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BUILD A SQUARE-WAVE GENERATOR

by B. J. Simpson

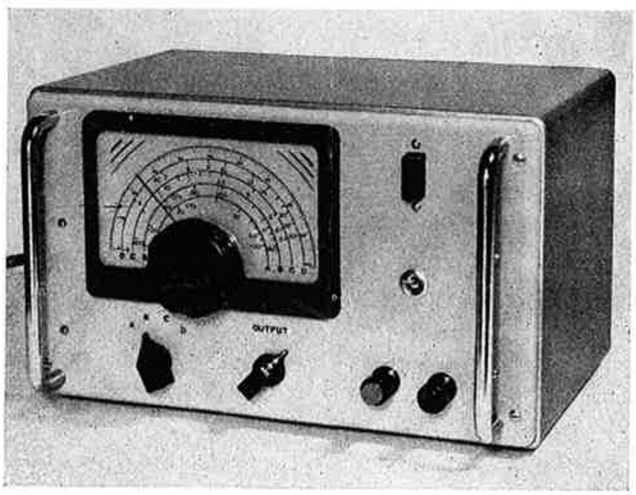
Introduction

This article starts with a true story, and the story has an interesting twist at the end; it is therefore a good true story. Names will be omitted to protect the innocent, as is the custom. The story concerns square waves and the RADIOTRONICS TR24 Transistorized Amplifier (May, June, 1962).

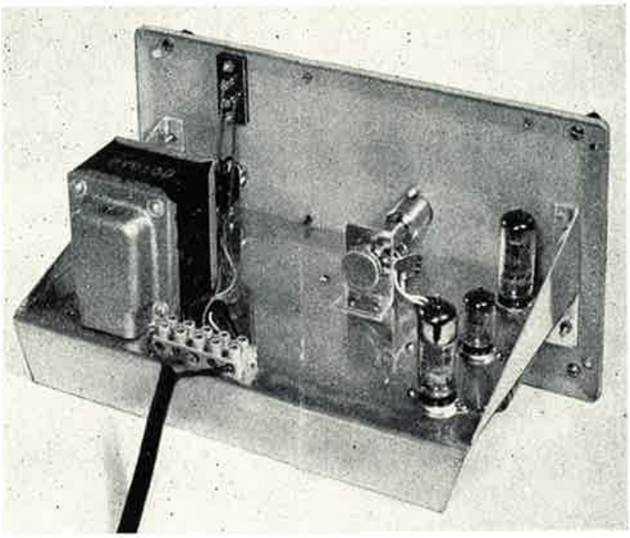
A friend of mine has been preparing for some time a version of the TR24 using printed circuit boards, with a view to making them available commercially a little later. So he built himself a model of the amplifier essentially as we had presented it, but with different layout. This was quite in order, as the unit is inherently very stable and non-critical as to layout.

However, when he came to check the unit "on the air," all he could get out of it was a "beautiful square wave," to quote. Recovering slightly from the shock and analysing the matter, it transpired that the square wave was the same frequency as the setting of the audio generator he had hooked up to the input to try the unit out. To cut the story short, he discovered, to his combined relief and annoyance, that a lead had fallen off the inside of the generator; it was pushing out a square wave instead of a sine wave.

Once discovered, the matter was soon resolved, and testing could proceed. The point of the story of course, is that he had already discovered quite a lot about the performance of the amplifier by inadvertently testing it with a square-wave input.



Photograph of the completed unit, showing the neat and pleasing appearance. This unit would be an asset to any workshop.



Rear view of the unit removed from the case, showing the general layout of components.

Square-Wave Testing

The technique of testing audio amplifiers with square waves has been well-known for many years. There is no doubt that nothing will tell one so much about an amplifier in such a short time. As a rule, the story that square-wave testing tells about the amplifier is in general terms, and this technique cannot therefore supplant other more time-consuming methods of measuring the performance. The technique is complementary.

This is not the time to delve into the matter of square-wave testing, as the subject has already been very well covered elsewhere. Let it suffice to point out that the method opens the way to rapid assessment of amplifier frequency response, phase shift, parasitics, resonant peaks, leakage capacitance, ringing in inductors, and all the other opponents of good performance.

Once some experience is gained in the interpretation of the output waveform from the amplifier, and its departure from the input waveform, a square-wave generator forms an exceedingly valuable tool. Unfortunately, most of the generators available are of very high quality and consequently expensive. A great deal of useful work can be done in the audio field using a much simpler instrument with a smaller frequency range. This unit has been developed especially for readers of this magazine, and will prove an interesting and valuable tool.

Circuit Techniques

As far as instruments go, there are two basic methods of producing square waves of variable frequency. The preferred method is to use a multivibrator or similar configuration to produce a waveform which is roughly square, and with the shortest possible rise times. That is, assuming a perfect square wave, the transition from peak negative to peak positive, or vice versa, should ideally occupy the shortest possible time. Rise times of a fraction of a microsecond are encountered. For audio work, exceedingly short rise times are not required. Subsequent shaping of the pulse is used to produce finally as perfect a square wave as possible. Where amplification within the instrument is required, an extremely wide bandwidth amplifier is required. All of these factors contribute to the high cost of a really good instrument.

The second method of producing square waves is to generate a sine wave at the required frequency, amplify it as necessary, and then to clip off the peaks of the waveform. This is the method often used in audio generators which have a square-wave facility. Consideration of this system will show that it is convenient where the sine wave is already present, and that it is easy to achieve equal mark-space ratio (equal duration of positive and negative halves of the square wave). The system is satisfactory for audio, but where the highest quality instrument is required, it will be seen that it is inherently much more difficult to achieve very short rise times, even if the sine wave is considerably amplified and then very deep clipping is applied.

Some instruments provide for a variation of the mark-space ratio, but here again, for audio work, this is hardly required. Nor do we need to take exceptional trouble to achieve an equal ratio; anything approaching equality will suffice.

The Circuit

In our case, it was decided to use a multivibrator system, in which frequency variation was achieved by altering the time constants of the circuit. The two triode sections of V1 form a free-running multivibrator, the frequency of which is determined by adjustment of the FREQUENCY control shown in the circuit diagram. A number of ranges is used to cover all the frequencies required, the RANGE switch selecting one of a number of fixed capacitors. The multivibrator is of the cathode-coupled type, and the circuit constants are adjusted to provide a roughly equal mark-space ratio.

The output from the 6CG7 is a typical multivibrator waveform, with sharp transitions from one state to the other, but with nothing like the flat-topped wave that we require. Clipping is therefore applied by V2, a 6AL5. Assuming no input from V1, the 6AL5 is so arranged that both diodes are biased off and will not conduct. When an input signal is applied, and its amplitude exceeds the bias on the diodes, then they will conduct on alternate half cycles, and so clip the top and bottom of the waveform. It will be seen that whilst the diodes present a very high impedance when not conducting, they present a very low shunting impedance when they are driven into conduction.

The waveform at pins 5 and 7 of the 6AL5 is now essentially what we want, a flat-topped wave with rapid transition time. Note that if the input to the clipper circuit falls, then a point will be reached where clipping no longer takes place and an unsatisfactory waveform will result. The circuit is therefore arranged so that a signal in excess of the clipping level will be applied to the 6AL5 at all frequencies, and with normal circuit variations.

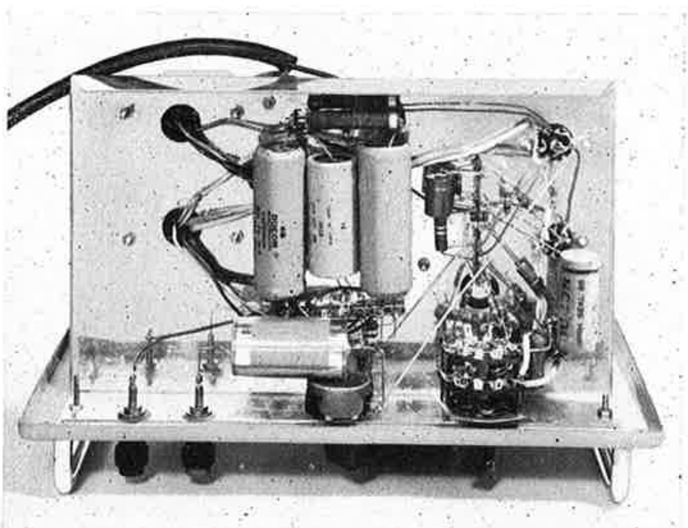
All that remains now is to convert the high impedance at the clipping point to a low impedance output. This is quite simply done by using a cathode-follower output stage. As no gain is obtained in this stage (in fact a slight loss), the maximum output amplitude is essentially determined by the clipping level in the 6AL5 stage.

The unit is powered by a voltage-doubler rectifier system using one of the simple mains transformers developed for use with silicon diodes. A simple RC filter is used and is adequate for the purpose. Two AWV 1N1763's are used as the rectifiers.

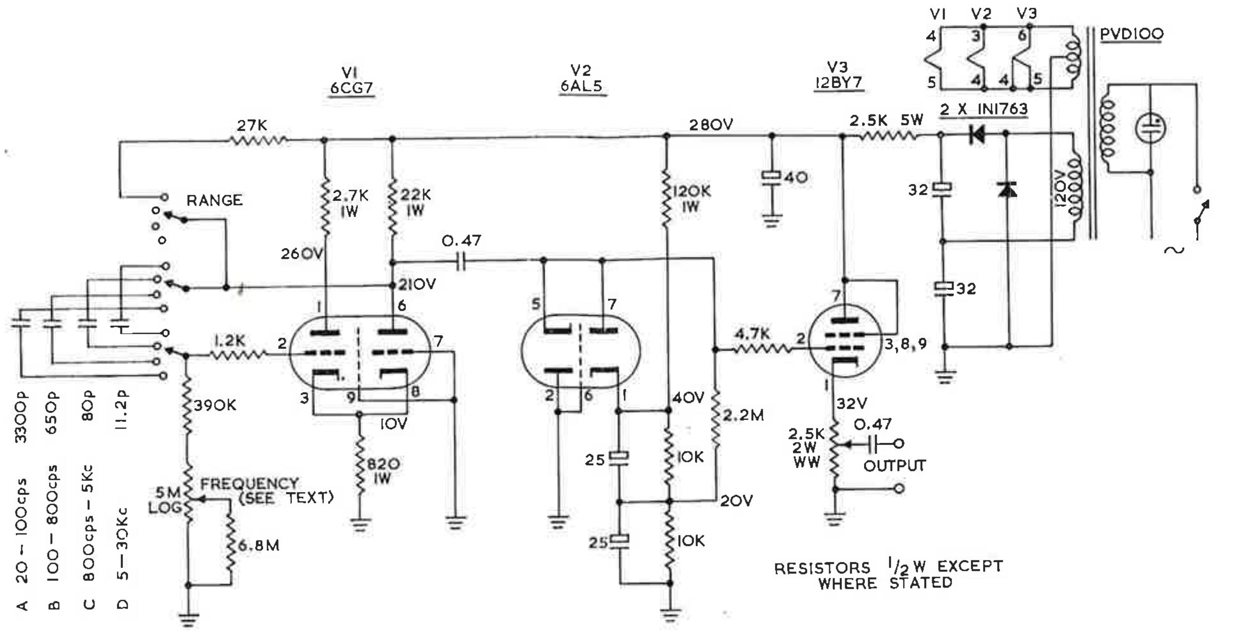
Construction

The unit is built into an Imhof 1490B case with two 10C chromium handles and BC504 aluminium bracket chassis. This is one of a number of cases on which we are standardising for our test instrument articles; the finish and appearance are well worth the extra cost, and make the units an asset to your workshop.

