly until, at a particular voltage known as the 'turnover voltage', the slope resistance falls to zero and then becomes negative.

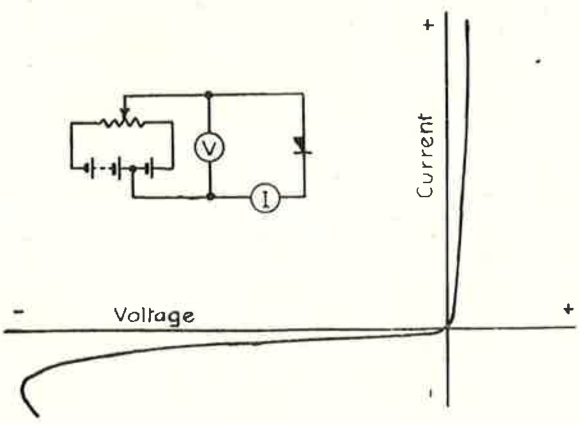


Fig. 1.

Measurements of rectification efficiency at various frequencies up to 100 Mc/s. indicate that the efficiency falls with frequency by an amount which depends upon the 'turnover voltage', the higher the 'turnover voltage' the lower the rectification efficiency. In view of this, crystals intended for use at very high frequencies have a maximum as well as a minimum 'turnover voltage' rating.

A typical current/voltage characteristic is shown in Fig. 1.

Rectification occurs at the junction between a metallic point and the surface of crystalline germanium. During manufacture this junction is treated to obtain optimum impedance characteristic and time stability. After assembly, the metal point is cemented to the germanium to prevent dislodgement by vibration, and the complete assembly is sealed to prevent ingress of moisture.

Externally the rectifier consists of a tough glass envelope with wired ends as shown in Fig. 2.

The rectifiers are very small in size, of low forward impedance, and are designed to be soldered directly into a circuit.

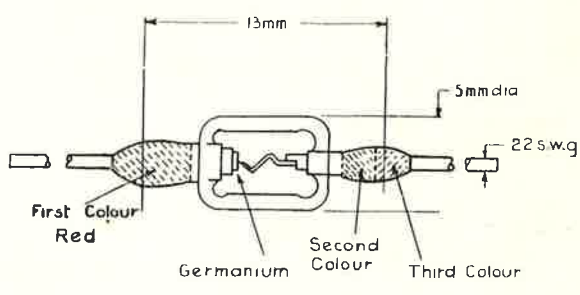


Fig. 2.

A characteristic of these fixed germanium crystal detectors is their remarkable property of withstanding severe mechanical shock or vibration. The construction gives a unit of high mechanical and electrical stability.

The estimated life of the rectifier is indefinite — in excess of 10,000 hours.

Data is given below of some Radiotron types with characteristics suitable for various purposes.

Type Gex. 33

Colour code: Red/Orange/Orange

Suitable for use as a rectifier feeding low impedance circuits, such as second detector in wide band amplifiers or instrument rectifier.

Reverse current at –10 volts:	Less than 1.0 mA
Forward current at +1 volt:	Greater than 3mA
Turnover voltage:	30/60 volts
Shunt capacitance:	1.0 $\mu\mu$F approx.

Type Gex. 44*

Colour code: Red/Yellow/Yellow

Suitable for use as a rectifier feeding medium impedance circuits, such as second detector in narrow band amplifiers, limiter in F-M circuits, or discriminator rectifier.

Reverse current at –10 volts:	Less than 100 μA
Forward current at +1 volt:	Greater than 1 mA
Turnover voltage:	Greater than 60 volts
Shunt capacitance:	1.0 $\mu\mu$F approx.

Type Gex. 55

Colour code: Red/Green/Green

Suitable for use as a rectifier feeding high impedance circuits, such as discriminator in F-M circuits with high output impedance, or rectifier in valve-voltmeter probe.

Reverse current at –10 volts:	Less than 10 μA
Forward current at +1 volt:	Greater than 1.0 mA
Turnover voltage:	Greater than 60 volts
Shunt capacitance:	1.0 $\mu\mu$F approx.

Type Gex. 99

Colour code: Red/White/White

Suitable for use as general non-linear circuit element such as current limiter, spark quench, channel switch, or modulator.

Reverse current at –10 volts:	Less than 1 mA
Forward current at +1 volt:	Greater than 1 mA
Turnover voltage:	Greater than 30 volts
Shunt capacitance:	Less than 3.0 $\mu\mu$F

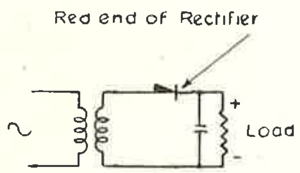


Fig. 3.—Connections for germanium rectifier.

In the colour coding system, RED is an indication of the negative end of the rectifier. Thus, the direction of easy flow of current is given when the RED end is joined to the *positive* end of the rectifier load.

* For the benefit of readers who may wish to use these Crystals in circuits obtained from other sources we would point out that type GEX44 is a close equivalent to the 1N34 and will operate successfully, when used within its ratings, in circuits designed around the 1N34.

Connections for Germanium Rectifier

Tests show that all types will withstand a steady forward current of **30 mA** and a recurrent peak of **100 mA** with safety. It is not expected that average variation of characteristics during life will exceed 10%.

There is little change in rectifier characteristics from 15° C. to 50° C., but above this temperature up to 100° C., both forward and back slope resistances decrease slowly.

It is intended to keep in stock the following types: GEX.33, GEX.44 and GEX.55.

An initial air-freight shipment of these types has now been received and orders can be supplied from stock.

Discontinued RCA Types

Type 3DP1-A — 3 inch cathode-ray tube with a radial deflecting electrode. Originally made during the war and now considered obsolete.

Type 3DP1-S2A — Similar to type 3DP1-A. Obsoleted by R.C.A.

Type 10 — Discontinued by R.C.A. However, power type 10-Y is still retained.

Type 559 — U-H-F diode. Now considered obsolete by R.C.A.

Type 850 — Screen-grid R-F power amplifier. Now considered obsolete by R.C.A.

IMPORTANT

Included with this issue is a pamphlet announcing details of the new subscription rate to Radiotronics.

Coincident with the increase in subscription rate Radiotronics will revert to monthly publication.

We would ask all our present subscribers to note particularly that it is essential that subscriptions for 1951 should be mailed immediately if they wish to avoid missing any of the early issues as each print will be strictly limited.

Electronic Keying Systems

By MACK SEYBOLD

RCA Tube Department

Amateurs have been using valves in keying circuits for the past twenty years. As a matter of fact, it was almost that far back when I operated c.w. on 80-metres with a pair of UX245's in parallel, to key the cathode circuit of an old 210. My recollection of that rig, even though it was a hay-wire bread-board dust-catcher, is very pleasant, and I never did get any key-click complaints from the nearby broadcast listeners or local hams.

Just to get an idea of where that keying circuit had come from, I recently thumbed back through the years of radio journals in my shack. To my amazement, I found dozens of keying circuits for the betterment of ham's estate.

Keying filters

Among the more complete articles on the subject, there was one by the late Ross Hull. This carefully prepared resume in February, 1929, *QST*, covered all of the major systems in use up to that time. Even to-day, I'll bet that more than half of the boys on the air are using keying systems that were described in Hull's article. The main components used are chokes and capacitors. In addition to eliminating the sparks at the key contacts, proper utilization of the components in the keyed stage or in the feed lines to that stage minimizes the production of transients in the radiated signal when the key contacts make and break. A typical filter for cathode keying is shown in Fig. 1.

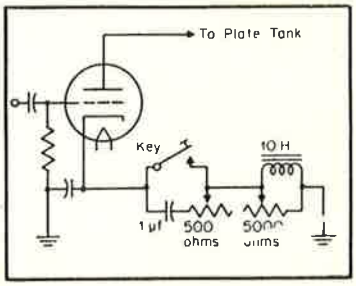


Figure 1

Sometimes the old choke and capacitor methods really do accomplish the desired results. But the tinkering required to clean up a thump or click condition with those systems, and their unreliability under varying conditions, are probably the fundamental causes for all of the additional work that has been done on keying circuits in subsequent years.

