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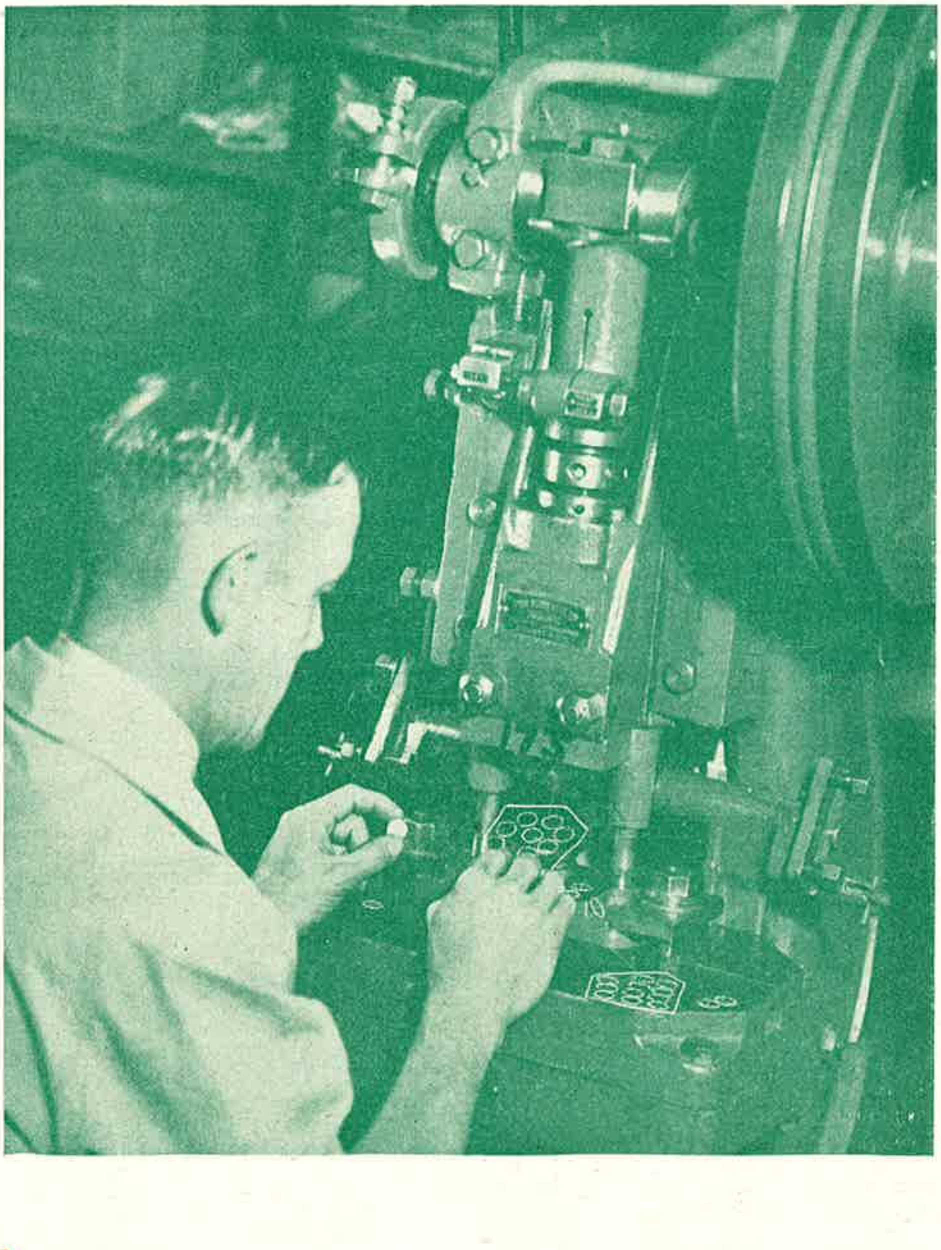
# RADIOTRONICS

Volume 17 August 1952 No. 8



An AWV Publication

PRICE 1/-

Registered at the General Post Office Sydney for transmission by post as a periodical.

# RADIOTRONICS

**Volume 17** **August 1952** **Number 8**

*By the way—*

Our cover illustration this month shows a press operator stamping out mica spacers used in valve assemblies.

One usually thinks of audio filters as devices with high-Q iron-cored chokes in them, and such components are expensive and the values critical. In this issue we present a simple peaked amplifier that uses only resistors and condensers, and yet will give you all the audio selectivity you can use. As a 1000 ∼ bridge amplifier it will prove particularly useful, also.

A number of enquiries have been received concerning the spiral bound Radiotron Valve Data Book, which was published last year. No copies of this book are available, but a completely revised second edition is expected to be published towards the end of this year. An announcement will be made later in Radiotronics when the second edition is printed.

Single issues of Radiotronics prior to 1952 are no longer available.

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Radiotronics is published twelve times a year by The Wireless Press for Amalgamated Wireless Valve Company Pty. Ltd. The annual subscription rate in Australasia is 10/-; in U.S.A. and dollar countries $1.25; and in all other countries 11/-. Price of a single copy 1/-.

Address all communications as follows:—

Amalgamated Wireless Valve Co. Pty. Ltd.,
Technical Publications Department,
G.P.O. Box 2516,
Sydney.

PRINTED BY CLOISTER PRESS (D. & W. SHORT), 45-49 GEORGE ST., REDFERN.

# Miniature Modulator

## General considerations

Where a medium-power audio amplifier in a minimum of space is required, such as a modulator unit for a mobile transmitter or a light-weight P.A. system, a combination of miniature valves used in a Class B amplifier is an attractive proposition.

Assuming a power supply of 300 volts, 40 mA is available, an investigation shows that the 12AU7, or alternatively a pair of 6C4's, will function quite satisfactorily as a Class B stage.

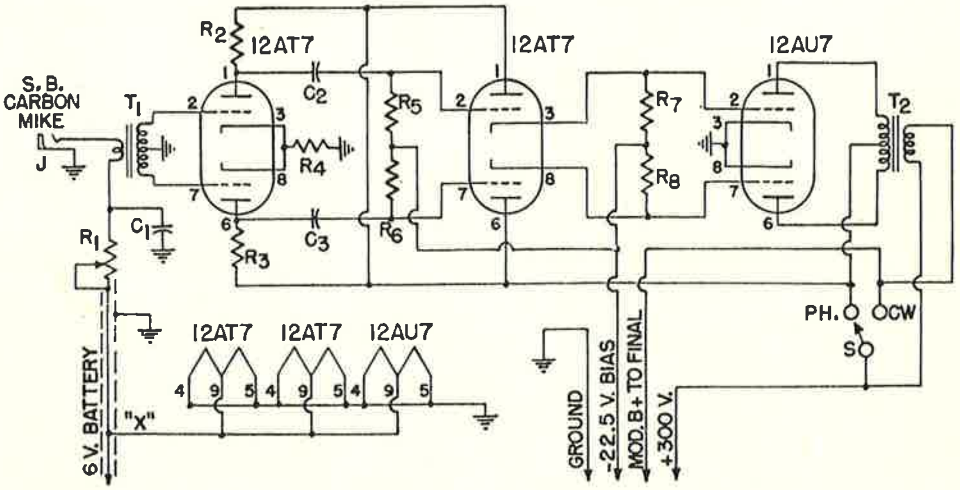
Strange as it seems, the 12AU7 will give a peak speech output of well over ten watts and, stranger still, at a distortion level well under that accomplished by a class-B operated 6N7, despite the fact that the 6N7 was originally designed for zero-bias class-B operation.

The net effect of this design is a high-quality modulator (including a voltage amplifier stage drawing less than 1 mA) that has a static drain of approximately 20 mA.

## Electrical details

With reference to the circuit diagram, Fig. 1, it will be noted that the entire modulator is push-pull throughout. Inasmuch as the class B stage and driver must be push-pull, it was deemed desirable to carry this through to the input circuit in the same fashion, to avoid a phase inverter and to simplify construction. Note that only three condensers and eight resistors are used in the entire unit.

A bias battery is specified in order to provide the proper grid bias voltage for the 12AU7 modulator and the 12AT7 driver. Under zero-signal conditions the bias voltage from either pin 2 or 7 of the 12AU7 to ground will be 15 volts, and the voltage measured across either $R_7$ or $R_8$ (the bias for the 12AT7 driver) will be 7 to 8 volts, when a 22.5 volt bias battery is used.