The general layout of the instrument is shown in the accompanying photographs. Because stray capacitances can make it difficult to achieve a good waveform at the higher frequencies, attention must be paid to the pin 6 plate circuit of the 6CG7, and the waveform path from pin 6 to the grid of V3. Keep these leads as short as possible and as far as is reasonable from other wiring and the chassis. The layout used in our model was designed to minimise strays and it would be a



Underside view of the square wave generator, showing the component layout. See text regarding critical layout.



Circuit diagram of the square wave generator described in this article. See text for further details.

good idea to follow it, at least as far as the three valves and the critical portion of the wiring are concerned.

Calibration of the unit when completed is a simple matter. The easiest way is to use an oscilloscope and a calibrated audio oscillator, carrying out the calibration by direct comparison of the frequencies. No adjustments should be necessary except trimming of the switched capacitor values as necessary to obtain the required frequency coverage.

The PVD100 mains transformer used in this unit has several secondary taps. In this case the blue and yellow leads (120 volts) were used, but this is not critical, so that if another suitable transformer is already to hand it can be used. Current consumption at the 280 volts B+ line is about 30 to 40 milliamps, depending on the operating frequency. The voltages shown in the circuit diagram are typical and may vary by up to 20% in individual units.

Care taken in the construction and layout will allow the full stated frequency range to be obtained without difficulty, and with sufficiently fast rise times for the type of application in mind. In the model an Eddystone 598 dial and drive were used, as it was felt that the additional cost of this unit over a simple slow-motion mechanism was justified. Further, it is extremely useful to have a scale calibrated in frequency rather than use an arbitrary scale and conversion chart. The appearance of the instrument and the ease of use well repay the extra time and money to make the unit a complete job.

The output of our model was not less than 25 volts peak-to-peak over all frequencies. Here again, the performance is more than adequate. Note that the output potentiometer must be a wire-wound component. The model used a 2.5 K ohms unit, but 2 K or 3 K would be suitable.

The switched capacitors, and all the signal wiring between pin 6 of the 6CG7 and pin 2 of the 12BY7, and the wiring between pin 2 of the 6CG7 and the FREQUENCY control, should have minimum capacitance to ground. Install the 5 megohm potentiometer so that the top (all in) position corresponds to the 180° position of the dial; in this way the usable portion of its travel corresponds to about 3.9 megohms, and a portion of the track is unused.

Using the Unit

Those readers who are not familiar with square-wave testing of audio amplifiers will be well repaid by the trouble taken to make a small study of the subject, and by learning to recognise the symptoms. Very useful in this regard is an old amplifier which can be used as a testing ground. Various effects could be tried out and the results noted for future use. Keep the con-

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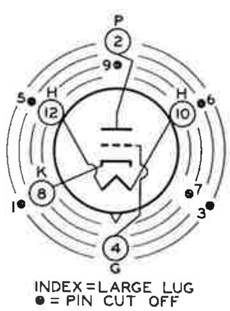
6CW4 RF AMPLIFIER IN VHF TV

This article describes the high-frequency performance of the 6CW4 nuvistor triode and its application as an rf amplifier in vhf television tuners. The performance of the 6CW4 is evaluated in an experimental neutralized grid-drive amplifier circuit, an experimental turret tuner, and a production-type switch tuner. Optimum noise-factor data and practical circuit considerations are also presented.

Design Features of the 6CW4

The 6CW4 is a high-mu, high-transconductance triode of the nuvistor design, featuring extremely small size and light weight. The cylindrical active elements are mounted coaxially on ceramic base wafers. Each element is supported by a tripod arrangement of leads which extend through the ceramic base wafer. One lead of each set is used as the external connector, as shown in the bottom view of Fig. 1.



PIN 1:	▲	PIN 7:	▲
PIN 2:	PLATE	PIN 8:	CATHODE
PIN 3:	▲	PIN 9:	▲
PIN 4:	GRID	PIN 10:	HEATER
PIN 5:	▲	PIN 11:	OMITTED
PIN 6:	▲	PIN 12:	HEATER

▲ Pin has internal connection and is cut off close to ceramic wafer--Do Not Use.

Fig. 1—Bottom view of pin arrangement for the 6CW4.

Sections of the shell which extend beyond the base wafer serve as indexing lugs for socket insertion. These indexing lugs also provide protection for the leads and ground the metal shell through the socket, making the use of a shield unnecessary.

Several important advantages are inherent in the unique nuvistor design. For example, high transconductance is obtained with a high transconductance-to-plate-current ratio, as shown by the transfer characteristics in Fig. 2. These characteristics are achieved with a considerable reduction of both plate-input and heater-input power. Interelectrode capacitances are comparable to those of conventional miniature triodes. In addition, accurate element spacing in the valves during assembly permits a high degree of uniformity of characteristics from valve-to-valve, especially with respect to interelectrode capacitances. As a result, valves can be replaced with a minimum of circuit adjustment.

The small size and short lead lengths of the valves make them particularly suitable for rf-amplifier applications in vhf television tuners. The high transconductance-to-plate-current ratio contributes to low valve noise factor. In addition, short-circuit input-impedance measurements indicate that these valves have higher input resistance than others having equivalent input capacitance and transconductance. These measurements approximate the input resistance of a completely neutralized triode in grid-drive operation. As shown by the following equation for power gain at vhf frequencies (impedance-matching losses neglected), increased input resistance results in increased gain.

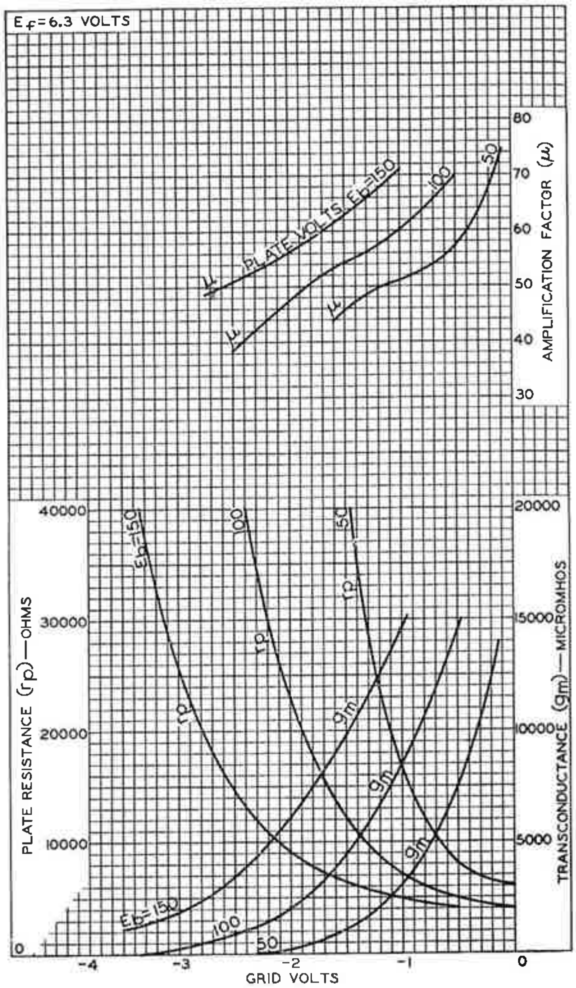


Fig. 2—Average characteristics for the 6CW4.

$$\text{Power Gain} = \frac{\mu^2 R_S R_L}{(r_p + R_L)^2}$$

where μ is the amplification factor,
R_S is the source resistance (matched to input resistance),
R_L is the load resistance, and
r_p is the plate resistance of the valve.

Neutralized Grid-Drive Amplifier

Fig. 3 is a circuit diagram of a neutralized grid-drive amplifier employing a 6CW4. In this circuit, input and output networks are matched to a 50-ohm signal generator and load impedance, respectively. A capacitive-bridge network is used for neutralization. The plate circuit, a double-tuned transformer-coupled network tuned to 200 megacycles, has a six-megacycle bandwidth. For determination of optimum noise factor, source admittance was varied by use of the "pi" input-matching network. Fig. 4 shows curves of noise factor and gain as functions of source conductance and susceptance for this circuit.

Noise-factor measurements were made by use of a coaxial temperature-limited diode noise generator. The noise power of the circuit, as indicated by a detected output signal, was increased 3 db by the introduction of rf noise signals from the noise generator. The noise generator used was calibrated to indicate noise factor directly when the noise power was doubled.

A substitution method was used for measurement of power gain. In this method, the voltage output of a signal generator was set to some arbitrary reference level. The amplifier was then placed in the circuit, and the signal-generator output was attenuated to reduce the amplified

CHANNEL NO.	NOISE FACTOR DB	TUNER VOLTAGE GAIN DB
2	3.9	47.9
6	4.0	45.3
7	5.3	45.0
13	5.5	44.0

Table I—Noise-factor and gain performance of an experimental turret tuner employing a 6CW4 at different television-channel frequencies.

CHANNEL NO.	NOISE FACTOR DB	TUNER VOLTAGE GAIN DB
2	5.1	45.0
6	5.0	43.5
7	4.7	44.5
13	5.5	45.0

Table II—Noise-factor and gain performance of a production-type switch tuner employing a 6CW4 at different television-channel frequencies.