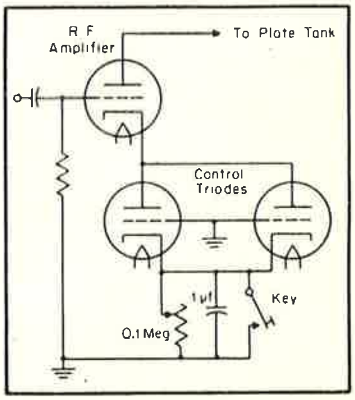


Figure 2A

Early electronic methods

The first complete references that I could find in which valves were used to make and break the circuit in the cathode lead of an amplifier were in *QST* for August, 1931. These references were, evidently, the source information for my earliest venture into electronic key-click control. One circuit is shown in Fig. 2A, and the *QST* article attributes the original idea to F. B. Kennell of RCA Communications. The immediate proponent of the system, W. H. Hannah, W2US, had remarkable success with the device in his amateur rig.

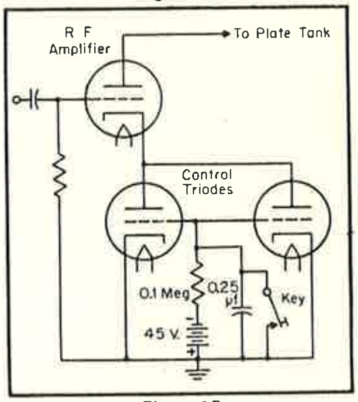


Figure 2B

The same issue showed the circuit (Fig. 2B) of C. W. Carter, W3AGT, for another electronic key-click control which was the basis for many series-controlled systems which were to follow.

Screen-grid transmitting valves

During the first few years after power pentodes and tetrodes became available for amateur use, the same general methods that previously had been used for keying triodes were employed. Some did utilize the screen grid as a keying control element by putting a key or a relay in series with the screen lead, but this method still required the use of key-click filters.

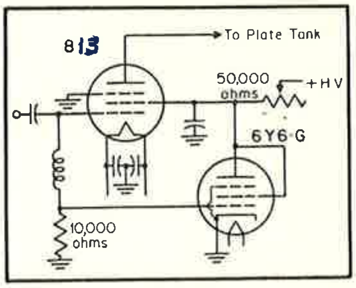


Figure 3

Reprinted from RCA Ham Tips (Jan.-March. 1950) by courtesy of Radio Corporation of America.

The beginnings of valve control of pentode-screen keying seem to have occurred in 1941. In the December *Radio* for that year, W. W. Smith, W6BCX, described "A Substitute for Safety Bias" which utilized a triode shunting the screen of an output valve. Later, F. T. Smith, W1FTX, built a transmitter (Feb. 1947 *QST*) in which a 6Y6-G, triode connected, controlled the screen of an 813. In these screen-shunting systems, an earlier stage is keyed, and the bias developed across the grid-leak of the final amplifier is the voltage which triggers the control valve. A diagram of the W1FTX final is shown in Fig. 3.

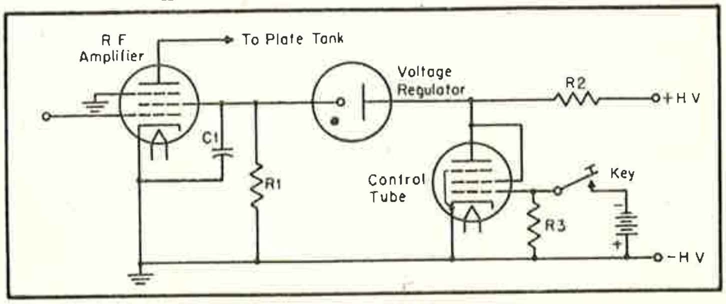


Fig. 4.

VR-valve keying

My solution to the problem of key-click control has been arrived at from a somewhat different approach, in that voltage-regulator valves are used. The characteristics of the VR valves are suited remarkably well for this application and the maximum ratings established for them are not exceeded in this new circuit.

The gaseous atmosphere within a VR valve limits the valve to two major operating states — conducting, and non-conducting. In the conducting state, the current-carrying medium is ionized gas, and the voltage drop between anode and cathode is constant throughout a range of current flow from 5 to 40 milliamperes. In the non-conducting state, when the voltage applied between the anode and cathode falls below the ionizing potential of the gas, the valve virtually is an open circuit. For the 0A3, the voltage required for valve operation is 75 volts. For the 0C3 and the 0D3 it is approximately 105 and 150 volts, respectively. For the miniature types 0A2 and 0B2 it is approximately 150 and 108 volts, respectively.

When one of these valves, say the 0C3, is placed in the screen supply lead of a pentode or tetrode, the valve will do one of two things: it will conduct or it won't. That, of course, is just what a mechanical key will do. In opposition to the key, however, there are no mechanical components to arc and spark when the circuit opens and closes. In addition, the ionizing and de-ionizing time of the gas within the VR valve causes an infinitesimal time delay which smooths the leading and lagging edges of a keyed character.

Operation

The VR valve in a keying circuit is an effective non-mechanical keying gap. Since the valve needs no filament supply, it can be placed at any convenient point in the screen supply line. The VR valve is controlled by an auxiliary vacuum valve which determines the "on" or "off" conditions. This control valve is activated by applying the correct potential to its grid No. 1 and, since the grid No. 1 is operated at a negative potential, very little current flows in the actual key circuit. Figure 4 shows the basic components of the system.

When the key is down, the bias voltage applied to the control valve prevents it from conducting and the VR valve and the r-f amplifier conduct current through R_2. The supply voltage, minus the drops across R_2 and the VR valve, is the effective screen-grid potential applied to the r-f amplifier and permits normal key-down operation.

When the key is up, the bias on the control valve drops to zero, making it conduct and, as a result, the voltage drop across it becomes lower than the required ionizing potential for the VR valve. The VR valve, therefore, stops conducting, the supply voltage is completely removed from the screen-grid of the r-f amplifier, and the transmitter goes off the air.

This procedure can be repeated as often as desired, as in c.w. work; the rapidity of keying is limited only by the ionizing and de-ionizing char-

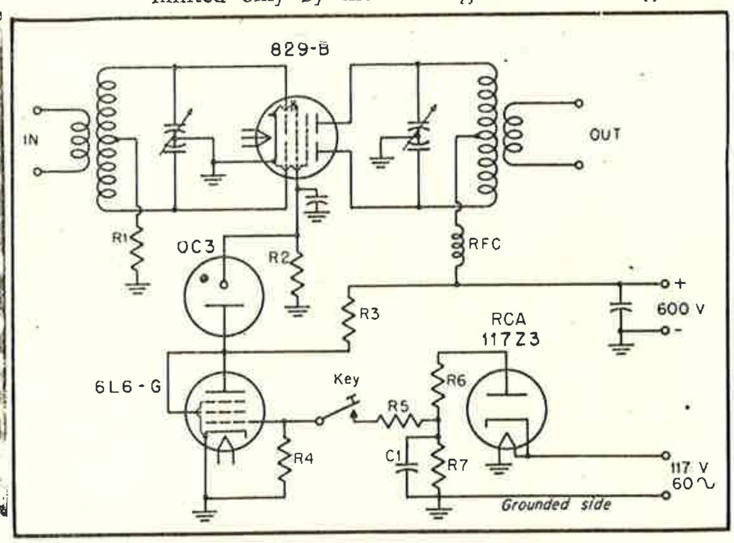


Fig. 5.

acteristics of the VR valve. In the circuit of Fig. 4, the resistors R_1 and R_3 are high in value, and are present merely to maintain each grid at a potential near zero when keying potentials are removed. C_1 is a conventional r-f by-pass capacitor.

