Radiotronics

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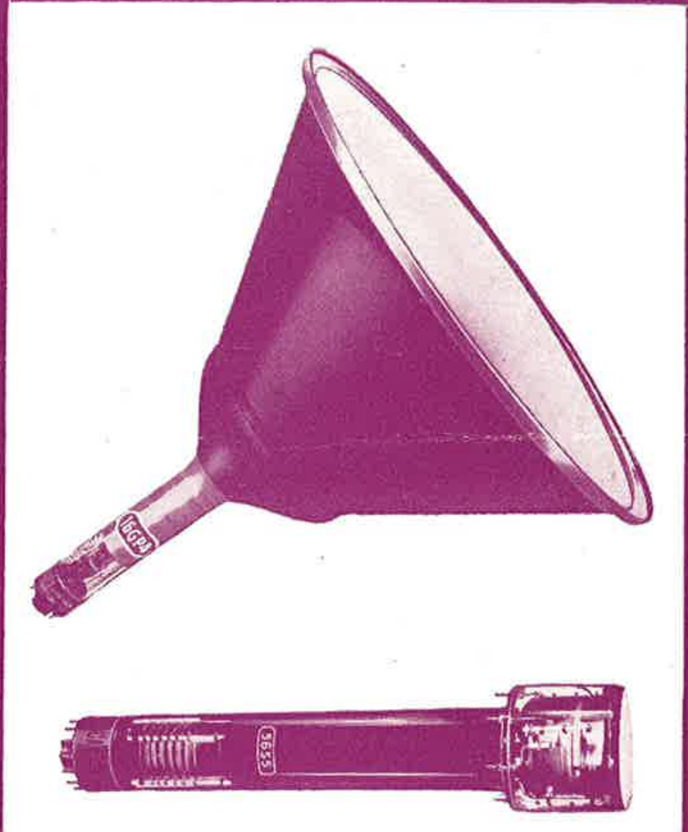

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RESEARCH • DESIGN • MANUFACTURE

NUMBER 143 JUNE 1950

RADIOTRONICS

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Technical Editor:
F. Langford-Smith, B.Sc., B.E.,
Senior Member I.R.E. (U.S.A.).

Asst. Tech. Editor:
Ian C. Hansen, Member I.R.E. (U.S.A.).

Sub-Editor:
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Circuit Laboratory Report A.518 **Number 3**

Low Distortion 25 Watt Amplifier Using 807 Tetrodes*

Performance

Radiotron amplifier A518 has been designed to deliver 25 watts output at 2% intermodulation distortion, or 35 watts with less than 10% amplitude distortion (Fig. 1). On public address amplifier standards the A518 would be rated as a 35 watt amplifier.

Intermodulation distortion of 2% is taken as being the limit for high quality output as Hilliard (Ref. 1) has stated that under the conditions used for testing this amplifier intermodulation distortion below 2% is not audible. This also agrees with tests by H. F. Olson (Ref. 5) on harmonic distortion.

The power output of the amplifier is maintained over the useable audio range from 40 to 12,000 c/s, the limits at which measurements could be made. At these limits the intermodulation distortion only rises slightly above that for a more restricted frequency range (100 to 2,000 c/s).

The frequency response has been deliberately restricted, as otherwise at the high frequency end it would be only 5 db. down at 100 Kc/s with intermediate peaks due to output transformer resonances. At the low frequency end the coupling capacitances have been adjusted to reduce the response as rapidly as possible below 10 c/s. Nevertheless the response is flat within ± 1 db. between 20 and 30,000 c/s. (Fig. 2).

The sensitivity (6 mV input for 25 watts output) makes the amplifier suitable for use with a low level output pick-up or microphone, and a 6AU6 valve has been used in a special triode condition which reduces noise from the first stage to a low level (54 db. below 25 watts with volume control at maximum, 75 db. down with volume control at minimum). Power supply filtering is sufficiently good to allow the amplifier to be used with a sensitive speaker in a quiet location without the hum being noticed.

Feedback from the voice-coil of the loud-speaker is provided for distortion reduction, improved frequency response and low output impedance (the mid-frequency output impedance of the 6.6 ohm transformer secondary is 1.5 ohms). The gain reduction due to feedback is 4 times (12 db.) and the stability is such that the amplifier can be operated into an open circuit (when the feedback increases to 26 db.) or a variety of capacitive or inductive loads at any input from zero to maximum without oscillation occurring.

A possible criticism of an amplifier which uses tetrode output valves and is expected to provide high quality reproduction is that although measured distortion may be low when the amplifier is operated into the correct load, the distortion may rise considerably when the impedance of the loudspeaker increases at the bass resonance and at high frequencies.

To check this the amplifier was set-up for intermodulation testing, delivering an equivalent power of 25 watts. The intermodulation distortion at this output level was 2%. Leaving the input voltage unchanged the load was increased from 6.6 to 20 ohms, and the intermodulation distortion dropped to 1.5%. The reduced distortion was obtained for a lower power output (approximately the same output voltage was delivered due to the negative voltage feedback) but the conditions of the test simulate the effect of speaker impedance variations. A three-to-one change in impedance was taken as being typical of the variation in better class speakers over the useful frequency range.

Circuit

The two low-level 6AU6 valves are triode-connected as described in Radiotronics 142 with load and bias resistors chosen to give minimum distortion. The advantages of this method of connection for low-level operation were discussed in the original article.

However, the other two 6AU6 valves are triode-connected in the more conventional manner described in Radiotronics 139, since this gives lower distortion at the higher level at which these valves operate. High capacity, low voltage electrolytic condensers are used for cathode by-passing in the first two stages to minimize hum and noise arising from heater-cathode leakage.