**CIRCUIT CONSTANTS**

(All resistors and capacitors ±20% tolerance unless specified otherwise)

| | |
|---|---|
| $C_1$ | 500 mf 15 volt electrolytic |
| $C_2$, $C_3$ | 1000 mmf 500 volt ceramic or mica |
| J | Open-circuit jack |
| $R_1$ | 250 ohm potentiometer |
| $R_2$, $R_3$ | 0.1 megohm, ½ watt |
| $R_4$ | 2200 ohm, ½ watt |
| $R_5$, $R_6$ | 0.47 megohm, ½ watt |
| $R_7$, $R_8$ | 10,000 ohm, ½ watt |
| S | SPDT toggle switch |
| $T_1$ | S.B. mike to push-pull grids |
| $T_2$ | Output transformer (see text) |

*Fig. 1. Circuit diagram of the Miniature Modulator.*

The static (resting) current of the 12AU7 in class B with 300 volts on the plate is approximately 15 mA.

Further economies in both current and weight can be realized in the driver stage by employing a device already well-known to readers of Radiotronics. By using a cathode-coupled driver (see Radiotronics, May, 1951, P. 99) no driver transformer is required and the driver itself adds only another 5 mA drain to the power supply.

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Note that the cathode current for the 12AT7 driver flows through the bias battery, and therefore this battery actually supplies a current in the order of a few milliamperes. In other words, the current does not tend to charge the battery, as in the usual bias case, but instead, tends to discharge it.

However, this current is so slight that normal shelf life may be expected from the battery. This battery has no drain on it during stand-by or complete off periods, as current is drawn from it only when high voltage is applied to the modulator.

The first 12AT7 tube acts as a push-pull voltage amplifier. Because carbon microphones have a wide variation in output voltage, this first stage was added so that adequate gain would be available regardless of the microphone used.

Voltage for the microphone is obtained from the car battery, and a single lead is used to provide filament voltage and microphone voltage. This lead should be made of heavy wire to avoid ohmic loss due to filament current, and it should be shielded to prevent undue noise pickup.

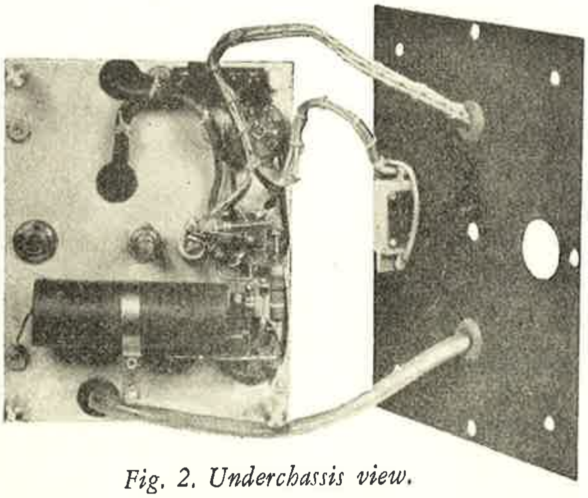

*Fig. 2. Underchassis view.*

Poteniometer $R_1$ (actually connected as a rheostat) serves as a gain control. Because it can only change the microphone current a small amount, it does not have a wide range of control, but it is useful for adjusting the level when different people use the microphone.

If the microphone has too much gain, it will be necessary to increase the value of $R_1$, or add a fixed resistance in series.

A phone-cw switch is provided which removes all high voltage from the modulator and shorts the secondary of the output transformer when the switch is in the c-w position.

An external switch must be provided to turn the filament circuit on and off. With the circuit shown this switch will also shut off the mike current. Some microphones incorporate switch contacts which may be used to control a relay for power switching. There are many possible control schemes and the refinements of the control system are left to the individual.

## Constructional details

The general nature of the mechanical work is shown in Figs. 2, 3, and 4. All of the parts, with the exception of the switch, are mounted on a piece of flat metal measuring $4\frac{1}{8}$ by $5\frac{1}{8}$ inches. The spacers which support this piece are $1\frac{1}{4}$ inches long.

Fig. 4 indicates how the parts are mounted on the flat chassis and Fig. 2 shows the wiring on the underside of the chassis.

The shaft on resistor $R_1$ is left long enough so that it projects through the front panel. The input jack is mounted on the chassis and a large hole cut in the front panel so that a mike plug can pass through. The switch is mounted on the front panel and the leads going to it are left a little long, as shown in Fig. 2, so that the chassis can be removed easily from the front panel.

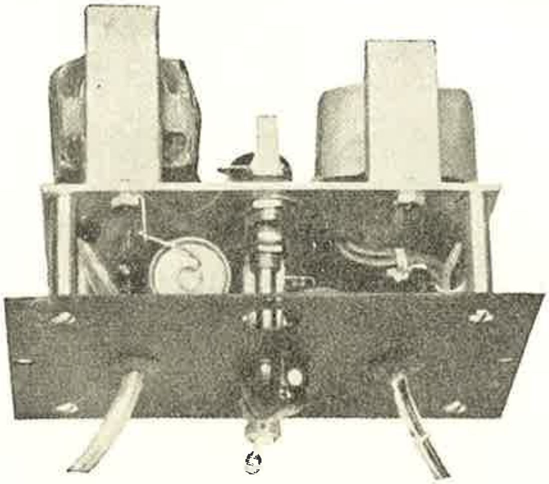

*Fig. 3. Uncased top view.*

The front panel is one of the removable 5 by 6 inch sides of a standard 4 by 5 by 6 inch cabinet.

Referring to Fig. 2, the input 12AT7 is the bottom tube, the 12AT7 driver is the middle tube, and the top tube is the 12AU7 output tube. These same tubes can be seen in Fig 4, and the order of the tubes is the same, looking from right to left.

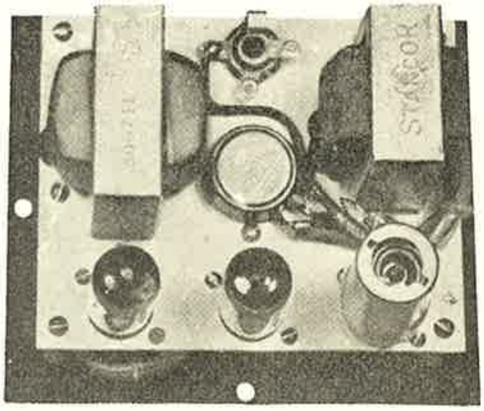

*Fig. 4. Uncased rear view.*

Note that the input 12AT7 uses a shield. Also in Fig. 4 transformer $T_1$ is on the right and transformer $T_2$ is on the left.

## Components parts

There are no critical components used in the mobile modulator and all parts may be plus or minus 20%, as indicated under circuit constants.

One part is worth discussing in more detail, however, and that is the output (modulation) transformer. Fundamentally, all that is required is a transformer with a primary plate to plate impedance of approximately 12,000 ohms and a suitable secondary impedance.

The prime considerations in choosing an output transformer for the mobile modulator are size, weight, efficiency and cost. A designer's concern over size, weight and cost is obvious, although concern over efficiency might not be.

If a transformer has a loss of 3 db (and this is not unusual) then one-half of the audio power is lost in the transformer. In other words, if 12 watts could be obtained from the tubes in a modulator stage, then only six watts would be available out of the transformer. This means you have only a six, not a twelve watt modulator.

In class B systems another important but frequently overlooked consideration is that of the design of the transformer itself. An improperly designed transformer can contribute a large amount of distortion to the output signal. While the efficiency depends upon the primary to secondary coupling, the distortion is controlled largely by the tightness of the coupling between the two halves of the primary winding.

Obviously, any transformer of the proper impedance and power rating will serve, within the limitations mentioned, as $T_2$. Some mechanical rearrangement may have to be made, depending on the size of the transformer selected.

### Testing

There is very little that need be done when the unit is finished. As mentioned previously, it would be wise to check the bias values, and a meter reading of the resting current would also be advisable.

Do not attempt to test the modulator with signal input unless it is connected to the final, or unless a dummy load is used. A 10 watt resistor across the secondary of the output transformer will serve as a dummy load.

### Other uses

Even though the mobile modulator has been designed for mobile service primarily, it will make an ideal modulator for emergency work. The power drain is small and the unit is compact and reliable.

This modulator may also be used in the home station if a change is made. Wire "X" should be disconnected from the hot lead so that the filaments may be energized by a 6.3 volt transformer. The hot lead can then go to a small 4.5 or 6 volt battery which will supply mike current.

Regardless of the use for which it is built, this high quality little modulator should find many uses around the shack.