VALVE TYPE	NOISE FACTOR DB	TUNER VOLTAGE GAIN DB
6BN4A	8.5	38.0
6FH5	7.5	41.0
6ER5	7.5	41.0
6CW4	5.5	45.0

Table III—Noise-factor and gain comparison of the 6CW4 with conventional miniature triodes in a television tuner at channel 13.

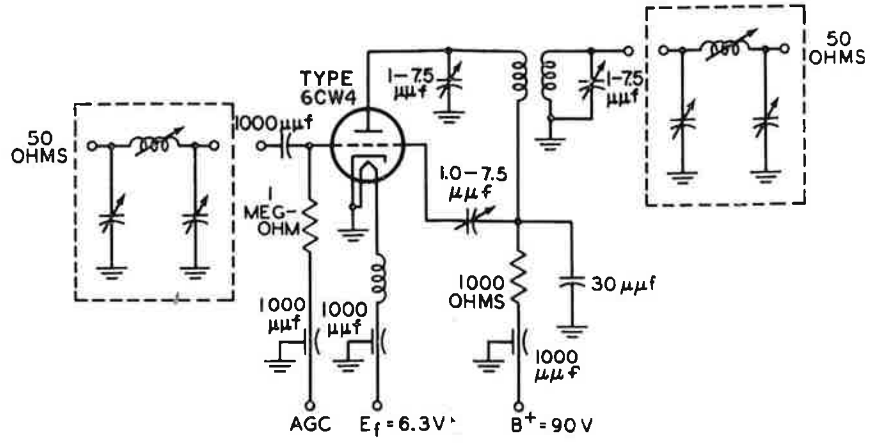


Fig. 3—Circuit diagram of a neutralized grid-drive amplifier employing the 6CW4.

signal to the original reference level. The difference between the dial readings of the signal-generator; which was calibrated in db, indicated the gain of the amplifier directly.

Turret Tuner and Switch Tuner

Fig. 5 is a circuit diagram of the rf-amplifier section of an experimental turret tuner. In this circuit, a 300-to-70-ohm balanced-to-unbalanced transformer input is used for impedance matching, and a mixer valve serves as the load for the rf section. Conventional tuner design techniques and components are used throughout. Noise factor and over-all tuner voltage-gain performance are tabulated in Table I.

Similar data are shown in Table II for a production-type switch tuner using a 6CW4. The circuit for this tuner is almost identical to those shown in Figs. 3 and 5; however, the load on the tuned circuits is increased to improve coupling on the lower-frequency channels.

A 50-ohm signal generator equipped with a balun to match to the tuner input was used for gain measurements of the turret tuner and switch tuner. Both tuner circuits incorporated a dummy if stage consisting of an if-amplifier with a 10,000-ohm grid resistor. Again, the substitution method outlined previously was used to measure gain. Table III compares the average performance of the 6CW4 at channel 13 with that of conventional miniature types in the same circuit.

General Considerations

The 6CW4 is designed to operate at relatively low plate voltages for best rf performance. Optimum tuner performance is obtained with a plate voltage in the order of 70 to 80 volts. Fig. 6 shows curves of noise factor and tuner gain as a function of plate power dissipation at various B+ values applied to the tuner. The B+ voltage is dropped slightly across a 1000-ohm resistor. At a given level of plate dissipation, performance improves as the plate voltage is reduced. (The limit of plate-voltage reduction is determined by the amount of plate current that can be drawn without the application of positive bias voltage.) The optimum plate voltage is determined as a

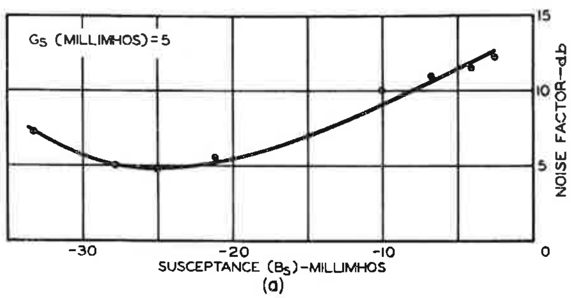


Fig. 4(a)—Noise factor as a function of susceptance for the amplifier shown in Fig. 3.

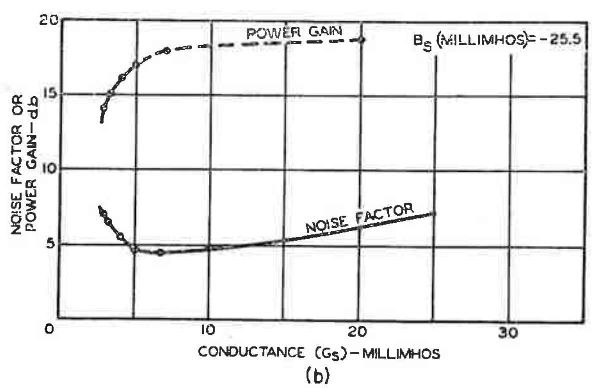


Fig. 4(b)—Noise factor and power gain as functions of conductance for the amplifier shown in Fig. 3.

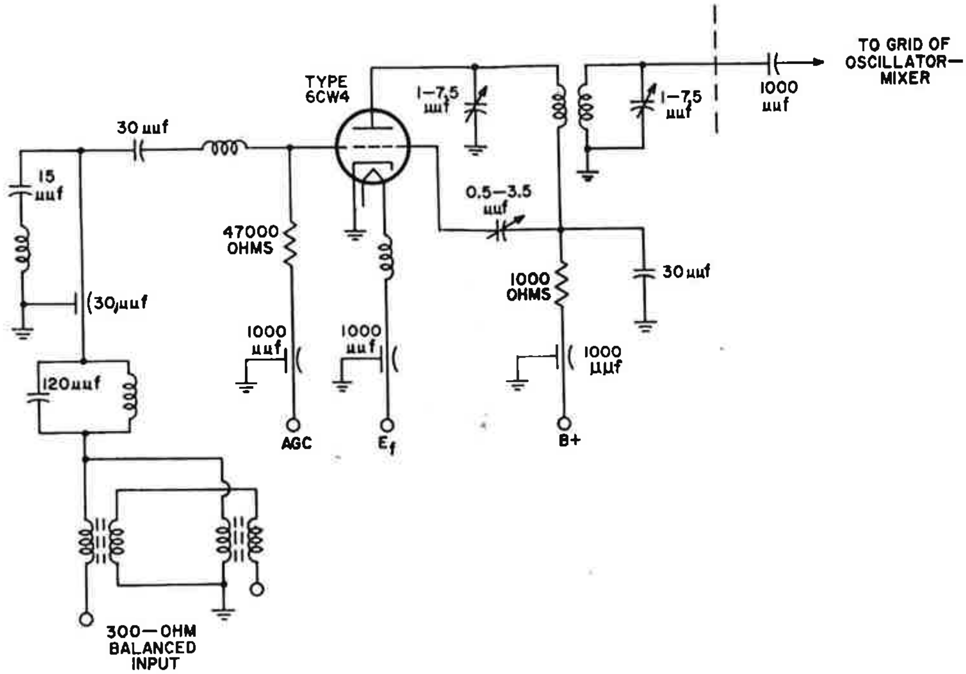


Fig. 5—Circuit diagram of the rf-amplifier section of an experimental turret tuner.

balance between performance and plate dissipation. Maximum signal-to-noise ratio at a given plate voltage is obtained at an applied bias of zero volts.

Under these conditions, the 6CW4 has a sharp-cutoff characteristic which is unsuitable in systems employing agc. Supply voltages available in television receivers are, however, considerably higher than the operating plate voltage of these valves. As a result, cutoff can easily be extended by the addition of a plate-voltage dropping resistor. Fig. 7 shows curves of gain reduction as a function of grid bias for different values of supply voltage. As indicated, the circuit designer determines the value of plate-voltage dropping resistor or series shunt combination which provides the desired cutoff characteristic.

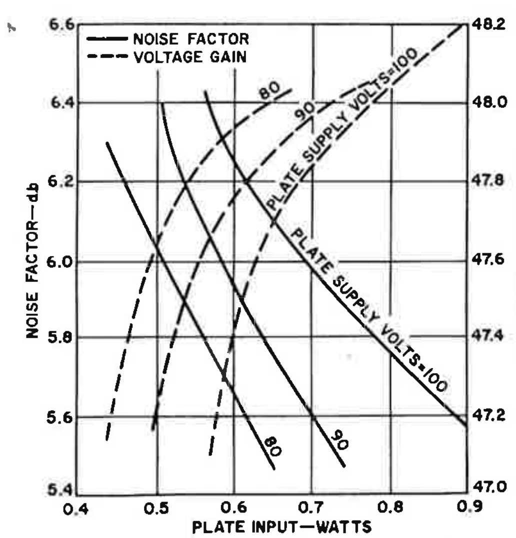


Fig. 6—Voltage gain and noise factor as functions of input power for various B+ voltages applied to the turret tuner.

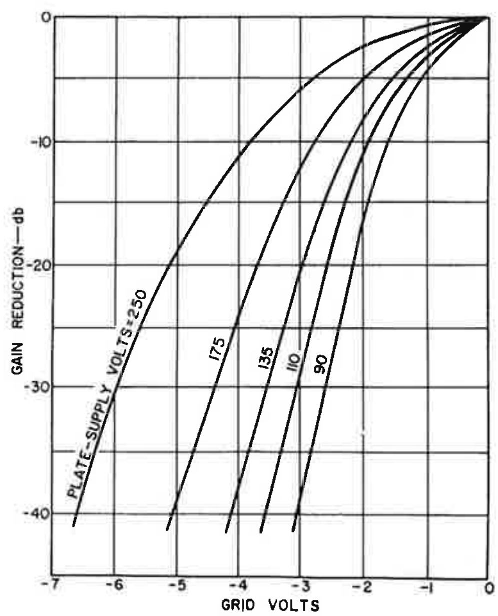


Fig. 7—Gain reduction as a function of grid-bias voltage for different values of supply voltage.