Transmitter circuits

A practical application of the system is shown in Fig. 5. Here an 829-B final is keyed with an 0C3

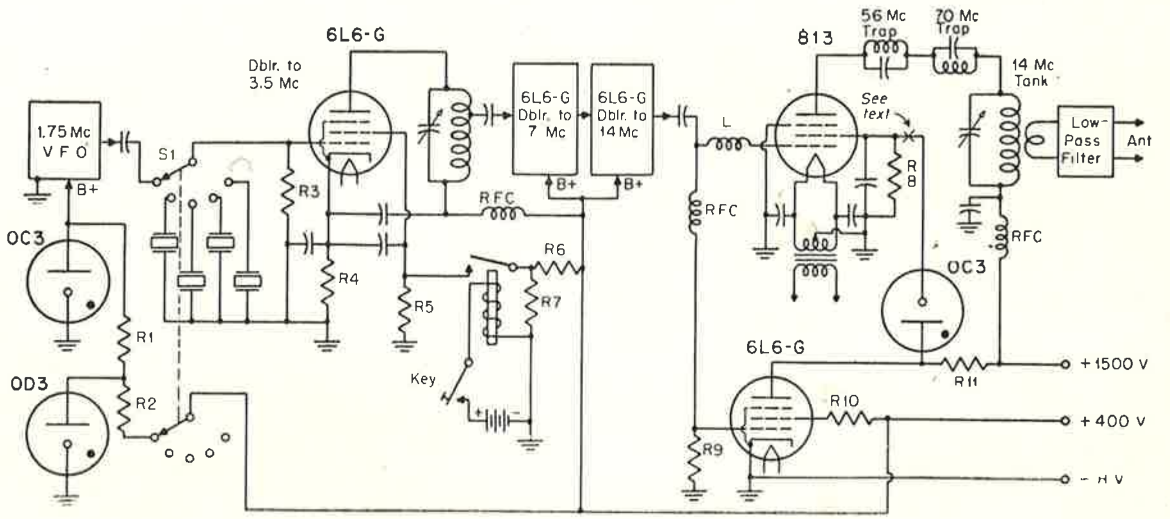


Fig. 6.

controlled by a 6L6-G. Cutoff bias for the 6L6-G is obtained from a 117Z3 as shown, or it can be taken from any type of supply capable of furnishing 125 volts of bias.

Another application of the VR valve keying circuit is shown in Fig. 6. This arrangement makes it possible to key a buffer stage or the oscillator so that break-in keying may be utilized. Bias for the control valve is obtained automatically at the correct time from the grid resistor of the final amplifier. The protection the control valve gives to the final amplifier, if excitation fails or when doubler and buffer stages are being adjusted, makes this keying system a valuable adjunct to a beam power or pentode final. (On higher-powered finals, a huskier beam power valve must be used as a control valve.)

One additional component has been added to the circuit by Bill Scherer, W2AEF. This addition is a 5-henry choke placed in series with the screen lead at the point marked "X". The choke was added to give a more rounded leading edge to the keyed character. Further details on this addition may be found in Scherer's article, "The Gold-Plated Special," in *CQ*, October, 1948.

For those who wish to design an amplifier operating under conditions other than those shown here, a detailed description of the system is given in "VR Tube Keying Circuits," which appeared in the May, 1948 issue of *CQ*. Another version of the system is described in "A TVI-Free Transmitter for 10 Meters" which was published in *CQ* for October and November, 1949.

If clickless keying is desired, the VR valve keying system is a straight-forward answer to a problem that has been confronting hams since the days of the spark transmitter.

PARTS LIST

Fig. 5
- C_1 = 3.0 μF, 150 working volts
- R_1 = 5800 ohms, 2 watts
- R_2, R_4 = 0.25 megohm, 0.5 watt
- R_3 = 10,000 ohms, 50 watts
- R_5 = 50,000 ohms, 0.25 watt
- R_6 = 100 ohms, 0.5 watt
- R_7 = 0.1 megohm, 0.5 watt

Fig. 6
- L = 0.5 henry, grid choke
- R_1 = 1250 ohms, 5 watts
- R_2 = 4400 ohms, 20 watts
- R_3 = 0.1 megohm, 0.5 watt
- R_4 = 1000 ohms, 10 watts
- R_5, R_7 = 0.25 megohm, 0.5 watt
- R_6 = 20,000 ohms, 5 watts
- R_9 : ~~R_8~~ = 7500 ohms, 5 watts
- R_8 : ~~R_9~~ = 0.5 megohm, 0.5 watt
- R_{11} : ~~R_{10}~~ = 35,000 ohms, 100 watts
- R_{10} : ~~R_{11}~~ = 50,000 ohms, 5 watts

R-F components and bypass capacitors are conventional.

VALVE DATA BOOK

The Radiotron loose-leaf valve data book previously available at 2/6 per copy, with additional sheets being distributed through Radiotronics has now been discontinued.

In its place a new valve data book is being prepared and it is expected to be on sale early in the new year.

Watch for a future announcement in Radiotronics.

Corrns. from P132.

Radiotron 6CB6 in Television Receivers

This Note describes the high-frequency characteristics of miniature pentode Radiotron 6CB6 and its use in television r-f tuners and video intermediate-frequency amplifiers. Input admittance data for the 6CB6 and design considerations for a video i-f amplifier system operating in the 40-megacycle region are given in detail.

Features of the 6CB6

The 6CB6 is, essentially, an improved 6AG5 with grid-plate transconductance increased 20 per cent., grid-plate capacitance decreased 30 per cent., and with the cathode and grid No. 3 connected to separate base pins. The high transconductance and reduced grid-plate capacitance of the 6CB6 make it possible to obtain high gain at high frequencies, while the separate grid-No. 3 connection makes possible the use of an unbypassed cathode resistor to reduce variations in input capacitance and input conductance with changes in bias. When this valve is used as an r-f amplifier in a television tuner, for example, it offers advantages over other high-gain low-cost valves because its lower grid-plate capacitance reduces oscillator radiation and its separate cathode and grid-No. 3 connections make possible the use of an unbypassed cathode resistor to minimize the detuning effects encountered in sets employing automatic gain control.

Although there are several other r-f pentodes having lower grid-plate capacitances than the 6CB6, the reduction in grid-plate capacitance of these types is accomplished by shielding the plate and is, therefore, accompanied by a large increase in output capacitance. Because the only capacitance in the tuned circuits of most television r-f and i-f amplifiers is that of the valve electrodes and associated wiring, a large increase in output capacitance causes a decrease in plate-circuit impedance and a consequent loss in gain. The maximum grid-plate capacitance of the 6CB6 is 0.020 μμF; its output capacitance is only 1.9 μμF.

Input admittance considerations

Because valve loading is one of the major factors limiting the gain of a television r-f stage, the input admittance data for the 6CB6 are of considerable design importance. Table I gives the short-circuit input admittance data for this valve at 100 megacycles.

Reprinted from RCA Application Note AN-143 by courtesy of Radio Corporation of America.

TABLE I

Short-Circuit Input Admittance Data at 100 Megacycles for Type 6CB6

Operating Conditions:		
Plate Voltage	200	volts
Screen Voltage	150	volts
Control-Grid Voltage	–2.5	volts
Plate Current	9.2	mA
Screen Current	2.28	mA
Transconductance	6200	μmhos
Short-Circuit Input Capacitance:*		
Tube Operating	10.14	μμF
Tube Cutoff	8.6	μμF
Tube Cold	8.15	μμF
Capacitance Increase (cold to cutoff)	0.45	μμF
Capacitance Increase (cutoff to operating)	1.54	μμF
Short-Circuit Input Conductance:*		
Tube Operating	460	μmhos
Tube Cutoff	87	μmhos
Tube Cold	81	μmhos
Conductance Increase (cutoff to operating)	373	μmhos
Grid-to-Plate Capacitance		
(measured at low frequency with valve cold)	0.0156	μμF

* Data for valve and socket, as measured on admittance meter; socket capacitance, 1 μμF; socket conductance, 2.55 μmhos

The short-circuit input conductance at a transconductance of 6200 micromhos is given in Fig. 1 for a frequency range of 20 to 100 megacycles. Fig. 2 gives the input conductance at 100 megacycles for a transconductance range up to approximately 7000 micromhos. The approximate input conductance at any frequency and any value of transconductance may be determined from Fig. 2 since, as shown in Fig. 1, input conductance is proportional to the square of the frequency. The change in input capacitance, which is approximately the same throughout the present range of television frequencies, may be determined from Fig. 3 for different values of transconductance.