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## A Peaked Audio Amplifier for Communications Receivers

**By G. D. Hanchett, Jr.,* W2YM.** *** C/o Tube Department, Radio Corp. of America, Harrison, N.J.**

The amateur bands are so crowded these days that a simple receiver without a narrow-band filter is no longer satisfactory for reliable c.w. amateur communication. Most modern receivers are equipped with crystal filters, but these are rather expensive and difficult to construct. For those amateurs who would like to obtain selectivity approaching that of a crystal-filter c.w. receiver, but at low cost, the peaked audio amplifier to be described is offered as a simple solution. When this amplifier is properly peaked, the bandwidth can be made as narrow as that obtained with the average crystal filter. The amplifier can be added to any type of receiver, regenerative or superheterodyne. The improvement is likely to be more marked following a regenerative receiver, although a great improvement can be made on a superheterodyne that has normal i-f selectivity. The single-signal feature of high i-f selectivity cannot be obtained with this audio filter (or any other), but it will still provide a marked improvement in reception.

The fundamental circuit upon which this amplifier is based is a simple twin-"T" resistance-capacitance bridge, as shown in Fig. 1-A. A bridge of this type has a null at a frequency $f_o = 1/2\pi RC$. If the bridge is connected to an audio amplifier, as shown in the block diagram of Fig. 1-B, a negative feed-back takes place at all frequencies except the frequency to which the bridge is tuned. Under these conditions, the audio amplifier has practically no gain at frequencies other than the null frequency the gain of the amplifier is maximum. By controlling the amount of negative feed-back, varying degrees of selectivity can be obtained.

The schematic diagram of the amplifier is shown in Fig. 2. The first tube of the amplifier is a 12AU7 (or 6SN7-GT). One triode section of this is used as an audio amplifier, and the other half is used to amplify the feed-back voltage. To isolate the headphones from the amplifier, a 6C4 (or 6J5-GT) output stage is used. If this output stage were not included, the impedance reflection from the headphones would load the amplifier and affect its characteristics. Since most receivers already have sufficient gain in the audio system, no additional gain is required, and the 6C4 is connected as a cathode follower.

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The capacitor $C_7$ is shunted by an adjustable mica trimmer, $C_8$, that is used to adjust the bridge for the proper null condition. Mica condensers, or other condensers with good power factors, are used in the bridge circuit because the ultimate selectivity depends upon the quality of these components.

## Construction

The amplifier can, of course, take practically any form. The unit built by the writer and shown in the photographs was mounted on a small metal chassis. A small heater-supply transformer was included, but the plate voltage was taken from the receiver. The plate drain of the unit is only a few milliamperes, and almost any receiver can stand this additional load without damage. The two tubes are mounted on the top of the chassis, and the selectivity (feed-back) control is brought out to the front of the chassis. The components given in Fig. 2 are for a peak frequency of 1000 cycles, selected because most headphones have a resonant frequency in this range. Other peak frequencies could be obtained, of course, by using other values derived from the equation of Fig. 1-A.

After the unit is connected to the output of a receiver, the selectivity control, $R_8$, should be adjusted for maximum feed-back (farthest from ground) and the amplifier peaked by adjusting $C_8$ for maximum "sharpness". Tune the receiver to a frequency on which there is no signal and increase the capacitance of $C_8$ until the amplifier tends to oscillate or "ring" from noise peaks. For the final setting, back off on $C_8$ until there is no tendency to oscillate. In the circuit shown, a slightly larger value for the combination of $C_7$ and $C_8$ is used than is theoretically indicated by the null-frequency formula. This increase makes the amplifier somewhat regenerative and, therefore, improves the selectivity of the amplifier. Tubes with various amplification factors were tried, but it was found that tubes with medium $\mu$ (about 20) gave the best results. Using higher-$\mu$ tubes ($\mu = 50$ to 100) made the system tend to oscillate.

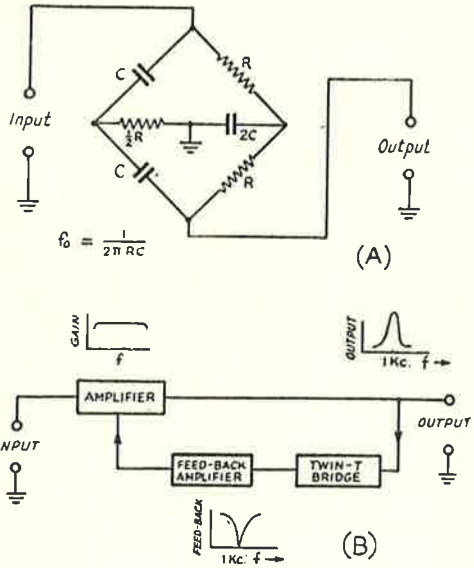

*Fig. 1.—A twin-"T" resistance-capacitance bridge is shown at A. No signal of frequency $f_0$ can pass from input to output when the components are properly matched.*

*A practical peaked amplifier can be built around the twin-"T" bridge by using the arrangement shown in B. The amplifier with a flat frequency characteristic will have a peaked characteristic centred on the null frequency of the twin-"T" bridge, because the amplifier is made highly degenerative at all other frequencies.*

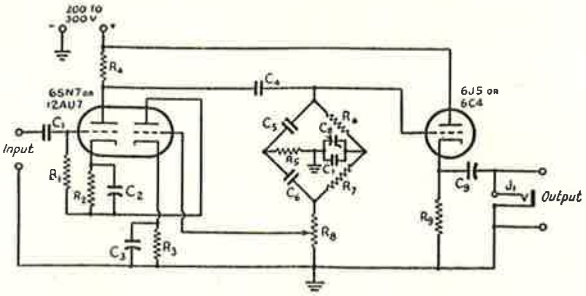

*Fig. 2.—Wiring diagram of the selective audio amplifier.*

**$C_1$, $C_4$ — 0.01-$\mu$fd. paper.**
**$C_2$, $C_3$ — 25-$\mu$fd. 25-volt electrolytic.**
**$C_5$, $C_6$ — 680-$\mu\mu$fd. mica.**
**$C_7$ — 1000-$\mu\mu$fd. mica.**
**$C_8$ — 600- to 1600-$\mu\mu$fd. adjustable mica trimmer.**
**$C_9$ — 0.1-$\mu$fd. paper.**
**$R_1$ — 1.0 megohm.**
**$R_2$, $R_3$ — 1500 ohms.**
**$R_4$ — 56,000 ohms.**
**$R_5$ — 0.1 megohm.**
**$R_6$, $R_7$ — 0.22 megohm.**
**$R_8$ — 2.0-megohm potentiometer.**
**$R_9$ — 10,000 ohms, 1 watt.**
**$J_1$ — Open-circuit jack.**
**All resistors $\frac{1}{2}$ watt unless otherwise specified.**

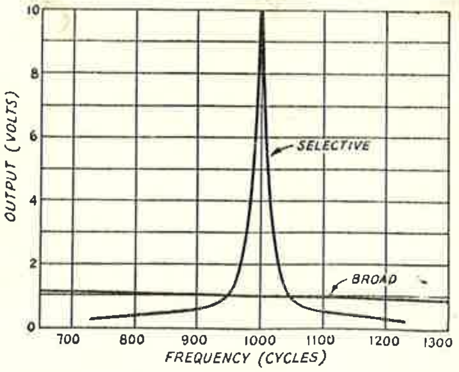

*Fig. 3.—The output of the amplifier for a 0.4-volt input signal. The gain and selectivity of the amplifier increase markedly as the selectivity control is advanced. At the maximum selectivity position the bandwidth is about 20 cycles.*

Fig. 3 shows the selectivity measured at minimum and maximum selectivity settings. The amplifier is reasonably flat over the entire audio range with the selectivity control at minimum, but it will sharpen up to a bandwidth of 20 cycles or so in the "maximum" position.

# BROADCAST AUDIO WIRING PRACTICE

**W. E. Stewart,**

**Manager, Broadcast Audio Engineering Section.**

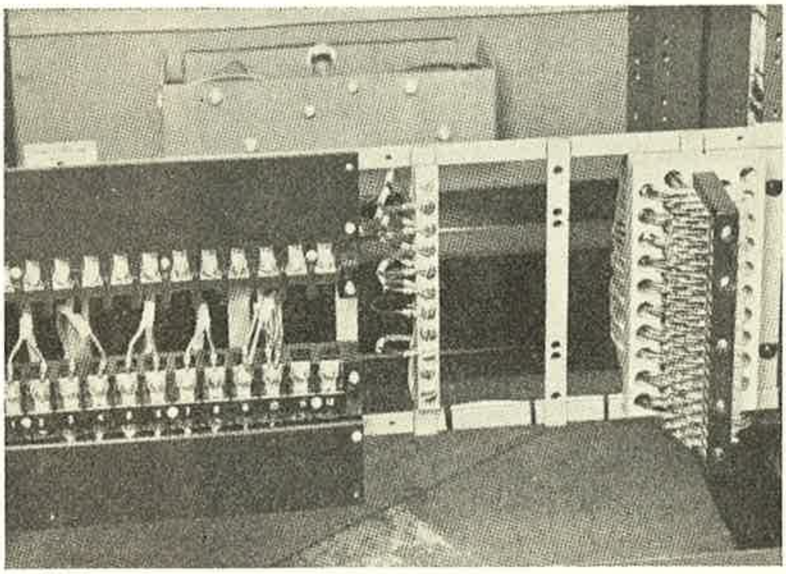

*Fig. 1. Photo of terminals at bottom of rack. Power terminals are at left, ground bus in centre and audio terminals at right.*

Almost every studio installation undergoes minor modifications from time to time, and the subject of proper installation practice is raised. Modern standards require careful elimination of noise and crosstalk from the programme circuits. It is not uncommon to spend many hours wiring in new components, only to find their performance reduced by the wiring itself. A tested and proven standard practice can avoid much wasted time.