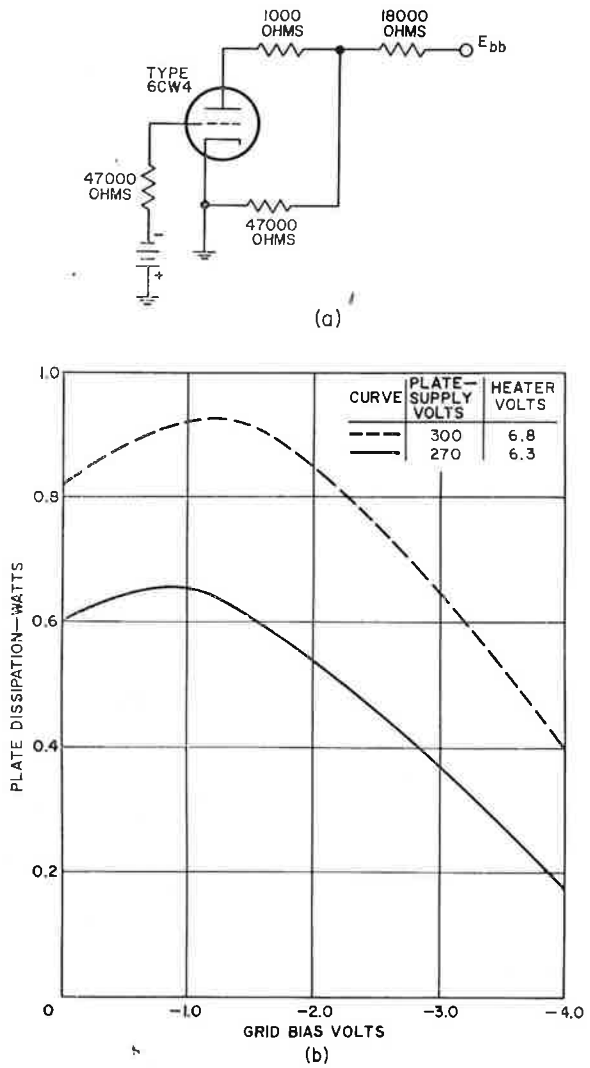


Fig. 8(a)—Typical circuit configuration using the 6CW4 and (b)—curves of plate dissipation as a function of agc bias voltage under normal and worst probable conditions of operation.

Use of the 6CW4 with a series dropping resistor at high supply voltages requires careful choice of valve-element operating values so that maximum ratings are not exceeded. Maximum plate dissipation, for example, does not necessarily occur when the valve is drawing maximum plate current, but can occur at some bias value other than zero. Fig. 8(a) shows a circuit configuration using a typical 6CW4. Fig. 8(b) shows curves of plate dissipation as a function of bias voltage obtained from the agc circuit. With the indicated circuit and voltage values, maximum plate dissipation (as shown by the solid-line curve) occurs at a bias voltage of slightly less than one volt.

With the values of resistors adjusted to either the upper or lower limits of their 10-percent tolerances to simulate the worst probable conditions of operation and with the high heater and B+ voltages indicated, the resultant maximum plate dissipation is shown by the dashed curved in Fib. 8(b). In this case, maximum plate dissipation occurs at a bias voltage slightly above 1 volt. The equipment designer has the responsibility of choosing values of circuit components which will prevent maximum ratings of the tube from being exceeded under the worst probable operating conditions.

As mentioned previously, optimum performance of the 6CW4 is obtained when operated with an applied bias of about zero volts. Under these conditions, the valve draws current, and is at some negative potential depending on the grid impedance to ground. Because automatic-gain-control systems generally present high impedances to ground, care must be taken to assure that a low grid bias voltage is applied to the valve. A common method is to return the agc terminal to a positive voltage through a large-value resistor to clamp the grid voltage at low signal levels.

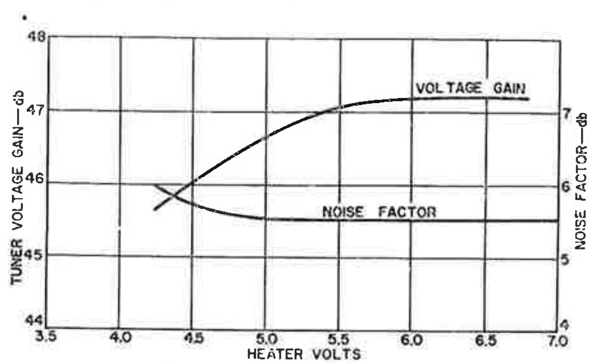


Fig. 9—Voltage gain and noise factor as functions of heater volts for the turret tuner.

Fig. 9 shows the variation of tuner noise factor and voltage gain produced by reduction in heater voltage. The curves show that noise factor is only slightly affected at reduced heater voltage, and that gain drops only about 0.5 db with a change in heater voltage from 6.3 to 5.0 volts.

(With acknowledgements to RCA)

CONTINUED

SINGLE-SIDEBAND

by B. J. Simpson

The first part of this article last month explained the nature of single-sideband and the reasons for its existence. It was shown that this technique is derived from the standard amplitude modulated system by the suppression of either one of the sidebands, and/or the suppression of the carrier. Where only the carrier is suppressed, the system is known as double sideband (DSB).

It is now necessary to examine the processes by which the carrier can be suppressed and/or one of the sidebands can be removed. Because the two processes can be used independently, and rely upon different considerations, they will be dealt with separately. Discussion of techniques for suppressing the carrier will be dealt with first.

Carrier Suppression

Suppression of the carrier is carried out in a balanced modulator or mixer, the active elements in which can be thermionic valves or semiconductors. Several different configurations are possible, but they all achieve the same basic result. The modulators in common use may be divided for ease of description into those which use thermionic valves and those which use semiconductor diodes. Finally, a balanced mixer using a special type of beam-switching valve will be described.

Push-Pull Balanced Modulator

The first of the thermionic valve modulators to be described is shown in Fig. 6. This configuration uses two tetrodes in an arrangement providing for parallel input of the rf signal and the modulating signal, and push-pull output. The output of a very stable rf source is fed to both grids in parallel, with the usual arrangements for biasing. The modulating signal M is fed through the usual matching and isolating transformer to the screens of the two tetrodes. The screens are bypassed to ground with respect to the rf signal by the capacitors shown. The output consists of a conventional push-pull rf output circuit, which is tuned in the usual way.

To examine the operation of this circuit, consider the rf input applied with no modulation present. Both grids are driven in phase, so that cancellation takes place in the output circuit, assuming exact balance, and no rf output will result. Carrier suppression of the order of 20 db can be obtained with this type of circuit using selected valves; on the other hand, a considerable increase in the carrier attenuation can be obtained by bias adjustments to equalise plate currents, balancing of grid and plate capacitances to ground, and similar measures.

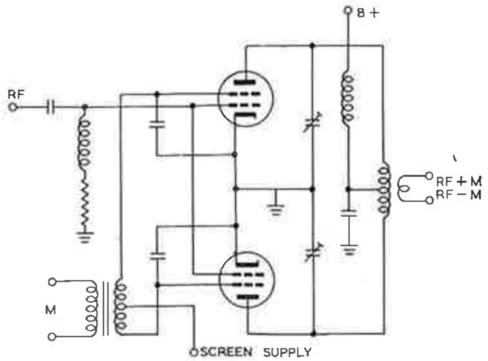


Fig. 6

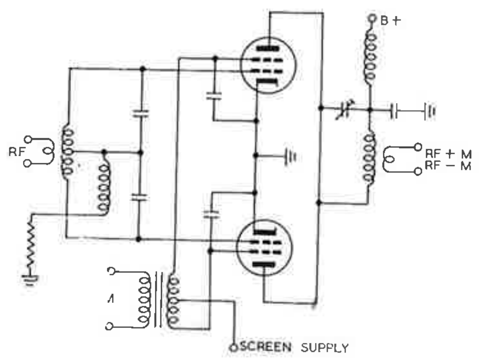


Fig. 7

Carrier suppression of about 35 db is usually required, so that some care in the design and adjustment of the circuit is required. Where a sideband filter is used (to remove one of the sidebands), additional carrier attenuation of perhaps 20 db is obtained.

When modulation is applied to the circuit, the modulating voltages applied to the two screens are 180 degrees out of phase. This causes the plate currents in the two valves to vary at a rate determined by the modulating frequency, but out of phase with respect to each other. The equal rf voltages in the two halves of the plate coil are now replaced by two unequal voltages, which are varying in amplitude at a rate depending on the modulating frequency.

This means that there will be a resultant rf voltage in the secondary of the plate coil. It will be seen that the modulating frequency is being used to create unbalance in the circuit. The resultant rf signal will be modulated by the audio input to produce sum and difference signals (sidebands) in the output. The modulator is balanced with respect to the carrier, but not with respect to the sidebands. The output will therefore consist of two sidebands with suppressed carrier. This is a double sideband (DSB) signal. If SSB is required, one of the sidebands must now be suppressed; this will be discussed later.

Parallel Output Balanced Modulator

A fairly obvious variation of the preceding circuit is shown in Fig. 7. In this circuit, instead of having the two valves fed with the rf signal inphase, with cancellation in the plate circuit, the two valves are fed out of phase, with cancellation in a single-ended output circuit. The operation of this circuit follows from the description of the previous one.

A word of explanation is required with both Fig. 6 and Fig. 7 as far as the screen supply is concerned. Both circuits are operated at low levels, the rf and modulating voltages usually being only a few volts. Dependent on the particular circuit and the valves used, the screen is sometimes operated at ground potential, sometimes with a small B+ voltage applied, and sometimes even with a small negative bias on the screens. The basic criterion of operation in both circuits is the degree of carrier suppression that can be achieved.

Another Variation

We have so far seen a circuit with push-pull output and one with push-pull input. There remains a circuit that has both input and output in push-pull. The basic configuration is shown in Fig. 8, and uses a couple of medium-mu triodes. The modulating signal is fed in antiphase to the two grids from a preceding phase splitter or centre-tapped transformer, whilst the rf input is applied to the common cathode connection. The in-phase injection of the rf signal results in cancellation of the carrier in the output, as previously explained.

A pre-set potentiometer is used to adjust the plate currents for maximum carrier suppression. This is a very simple circuit, not only to build, but also to operate.

An Aside

From the foregoing remarks, readers will already have deduced that there are other differences in SSB and DSB transmitters besides the suppression of the carrier and/or one of the sidebands. The reference to low level operation of the three modulators already mentioned provides a clue to a further difference. A little reflection will show that if the carrier is not required, there is little point in amplifying it right through to the final stage.