When the grid bias of any r-f or i-f amplifier is varied in order to vary the gain, both the input capacitance and the input conductance of the valves vary also, and the shape of the pass band is changed. In a television receiver employing automatic gain control (agc), the r-f response as well as the i-f response will vary. The increase in input capacitance of the 6CB6 at 50 megacycles with the cathode at r-f ground potential is approximately 1.5 $\mu\mu$F between the value of 8.6 $\mu\mu$F at cutoff and the value of 10.1 $\mu\mu$F at a transconductance of 6200 micromhos. The corresponding increase in input conductance is about 100 micromhos (from 87 micromhos at cutoff). In order to compensate for these changes in input capacitance and input conductance, an unbypassed cathode resistor can be used with the 6CB6 because of its separate grid-No. 3 connection. Figs. 3 and 4 were taken to determine the optimum value of cathode resistor and show that a 47-ohm cathode resistor is very nearly optimum. Although the data for Figs. 3 and 4 were taken at 50 megacycles, the optimum cathode resistor will be about the same for all

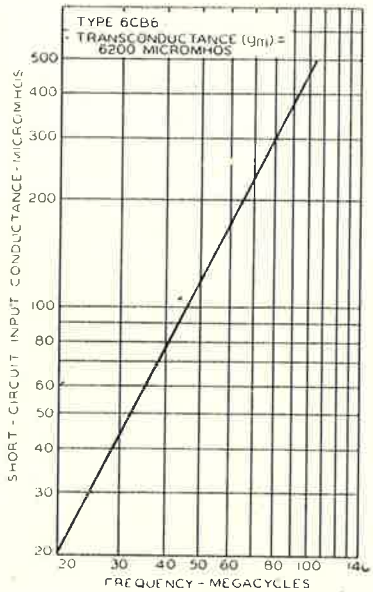


Fig. 1.—Change of short-circuit input conductance with frequency.

currently used television r-f and i-f frequencies. Because a 47-ohm cathode resistor is too small in value to provide proper bias for the valve, it is necessary to supplement this bias either with fixed bias or with additional cathode bias supplied by a 130-ohm bypassed resistor.

Intermediate-frequency amplifier design

The main requirements of a suitable television i-f amplifier valve are high transconductance for high gain and low grid-plate capacitance for low feedback. The combination of reduced grid-plate capacitance and high transconductance of the 6CB6 makes it possible to obtain a higher gain with this valve than with other similar types. This valve, therefore, is suitable not only for the 20-megacycle i-f band, but also for the RMA-recommended 40-megacycle band. The remainder of this Note considers some of the factors important to the design, construction, and operation of a stagger-tuned video i-f amplifier system operating with a picture i-f carrier frequency of 45.75 megacycles and a sound i-f carrier frequency of 41.25 megacycles.

Damping resistor considerations

In the design of a stagger-tuned i-f amplifier, the values of the damping resistors required to obtain the desired pass band are affected by the input conductance of the valves used. At high frequencies, the valve input-conductance components due to transit-time effects, valve lead inductance, and feedback through the grid-plate capacitance from the plate circuit are effectively in parallel with the tuned grid circuit. The conductance components due to transit-time effects and valve lead inductances are positive in sign and vary with the square of the

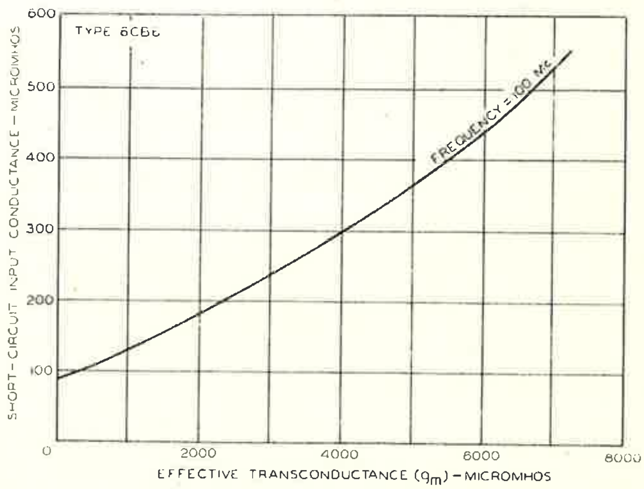


Fig. 2.—Change of short-circuit input conductance with transconductance.

frequency. The input-conductance component due to feedback through the grid-plate capacitance, measured at the grid-circuit resonant frequency, may be either positive or negative. It is positive for a valve with the plate circuit tuned to a frequency lower than that of the grid circuit and negative for a valve with the plate circuit tuned to a frequency higher than that of the grid circuit. If the grid-plate capacitance is high, the input conductance will vary rapidly over the band and the grid and plate circuits will not be independently tunable. Because bandwidth depends to a considerable extent on input conductance, the short-circuit input-conductance curves given in Figs. 1 and 2 can be used together with the formulas given in the Appendix to obtain an approximate value for the damping resistor. These curves, however, do not show the input-conductance component due to feedback through the

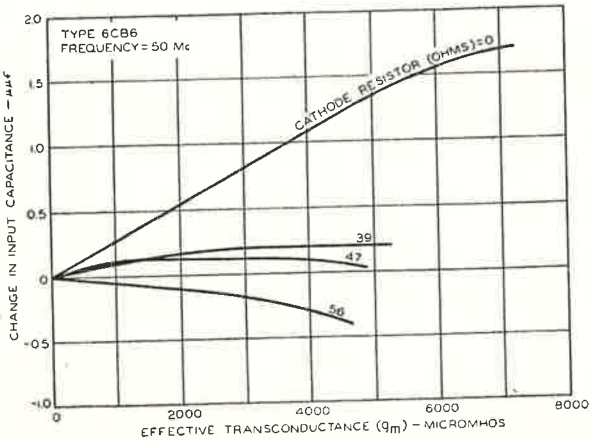


Fig. 3.—Change of short-circuit input capacitance with transconductance for various values of unbypassed cathode resistors.

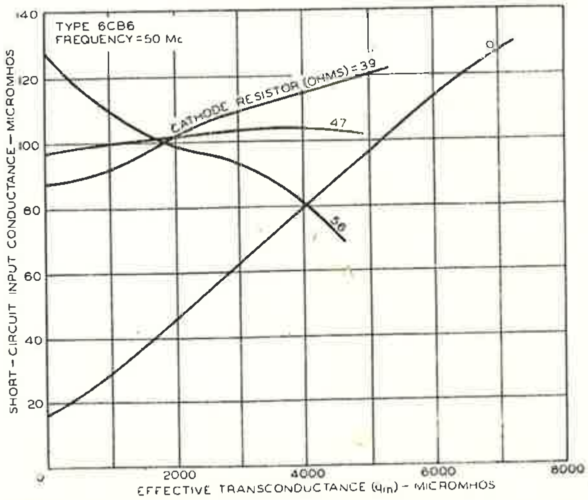


Fig. 4.—Change of short-circuit input conductance with transconductance for various values of unbypassed cathode resistors.

grid-plate capacitance and, consequently, exact determination of the value of the damping resistors is a trial-and-error process with the calculated values serving as a guide.