The phase-splitting circuit is one described by Scroggie (Ref. 2, 3) which is chosen for its self-balancing action. So long as the values of the 1 megohm resistors are within 10% of each other, output distortion measurements are substantially unchanged from those quoted.

The high frequency response of the amplifier is controlled by the 0.002 μF capacitors connected from each 807 plate to the centre tap of the output transformer primary, by the 350 μμF capacitor from

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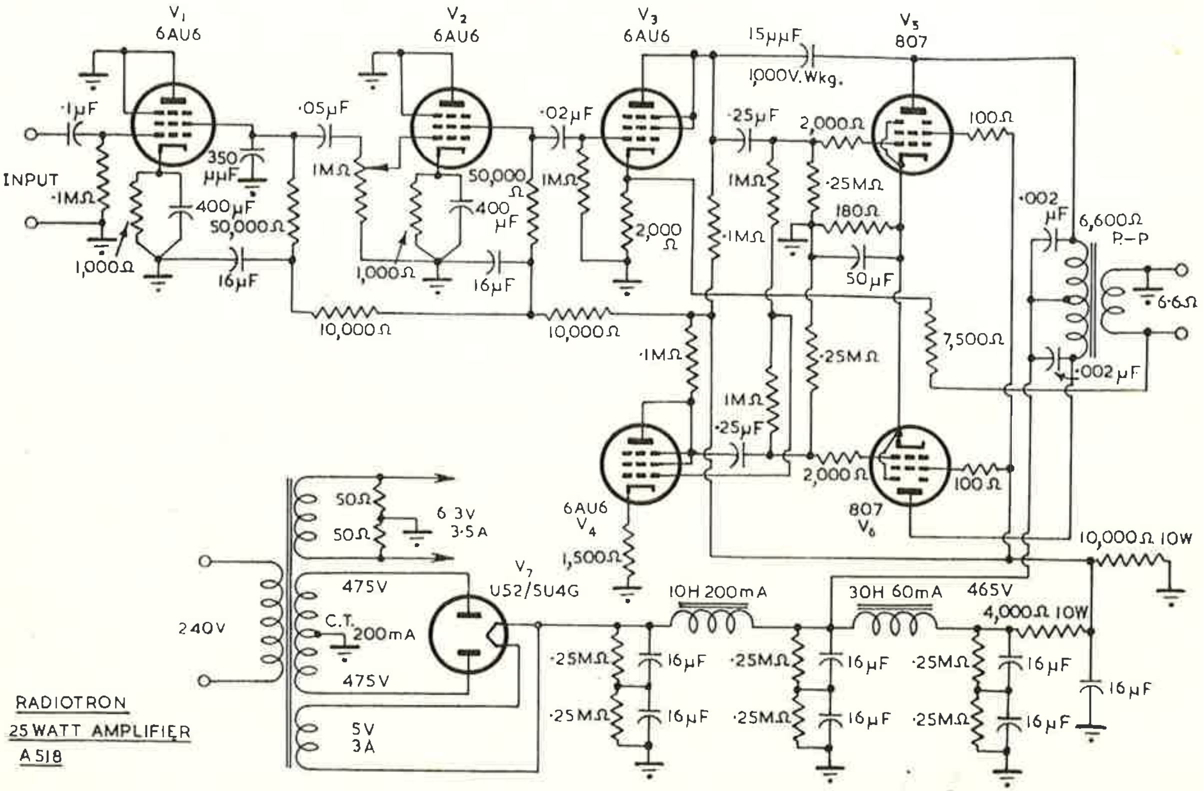
* Contributed by the Circuit Design Laboratory, Valve Works, Ashfield.



Circuit of Radiotron 25 watt amplifier.

the anode of the first 6AU6 to ground and by the 15 μμF capacitor between the plate of one 807 and its 6AU6 driver. The 0.002 μF capacitors prevent the impedance of the output transformer from rising to high values at frequencies at which distributed capacitances resonate with leakage inductance. The 15 μμF capacitor provides an internal feedback loop which rapidly decreases the gain outside the useable a-f range and thus prevents the amplifier from becoming unstable when working into an open-circuited load. The 350 μμF capacitor is used to minimize the effects of a slight peak in the response of the feedback section of the amplifier, and to sharpen the cut-off outside the desired range.

A capacitive input filter is used because the variation in current drain between no-load and full-load is slight, and by using capacitive input a lower voltage high tension winding is required on the power transformer.

The output transformer used was a special type* as designed for the Radiotron amplifier A515 (Radiotronics 128 and 130). However, since the plate-to-plate load of the output valves required for this amplifier is 6,600 ohms, the load was reduced to 6.6 ohms on the nominal 10 ohm secondary of the original 10,000 ohm plate-to-plate transformer. Since there is only 12 db. of feedback applied to the amplifier it is not anticipated that the output transformer design will be critical, but a well designed transformer, such as the one used in the developmental model, is required.

If a different secondary impedance is specified it will be necessary to alter the value of the 7,500 ohm feedback resistor in the ratio of the square root of the impedance change. For example, with a 10 ohm secondary winding the resistor would become

$$7{,}500 \; \frac{10}{6.6} \text{ ohms} = 9{,}700 \text{ ohms.}$$

A resistor having a nominal value of 10,000 ohms would be satisfactory in this instance.

Tests were carried out with a number of 807 valves chosen at random and the variation in distortion was so small that balancing of the plate current (as in Radiotron amplifier A515) was not considered necessary.