In a conventional AM transmitter, modulation is most commonly applied in the final power

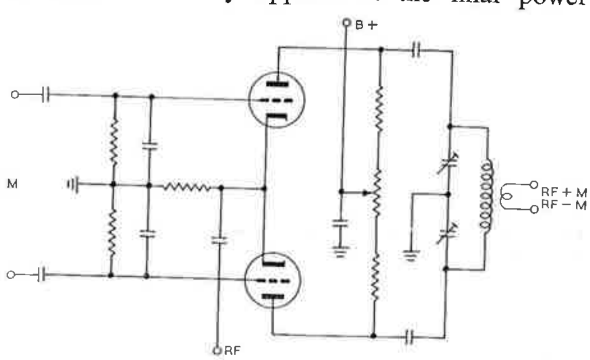


Fig. 8

amplifier stage, requiring among other things a high-power modulator. The practice with SSB and DSB is a little different in that modulation and suppression of the carrier is usually carried out at comparatively low levels, often with ordinary receiving-type valves. The sideband(s) are subsequently amplified in a linear rf amplifier. Here the choice lies between a class A or a class B amplifier, but in practice the low efficiency of the class A circuit leads to the use of a linear class B amplifier.

The class C amplifier is unsuitable for this type of application, but the (usually) added cost of the class B amplifier is likely to be more than offset by the lower cost of the modulating circuits. In any case, the overall system efficiency is considerably higher, as explained in the earlier part of the article.

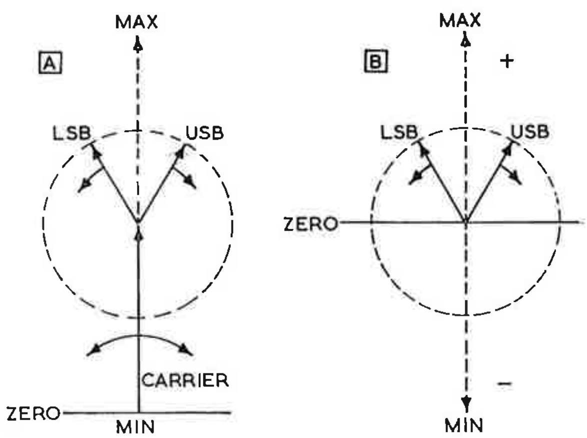


Fig. 9

More Vectors

In Fig. 3 of the earlier part of this article, I attempted to show in a simple vector diagram the process of modulating a carrier wave, assuming 100% modulation. It was seen that the power in the two sidebands was half the carrier power. It was also seen that the minimum resultant was zero and the maximum resultant twice the carrier power. Bearing that diagram in mind, it may now be interesting to see what has happened now that the carrier has been virtually removed.

An amplitude-modulated wave is shown in Fig. 9A. The two sideband vectors are shown as half the length of the carrier vector, and because their frequencies differ from that of the carrier, they can be regarded as revolving around the carrier vector. The lower sideband is shown revolving anti-clockwise and the upper sideband clockwise. To complete the picture, the sum resultant of both the carrier and the two sidebands is assumed to revolve around the origin of the carrier vector at the carrier frequency.

At peak modulation, the resultant will be twice the carrier amplitude, as shown, whilst 180 degrees later, the sideband vectors will have rotated such that their resultant is equal in amplitude but opposite in polarity to the carrier. The resultant at this instant will be zero.

Turning now to Fig. 9B, here is the equivalent diagram of the suppressed carrier system. Here the zero reference point coincides with the origin of the sideband vectors, because there is no output when no modulation is applied. The two sideband vectors rotate as before. Two conditions obtain where the sideband vectors coincide, resulting in maximum and minimum resultants, or positive and negative peaks. When the vectors oppose each other, on the zero reference line, then the resultant output is also zero.

The fact that the sideband resultant swings both negative and positive with respect to the zero reference line during one modulation cycle is of great importance, and is made use of later when the transmitted intelligence has to be recovered in the receiver.

Ring Modulator

It is now necessary to look at the variety of balanced modulators which is available, using semiconductor diodes. The well-known advantages of semiconductors make these circuits very popular. These diode modulators, originally using copper-oxide rectifiers, have been in use in carrier telephony circuits for many years. Until the growing interest in SSB, they were almost unknown elsewhere.

The first circuit to be mentioned is the balanced ring modulator, shown in Fig. 10. Whilst this is the most efficient type of diode modulator, it requires close matching of the

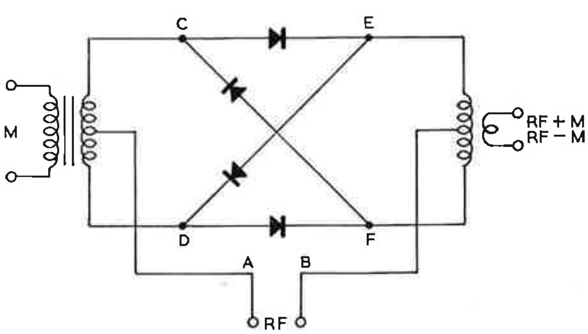


Fig. 10

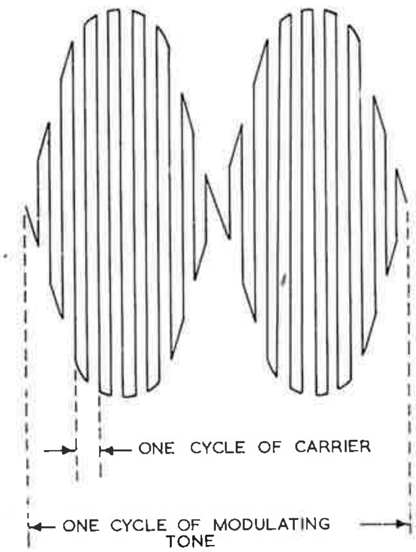


Fig. 11

characteristics of the diodes used, or good carrier suppression will not be obtained.

Turning to the operation of the modulator, it must first be shown that the application of the rf input alone does not result in an output signal. The carrier balancing system can be traced by following the reference letters inserted in the diagram. When the input is positive at point A, current will flow from B, through both paths EC and FD and so to A. When the rf input polarity reverses, current will flow from A, through both paths CF and DE, and so to B. In each case, equal and opposite rf signals will be present in the primary of the output transformer, and no output will result, assuming perfect balance.

When a modulating signal is applied to the modulator, the circuit will be unbalanced with respect to the sum and difference frequencies. This arises from the fact that the diodes will be differentially biased by the audio signal, which is applied in push-pull fashion. When the audio signal polarity is such that any diode is receiving a reverse bias, then the diode will not conduct for the rf signal until the amplitude of that signal exceeds that of the audio signal at that instant. At the same time the audio signal applied to the corresponding diode in the other arm of the modulator will receive a forward bias from the audio signal, assisting conduction.

It is the practice with this type of modulator to use an rf signal several times the amplitude of the modulating signal. This assists in securing good switching action, and also tends to reduce distortion in the modulator output. The resulting output waveform from the modulator will resemble that shown in Fig. 11. As one may expect from the pulsed nature of the output and the waveform shown in Fig. 11, harmonics are generated in the modulator. The most serious effect is the generation of two further sidebands disposed around the second harmonic of the carrier frequency. There are also other unwanted higher-order products. These unwanted outputs are removed in the following tuned stages of the transmitter, additional filters being used if necessary.

Shunt Diode Modulator

Shunt-balanced modulators employing diodes are also used, and they fall into two basic types, those using four diodes and those using only two. The four-diode shunt modulator is shown in Fig. 12, and, as one may expect, it has a strong superficial similarity to the ring modulator.

The operation of this circuit is quite different. The four diodes, in the absence of a modulating signal, conduct in response to the rf input; all four conduct on one half cycle of the rf input and present a high impedance to the other half cycle. This gives rise to two conditions, one where the rf input and output circuits are virtually short-circuited (hence the term shunt modulator), and the other where the impedance inserted into the rf input circuit is very high (diodes not conducting). In both cases the rf input at the carrier frequency is effectively zero.

As in the previous case, the diodes in the two arms of the bridge configuration will be differentially biased by a modulating signal, unbalancing the modulator with respect to the wanted sum and difference signals which form the sidebands. The output waveform of the modulator will be similar to that shown in Fig. 11, except that alternate half cycles of the output will not be present; see Fig. 13. Output currents flow for only half the modulated carrier frequency. When the output is fed to a tuned circuit, the missing half cycles will be supplied by the familiar "flywheel" effect of a resonant circuit.

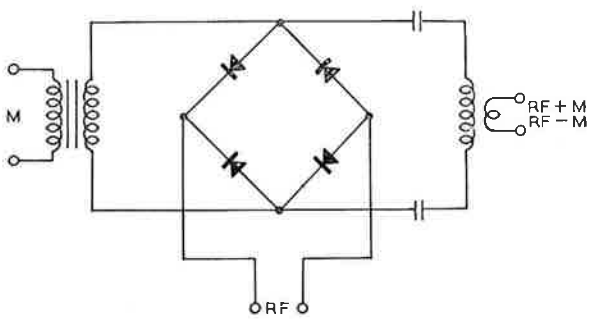


Fig. 12

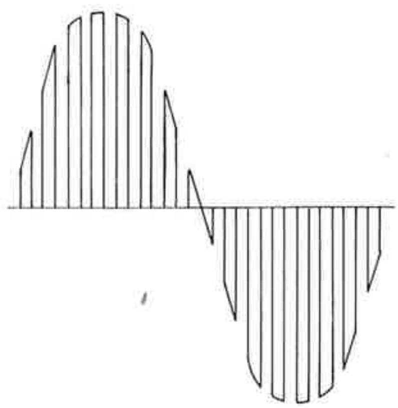


Fig. 13

Another shunt-type modulator is shown in Fig. 14. This circuit uses two diodes which form part of a bridge configuration, the other two arms of the bridge consisting of resistive elements. As before, the two diodes conduct only to alternate half cycles of the rf input in the absence of a modulating signal. Balance is achieved by means of the adjustment in the resistive arms of the bridge. A preset capacitor is sometimes used between either A or B and ground to ensure equal rf voltages at points A and B, the appropriate connection being determined experimentally. A balanced rf input signal is required for this type of modulator.

Once more, when the modulating signal is applied, the two diodes will be biased in different directions, and so the bridge will become unbalanced. As before, rf output will be obtained from the circuit, consisting of the required sum and difference signals, that portion of the carrier that "leaks through" the modulator, and the unwanted products previously mentioned.

The use of shunt modulators inevitably imposes a wide variation of load conditions on the rf source. The source must therefore have good regulation and frequency stability under load.