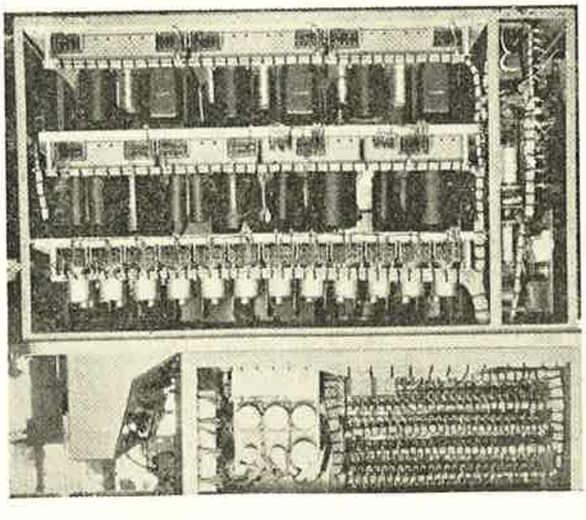

*Fig. 2. View of wiring in a control desk. A.C. circuits are below the shelves, and audio above.*

There are two basic philosophies employed in practical approaches to the noise problem. In one system every circuit shield is carefully isolated from its neighbours and grounded at one point only. In the other, all the shields of one unit (such as a rack) are put in such close contact that a brute-force ground is provided for any stay currents that might be present. This latter approach is taken in RCA equipment with modifications as follows:—

Every rack, cabinet or desk is wired as a unit to terminal boards. The terminal boards are placed as near as possible, consistent with accessibility, to the point where the external circuits enter the unit. See Figs. 1, 2, and 3 for examples.

In a rack, as viewed from the back, all audio cables are run on the right side of the rack; and all signal, a.c. and d.c. power cables are run on the left side. All audio circuits are twisted pair conductors shielded with a tinned copper braid. Separate cables are formed for:

(a) Microphone outputs, preamplifier outputs and other audio circuits with levels below — 20 VU.
(b) Mixer, line and channel circuits up to ± 30 VU.
(c) Loudspeaker and other lines above ± 30 VU.
(d) At times further subdivisions are made for convenience in bulk or because levels are widely separated.

Each cable is bound with lacing cord so the shields are in tight contact for their entire length. Where two audio cables cross or join, they should either be definitely insulated or bound together. It is better to have tight contact than to risk an intermittent noise source made by casual contact.

The ends of the individual shields are terminated either with "wedge-on" collars or with plastic tape. The shields are grounded to a main ground bus near the terminal block. A shielded ground lead is run from each amplifier chassis to the ground bus.

The a.c. and d.c. power circuits are handled similarly. All a.c. circuits should be in twisted pair, shielded cable. The a.c. current should be balanced in each pair. That is, one pair should not be used for one side of a circuit and a second pair for the other side. If more than one pair is needed for the load, two or more pairs should be used with part of the load on each.

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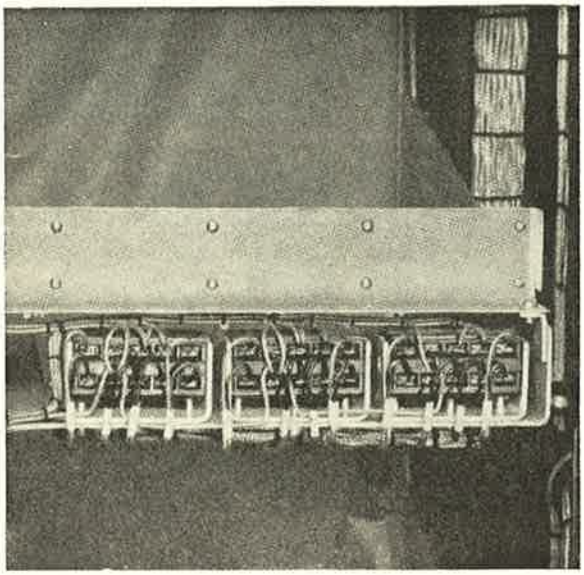

*Fig. 3. Typical plug-in shield wiring. Cover will be turned down to protect wiring after circuits are checked.*

Plus and minus plate potentials should be carried in single conductor shielded cable. Shields are tied off and grounded the same as the audio circuits.

Signal circuits do not require shielded wire.

The frames of jacks should be tied together and grounded with a shielded wire the same as amplifier chassis.

In installing the equipment in a studio or control room the following rules have been found useful:

The pairs run in conduits should be grouped in the same general way as the cables in the racks. The audio conduits should be kept free from grounds to power conduits or power circuits. Low level audio circuits (less than –30 VU) should have the shields insulated from the conduits and from each other. Splices should be avoided. Low level conduits should be well spaced from power conduits.

Signal and telephone circuits should not be run in the same conduit with programme or power circuits. Telephone leads should be twisted pair. Power and audio grounds should consist of separate, heavy shielded leads to the main station ground.

TV circuits in general should be considered high level circuits and should therefore be kept away from low level audio circuits. In particular, pulsed lamp circuits should be routed as far away from projector photocell and preamplifier circuits as possible. Shields should be insulated from ground and the audio circuit and shield grounded only at the point of lowest level.

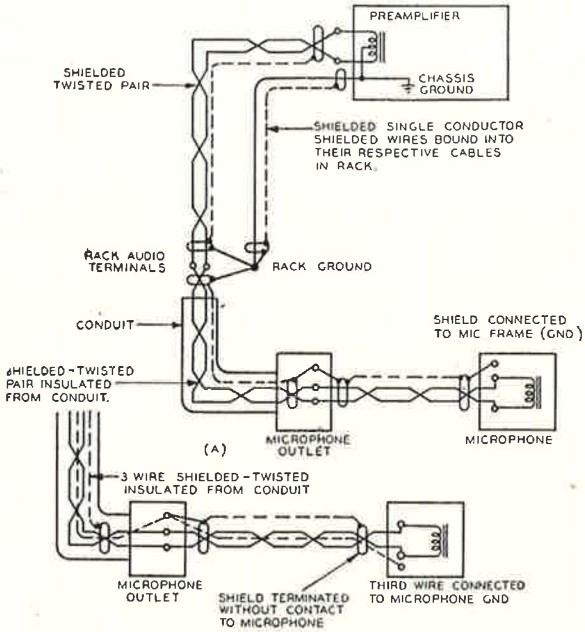

*Fig. 4. Sketch showing typical microphone grounding practices.*

Typical good practice for microphones is shown in Fig. No. 4a. In this case two conductor shielded wire, with insulation over the shield, is used for the conduit run and the microphone cord. Fig. No. 4b shows somewhat better practice in which 3-conductor shielded, insulated cable is used for the conduit run and microphone cord. This latter practice removes any ground current from the shield.

Turntable pickup circuits should be handled like microphones, with particular care being taken to

*(Continued on page 135)*

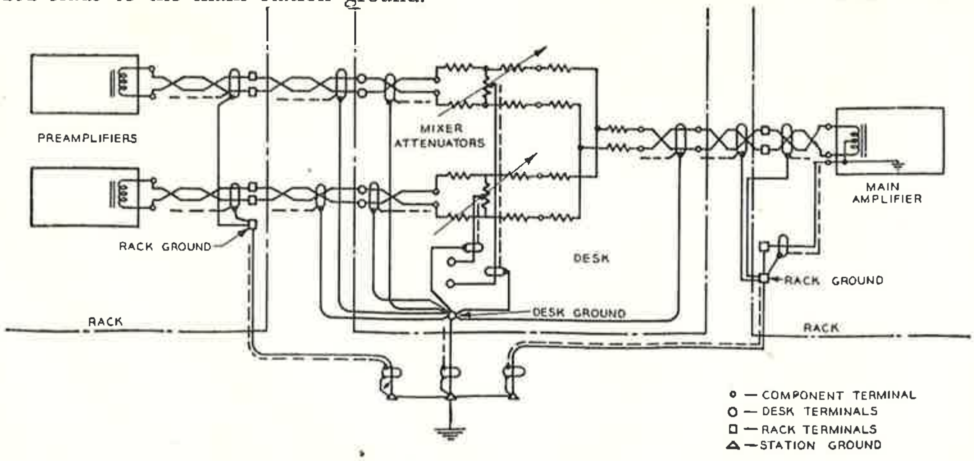

*Fig. 5. Diagram showing good grounding practices for inter-connections of preamplifier and mixer circuit.*

# INEXPENSIVE SQUARE-WAVE GENERATOR

By G. W. Gray,

RCA Laboratories Division, Princeton, N.J.