Circuit considerations

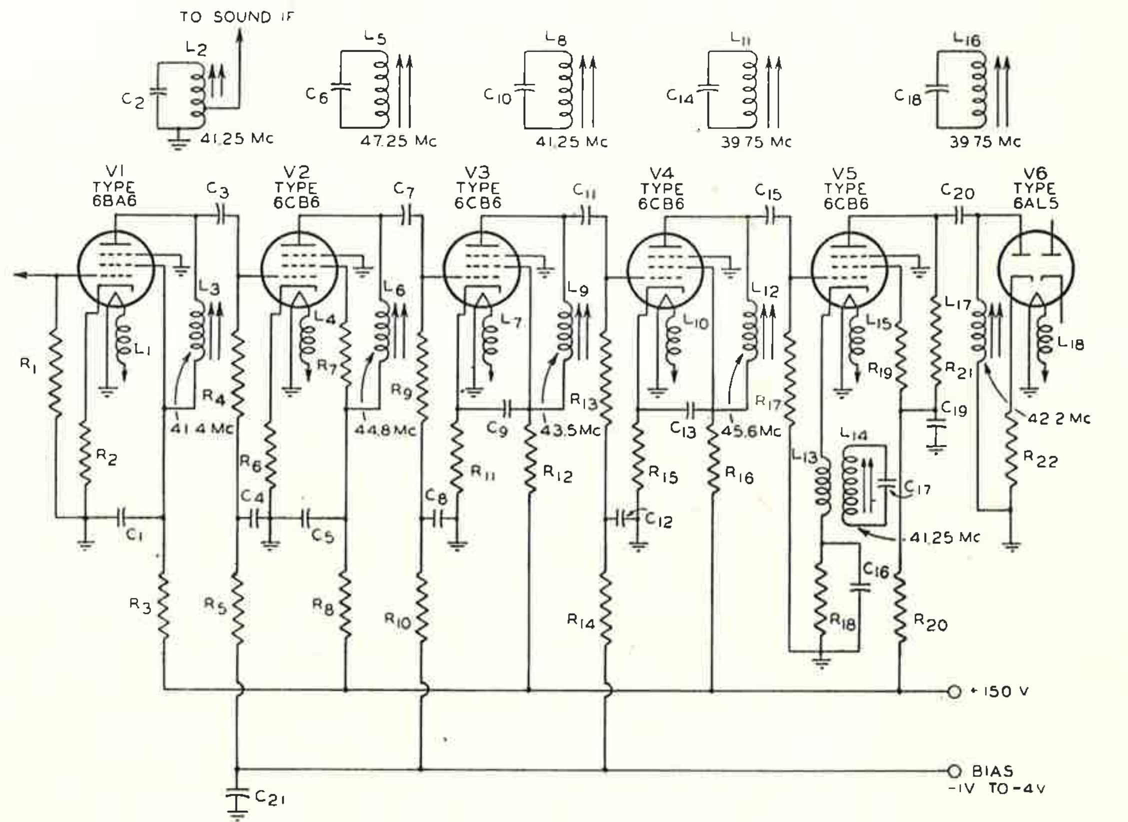
A circuit diagram of a 40-megacycle video i-f amplifier using 6CB6's is shown in Fig. 5. The diagram includes the plate circuit of a converter stage and a 6AL5 video detector. Each stage is tuned by adjusting its inductance for resonance with the valve and circuit capacitance. Because valve capacitance will vary slightly from valve to valve, when valves are changed retuning is necessary to obtain the same band-pass characteristics.

If the screen grid of an r-f amplifier valve is not at r-f ground, the effective grid-plate capacitance of the valve will be much higher than the value measured at low frequencies and regeneration may be encountered. At frequencies of 30 megacycles and higher, however, it becomes quite difficult to ground the screen grid effectively because of the inductance of the screen-grid and bypass-capacitor leads. In many cases, therefore, it may be necessary to adjust the lead inductances so that they are in series resonance with the bypass capacitor in order to ground the screen grid effectively. In the amplifier circuit shown in Fig. 5, effective grounding could have been accomplished by selecting a suitable value of screen bypass capacitor and, depending upon its physical construction, by adjusting its lead lengths for series resonance. This method of preventing regeneration, however, was not needed in this amplifier because a 10-ohm unbypassed resistor is used in series with the screen grid in the high-impedance stages (V2 and V5) and the screen grid is bypassed to the cathode in the low-impedance stages (V1, V3, and V4).

In Fig. 5, two values are given in the parts list for the damping resistor in the grid circuits of V2 and V5. The first value is the d.c. resistance as given by the manufacturer's color code. The second value is the high-frequency resistance as measured on a *Q* meter at 42 megacycles. Allen-Bradley half-watt resistors were used, but it is not known whether the relation found between the d.c. resistance and the high-frequency resistance of Allen-Bradley resistors would be consistent for all resistors of the same value. The high-frequency resistance of resistors made by other manufacturers will, of course, also show variations from the d.c. resistance value. The effect of the damping resistors, therefore, requires close individual checking when a stagger-tuned amplifier for frequencies in order of 40 megacycles and higher is constructed.

Overall video gain

The overall video gain of the amplifier circuit given in Fig. 5 is measured in the following manner. A 43.5-megacycle unmodulated signal is applied to the control grid of the first i-f stage, an r.m.s.-reading vacuum-tube voltmeter is connected to the plate of the video detector (across *L17*), and a d.c. voltmeter is connected across the video detector load (*R22*). The a.c. voltage at the detector plate and the d.c. voltage across the detector load are measured. Then, the meter across *L17* is removed and the output of the signal generator reduced until the original d.c. voltage across the detector load is obtained. The overall gain is calculated by dividing the voltage measured across *L17* by the output voltage of the signal generator. The overall gain at a fixed bias of one volt was 14280: an average stage gain of almost 11. Approximately the same gain can be obtained with the 6CB6 in amplifiers operating at lower intermediate frequencies including the 21.25-to-25.75-megacycle band.



C1: 5000 μμf, ceramic
C2: 22 μμf, ceramic
C3: 150 μμf, mica
C4: 5000 μμf, ceramic
C5: 470 μμf, mica
C6: 75 μμf, ceramic
C7: 150 μμf, mica
C8: 5000 μμf, ceramic
C9: 5000 μμf, ceramic
C10: 22 μμf, ceramic
C11: 150 μμf, mica
C12: 5000 μμf, ceramic
C13: 5000 μμf, ceramic
C14: 25 μμf, ceramic
C15: 150 μμf, mica
C16: 47 μμf, mica
C17: 100 μμf, ceramic
C18: 25 μμf, ceramic
C19: 5000 μμf, ceramic
C20: 150 μμf, mica
C21: 0.1 μf, paper
L1, L4, L7, L10, L15, L18: Heater chokes, 15 turns #22 enam. on 1/4" form
L2, L8, L11, L16: 7 turns #14 enam., 1/2" dia. 5/8" long concentric with and 5/8" from cold end of L3, L9, L12, L17, respectively
L3, L9, L17: 11 turns #30 enam. on 1/4" form
L6, L12: 10 turns #30 enam. on 1/4" form

All capacitors 500 volts.

L13: 2 turns #22 enam. on 1/4" form
L14: 4 turns #22 enam. on 1/4" form 3/8" long 1/4" from cold end of L13
L5: 3 turns #14 enam., 1/2" dia. 1/2" long concentric with and 5/8" from cold end of L6
R1: 1000 ohms
R2: 68 ohms
R3: 470 ohms
R4: 15000 ohms (12000 ohms at 42 Megacycles)
R5: 1000 ohms
R6: 47 ohms
R7: 10 ohms
R8: 470 ohms
R9: 3900 ohms
R10: 1000 ohms
R11: 47 ohms
R12: 470 ohms
R13: 2400 ohms
R14: 1000 ohms
R15: 47 ohms
R16: 470 ohms
R17: 82000 ohms (32000 ohms at 42 Megacycles)
R18: 180 ohms
R19: 10 ohms
R20: 470 ohms
R21: 6800 ohms
R22: 22000 ohms

All resistors 0.5 watt.

Fig. 5. Schematic diagram of 40-megacycle 4-stage stagger-tuned video i-f amplifier system.

Amplitude characteristics

When an unmodulated signal is impressed on the grid of the first i-f stage, a d.c. voltage is developed across the video detector load. In Fig. 6, the d.c. voltage obtained with different values of fixed grid bias is plotted against input signal strength to show the amplitude characteristics. These characteristics are very nearly straight lines. A straight-line amplitude characteristic is desirable in order to obtain undistorted video signals from the i-f amplifier.