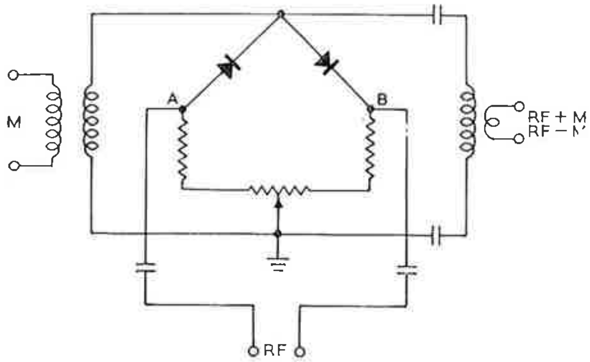


Fig. 14

Poor frequency stability will lead to a frequency-modulated component in the output. Good matching of the diodes used in diode modulators is also required, although they have the advantage of good long-term stability once they are selected and installed.

Series Balanced Modulator

Single-sideband operation requires not only the suppression of the carrier, but also the removal of one of the sidebands. One of the ways of removing a sideband is the use of a filter, which generally follows the modulator in the circuit arrangement. Correct termination of the filter can sometimes be difficult using the circuits already described, and it is cases like this where a series-balanced modulator may be beneficial.

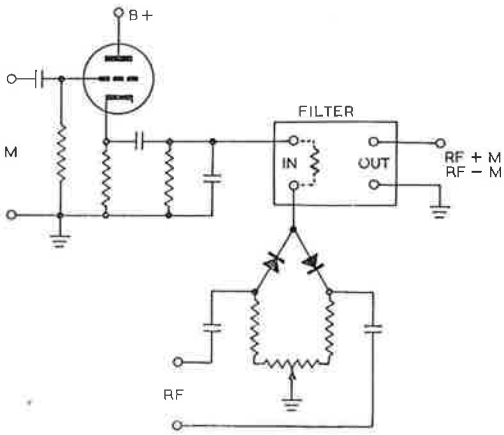


Fig. 15

A typical arrangement of a series-balanced modulator is shown in Fig. 15. Here the circuit is developed around the two-diode shunt modulator previously discussed, but the four-diode circuit could also be used. The filter input impedance with its correct termination resistor are shown diagrammatically by the dotted resistor shown. The method of operation of this circuit follows from what has already been said about the diode shunt modulators.

The circuits shown here are typical of practice. They do not include measures that may have to be taken to obtain correct matching of impedances or other practical considerations. They are representative, and will be found in widespread use under the basic configurations shown.

Beam Deflection Modulator

As foreshadowed earlier, there is yet one more balanced modulator to be mentioned, using a special beam deflection valve. This circuit is fairly recent in origin, and appears to be gaining in popularity. Typical of the valves used in this circuit is the 7360. An extensive description of this valve and the circuits in which it could be used was given in "Radiotronics," Vol. 25, No. 6, June, 1960.

Typical of the application of the 7360 is Fig. 16. The diagram of the valve shows it to be in essence a tetrode with two separate plates, and containing two deflection plates. The total beam current flowing in the valve is determined by the potential on the control grid, whilst the sharing of the total current between the two plates is determined by the potentials on the deflection electrodes. Each plate current is a function of two simultaneous voltages. Important to the operation of the valve is the fact that with no potential on the deflection electrodes, the single beam current strikes neither plate, so that no plate current flows.

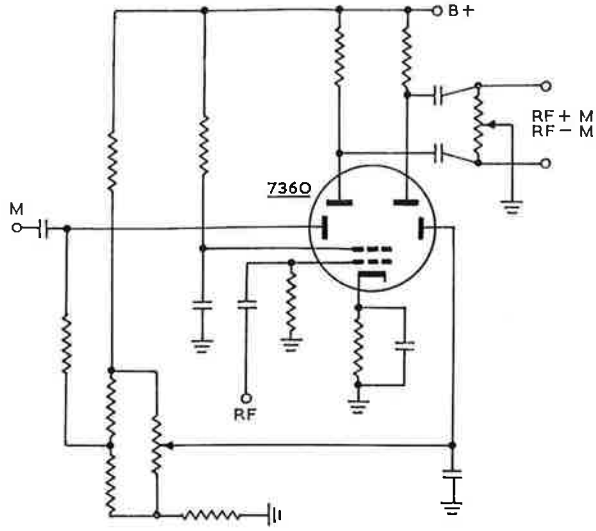


Fig. 16

This means that if the rf input is applied to the control grid, and no signals are applied to the deflection plates, the required balance state is achieved, and no rf output will result. If now an audio modulating signal is applied across the deflection electrodes, the beam will be switched between the two plates, unbalancing the modulator, and resulting in an output which contains the required sum and difference frequencies. Note that the circuit shown here provides a balanced rf output from unbalanced rf and audio inputs.

Centring of the electron beam under quiescent conditions is achieved by means of a circuit which provides equal dc potentials to the two deflection electrodes, and provision is made for the balancing of these potentials. In the same way, provision is made for the balancing of the rf output with respect to ground.

Summary

We have now seen the reasons for single-sideband operation and the advantages to be gained. We have also discussed typical methods of producing sidebands without the carrier, or at least with the carrier suppressed to the extent where its amplitude is negligible. The next part of this article will deal with the methods of suppressing one of the sidebands, which will in turn be followed by a description of the methods used to recover the intelligence at the receiving point.

SQUARE-WAVE GENERATOR (Continued)

necting leads between the generator and the amplifier under test as short as possible. Remember that a square wave contains harmonics in considerable amplitude up to at least 20 times the fundamental frequency.

This means that when using the unit at say, 5 Kc, precautions suitable to operation at 100 Kc at least must be taken. For the more studious readers, two articles may be worth looking up. They are "Square Wave Analysis at Audio Frequencies," Audio Engineering, May 1947, and "Amplifier Testing by Means of Square Waves," Communications, February 1939. If there is sufficient demand, we will prepare and publish an article on the subject.

A word of warning must be given, and it is that a square-wave generator is a wonderful instrument for separating the good from the bad in amplifiers. Often an amplifier which sounds quite good, and seems good on paper, is shown to have several faults when tested with square waves. The instrument may show shortcomings in an amplifier that has been treasured for years. But if you are trying to do the best you can in the audio field, this unit cannot fail to help.

7587 NUVISTOR TETRODE IN RF AND IF APPLICATIONS

This article discusses the use of the 7587 nuvistor tetrode in small-signal high-frequency circuits. Input-admittance data are given for frequencies from 20 to 150 megacycles, and a small 60-megacycle wide-band amplifier is described.

Design Features

The all-metal-and-ceramic 7587 sharp-cutoff nuvistor tetrode uses a concentric cylindrical open-ended cantilever construction. The use of a top cap for the plate connection provides excellent input-to-output isolation, a low grid-No. 1-to-plate capacitance of 0.01 picofarad, and a low output capacitance of 1.4 picofarads. In addition, the valve construction provides extremely low residual-gas currents, freedom from buildup of leakage paths, and a rugged internal structure. The low heater drain of 150 milliamperes, the high transconductance of 10,600 micromhos at 10 milliamperes of plate current, and its small size make the 7587 particularly useful for general industrial and military application.

The bandwidth figure of merit GB for a tetrode is given by

$$GB = \frac{g_m}{2\pi\ (C_{in} + C_{out})}$$

The 7587 has a cold input capacitance C_{in} of 6.5 picofarads, a cold output capacitance C_{out} of 1.4 picofarads, and a transconductance g_m of 10,600 micromhos. Substitution of these values in the above formula produces a nominal figure of merit of 214 megacycles. However, the input capacitance for a valve in a socket under operating conditions is approximately 9 picofarads. Consequently, the actual figure of merit for an operating valve is 162 megacycles.

Input Admittance

When the current through a valve is varied, as it may be for the purpose of controlling the gain of an amplifier stage, variations in the input conductance and in the input capacitance affect the gain-vs-frequency characteristic of the circuit connected to the input of the valve.

Table I shows the values of short-circuit input capacitance and short-circuit input conductance of the 7587 for conditions of normal operation, cutoff, and with the valve cold. These values were measured in a socket at a frequency of 60 megacycles. The capacitance values are nearly independent of frequency up to approximately 150 megacycles. Theoretical considerations indicate that the conductance should increase with the square of the frequency. The measured data shown in the curve of Fig. 1 indicate a somewhat more rapid increase with frequency. This difference results from the series inductance in the measurement circuit.

Operating Condition	Input Capacitance (pf)	Input Conductance (μmhos)
Valve operating (I_b = 10 ma)	9.0	100
Valve cut off (I_b = 0)	7.7	18
Valve cold (no heater voltage applied)	7.1	17
Change from cutoff to I_b = 10 ma	1.3	82
Change when heater voltage is applied	0.6	—

Table I—Variation of Short-Circuit Input Capacitance and Input Conductance of the 7587 at 60 Megacycles.

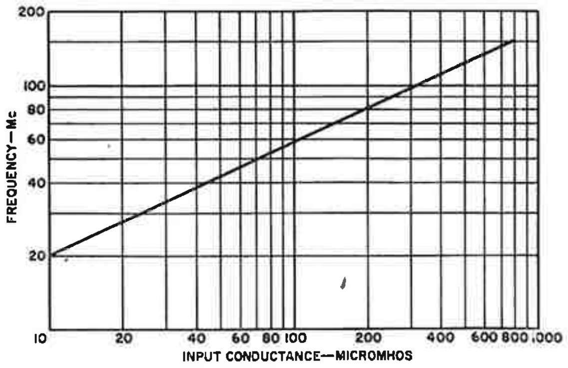


Fig. 1—Relationship between operating frequency and input conductance of 7587 at typical operating conditions (g_m = 10,600 μmhos; I_b = 10 ma).

The variation of input capacitance and input conductance with operating conditions can be reduced substantially if the valve is used in a circuit which includes an unbypassed cathode resistor. Figs. 2 and 3 show data for the 7587 measured in such a circuit. As shown in Fig. 2, the value of unbypassed cathode resistance which provides minimum change in capacitance between cutoff and maximum transconductance is about 18 ohms. The value which provides minimum change in conductance is about 33 ohms, as shown in Fig. 3. The circuit designer should select a resistance value between these limits which will have the smallest effect on the bandpass characteristic of his particular system.

The data shown in Figs. 2 and 3 were measured at a frequency of 60 megacycles. The choice of resistance values for optimum results is not affected appreciably by changes in frequency, but the magnitude of the conductance values varies approximately with the square of the frequency, as mentioned previously.