A cathode-controlled multivibrator employing three tube envelopes gives square waves from 50 c/s to 1 Mc/s with only 2-percent. tilt in the negative half. An additional circuit is described in which a pair of pentodes replaces cathode resistors to produce square waves.

In recent years there have been many articles in the literature concerning the use of square-wave techniques for testing amplifier circuits; however, no simple and inexpensive circuit for producing good-quality square waves is shown.

In producing such a generator it would appear that the use of clipper stages following a sine-wave oscillator is ruled out, since clipping at the bandwidth required for television testing requires several stages with carefully adjusted peaking coils and biasing. Thus a multivibrator-type square-wave generator is indicated. However, ordinary multivibrators show some serious shortcomings, the most serious of which is the grid current they draw.

This grid current is required since it is, by rapidly charging a capacitor with grid current and then slowly letting it discharge, that the frequency of oscillation is determined. Since the grid current is produced during the switching action it slows down the speed of operation of the circuit just the same as shunt capacitance to ground would do. In other words, it matters very little whether current must be put into a shunt capacitance to cause a voltage change or whether a series capacitance must be charged with grid current. Thus, it is difficult to design ordinary multivibrators with fast enough switching actions to produce square waves suitable for testing television-type amplifiers.

Another effect of such grid current is the production of overshoot on the negative-going portion of the square wave at the plate. It is caused by grid current flow as the conducting tube momentarily draws more than zero bias current when its grid is driven positive. This overshoot must then be removed by a clipping stage since it appears much like the square-wave response of certain types of overcompensated amplifiers. A further shortcoming of conventional multivibrators is that to change the frequency and keep the square wave symmetrical it is necessary to change two capacitors in the circuit. The switching problem is cumbersome because neither side of the capacitors is customarily grounded

## Cathode-controlled multivibrator

A cathode-controlled multivibrator circuit has been developed that overcomes these difficulties. In the circuit shown in Fig. 1, each plate is coupled to the other tube's grid by a clamp circuit such that the positive peak of any recurrent waveform is clamped at zero bias. This coupling circuit must have a time constant long compared with the period of oscillation desired, but otherwise has no effect on the period. The gain from grid to plate of each tube is less than unity because, as will be shown, to produce good-quality square waves it is necessary to have the cathode resistor considerably larger than the plate resistor.

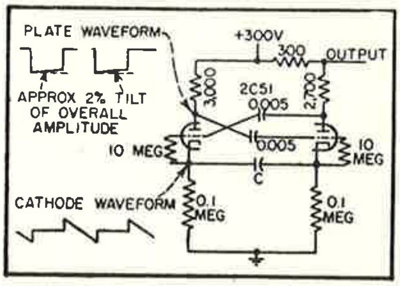

*Fig. 1.—Single-tube (double-triode) square-wave oscillator circuit.*

If capacitor *C* between the cathodes is removed, the circuit will not oscillate because even though the phase of the feedback is regenerative the loop gain is less than unity. However, when capacitor *C* is added, another loop is completed as follows. If the cathode of one tube goes slightly negative, the signal is coupled to the other cathode by *C*, which in turn produces an amplified negative signal on its plate. This signal is coupled to the grid of the first tube, which drives the cathode even more negative. This loop may easily have enough gain to cause oscillation. In addition, the grid-plate loop aids the regeneration even though by itself it could not cause oscillation.

Figure 1 shows the plate and cathode voltage waveforms as a function of time. The grid waveform of each tube is exactly the same as that of the plate to which it is coupled since the coupling circuit has a long time constant. From these waveforms it can be seen that the period of oscillation is determined by capacitor *C* discharging through the cathode resistor of the cut-off tube until the cathode reaches a low enough voltage to start

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conducting. Now the regeneration of the circuit is such as to turn the cut-off tube on and the conducting tube off.

This switching can occur rapidly because there is no grid current drawn to slow down the switching action and there are two positive feedback loops both acting in the same direction. After the switch has occurred, the off tube is doubly cut off since its grid has been driven negative by the plate of the conducting tube and its cathode has been carried positive by capacitor *C* coupling the positive impulse from the cathode of the conducting tube. Capacitor *C* now discharges until the cathode of the cut-off tube becomes low enough to start conduction, at which time the circuit flips over and the same cycle repeats again.

As may be seen in the oscillogram of plate voltage, there is a slight amount of tilt in the negative half of the square wave. This is due to a variation in the discharge current through capacitor *C* as the voltage varies across the cathode resistor of the cut-off tube. Since this current must come from the cathode of the conducting tube, the variation shows up as a slope in the voltage at that point. The grid of the conducting tube is at a fixed potential so the change in cathode potential that results from the change in current from the cathode results in an even greater slope on the plate voltage.

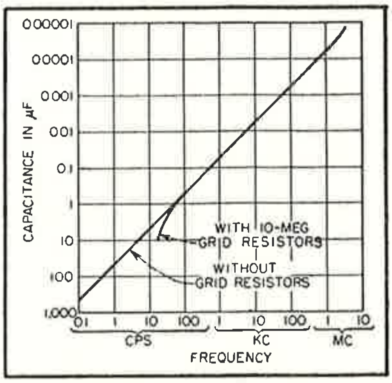
To minimize this effect the cathode resistor should be made as large as possible compared to the cathode self-impedance of the conducting tube. The major term in the cathode self-impedance of the conducting tube is a factor $(1/g_m) + (R_L/\mu)$, where $g_m$ is tube trans-conductance, $\mu$ is amplification factor of tube and $R_L$ is impedance in plate of tube.

If the cathode resistor is made too large the $g_m$ of the tube is lowered excessively because the plate current of the tube is reduced too much and the ratio between cathode resistor and cathode self-impedance is reduced. Optimum cathode resistor size is indicated by minimum tilt in the negative half of the square wave. With a 2C51 tube it is possible to reduce the tilt to 2 percent. of the overall square-wave amplitude while with a 12AT7 (as a suggested Australian alternative) tube the tilt will be 5 percent.

## Frequency control

The frequency of this cathode-controlled multivibrator as a function of capacitor *C* is shown in Fig. 2. With the 10-megohm resistors from grid to cathode, the time constant of the coupling circuit is only 1/20 second so that at about 60 cycles per second the curve departs from linearity. This range of linear relation between frequency and capacitance may be increased by removing the 10-megohm grid resistors since they are really not needed. Leakage through the cathode-coupling capacitors and gas current from the tube will both tend to make the grid go positive so that the circuit is effectively the same without the resistors except that the time constant is much longer. This effect is shown in Fig. 2 where, without the grid resistors, the curve extends linearly much further.

If it is necessary to have the negative half of the square wave flatter than is obtainable with the circuit of Fig. 1 the cathode resistors may be replaced by pentodes as shown in Fig. 3. With this circuit the discharging of the frequency-determining capacitor *C* must be a constant-current discharge since a pentode is a constant-current device. With the discharge current constant the plate current of the conducting tube is a constant and thus there is no tilt in the negative half of the output square wave.



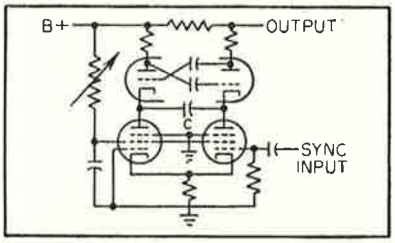


*Fig. 2.— (Top) Calibration of frequency versus capacitance of C in Fig. 1.*

*Fig. 3.— (Bottom) Three-tube version of the square-wave oscillator in which pentodes replace cathode resistors. This circuit also provides for synchronizing.*

This circuit has the additional advantages that: (1) a synchronizing pulse may be inserted on the grid of one of the pentodes; (2) changes in the plate current of the pentodes will produce small changes in frequency; (3) if the plate current of the pentodes is made unequal it is possible to make the square wave quite unsymmetrical so that the circuit becomes a pulse generator.

This cathode-controlled multivibrator circuit overcomes the difficulties inherent in more conventional circuits because no capacitor is required to change charge during the switching cycle, except for the unavoidable shunt-wiring capacitances. The circuit is consequently fast in action and has no undesirable transients such as over-shoot. Also the problem of changing frequency is reduced to changing one capacitor to vary the frequency range by a factor greater than ten million.