Mechanical details

The experimental amplifier used a separate chassis for the power supply unit, and unless this is done it will probably be difficult to duplicate the low hum level quoted.

The voltage gain of the amplifier from input to 807 plate is greater than 35,000 times so that care should be taken when laying-out the amplifier to obtain adequate separation (or alternatively sufficient shielding) between input and the output of the final stage.

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* Supplied by Ferguson's Radio.

Measurements

Most of the measurements presented are self-explanatory, but since no method of making intermodulation measurements has yet been standardized, the details are set out here.

If two or more frequencies are passed simultaneously through an amplifier with a non-linear input-output characteristic intermodulation will take place, i.e. the two signals will combine with each other to produce new frequencies not present in the input. For example, if frequencies of 100 and 1,000 c/s are applied to the input, intermodulation products of 1100, 1200, 1300, 1400 etc. ("superior" products) and 900, 800, 700, 600 etc. ("inferior" products) may be created as sidebands of the 1000 c/s tone. Sidebands of the 100 c/s signal will also be formed. Since these frequencies are not harmonically related this type of distortion is more objectionable than "harmonic" distortion as normally measured, because in the latter case the harmonics usually generated are separated by consonant musical intervals.

Following Hilliard's recommendation (Ref. 4) two signals were applied to the amplifier, the amplitude of the lower frequency signal being four times that of the higher frequency one. This has the effect of increasing the measured intermodulation when the high frequency signal is considered as a carrier. The products were tuned-in separately on a wave analyzer and recorded, and the percentage total intermodulation calculated from the formula:—

Percentage Intermodulation Distortion =

$$\frac{(E_{f_1-f_2}+E_{f_1+f_2})^2+(E_{f_1-2f_2}+E_{f_1+2f_2})^2+(\ldots\ldots\ldots\ldots)}{E_{f_1}}\times 100$$

where f_1 = higher frequency f_2 = lower frequency and $E_{f_1-f_2}$ etc. = intermodulation products.

The readings taken with frequencies of 100 c/s and 2,000 c/s are presented in full with all products greater than 0.05% of the 2,000 c/s carrier recorded. The numbers at the top of the columns show the order of the intermodulation product and the sign denotes addition or subtraction. For example at an indicated power output of 10 watts, and at a frequency of (2,000 + 100) c/s the intermodulation product is 0.25% of the 2,000 c/s output and at (2,000 − 100) c/s it is 0.2%; at (2,000 + 2 × 100) c/s i.e. 2,200 c/s it is 0.7% and at 1,800 c/s it is 0.75%.

For combinations of frequencies other than 100 c/s and 2,000 c/s only the calculated r.m.s. sum is presented. The reason for including tests with frequencies of 40 c/s and 12,000 c/s is that many amplifiers are incapable of maintaining their mid-frequency output to the extremes of the a-f range. At low frequencies insufficient output transformer primary inductance may reduce the output, and at high frequencies capacitances may upset load conditions with the same result.

Although some commercial intermodulation testers give an indication of the combined intermodulation products after removing the carrier, it was considered preferable to measure the components separately and combine them mathematically. In accordance with the method described in I.R.E. Standards on Radio Receivers 1948, only the intermodulation products of the lower frequency signal with the high frequency signal considered as a carrier were measured. For example, in the case of signals of 100 c/s and 2,000 c/s there are products with frequencies of 1,900 and 2,100 c/s, 1,800 and 2,200 c/s, 1,700 and

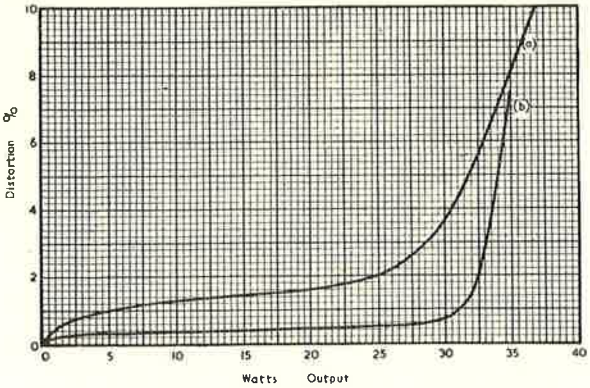

Fig. 1. Distortion vs. Output.
Curve (a) Intermodulation distortion vs. equivalent output.
Curve (b) Harmonic distortion vs. actual output.

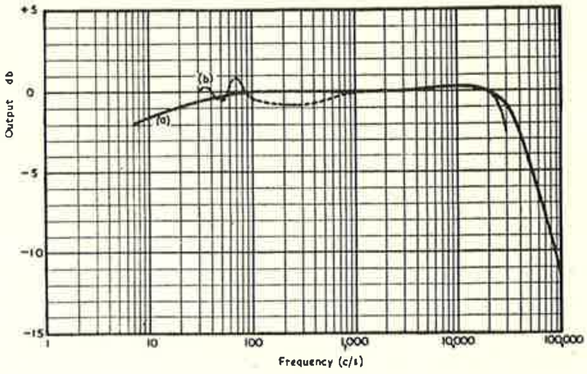

Fig. 2. Frequency response.
Curve (a) Resistive load
(0 db. = 15 watts)
Curve (b) Speaker load
(0 db. = 2.5 volts)

2,300 c/s etc., grouped around f_1, but equally there are products of 1,900 and 2,100 c/s, 3,900 and 4,100, 5,900 and 6,100 c/s etc., grouped around f_2. This second set of products has been ignored, except for the first two which are included in each group. Because of the difference in level between the two signals, the omission of the second group of intermodulation products has little effect on the total measured distortion.