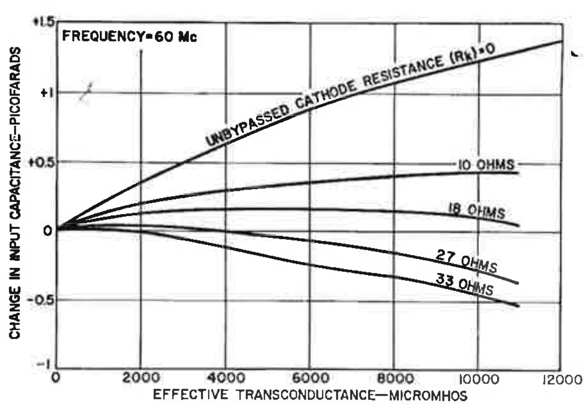


Fig. 2—Variation of short-circuit input capacitance at 60 megacycles as a function of transconductance for several values of unbypassed cathode resistance.

The amount of unbypassed cathode resistance needed to minimize the variation of capacitance and conductance with operating conditions is less than the total value of cathode resistance suggested in the published "typical operating conditions" for the 7587. Consequently, either an additional bypassed section of cathode resistor or an external source of bias is required. In high-gain circuits with automatic gain control, the initial biasing voltage developed from noise in the system under small-signal conditions is often sufficient to supply enough initial bias to prevent excessive plate and screen-grid currents.

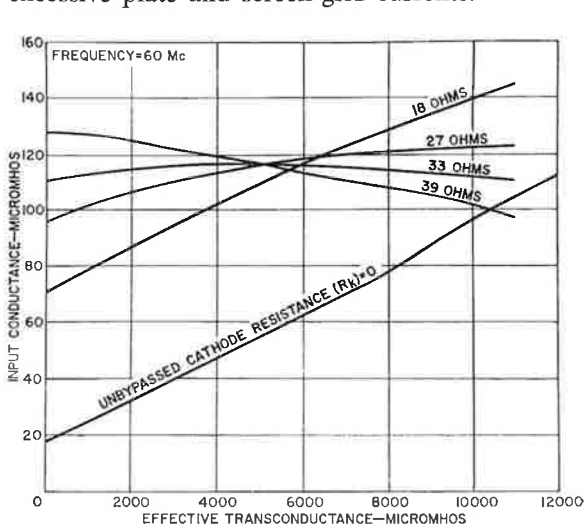


Fig. 3—Variation of short-circuit input conductance at 60 megacycles as a function of transconductance for several values of unbypassed cathode resistance.

IF Amplifier Design

The simple five-stage 60-megacycle if amplifier shown in Fig. 4 demonstrates the capabilities of the 7587. This amplifier consists of staggered single-tuned stages (a flat-staggered quintuple[1]), and has a bandwidth of 8 megacycles.

As shown in Fig. 5, the first stage consists of two 7586 nuvistor triodes in cascode arrangement, followed by four 7587 nuvistor tetrodes and a diode-connected 7586 triode used as a detector. The cascode input stage was chosen to take advantage of the low noise figure inherent in this configuration. The tetrodes are single-tuned and staggered. The first two tetrodes V_3

[1] G. F. Valley, Jr., and H. Wallman, *Vacuum-Tube Amplifiers*, 1st Ed., McGraw-Hill Book Company, Inc., New York, N.Y., 1948, pp. 180-186.

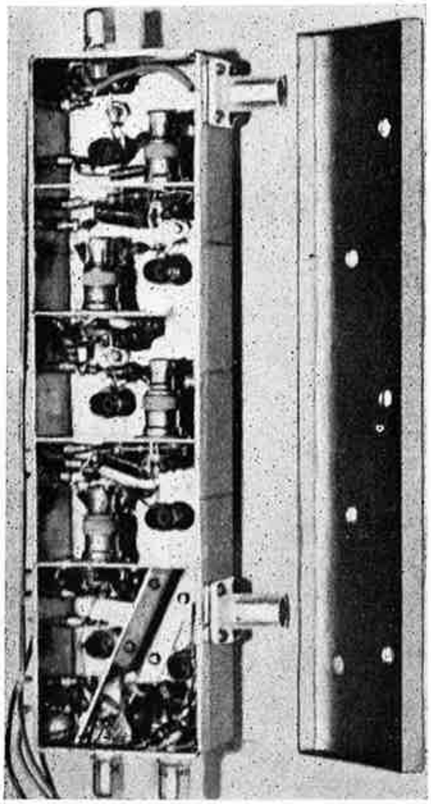


Fig. 4—Photograph of an experimental 5-stage 60-megacycle if amplifier using 7587.

and V_4 are gain-controlled. Unbypassed cathode resistors are mandatory to preserve the proper shape of the bandpass characteristic as the gain is varied. V_5 and V_6 have fixed gain; however, small unbypassed resistors have been added to provide some degeneration to improve stability.

The damping of each tuned circuit in a stagger-tuned amplifier is dependent upon the individual bandwidth and frequency required to achieve an over-all flat bandpass of desired width. In this amplifier, the plate-load resistor of the preceding stage and the input conductance of the tuned grid-No. 1 circuit are placed in parallel to achieve the proper bandpass. The value of damping resistance calculated for each stage can be only a first approximation because the short-circuit input-conductance data do not show the input-conductance component resulting from feedback through the grid-No. 1-to-plate capacitance. This feedback component, when measured at the grid-No. 1-circuit resonant frequency, is positive when the plate circuit is tuned to a frequency lower than that of the grid-No. 1 circuit and negative when the plate circuit is tuned to a frequency higher than that of the grid-No. 1 circuit.

Each stage is tuned by adjustment of its inductance for resonance with the valve and stray circuit capacitance. Because of the uniformity of nuvistor characteristics from valve to valve, a minimum amount of adjustment is necessary to retain the proper bandpass characteristics when valves are interchanged.

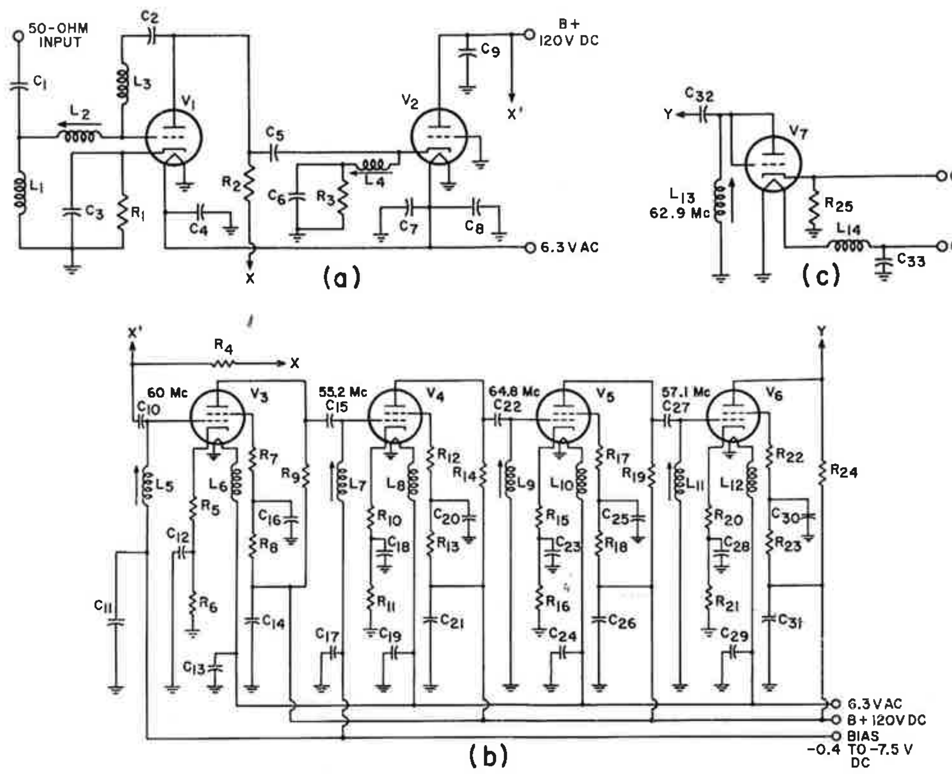
The effective grid-No. 1-to-plate capacitance of an rf amplifier valve is much higher than the value measured at low frequencies if the screen grid is not at rf ground potential. It becomes difficult to ground the screen grid effectively at frequencies above 30 megacycles because of the inductance of the screen-grid and bypass-capacitor leads. Unbypassed 10-ohm series resistors were used to "swamp out" any possible high-Q series resonances that might be caused by the bypass capacitor. The lead inductances can be adjusted to resonate in series with the bypass capacitor to effectively ground the screen grid.

The small size and double-ended construction of the 7587 simplify the circuit layout because the chassis can be made to act as a physical barrier between input and output. Experience has shown that high packaging densities (small over-all size) can be readily achieved with negligible instability provided proper bypassing and decoupling are used. The basic amplifier chassis shown in Fig. 4 measures 1-7/8 inches by 1-1/4 inches by 7-5/8 inches. The over-all length is increased slightly to 8-1/4 inches when the cascode 7586 nuvistors are inserted in one end. This unit is not as small as possible; other units using more valves have been built even more compactly.[2] In this circuit, however, each stage was "compartmentalized" to minimize the external feedback paths by taking advantage of the double-ended feature of the 7587. The plate, bias, and heater-voltage supply or bus lines are strung "outboard" along one side of the chassis.

Gain and Bandpass Measurements

The over-all amplifier gain of the circuit shown in Fig. 5 was measured with a 60-megacycle unmodulated signal applied to the input grid of the first stage. A dc voltmeter was connected across the detector load resistor and an rf vacuum-tube voltmeter was connected across the tuned inductance in the detector plate circuit. With the signal on, both rf and dc voltages were measured. The rf meter was then removed, and the level of the input signal was adjusted until the original dc voltage was obtained across the detector load. The rf output voltage was then divided by the adjusted input-signal voltage to calculate the over-all gain. The voltage gain of the 60-megacycle amplifier was 104,000; the average gain per stage was greater than 10.

[2] "Nuvistors Improve Performance of Beacon IF Strip," *Electrical Design News*, Vol. 5, No. 8, August 1960, p. 34.