Figure 4 shows the circuit of a complete square-wave generator. In order to provide a low-impedance output at a reasonable level a cathode-follower output is used. To minimize the load changes on the power supply, two cathode followers driven out of phase are used so that the current drawn by the two is more nearly constant. This is necessary since it is virtually impossible to filter a power supply for low-frequency square waves by means of capacitors, yet for this small generator an electronically regulated supply is an unwarranted complexity. The extra output can be differentiated and is handy as a synchronizing signal that is not affected by the setting of the gain control. Although the rectifier tube, a 6X4, is nominally rated for 6.3-volt heater the power transformer used has only a 5-volt winding for the rectifier. Owing to the low-current requirement of the generator this has proved adequate.

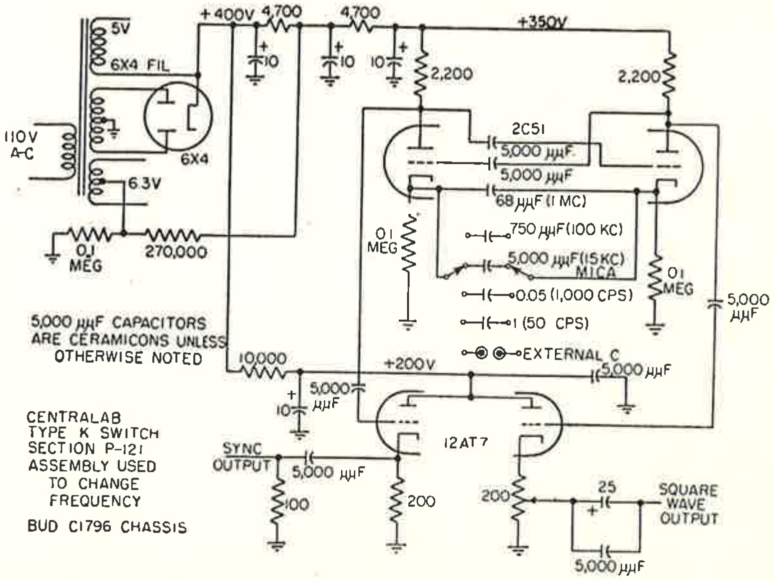


*Fig. 4.—Complete square-wave generator circuit. Balanced cathode-follower output (12AT7 tube) equalises power supply drain and simplifies filtering.*

The total power consumption for the complete generator is about 20 watts at 120 volts input, the overall size is 4 × 5 × 6 inches, the output amplitude is 3 volts across 200 ohms, and the frequency range is from about 30 cycles to 1 megacycle. Any of five predetermined frequencies is available directly by means of the switch on the front panel. Any other frequency may be obtained by connecting the proper size capacitor across the two binding posts and turning the frequency switch to the external-capacitor tap. To find the proper value of capacitor for the frequency desired a calibration chart can be attached to the front panel of the square-wave generator.

Since the calibration is so nearly linear, as given in Fig. 2, one decade of the graph may be expanded in order to obtain increased accuracy. For any other frequency range it is only necessary to multiply the capacitance scale by the inverse of the factor of ten by which the frequency scale need be multiplied.

To make the calibration chart for any particular square-wave generator, it is only necessary to measure the frequency resulting from the use of a known value capacitor. This will give one point for a graph that is logarithmic in both directions. By drawing a straight line with a slope of minus one on the graph the proper calibration will be obtained. In passing, it might be noted that this high degree of linearity between capacitance and frequency offers the possibility of making a wide-range capacitance meter with this circuit.

Oscillograms show that no over-shoot or other undesirable transients are present. Also the rise time may be measured and is approximately 0.05 micro-second, which is fast enough for testing television video amplifiers. Since the multivibrator does not change the charge on any capacitor while switching, the rise time is virtually independent of frequency instead of being some fraction of pulse duration as is the case with most multivibrators. To produce very low frequencies it is necessary to use large capacitances obtained practically only from electrolytic capacitors. Since the polarity of the voltage across the capacitor reverses, it is necessary to use two capacitors with their negative ends connected together and the two positive ends connected across the cathodes of the multivibrator.

The square-wave generator described is adequate for most purposes. The major limitation is probably the lack of means for synchronizing from some external source. If this feature is required too, it may be added by using the circuit of Fig. 3.

## BROADCAST AUDIO WIRING PRACTICE

*(Continued from page 132)*

keep the motor power circuits and their shields away from the audio circuits.

The input to mixer circuits is usually at comparatively high level, but the output is frequently very close to microphone level, and the circuits should be treated in the same way. Fig. No. 5 shows typical good grounding practice in this respect. Unbalanced circuits may be used, but are usually more difficult to handle if there is noise present. It will be noted that the only ground to this part of the system is at the point of lowest level and that all the circuits are balanced to ground. The centre taps of the mixer attenuators are only tied to ground if special noise difficulty is encountered and tests indicate improvement. This occasionally happens on circuits which connect to remote lines or studio equipment with separate ground systems.

# New Principle for Electronic Volume Control

By Harold E. Haynes,

R.C.A. Victor Division, Radio Corporation of America.

The principle described is a radical departure from those heretofore used in compressors. The features of this compressor are extremely low thump, very fast action (if desired), low distortion and freedom from the need for special circuit components or selected tubes. Fundamental circuits are discussed, and performance obtained with a complete compressor embodying the system is presented.

A volume compressor is an automatically actuated variable-gain amplifier, used for reducing the dynamic range of programme material. The timing characteristics of the voltage derived from the signal for actuating the variable-gain amplifier are customarily such as to provide a very rapid gain reduction whenever the signal level rises abruptly, but to increase gain relatively slowly when the signal level drops. Very short acting times, less than one millisecond, are often used in order to minimize unwanted initial peak amplitudes on sounds having sudden large increases in envelope amplitude, such as certain spoken syllables.[1] If a change of gain is accompanied by a shift in d.c. axis of the wave, a spurious aperiodic signal, commonly called "thump", will be produced. The d.c. component of this shift will, of course, be filtered out by the low-frequency cutoff characteristic of the system; nevertheless, to the extent that the gain-reducing action can be considered instantaneous, this shift is a step-function and contains energy at all frequencies. The more rapid the attack and the better the low-frequency response of the system, the more objectionable will be the thump.

## Background

Brief mention of a few commonly used methods of varying gain will serve to point out their shortcomings as far as balance, or tendency to produce thump, is concerned. The most common type of compressor employs as a variable-gain device some nonlinear electrical element, an element in which the two electrical quantities employed as input and output are related by a curved characteristic. This type of element is utilized in such a way that the slope of the characteristic at the operating point determines the gain of the circuit in which it is connected (which in general may be either greater or less than unity). Variations in gain are produced by super-imposing upon the input signal an adjustable control signal, the amplitude of which determines the operating point. Examples of this type of variable-gain device are nonlinear semiconductors, such as Thyrite, and vacuum tubes as usually used in compressors and limiters.

In the latter class is the familiar "variable mu" or "exponential" pentode, in which various points on a curve of transconductance vs. grid voltage are selected by adding a control voltage to the signal in the grid circuit. It is clear that in this case, as with all others in which the control effect is merely a bias superimposed upon the signal, there will inevitably be an output component produced by a change in gain, and hence a thump.

There are other vacuum-tube variable-gain circuits in which the controlling voltage is not superimposed upon the signal. One example is the "loading-tube" circuit, in which the plate impedance of a tube is shunted across a relatively high-impedance signal source, and the value of this impedance is changed by varying the grid voltage. Here a family of curves of plate current vs. plate voltage exists, their slopes varying as a function of grid voltage. Unfortunately, however, changing from one curve to another causes a change in plate current, so that the same fundamental problem presents itself, as before. A generalization may be made to the effect that a change in any tube characteristic causes a change in plate current; hence, circuits of this class also suffer to a greater or lesser extent from an inherent tendency to thump.

The obvious and almost universal remedy is the use of push-pull circuits, in which signal is applied to the two variable-gain elements out of phase, while

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gain-control voltages are applied in phase. Recombining the outputs in push-pull fashion then makes the signal components in phase and the gain-control components out of phase, so that they tend to cancel. A great reduction in thump is thereby obtained, but it is apparent that in order for perfect cancellation to occur under all conditions the characteristic curves of the two elements must be identical at every point in their operating ranges. An estimate of the degree of similarity required if the thump level is to be negligibly low may be made on the basis of the following arbitrary but not unreasonable assumptions: (1) that a change of gain between any two values within a range of at least 10 db should produce thump of the order of 40 db below signal level; and (2) that signal level is limited to 5% modulation of plate current by considerations of nonlinear distortion. These values lead to the conclusion that the plate currents must be equal (or differ by a constant amount) within something of the order of 0.05% throughout the operating range. Obtaining and maintaining the degree of similarity of characteristics necessary for such high-quality performance, though sometimes adequately accomplished, is expensive and time-consuming, and it frequently entails special selection and aging of tubes, plus frequent checking of those in service.