In the performance figures quoted, it will be seen that intermodulation distortion is given as a function of "indicated power" and "equivalent power". The reason for the $^{25}/_{17}$ ratio between these two sets of figures was given in some detail in Radiotronics 130, but briefly it can be said that the equivalent power should be used for comparison with amplifiers tested with a single sine wave. The ratio of $^{25}/_{17}$ is introduced because amplifier overload is a function of peak voltage output, so that when two signals are present overloading is proportional to the sum of the signals, whereas power output is proportional to the square root of the sum of the squares of the output signals.

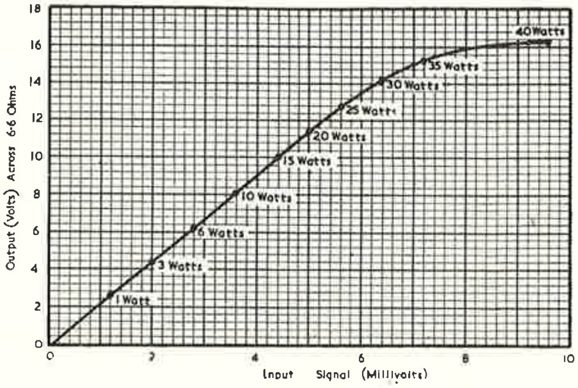

Fig. 3. Linearity curve.

Sufficient information about the intermodulation measurements has been presented to make it evident that all such tests are made under a purely arbitrary set of conditions and it is emphasized that intermodulation distortion figures taken on different amplifiers must not be compared unless it is known that all conditions of measurement are identical.

From Fig. 1 and the table of harmonic distortion vs. output, it might be thought that no signs of overloading could be detected from harmonic distortion measurements until the output exceeded 30 watts, whereas from intermodulation measurements overloading is noticeable from 25 watts upwards. However, up to an output of 25 watts the only harmonics with an amplitude above 0.05% are the second and third, whereas at 27 watts output every harmonic up to the twelfth exceeds this figure. From this it seems probable that if intermodulation testing facilities are not available, the overload point of a good amplifier should be taken as the output at which high order harmonics suddenly appear.

Intermodulation and harmonic distortion measurements are different tests for the same characteristic — amplifier non-linearity. A point of interest is that over the majority of the power output range of the amplifier the intermodulation distortion figures are approximately four times the harmonic distortion figures.

Operating test.

The amplifier has been demonstrated with an Altec-Lansing 15″ speaker, a good quality pick-up and a number of good recordings, including one specially made for demonstration purposes.

Even when the special demonstration record was in use, the large majority of the listeners preferred the reproduction with a 5 Kc/s low-pass filter following the pick-up, and none considered that any improvement was obtained by increasing the response above 8 Kc/s since the added record noise and distortion products marred the reproduction. If either speaker or pick-up have a reduced high-frequency response it is possible that filters would be preferred with higher cut-off frequencies.

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2. Scroggie, M. G., "The See-Saw Circuit", Wireless World 51, 7 (July 1945), page 194.
3. Carpenter, R. E. H., "'See-Saw' or 'Paraphase'?", Wireless World 51, 9 (Sept. 1945), page 263.
4. Hilliard, J. K., "Intermodulation Testing", Electronics 19, 7 (July 1946), page 123.
5. Olson, H. F. (book), "Elements of Acoustical Engineering" (2nd ed., 1947, D. Van Nostrand Company Inc.).

PERFORMANCE FIGURES.

1. Sensitivity

The figures quoted are the 1,000 c/s voltages required at the respective control grids to give an output of 25 watts at the transformer secondary. No feedback was applied and the 7,500 ohm feedback resistor was returned to ground instead of to the voice coil.

Valve	**Input Voltage**
V1	1.4 mV
V2	40 mV
V3 and V4	1.1 V
V5 and V6	15 V
With feedback, V1	5.6 mV (Feedback = 12 db)

2. Harmonic distortion

(a) *At 1,000 c/s*

Output (watts)	0.5	5	15	25	30	35
Distortion (%)	0.2	0.3	0.4	0.5	0.7	7.5

(b) *At 25 watts output*

Frequency (c/s)	40	1,000	5,000
Distortion (%)	0.8	0.5	0.6

3. Intermodulation distortion

(a) 100 c/s and 2,000 c/s. 4:1 ratio of low frequency to high frequency signal amplitudes.

		Sign of			Order of intermodulation product.					
Indicated watts	Equivalent watts	I.M. product	1 %	2 %	3 %	4 %	5 %	6 %	7 %	Total %
0.5	0.74	+	0.05	0.2						0.4
		—	0.05	0.2						
1.0	1.5	+	0.05	0.25						0.6
		—	0.05	0.35						
5.0	7.4	+	0.05	0.55						1.2
		—	0.05	0.6						
10	15	+	0.25	0.7						1.5
		—	0.2	0.75						
17	25	+	0.4	0.85	0.05					2.0
		—	0.35	0.9	0.05					
20	29	+	0.7	1.2	0.2	0.2	0.05	0.05		3.1
		—	0.7	1.4	0.2	0.1	0.05	0.05		
25	37	+	1.5	4.0	1.0	1.5	0.5	0.2	0.05	10
		—	2.0	4.5	1.0	1.5	0.5	0.3	0.05	

(b) 4:1 ratio.

Indicated watts	Equivalent watts	40 c/s & 2,000 c/s %	100 c/s & 2,000 c/s %	100 c/s & 12,000 c/s %
0.5	0.74	0.5	0.4	0.5
1.0	1.5	0.6	0.6	0.6
5.0	7.4	1.2	1.2	1.2
10	15	1.7	1.5	1.7
17	25	2.5	2.0	2.4
20	29	3.3	3.1	3.2
25	37	15.0	10.0	12.0

4. Noise level

(a) *Including hum.*

Gain control full on:—
noise output = 54 db. below 25 watts.