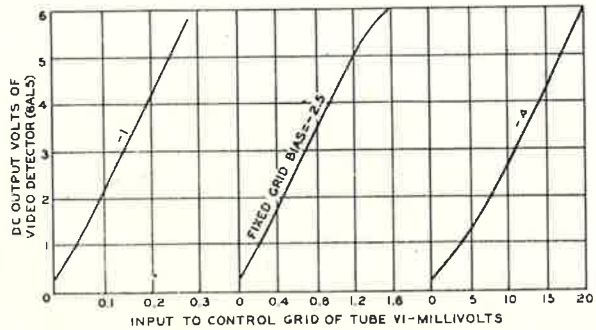


Fig. 6.—Amplitude characteristic of video i-f amplifier of Fig. 5.

Rejection considerations

It is much more difficult to obtain sufficient rejection of the associated sound and adjacent-channel frequencies in a 40-megacycle system than in a 20-megacycle system. Because the ratio of mean frequency to bandwidth is almost twice as great for the 40-megacycle system as for the 20-megacycle system, the absorption traps for the higher-frequency system must have twice the Q required for the lower-frequency in order to have the same degree of rejection. In practice, it is almost impossible to obtain such high Q and still maintain a convenient physical coil size. The unloaded Q of the traps used with the amplifier described in this note was approximately 300. Even with such high-Q traps, it was found necessary to use two traps at 41.25 and 39.75 megacycles and to accept less than the desired rejection. For the least effect on the desired portion of the response curve and the smallest "after-response" due to the rejection traps, each trap, in addition to having high Q, should be coupled to a circuit which is relatively close to it in frequency.

Band-pass data

The band-pass characteristics of the amplifier are obtained by applying the i-f signal to the grid of the 6BA6 mixer stage and varying the input-signal voltage to obtain a constant d.c. voltage across the video detector load throughout the pass band. Band-pass characteristics of the system are shown in Figs. 7 and 8. Fig. 7 is the response before rejection traps were added. These curves show a slight change in response with increasing grid bias, but the change is not enough to have any discernible effect on picture quality. Fig. 8 shows the response after the addition of rejection traps. Obviously, the rejection at the frequency of the adjacent picture carrier and associated sound carrier is poor and the "after-responses" are quite pronounced. If sufficient space is available, additional rejection and smaller "after-responses" may be obtained by increasing the wire size and coil diameter of the rejection traps, thereby increasing their Q. Further improvement could be obtained by coupling to the converter stage or to the video detector stage with a double-tuned circuit and increasing the selectivity at the edge of the pass band. If a reduction in bandwidth is not too objectionable the overall bandwidth may be reduced, making it easier to trap out the undesired frequencies. Finally, if cost is not a major factor, the tuned circuits may be made of band-pass circuits such as overcoupled double-tuned circuits or m-derived networks.

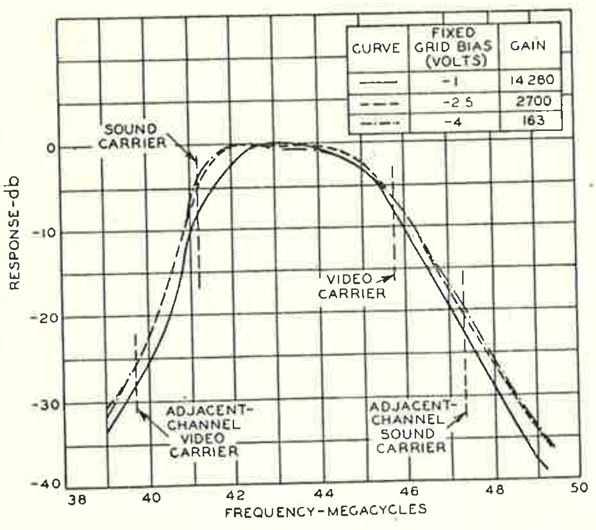


Fig. 7.—Frequency response of video i-f amplifier before addition of rejection traps.

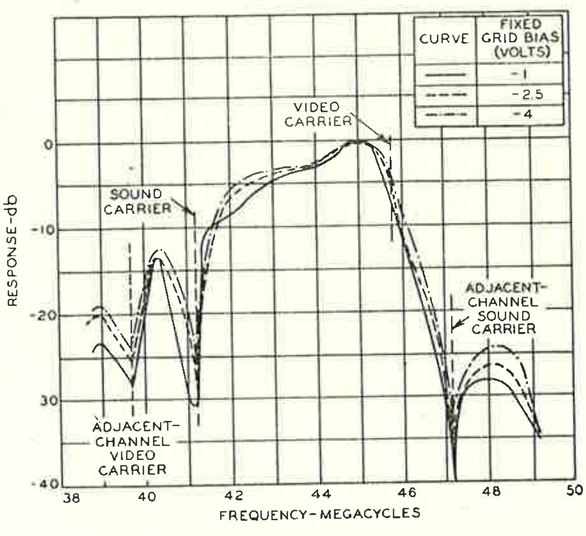


Fig. 8.—Frequency response of video i-f amplifier after addition of rejection traps.

APPENDIX

The following formulas* apply to the calculation of the resonant frequencies and the values of 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.

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

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

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 furnishes the required resonant frequency and dissipation factor for each stage.

It will be noted that two sets of formulas are given for a four-stage amplifier, one set for an amplifier consisting of two pairs of identically tuned stages (staggered pairs), and one set for an amplifier consisting of four individually tuned stages (staggered quadruple). Design problems for an amplifier using staggered pairs are simplified because only two frequencies are required and because the dissipation factors for all stages are the same. An amplifier employing a staggered quadruple, however, will have slightly higher gain and a flatter response curve.

Formulas for Four-Stage Stagger-Tuned System (Staggered Pairs)

Stage	Resonant Frequency	Dissipation Factor
A	$f_o + 0.35\ \Delta f$	$0.71\ \delta$
B	$f_o - 0.35\ \Delta f$	$0.71\ \delta$
C	$f_o + 0.35\ \Delta f$	$0.71\ \delta$
D	$f_o - 0.35\ \Delta f$	$0.71\ \delta$

Formulas for Four-Stage Stagger-Tuned System (Staggered Quadruple)

Stage	Resonant Frequency	Dissipation Factor
A	$f_o + 0.46\ \Delta f$	$0.5\ \delta$
B	$f_o - 0.46\ \Delta f$	$0.5\ \delta$
C	$f_o + 0.19\ \Delta f$	$0.92\ \delta$
D	$f_o - 0.19\ \Delta f$	$0.92\ \delta$

Formulas for Five-Stage Stagger-Tuned System (Staggered Quintuple)

Stage	Resonant Frequency	Dissipation Factor
A	$f_o + 0.48\ \Delta f$	$0.31\ \delta$
B	$f_o - 0.48\ \Delta f$	$0.31\ \delta$
C	$f_o + 0.29\ \Delta f$	$0.81\ \delta$
D	$f_o - 0.29\ \Delta f$	$0.81\ \delta$
E	f_o	δ

* G. E. Valley and H. Wallman, VACUUM TUBE AMPLIFIERS (1st ed.; New York: McGraw-Hill Book Co. Inc., 1948), p. 186.

The 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 and includes the equivalent shunt resistance of the tuned circuit (measured or estimated) and the input conductance of the valve (from Fig. 1).
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 valve 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 overall gain (M) of the stagger-tuned system is $M = m^n$
where n = number of stagger-tuned stages.

COMMUNICATION RECEIVER DESIGN

(Continued from page 105)

unit away from hot valves, etc., will reduce the effects of temperature drift to a minimum. Compensation for drift, by means of special fixed condensers, may be desirable in the oscillator circuit.

The tuning unit, by virtue of the fact that it is closely connected with the tuning dials, will normally occupy the central front portion of the receiver chassis. The remaining stages may then be positioned to either side and to the rear of this unit. Of these the most critical will be the i-f stages which should closely follow one another from the mixer anode in order to avoid feed-back and other troubles. The aerial and r-f circuits should be well isolated or screened from the i-f circuits (particularly from the first i-f stage), in order to prevent direct i-f pickup, etc.