## Principles of the new system

A means of varying gain by employing vacuum tubes, but one which is not based upon monlinear characteristics, was thus sought, and an approach which proved fruitful is described. It is based upon the principle of keying a transmission device between gain values of zero and some fixed value, at a high frequency, and obtaining different effective-gain values by controlling the relative durations of "off" and "on" periods. Otherwise expressed, this means amplitude modulating the signal with a high-frequency rectangular wave or series of rectangular pulses of varying duty factor. Of course, such a modulated signal contains high-frequency components not present in the original, but by proper choice of modulating frequency, these can readily be made inaudible and easily separable from the signal by filters.

The action is illustrated in Fig. 1. A sinusoidal signal of frequency $f_s$ is shown modulated by a rectangular wave of frequency $f_k$, in which the duty factor is $k$. It can be shown that the modulated wave contains the original signal multiplied by the factor $k$, plus an infinite number of modulation products of frequencies $nf_k \pm f_s$. It follows that if the maximum signal frequency component to be accommodated is, for example, 15 Kc/s, the lowest sideband will be $f_k$—15 Kc/s. This sideband should be substantially higher than the maximum signal frequency, to facilitate removal of the sidebands. (It is pointed out later that the keying pulses should be as nearly rectangular as possible; hence, it is desirable to use the lowest permissible keying frequency, in order to minimize circuit difficulties).

With the unwanted components of the modulated wave filtered out, there remains only the desired signal multiplied by $k$; hence, if the value of $k$ can be varied in accordance with an appropriate control voltage, compression involving only linear electrical elements will have been accomplished.

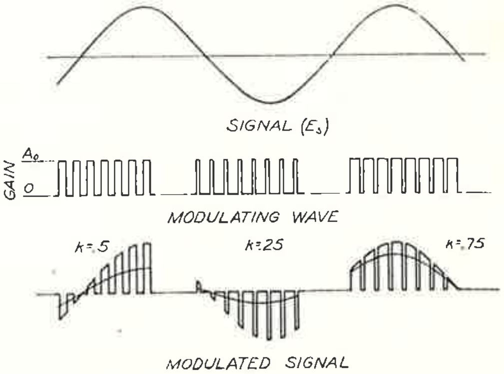

*Fig. 1. Sinusoidal signal of frequency $f_s$ modulated by a rectangular wave of frequency $f_k$.*

Since the keying frequency must be at least 30 Kc/s, a vacuum-tube circuit appears to be the only promising type of keying device. Hence, the same objection that was raised previously to tube circuits may at first seem valid, namely that a d.c. component of plate current, which will change with changes in gain, will still be required. The important distinction here is that the tube will need to operate only *at one mean value of plate current* (corresponding to "on"), and *at cutoff* (corresponding to "off"). Thus, any two tubes can be used in push-pull, and substantially perfect balance can be obtained at their single operating points. They can, and should be, linear devices, and as such will permit relatively large signal amplitudes without objectionable distortion. Furthermore, their linearity may be enhanced by means of negative feedback, an expedient which would tend to nullify the gain-changing properties of conventional circuits.

## Circuit methods

Figure 2 shows the basic circuit of such a keyed amplifier. Two cathode followers are connected in push-pull, with positive keying pulses introduced in the cathode circuit. Pulse amplitude is sufficient to cut off plate current completely even when peak signal amplitude occurs. Additional positive bias voltages, $E_{c1}$ and $E_{c2}$, permit desirable operating points to be selected, one of them being adjustable to permit balancing.

It is apparent that the keying pulses must have negligible rise and fall times, in order that the tubes will not be operating at points on their characteristic other than the desired one during an appreciable fraction of the time. This means that a minimum of capacitance loading should be permitted at any point in the pulse circuit. Therefore, resistors $R_3$ and $R_4$ are inserted to isolate the output transformer from the pulse circuit. Unwanted modulation products are removed by a simple low-pass filter following the output transformer.

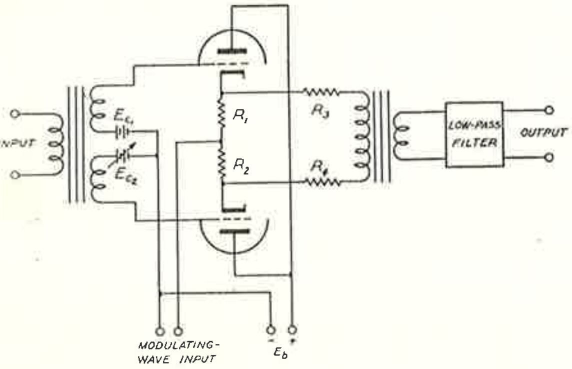

*Fig. 2. Basic circuit of vacuum-tube keyed amplifier.*

An essential adjunct to the keyed amplifier, when used in a compressor, is a source of pulses of controllable duration and of approximately constant frequency, having the requisite relation between duration and control voltage. It can be shown that in a compressor driving control voltage from output, as is customary, and having a slope of $\frac{1}{2}$ on a decibel basis (2:1 compression), numerical gain should be inversely proportional to control voltage; hence, a pulse generator was developed in which the "on" (negative) pulse width closely approximates this relation. Figure 3 shows the basic circuit of the pulse generator. A 45 Kc/s square wave, generated by a multi-vibrator, is differentiated by $C_1$ and $R_5$, to produce a series of alternate positive-voltage and negative-voltage pulses of very short duration. The negative pulses cause capacitor $C_2$, which is also connected to the grid of sharp cutoff pentode $V_3$, to be charged negatively once for each cycle, through diode $V_2$. $C_2$ discharges toward zero through $R_6$, which is connected to the source of control voltage, the latter being variable from zero to a relatively large positive value. Thus, the plate current of $V_3$ is cut off for a portion of each interval between pulses which becomes smaller as the value of the control voltage is increased. It is these periods of cutoff which eventually become "on" pulses for the keyed amplifier, their duration relative to the pulse period being the factor *k*. The time constant of $C_2$ and $R_6$ is made about equal to the pulse spacing (22$\mu$sec), and the potential to which $C_2$ is charged by the negative pulses is about ten times the cutoff grid voltages of $V_3$; hence $V_3$ draws no plate current unless the control voltage has a substantial positive value. This means that the significant part of the discharge curve of $C_2$ is reasonably linear, and it can be shown that this causes the duration of the cutoff period, and hence the value of *k*, to be nearly inversely proportional to the control voltage, as desired. The rapidity with which the value of *k* can be changed, and hence the speed of action of the compressor, in practice is limited only by the properties of the circuit by which gain-controlling voltage is derived.

The plate-current pulses of $V_3$, which are roughly rectangular because of its sharp cutoff characteristic, produce voltage pulses which are further shaped by subsequent amplifier and limiter stages so as to have very short rise and fall times, and applied to the amplifier circuit of Fig. 2.

These two basic circuits, with the addition of conventional means of deriving control voltage proportional to compressed output, and having the desired timing characteristics, constitute a complete compressor. Since this type of control circuit is well known, and for the present application need be little different from those for other compressors, this subject will not be discussed further.

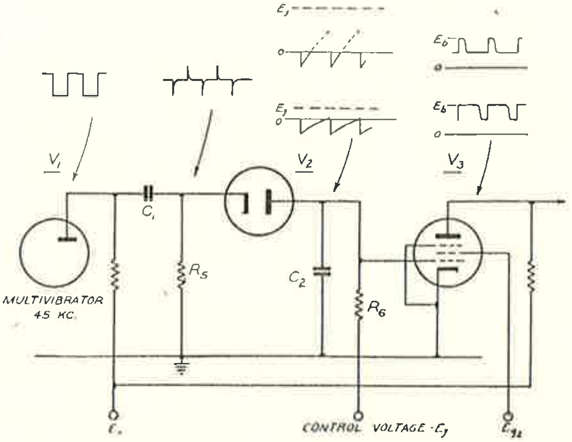

*Fig. 3. Basic circuit of pulse generator.*

## Performance

A complete compressor based upon these circuits has been built and is illustrated in Fig. 4. Its operating characteristics were made to conform to those of existing compressors so that its performance could be easily evaluated. Although this model is somewhat more complex than the simplest compressors, no special tubes or other special components are used. It affords a useful gain-reduction range of more than 15 db. Performance, especially with regard to thump, is excellent, both with respect to the degree of balance obtainable and to the long-time stability of this balance.