Gain control full off:—
noise output = 75 db. below 25 watts.

(b) *Hum level.*

Gain control full on:—
50 c/s – 0.8 mV
100 c/s – 0.4 mV } Total = 0.13 microwatt
150 c/s – 0.2 mV

Gain control full off:—
50 c/s – 0.6 mV
100 c/s – 0.4 mV } Total = 0.1 microwatt
150 c/s – 0.4 mV

5. Voltage and current analysis.

(Measured with 500 ohm voltmeter, without signal input. All voltages measured from ground. Meter scale indicated in brackets.)

	Plate	Screen	Cathode
V1 (6AU6)	110 V (400V)	—	2.0 V (10V)
V2 (6AU6)	115 V (400V)	—	2.0 V (10V)
V3 (6AU6)	105 V (400V)	—	2.2 V (10V)
V4 (6AU6)	110 V (400V)	—	2.3 V (10V)
V5 & V6 (807)	453 V (1000V)	291 V (400V)	21 V (100V)
V7 (U52/5U4G)	475 V r.m.s. per plate.		

Current	Plate	Screen
807 (2 valves) no output	110 mA	4.5 mA
807 (2 valves) 25W output	110 mA	9.5 mA

Total H.T. Current .... 150 mA

The Elements of a TV System*

A brief review of the Functions of the Most Important Parts of the U.S.A. TV System, with an Explanation of the Reasoning Behind the Choice of Standards, Types of Transmission, Shape of Synchronizing Pulse, etc.

by JOHN H. ROE

Supervisor, TV Systems Engineering Group,
Engineering Products Dept.

PART III — CONCLUSION

The constant search for means of immunization against the effects of noise has brought about the development of automatic frequency control (afc) of the scanning circuits in television receivers. In triggered circuits, each scanning line (and each field) is initiated individually by a pulse in the incoming signal. Contrastingly in an afc system, scanning generators are governed by stable oscillators which in turn are controlled by voltages obtained from phase comparison of the incoming sync. pulses with the scanning signals themselves. The time-constant of the comparison circuit is usually made long compared to the period of scanning so that random noise pulses have very little effect on the resulting control voltage, and correspondingly little effect on the scanning frequency. The fact that such afc circuits are keyed circuits provides a further immunization factor by eliminating the possible effect of all noise pulses except those which coincide with the short keying intervals. The use of afc scanning circuits makes possible accurate synchronizing of a receiver under such bad conditions of noise that the masking of the picture by the noise renders it completely unusable. Thus failure to synchronize may be largely eliminated as a limiting factor in picture reception.

AFC may be used with both vertical and horizontal scanning circuits, but so far is being used commercially for horizontal circuits only. One reason for not using afc with the vertical circuits is that the time-constant must be very long to provide a stable control voltage. As a result, the circuit will not recover from an extended interruption of the incoming signal until an intolerably long time has elapsed. The frequency of the oscillator drifts during an interruption, and may not recover for a large number of seconds after the signal returns. During the period of recovery, the raster rolls over continuously at a decreasing rate until control is restored. The time-constant of the horizontal circuit, on the other hand, may be short enough so that recovery takes place in less than one field. Triggered scanning circuits, of course, recover from signal interruptions very rapidly, but they do not have the same high immunity to noise that the afc circuits have.

As a result of the use of afc circuits in receivers, a high degree of frequency stability is required in the horizontal sync. and blanking signals. Frequency modulation of the horizontal pulses is intolerable because it causes the right and left hand edges of the blanked raster in the receiver, as well as vertical lines in the scene, to assume the same shape as the modulating wave. As shown in Fig. 10 the border of the complete raster in the receiver is rectangular, but frequency modulation of the horizontal sync. and blanking will distort the shape of the border produced by blanking. Frequency modulation by a 60-cycle sine wave is illustrated.

Horizontal retrace begins along a straight vertical line regardless of timing and since this retrace is controlled by a stable oscillator in the receiver which is not responsive to short-time changes in sync. timing, the presence of variations in sync. timing and of corresponding changes in blanking pulse timing, will show as a displacement of the edges of the blanked raster. The frequency stability of the sync. generator must therefore be at least equal to the stability of the oscillators used in afc receivers. The maximum rate of change of frequency allowable in a sync. generator has been specified by RMA as 0.15% per second. This is a rather strict tolerance as indicated by the fact that it allows a total displacement of only 1/32 of an inch (approx.) in a period of one field in a picture 10 inches wide.

Film projection

The use of standard sound motion picture film for television programme material offers a special

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* Reprinted from RCA Broadcast News by courtesy of the Radio Corporation of America.

problem which arises from the difference in the picture repetition rates used. For reasons explained previously, the rate used for television is 30 frames and 60 fields per second. The standard speed for sound film, both 16mm and 35mm, is 24 frames per second, and since each frame is projected twice, the picture rate is 48 per second. The basic problem of reconciling the frequency difference has been met by using special projectors for television in which alternate frames of the film are projected twice and the remainder are projected three times. In this way, 60 pictures are obtained in place of the usual 48, but the average speed of the film through the projector is unchanged; hence the sound take-off is entirely normal.