To facilitate relay switching for communication purposes it is an advantage to bring out the connections to the B switch to a suitable termination at the rear of the receiver. The provision of fuses in the mains and/or B circuits will often save service expenditure at a later date!

If the power supply is adequate, another desirable feature is to bring out B and filament connections to a socket at the rear of the receiver to enable a v.h.f. converter, etc., to be used in conjunction with the receiver.

Low Noise

RADIOTRON 5879

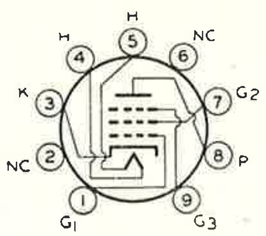
Sharp-Cutoff Pentode

Type 5879 is a new, sharp-cutoff pentode of the 9-pin miniature type intended for use as an audio amplifier in applications requiring reduced audio noise.

Utilizing a single-ended structure which is relatively short and rigidly mounted to minimize microphonics, a controlled getter deposit to minimize internal leakage, and a double-helical-coil heater to minimize hum—all features contributing to reduced audio noise—the 5879 is especially recommended for service in the input stages of medium-gain public-address systems, home sound recorders, and general-purpose audio amplifiers.

SOCKET CONNECTIONS
Bottom View

Pin 1: Grid No. 1
Pin 2: No connection
Pin 3: Cathode
Pin 4: Heater
Pin 5: Heater
Pin 6: No connection
Pin 7: Grid No. 2
Pin 8: Plate
Pin 9: Grid No. 3



GENERAL DATA

Electrical:

Heater, for Unipotential Cathode:		
Voltage (a.c. or d.c.)	6.3	volts
Current	0.150	amp
Direct Interelectrode Capacitances (with no external shield):		
Pentode Connection:		
Grid No. 1 to Plate	0.11 max.	μμF
Input	2.7	μμF
Output	2.4	μμF
Triode Connection:		
(Grids No. 2 and No. 3 Connected to Plate)		
Grid No. 1 to Plate	1.4	μμF
Grid No. 1 to Cathode	1.4	μμF
Plate to Cathode	0.85	μμF

Mechanical:

Mounting Position	Any
Maximum Overall Length	$2\frac{3}{16}''$
Maximum Seated Length	$1\frac{15}{16}''$
Length from Base Seat to Bulb Top (excluding tip)	$1\frac{9}{16}'' \pm \frac{3}{32}''$
Maximum Diameter	$\frac{7}{8}''$
Bulb	T-6½
Base	Small-Button Noval 9-Pin

Class A_1 AMPLIFIER

Pentode Connection

Maximum Ratings,

Design-Centre Values:

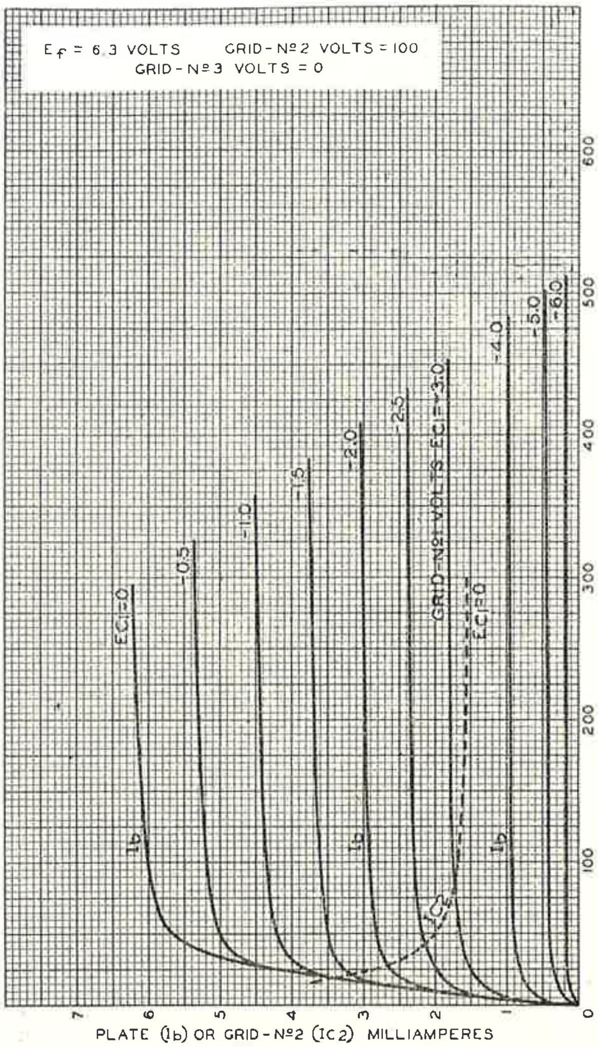
Plate Voltage	300 max.	volts
Grid—No. 2 (Screen) Voltage	150 max.	volts
Grid—No. 2 Supply Voltage ..	300 max.	volts
Grid—No. 2 Input	0.25 max.	watt
Plate Dissipation	1.25 max.	watts
Grid—No. 1 (Control-Grid) Voltage:		
Negative Bias Value	50 max.	volts
Positive Bias Value	0 max.	volts
Peak Heater—Cathode Voltage:		
Heater negative with respect to cathode	90 max.	volts
Heater positive with respect to cathode	90 max.	volts

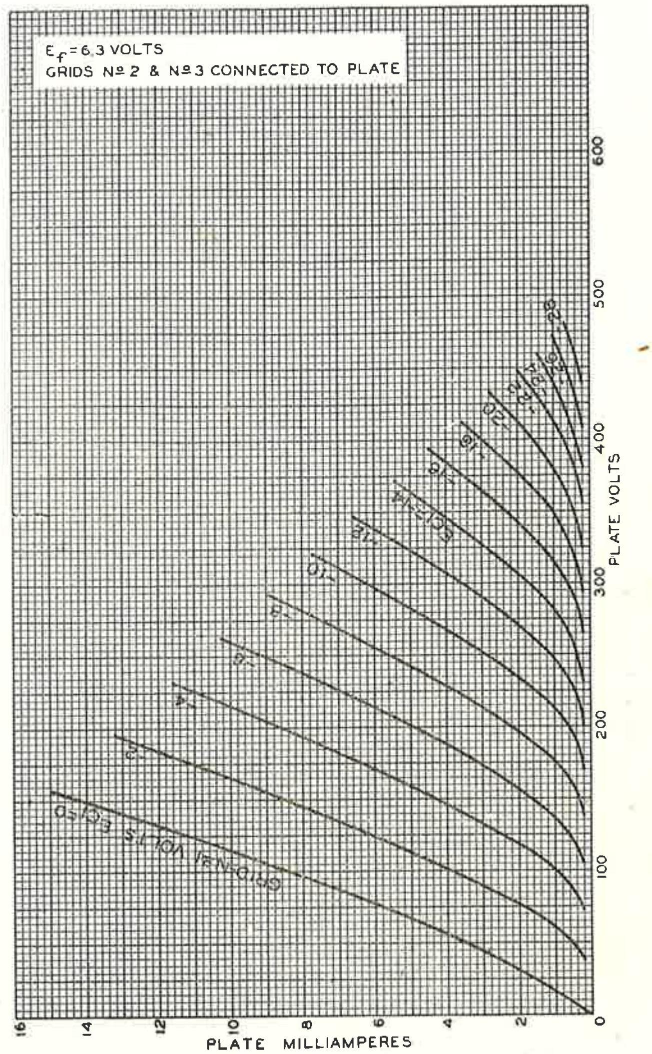
Characteristics:

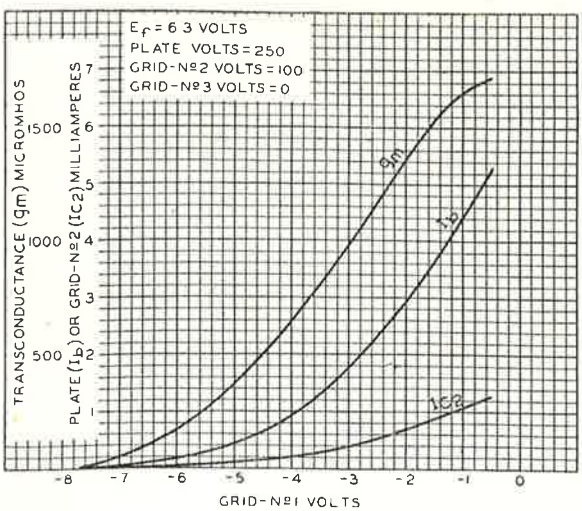
Plate Voltage	250	volts
Grid No. 3 (Suppressor)	Connected to cathode at socket	
Grid—No. 2 Voltage	100	volts
Grid—No. 1 Voltage	−3	volts
Plate Resistance (Approx.)	2	megohms
Transconductance	1000	μmhos
Grid—No. 1 Bias (Approx.) for Plate Current of 10 μamp	−8	volts
Plate Current	1.8	mA
Grid—No. 2 Current	0.4	mA

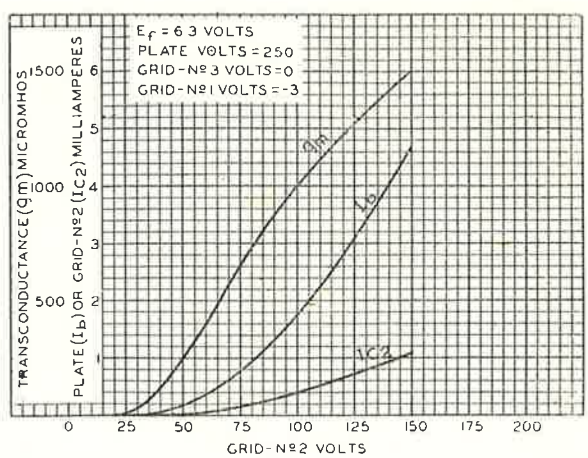
Maximum Circuit Values:

Grid—No. 1—Circuit Resistance
2.2 max. megohms


Ef = 6.3 VOLTS GRID-Nº2 VOLTS = 100
GRID-Nº3 VOLTS = 0
GRID-Nº1 VOLTS EC1 = -3.0
EC1 = 0
-0.5
-1.0
-1.5
-2.0
-2.5
-4.0
-5.0
-6.0
Ib
IC2
EC1 = 0
PLATE (Ib) OR GRID-Nº2 (IC2) MILLIAMPERES


Ef = 6.3 VOLTS
GRIDS Nº 2 & Nº 3 CONNECTED TO PLATE
GRID-Nº1 VOLTS EC1 = 0
PLATE VOLTS
PLATE MILLIAMPERES


Ef = 6.3 VOLTS
PLATE VOLTS = 250
GRID-Nº2 VOLTS = 100
GRID-Nº3 VOLTS = 0
TRANSCONDUCTANCE (gm) MICROMHOS
PLATE (Ib) OR GRID-Nº2 (IC2) MILLIAMPERES
gm
Ib
IC2
GRID-Nº1 VOLTS


Ef = 6.3 VOLTS
PLATE VOLTS = 250
GRID-Nº3 VOLTS = 0
GRID-Nº1 VOLTS = -3
TRANSCONDUCTANCE (gm) MICROMHOS
PLATE (Ib) OR GRID-Nº2 (IC2) MILLIAMPERES
gm
Ib
IC2
GRID-Nº2 VOLTS

Triode Connection

Grids No. 2 and No. 3 Connected to Plate

Maximum Ratings,
Design-Centre Values:

Plate Voltage	250 max.	volts
Total Plate Dissipation	1.5 max.	watts
Grid—No. 1 Voltage:		
Negative Bias Value	50 max.	volts
Positive Bias Value	0 max.	volts
Peak Heater—Cathode Voltage:		
Heater negative with respect to cathode	90 max.	volts
Heater positive with respect to cathode	90 max.	volts

Characteristics:

Plate Voltage	100	250	volts
Grid—No. 1 Voltage	–3	–8	volts
Amplification Factor	21	21	
Plate Resistance (Approx.)	17000	13700	ohms
Transconductance	1240	1530	µmhos
Total Plate Current	2.2	5.5	mA

Maximum Circuit Values:

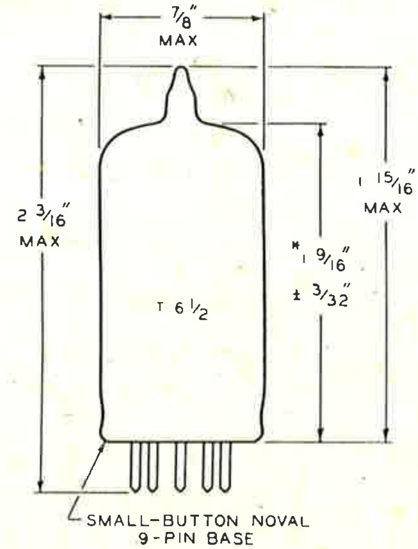
Grid—No. 1—Circuit Resistance 2.2 max. megohms

INSTALLATION and APPLICATION

The base pins of the 5879 fit the noval 9-pin socket. The socket may be mounted to hold the valve in any position.

It is recommended that pins No. 2 and 6 be grounded in all applications. Grounding of these pins will effectively shield grid No. 1 and plate from heater and help to reduce hum level when an a.c. heater supply is used.

DIMENSIONAL OUTLINE



* MEASURED FROM BASE SEAT TO BULB-TOP LINE AS DETERMINED BY RING GAUGE OF 7/16" I.D.

Operating Conditions as Resistance-Coupled Amplifier for Maximum Voltage Output

Plate-Supply Voltage	90			180			300			volts
Plate Load Resistor	0.1	0.22	0.47	0.1	0.22	0.47	0.1	0.22	0.47	megohm
Grid—No. 2 Resistor	0.15	0.40	1.0	0.20	0.53	1.1	0.2	0.52	1.2	megohms
Grid—No. 1 Resistor (of following Stage)	0.22	0.47	1.0	0.22	0.47	1.0	0.22	0.47	1.0	megohm
Cathode Resistor	2200	3800	7400	1400	2300	3700	1100	1600	2500	ohms
Grid—No. 2 By-pass Capacitor•	0.08	0.065	0.04	0.08	0.07	0.07	0.1	0.1	0.1	µF
Cathode Bypass Capacitor•	4.4	3.2	2.0	5.85	4.45	3.5	6.8	5.45	4.3	µF
Blocking Capacitor•	0.013	0.006	0.003	0.013	0.006	0.003	0.013	0.006	0.004	µF
Peak Output Voltage□	28	30	30	59	62	59	110	113	110	volts
Voltage Gain†	32	44	57	46	62	66	53	64	76	

Operating Conditions as Resistance-Coupled Amplifier for Maximum Voltage Gain

Plate-Supply Voltage	90			180			300			volts
Plate Load Resistor	0.1	0.22	0.47	0.1	0.22	0.47	0.1	0.22	0.47	megohm
Grid—No. 2 Resistor	0.35	0.80	1.9	0.35	0.80	1.9	0.35	0.80	1.9	megohms
Grid—No. 1 Resistor (of following Stage)	0.22	0.47	1.0	0.22	0.47	1.0	0.22	0.47	1.0	megohm
Cathode Resistor	1700	3000	7000	700	1200	2500	300	600	1200	ohms
Peak Output Voltage□	17	21	25	28	31	32	32	37	42	volts
Voltage Gain*	39	59	75	56	87	122	68	109	152	

• The grid—No. 2 and cathode by-pass capacitors, and blocking capacitors have been chosen to give output voltages at 100 cps (f_1) which are equal to 0.8 of the mid-frequency value. For any other value of (f_1), multiply the value of cathode and by-pass blocking capacitors by $100/f_1$.

* At an output voltage of 1 volt r.m.s. and grid—No. 1 bias of 1 volt.

□ This peak output voltage is obtained across the grid resistor of the following stage at any frequency within the flat region of the output vs frequency curve, and is for the condition where the signal level is adequate to swing the grid of the resistance-coupled amplifier valve to the point where its grid starts to draw current

† At an output voltage of 5 volts r.m.s.

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