Two chief methods have been used for observing and measuring the effects of unbalance in compressors. One which is typical of actual operating conditions, described by Maxwell[2] consists of abruptly raising the level of a relatively high frequency sine-wave input signal to the compressor and observing on an oscilloscope the transient appearing in the output. Although it depicts the thump phenomenon very graphically, this test is open to the objection that it takes into account balance conditions at only two specific points in the gain-reduction range. Also, in a well-balanced compressor the transient amplitude is too small to be conveniently observable. A second method, often built into compressors as a balance check, measures cross modulation between the gain-control circuit and the signal circuit by applying a sinusoidal test voltage in the gain-controlling circuit. A single test of this kind includes the effects of unbalance at all points in the gain range swept, and if the test conditions are suitable it affords a good overall evaluation of balance.

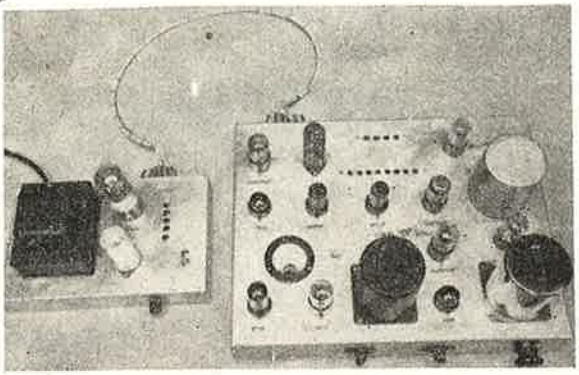

*Fig. 4. Complete compressor based on the circuits shown in Figs. 2 and 3.*

Measurements of both types have been made on the pulse-modulation compressor. In the first type, a 250-cycle low-pass filter was used, following the compressor, to reduce the carrier amplitude, and thereby make the transient more easily seen. For a 10-db increase in input (5-db gain reduction), signal-to-thump ratios of 50 to 60 db were obtained.

The cross-modulation method is felt to be preferable for specifying unbalance, because it does detect unbalance at all points in the range used. A figure of merit called "signal-to-unbalance ratio" is proposed to describe the performance of a compressor when tested in this manner. It is expressed in decibels and is defined as follows: Signal level is the maximum output level, with 10 db of gain reduction, at which some satisfactorily low value of total harmonic distortion of a 1000-cycle signal is produced. In the present case, this value is taken as 0.5%. Unbalance level is the output produced, in the absence of signal, by a 60-cycle control voltage which varies the gain reduction throughout the range of 0 to 10 db.

Using this cross-modulation test method, excellent signal-to-unbalance ratios have been obtained, along with freedom from the need for special tube selection and from the necessity for frequent rebalancing. Tests have shown that, except for the possible rejection of perhaps 10% of samples, tubes selected at random for the variable-gain stage (6SN7GT) will all produce optimum signal-to-unbalance ratios of 55 db or more. Operation over periods of a few hundred hours has indicated that the balance does not deteriorate more than 10 db during this length of time, and that the original figure can be readily regained by rebalancing. Although unregulated heater and plate supplies were used, line-voltage variations of 10% also increase the unbalance only about 10 db.

By adoption of pulse-modulation techniques, it has thus been possible to construct a compressor whose performance regarding thump is equal or superior to that of any now used in the most exacting applications, without the need for specially selected tubes or other components. Its moderate added complexity is felt to be of secondary importance in the light of its very significant advantages.

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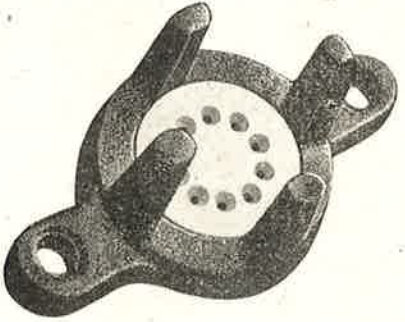
2. **Donald E. Maxwell, "Dynamic performance of peak-limiting amplifiers," Proc. IRE, 35: 1349-1356, Nov., 1947.**
3. **F. E. Terman, Radio Engineers' Handbook, 1st ed., McGraw-Hill, New York, 1943, pp. 22, 23.**

## RADIOTRON MINIATURE VALVE PIN STRAIGHTENERS

The following points should be remembered when using these useful Radiotron tools.



The pins on the button of a miniature valve are moulded into the glass on a jig of high precision, which enables a predetermined amount of strain to be later introduced into this button.

The Pin Straighteners are only made to approximate this precision jig and are intended to take care of any pins which have become bent due to mishandling.

It is therefore essential that the user of these Pin Straighteners appreciate this fact and only uses sufficient pressure on the valve to bring the pins back to their approximately normal position.

Remember—it is not necessary and frequently dangerous to force the valve "home" in the Straightener.

The pins should only enter a distance sufficient to give the required straightening, a condition which can be judged by the amount of pressure applied to the valve.

# *New* RCA Releases

## RADIOTRON 6181
## UHF POWER TETRODE WITH CERAMIC-METAL SEALS

Just announced by RCA is a new "1 kw" power tetrode for use in uhf television transmitters. Designated as the 6181, this new forced-air-cooled power tetrode can be operated with full plate voltage and plate input at frequencies as high as 900 megacycles per second, and is capable of delivery a synchronizing-level power output of 1200 watts in broad-band television service.

Featured in the design of the 6181 is a coaxial-electrode structure providing low-inductance, large-area, rf electrode terminals insulated from each other by low-loss ceramic bushings.

The coaxial-electrode structure of the 6181 is designed especially for use with high-power circuits of the coaxial-cylinder cavity type. The design provides for easy insertion into the cylinders, and permits effective isolation of the plate from the cathode.

The 6181 has an indirectly heated, low-temperature, coated cathode of the matrix type for long service. It has a 120-volt heater electrically shielded from the rf input and output circuits.

## RADIOTRON 6012
## GAS THYRATRON

The new gas thyratron 6012, just announced by RCA, is designed especially for motor-control and low-power inverter service in circuits operating at 60 cps. It is conservatively rated to withstand a maximum peak inverse anode voltage of 1300 volts, a maximum peak cathode current of 5 amperes, and a maximum average cathode current of 0.5 ampere.

Operating features of the 6012 include a negative-control characteristic which is essentially independent of the ambient temperature over the range from —75° to + 90°C, low preconduction currents, low control-grid-to-anode capacitance, and low control-grid current.

The 6012 employs a compact design in which special attention has been given to features which improve its strength not only against shock but also against vibration. Use is made of a button stem to strengthen the mount structure and to provide relatively wide inter-lead spacing for reduction in susceptibility to electrolysis. These features all contribute to the dependability of the 6012 and to its suitability for industrial applications.

This new gas thyratron has a maximum overall length of $4\frac{1}{4}$ inches and a maximum diameter of $1\frac{3}{4}$ inches.

## RADIOTRON 5822
## WATER-COOLED IGNITRON

Type 5822 is a water-cooled, steel-jacketed, mercury-pool-cathode tube of the ignitron family. It is intended especially for use in frequency-changer resistance-welding service.

In the frequency-changer method of resistance welding, three-phase 60-cycle power is converted to single-phase power having a frequency of 5 to 12 cycles per second. This method offers appreciable reduction in kva demand in comparison with that required in single-phase welding. Furthermore, the three-phase circuit balances the power load and permits improved results in welding aluminium, magnesium, and their alloys.

Featured in the design of the 5822 is (1) an internal structure which permits a deionization time such that satisfactory operation is obtained under the severe conditions of commutation encountered in frequency-charger resistance-welding service; (2) a spiral channel in the jacket wall through which water is circulated to provide uniform cooling; and (3) an ignitor suitable for intermittent operation.

## RADIOTRON 6AF4
## U-H-F MINIATURE TRIODE

Type 6AF4 is a u-h-f oscillator triode of the 7-pin miniature type designed for use as a local oscillator in u-h-f television receivers covering the frequency range of 470 to 890 megacycles per second. This tube has good frequency stability, low interelectrode capacitances, low lead inductance, and low r-f lead resistance.

Design features of the 6AF4 include silver-plated base pins to minimize losses due to skin effect at the ultra-high frequencies, short internal leads to reduce lead inductance and lead resistance, and a short mount structure utilizing small parts to provide low interelectrode capacitances. In addition, the 6AF4 has double base-pin connections for both plate and grid. These double connections are positioned to facilitate operation of the tube with either series — or parallel-resonant lines and to provide greater flexibility in circuit connections.

## Radiotron Designers Handbook
## 4th. Edition

We have been advised by the printers, that owing to circumstances beyond their control deliveries will not commence until October.