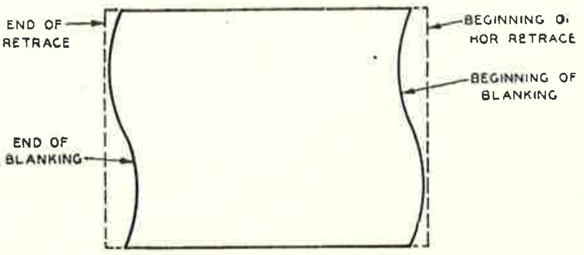

Fig. 10. Effect of frequency modulation of horizontal sync. and blanking on shape of raster in receiver with AFC of horizontal scanning.

Another problem also presents itself in the use of intermittent film projectors for television. The vertical scanning period occupies from 92% to 95% of the total period. If the projected image is to be thrown on the pickup tube during the scanning period at all, it must be for the entire time so that all parts of the area will be subject to the same lighting conditions. Such an arrangement would leave only the vertical retrace period (5% to 8% of the total, or approximately one thousandth of a second) in which to pull down the film to the next frame. 35mm film will not stand up under accelerations produced by sprocket hole pull-down in such a short period; hence some other scheme must be used. The method which has been adopted for use with intermittent projectors makes use of the storage property of certain kinds of pickup tubes, such as the iconoscope. The frame of film is projected with very intense illumination during the vertical blanking period only, while neither the pickup tube nor the receiver is being scanned. Then the light is cut off and the pickup tube is scanned in the absence of any optical image from the film. The signal generated during this scan results from charges stored on the sensitive surface during the preceding flash of light. While the light is cut off during the scan there is ample time to pull the film down before the next flash of light, without exerting destructive forces. The pulses of light may be obtained by chopping the output of a continuous source with a rotating disc, or (with a special type of arc lamp) by pulsing the source itself by electronic means. The storage properties of pickup tubes for this purpose must be sufficiently good so that dissipation of the stored charges is negligible between light pulses. Appreciable dissipation causes loss of contrast at the bottom of the picture.

Another solution to the problem of film projection in television is the use of a continuous projector, a type which produces a stationary image from continuously moving film by means of moving mirrors or lenses. This solution has not been accepted commercially so far because of practical difficulty in making the optical system sufficiently accurate to stop motion of the image completely.

The film problem in England, Europe, and other areas where 50 cycle power systems are standard, and where the television field frequency is also 50 cycles per second, is simpler in one respect, namely, that it is not necessary to use the two-three ratio for projection of alternate frames of film. Instead, the film is projected as it is in theatres where each frame is projected twice. No attempt is made to compensate for the difference between the 24 frame taking speed and the 25 frame projection speed. The results are an approximate 4% increase in the apparent speed of motion of objects in the scene (which is probably negligible) and a slight rise in the pitch of all sounds. This latter effect is the more objectionable of the two, though generally it is not noticeable in speech and many other ordinary sounds. The change in pitch is undoubtedly noticeable to the trained musician in the case of musical sounds and must produce an unpleasant mental reaction to the music. However, no easy solution to the problem is known, and the situation is accepted without serious complaint. The other aspects of the film problem are not affected by the use of 50 fields instead of 60.

Modulation

The choice of amplitude modulation for television transmitters was made after comparison of results of a-m and f-m transmissions in field tests in the New York area. The results indicated clearly that f-m is not suitable for use in television broadcast transmitters or in any television radiating system where multipath transmission is encountered. The reason may be understood easily. Multipath transmission in any case means that signals arrive at the receiving antenna from two or more directions, one of which is usually the direct path from transmitter to receiver, and the others indirect paths along which the signal is reflected by objects which are off to the side of the direct path. In the a-m case, a reflected signal, which arrives after the direct signal, produces a single ghost or other repetition of the scene displaced to the right by a distance equivalent to the increase in delay over the longer reflected path. The intensity of the ghost depends on the relative strengths of the two signals. In the f-m case, the delay in the reflected signal can mean that two distinct carrier frequencies arrive at the receiver simultaneously. When this happens, the resulting beat between the two frequencies appears in the picture in the form of a moiré pattern, or multiple repeat after each object. The frequency of the repeats, or spacing of the moiré, is a function of the contrast between adjacent areas in the scene; hence it varies constantly with changes in the scene. The most objectionable moiré is produced by the blanking pulses because they usually represent the largest possible contrast. Where multipath transmission is not present, as in the case of point-to-point relaying systems using highly directive antennas, f-m may be used with excellent results.

American and British standards are at variance in the matter of polarity of transmission. In Great Britain, positive transmission is used. Positive transmission means simply that the carrier is modulated so that an increase in picture brightness brings about an increase in carrier amplitude. In negative transmission, adopted as standard in this country, an increase in picture brightness brings about a decrease in carrier amplitude. Thus sync. peaks represent maximum carrier. The principal points presented in favour of negative transmission are these:

1. An improvement in efficiency is realizable with negative transmission in the case of high level grid modulation. As stated previously, sync. pulses in the negative system represent maximum carrier. Because they are rectangular pulses, any saturation in an amplifier cannot affect their wave shape, but simply reduces their amplitude. Therefore it is possible to utilize the upper non-linear end of the modultion characteristic for the sync. pulses provided they are pre-emphasized so that the ratio of sync. to picture is correct in the modulated carrier. By using this non-linear part of the modulation characteristic for sync., the entire linear portion of the characteristic is reserved for the picture signal. In the positive system, on the other hand, where the picture whites represent maximum carrier, the non-linear end of the characteristic cannot be used at all without compressing the whites in the picture signal. Furthermore, 25% of the linear characteristic is unavoidably absorbed by the sync. pulses
2. Noise peaks which produce an increase in carrier will produce white spots, which may bloom (spread to abnormally large size) in the positive system, while they produce black spots in the negative system. Such black noise would normally be less objectionable than the white noise. In other words, the kinescope itself acts as a noise limiter in the negative system.
3. The average carrier power rating of the transmitter for given peak output is less in the negative system than it is in the positive system.
4. The signal produced in a negative transmission system lends itself to the use of simple a-v-c in receivers because sync. pulse peaks represent constant carrier level. Either peak-sensitive or keyed a-v-c circuits may be used which make use of this constant level as a reference. No such simple reference is available in the positive system.
5. Some receivers use an intercarrier sound system, i.e., a system in which the sound is obtained from the beat between the a-m picture carrier and the f-m sound carrier. This requires that the picture carrier never be driven to zero. In the positive transmission system, this requirement means that the black-reference carrier level must be raised at a further sacrifice of efficiency. In the negative system, only a slight reduction in video modulation amplitude is necessary.