C_1 C_2 C_5 C_{10} C_{15} C_{22} C_{27} C_{32} = 390 $\mu\mu f$
C_3 C_4 C_6 C_8 C_{12} C_{16} C_{18} C_{20} C_{23} C_{25} C_{28} C_{30} = 500 $\mu\mu f$
C_7 C_9 C_{11} C_{13} C_{14} C_{17} C_{19} C_{21} C_{24} C_{26} C_{29} C_{31} C_{33} = 1000 $\mu\mu f$, feedthrough
L_1 = 4.8 μh
L_2 L_4 = 0.5 μh, ferrite core
L_3 L_6 L_8 L_{10} L_{12} L_{14} = 5 μh
L_5 L_9 L_{13} = 8 turns No.24 AWG close-wound on 1/4-inch round form (ferrite core)
L_7 L_{11} = 9 turns No.24 AWG close-wound on 1/4-inch round form (ferrite core)
R_1 = 150 ohms
R_2 R_{14} = 4700 ohms
R_3 = 220 ohms
R_4 R_{24} = 1800 ohms
R_5 R_{10} = 27 ohms
R_6 R_{11} = 39 ohms
R_7 R_{12} R_{17} R_{22} = 10 ohms
R_8 R_{13} R_{18} R_{23} = 27000 ohms
R_9 = 6200 ohms
R_{15} R_{20} = 18 ohms
R_{16} = 47 ohms
R_{19} = 2200 ohms
R_{21} = 68 ohms
R_{25} = 18000 ohms

Fig. 5—Schemetic diagram for amplifier shown in Fig. 4: (a) 7586 preamplifier; (b) 7587 if amplifier; (c) 7586 detector used for measurements.

For measurement of the bandpass characteristics of the amplifier, the input-signal level was varied to maintain a constant dc voltage across the detector load throughout the passband. Fig. 6 shows the gain-bandpass characteristics of the amplifier of Fig. 5 for gains of 705, 11,100, and 104,000.

Maximum Operating Temperature

Good design practice for reliability and life requires consideration of the maximum operating-temperature ratings for all components. The 7587 can operate at a maximum shell temperature of 150 degrees Centigrade, measured at the gussets near the base of the valve. The 150-degree limitation permits chassis temperatures of up to 75 degrees Centigrade at full dissipation. Temperature measurements are made with a small thermocouple welded to a gusset of the nuvistor. The weld can be made by discharging a capacitor through the junction of the thermocouple and the valve shell. A 200-microfarad capacitor charged to about 75 volts welds a thermocouple wire having a diameter of 0.005 inch.

The 7587 is designed so that the metal shell, particularly the lug contact to the socket, conducts heat away from the valve. When printed-board or other poor-heat-conducting materials are

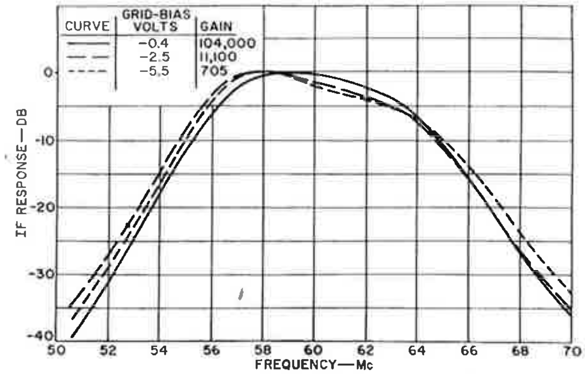


Fig. 6—Gain-bandpass characteristics of amplifier shown in Figs. 4 and 5.

used for the chassis, care should be taken to insure that the shell temperature does not exceed 150 degrees Centigrade. Operation at higher temperatures may increase grid current and impair life. When the valves are used at full dissipation on a printed board, the metallic portion of the chassis should be connected to a suitable heat sink.

(With acknowledgement to RCA)

A USEFUL POWER SUPPLY

A transistor power supply for use in hybrid and other equipments, this basic circuit will power transistorized units from the 6.3-volt heater supply.

Introduction

We have often been asked for a power supply unit to drive small transistorized units, sometimes as separate units, and sometimes in conjunction with valve equipment. Typical cases would be a self-powered preamplifier, or a preamplifier to be used with a valve amplifier as an add-on attachment. There are many similar cases.

This basic circuit has been developed to meet this need. It will supply up to 9 volts at currents up to 100 milliamps, and so will meet most of the requirements in this field.

The Requirement

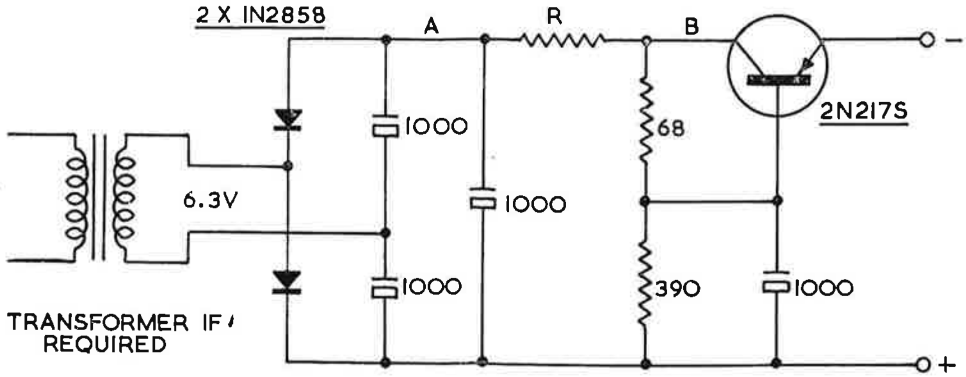
In considering the requirement for this circuit, two points were immediately obvious. One was that any valve unit will almost certainly have some spare capacity in the heater supply arrangements; the heater windings are rarely loaded right up to the ratings of the transformer. This seemed therefore to be a good solution to the primary source of power. Where self-powered units were required, then the 6.3 volts could be obtained from a small heater or filament-type transformer.

The second basic requirement was that the ripple voltage should be very low. It is no use trying to drive a preamplifier from a power source having a high ripple content.

As far as voltages were concerned, most of the units which would use this circuit require 6 or 9 volts supply, or at least something up to 9 volts. A maximum current rating of 100 milliamps was considered more than adequate.

The Circuit

The circuit chosen was a conventional voltage-doubler, operating with a 6.3-volt input to produce light-load voltages of the order of 17 volts at the rectifier output. A choice of quite high capacitors was made, both for the voltage-doubling units and for the main filter capacitor. The rest of the circuit consists of a series resistor R, of which more later, and a dynamic filter.



Circuit diagram of the power supply discussed in this article. See text for values of resistor R and application data.

The dynamic filter uses an AWV 2N217S transistor, with a sleeve-type heat sink. The maximum ratings of the power unit are within the ratings of the 2N217S when used in this way. If a small current only is required, an AWV 2N408 could be used at the saving of a few pence. This type of filter has now appeared in these pages several times; readers will remember that the device is in effect a capacitance multiplier system, in which the 1000 mfd capacitor connected between the base of the 2N217S and ground produces across the output of the unit an equivalent capacitor equal to the physical value of the base-to-ground unit (1000 mfd.) multiplied by the gain of the transistor.

It would in theory be possible to adjust the output voltage of the circuit by suitably adjusting the bias on the 2N217S. This raises the possibility of exceeding the dissipation rating of the transistor, and the circuit was therefore so arranged that the dissipation was always within the ratings with any variation of resistor R.

Resistor R may be adjusted to give the required output voltages, depending on the loading conditions. The accompanying table gives the value of the series resistor for four operating conditions. The value for intermediate conditions may be readily interpolated from these figures, at least within the limits of experimental accuracy.

Performance

The table already referred to also gives details of ripple voltage readings taken on the circuit, and other data. The table was set up for currents of 10 and 100 milliamps at output voltages of 6 and 9 volts. As one may expect, the ripple content in the output rises as the current drawn is increased. But the ripple content is still very low, being 72 db down on the output voltage in the worst condition (9 volts at 100 ma). At the 10 milliamp figure, the ripple is 86 db and 84 db down respectively at the 6-volt and 9-volt output figures. This is considered satisfactory for most applications.

Using the Circuit

To use this circuit, there are only two things to do, provide the 6.3-volt (nominal) input for the circuit, and adjust the series resistor R to a suitable value. The provision of the primary input power requires a word of warning. This

OUTPUT VOLTS	LOAD CURRENT	RESISTOR R	RIPPLE VOLTAGE	VOLTAGE A	VOLTAGE B
9	100 MA	22	2.4 MV	14.25	11.0
9	10 MA	100	0.5 MV	16.75	10.75
6	100 MA	75	1.0 MV	15.5	7.5
6	10 MA	250	0.3 MV	17.0	7.25

Table showing ripple and dc voltages for four operating conditions of the power supply, with values of resistor R.

circuit as shown here must not be connected directly into a valve unit where the heater supply is grounded. This almost inevitably means that there are only two alternatives when the circuit is used with a valve unit.

One of the alternatives is to use an unused heater winding on the mains transformer; this is a rare possibility. The other alternative, which is the one which will generally be used, is to take the heater supply through a small 1:1 ratio transformer to provide isolation. This will remove any possibility of shorting part of the heater supply through one of the rectifiers, and will allow us to ground either side of the output, as may be required.

A suitable transformer, having regard to the small current drain, can easily be made up using a discarded audio output transformer of about 2-watt rating or higher. A core of this size would probably require something of the order of 10 to 20 turns per volt, that is, 60 to 120 turns for both the primary and secondary windings. This is neglecting efficiency, which would typically be about 80%. This means that the secondary turns should be increased to perhaps 75 to 150 turns respectively. The transformer would be layer-wound with thin paper interleaving between windings, and a slightly heavier interleaving between primary and secondary. It is impossible to be specific on the question of this transformer because core sizes and materials vary so much; these remarks should therefore be regarded merely as an indication of what may be required.

Where this circuit is required for use with a self-powered unit, then the most likely source of primary power would be a small heater or filament transformer.

Many adaptations of this circuit are possible, but be careful not to exceed the dissipation rating of the 2N217S if experimenting. This rating is easily calculated by taking the product of the current drawn from the supply multiplied by the voltage drop across the transistor. For example, referring again to the table, with a 9-volt output at 10 ma, the input voltage to the 2N217S is 10.75 volts, so that the voltage drop across the unit is 1.75 volts. Multiplying this figure by 10 ma, we get the transistor dissipation as 17.5 mw.

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A demonstration of transistorized high-fidelity equipment will be given at the November meeting of the Sydney Division of the Institution of Radio Engineers (Aust.), to be held at Science House, Gloucester St., Sydney, on the evening of Thursday, November 29th. Details of the equipment demonstrated will be published in full in later issues of "RADIOTRONICS".

Editor .... **Bernard J. Simpson**

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