Plate modulation, with its high efficiency and freedom from distortion, is not suitable for television because it is impractical to develop the necessary large amount of power in the modulator in the low impedance required by the broad band. In transmitters where high-level modulation is used the modulating signal is therefore applied to the grid circuit of the final r-f amplifier with consequent economy in the power requirements of the modulator.

Polarization

The question as to which polarization of the carrier waves is better, in the portion of the spectrum used for television, is probably impossible to answer conclusively from a theoretical study alone and was, therefore, investigated experimentally. A paper on this subject by Wickizer describes an investigation carried out at three frequencies, 49.5, 83.5 and 142 mc. around New York City. This investigation indicated preponderantly higher signal strength for horizontal polarization than for vertical. In some cases there was evidence that vertical polarization was preferable within a short radius of the transmitter, but the percentage of the total service area over which this was true was very small. The largest ratio of horizontal/vertical signal strength measured in this study was 9.8 db, the average being a little over 4 db. Thus, though the difference is substantial on the average, it is not great enough to make vertical polarization unusable.

There are other advantages to horizontal polarization among which are the following. Multipath reflections in general are produced by vertical surfaces such as the sides of buildings, cliffs, and grooves of trees. Both theoretical and experimental evidence shows that horizontally polarized waves are reflected from such surfaces less than those that are vertically polarized.

Investigation of the character of man-made noise signals has shown that relatively few have appreciable horizontally polarized components. Consequently there is less tendency to pick up noise in horizontal receiving antennas.

The construction of horizontal antennas is somewhat easier, in both transmitting and receiving cases, than the construction of vertical antennas. Horizontal dipoles are simple in construction and are easily balanced with respect to the earth's surface, the roofs of buildings, and supporting structures. Proper balancing of vertical dipoles is more difficult and they are usually abandoned in favor of vertical quarter wave radiators with artificial ground planes which are bulky and often difficult to handle.

The horizontal directivity of horizontal dipoles is a substantial aid in reducing the pickup of interfering signals at a receiving location.

These are probably the most important considerations which led to the adoption of horizontal polarization in this country.

Single side-band transmission

The development of what is usually called single side-band tansmission is probably one of the most valuable contributions to the television art, for it has made possible much more efficient use of the available channels, or from another standpoint, it has made possible the use of much smaller channels than would be possible otherwise.

It is well known that amplitude modulation of a wave produces a band of frequencies about the carrier as a centre, the boundaries of which are the sum and difference frequencies of the carrier and the maximum frequency of the modulating signal. In the broadcasting of sound by a-m it is considered desirable to include both the carrier and the upper and lower side-bands. This requirement makes necessary a total bandwidth equal to twice the highest modulating frequency. For example, for transmission of 10 kc. sound, a bandwidth of 20 kc. would be required.

In television transmission, it has been shown that partial suppression of one side-band does not detract from the quality of the picture in any way, but actually improves the results by making more space available for the other side-band. Side-band suppression is equivalent to moving the carrier toward one side of the transmission channel. In the RMA standards, the carrier is located 1.25 mc. from the low end of a 6 mc. channel, and approximately 4.5 megacycles above the carrier are allowed for the upper side-bands. The remaining part of the channel is allocated to the transmission of the sound. If all of both side-bands were to be transmitted, the channel would have to be increased in width by at least 3 mc., thus making a total channel width of 9 mc. or more. Such an increase in the width of television channels would work a serious hardship by reducing the number of channels available in a field where there is already considerable evidence of scarcity.

Single sideband transmission also permits economy in receiver design by allowing the use of a narrower i-f band. The i-f cutoff may be less abrupt so that phase shift in the upper end of the video frequency band is less severe.

Suppression of the lower side-band is accomplished in the transmitter by either of two means. In the case of high-level modulation, it is done in a filter having the required characteristics located in the antenna transmission line. In transmitters using low-level modulation, it may be done by proper tuning of the linear amplifiers following the modulated amplifier.

References

The preceding discussion is necessarily brief and cannot serve as much more than an outline for further reading. There are many papers dealing more comprehensively with the details and problems associated with the various parts of the television system. References to some of these are included in the following bibliography. Most of the papers referred to also include reference to others which, *in toto*, comprise a comprehensive list.

One book deserves special mention as a reference covering much of the engineering background of our television system. It is entitled, "Television Standards and Practice" (McGraw-Hill Publishing Co., 1943), and is essentially an abridged version of the proceedings of the National Television System Committee as edited by Donald G. Fink. It includes a statement of the standards recommended by the Committee to the Federal Communications Commission, discussion of the investigations on which the recommendations were based, and references to pertinent papers.

Acknowledgment

The author wishes to acknowledge the helpful criticisms and suggestions made by G. L. Beers, P. J. Herbst, H. N. Kozanowski, and W. J. Poch.

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Television Antennas and Transmission Lines*

By JOHN R. MEAGHER

Television Specialist, RCA Renewal Sales

PART III — CONCLUSION

Why does there appear to be so much contradictory information about television antennas?

In the first place, there are wide differences of opinion based on individual experience under different conditions.

For instance, technician "A" in a *strong* signal area is convinced that a certain antenna has broad-band response because it gives satisfactory reception on all of the TV areas.

But technician "B" in a *weak* signal area is convinced that this same antenna has narrow-band response, because he must use several of these antennas, each cut for a particular channel to obtain sufficient signal on each station.

Here are two different opinions. Is this a wide-band antenna, or a narrow-band antenna? How should the manufacturer rate it?

In the second place, almost all of the practical information that is available on antennas applies only to the resonant frequency. This information includes the widely-known values of dipole

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* Reprinted from RCA Radio Service News (March-April 1949) by courtesy of Radio Corporation of America.

characteristics, such as:
(a) impedance
(b) gain from the use of a reflector
(c) change of impedance, due to the reflector
(d) and the directivity pattern.

All of these characteristics become entirely different when the antenna is used to receive channels at other than the resonant frequency, and this is exactly the condition that applies in television, because in probably 80% of all TV installations, a single antenna is used to receive two or more stations on different frequencies.

Yet we continue to think and to talk about television antennas in terms of characteristics that apply only at the resonant frequency.

To illustrate this point, assume that a technician stops to admire an antenna installation. He sees that it is a plain dipole cut for one of the low channels, so he classifies it as having 70-ohm impedance. The transmission line is 70-ohm coax, so he is satisfied that it matches the antenna correctly for maximum power output. He knows that a plain dipole has a figure "8" reception pattern, with best reception at right angles to the rods. This dipole is broadside to the direction of the stations, so he is satisfied that it is oriented for best signal pickup.

Of course, the technician is correct on all of these points, providing the owner happens to be looking at the particular channel for which the antenna is resonant.

But suppose the owner looks at another low-band channel: The antenna impedance is no longer 70-ohms; it may be quite different. So the 70-ohm coax does not match the antenna.

Or suppose the owner looks at a high-band channel: The antenna impedance is not 70-ohms, it may be several hundred ohms; so again the 70-ohm coax does not match the antenna. The antenna does not have a figure "8" pattern; it has four major lobes at about 40° angles to the rods, so the antenna is not oriented for maximum signal on these channels.

In the third place, it does not make sense to consider the "gain" of an antenna without considering the loss in the type of transmission line that must be used with that particular antenna. We are *not* interested in the voltage up at the terminals of the antenna; we *are* interested in the voltage down at the input terminals of the receiver.

At frequencies used in radio and short-wave work, the signal loss in low-impedance coax is seldom a serious factor. But at the higher frequencies used in television, up to 216 megacycles, the gain achieved on one channel by matching a low-impedance antenna to a coax line may be outbalanced by the high losses in the coax, compared to the much smaller losses in 300-ohm ribbon line.

An additional reason for contradictory information is the fact that wide-band television antennas are not developed at an office desk with the aid of a handbook. They are developed by experimental methods with numerous measurements and comparisons every step of the way. Consequently there is disagreement in many cases between those who have become experts by reading the handbooks, and those who are actually conducting antenna measurements and comparisons under controlled conditions.

Comparison of TV antenna characteristics over all the TV channels requires a carefully chosen location, special set-up, special signal generators, special loads, and special measuring equipment; plus specialized knowledge and experience. Only a few agencies in the country are equipped at present for this work.

What is being done to clarify the situation?

This article is one step in an effort to provide impartial answers to questions about TV antennas.

We are not including any "how-to-build-it" information. Frankly, the best plan for anyone who wants to build antennas is to copy a design that has proved satisfactory for the particular set of conditions.

What is the best antenna for television?

No one antenna is "best" for all TV installations. The best antenna for a particular location depends on many factors.

What factors are involved in selecting the best antenna for a particular installation?

We will name seven factors as shown in Table 1.

1. The number of TV stations that are to be received, and their frequencies.
2. The field strength of each station at the receiving site.
3. The direction of each station from the receiver.
4. The reflection conditions (echoes or ghosts), and the direction of such reflections, for each station.
5. The interference conditions, r-f and electrical for each station.
6. The type and impedance of transmission line.
7. The price that the owner is willing to pay for material and labor to get good reception on each station.

Local Factors	Determine Reflections	**Antenna Requirements**
Stations, Frequencies		Lo-gain, Wide Band,
Field Strength		Hi-gain, Narrow Band,
Direction		One Antenna, or
Reflections		Several Antennas
Interference		Band Width
Line Impedance		Plain or Folded Dipole
Cost		Type and Height

Table 1.

How are these factors related to the antenna?

The number of stations, and their frequencies determine the bandwidth that the antenna must cover.

The signal strength of each station determines whether low-gain broad-band, high-gain narrow band, or a combination of such antennas must be used.

Direction, reflection, and interference conditions determine whether a single antenna, or several antennas are required so that each can be oriented for best signal and least reflections.

In addition, the line and receiver impedances may often determine whether plain dipoles or folded dipoles are required.

The cost naturally influences the final choice and may necessitate a compromise. The cost also determines the height which is usually a very important factor in weak-signal areas.

Let's consider a centre-city location with three stations, all strong, all in different directions, and severe reflections on all. What is the best antenna?

The safest answer, without making a survey at the particular location, is to use three separate antennas with reflectors, and three separate transmission lines running to a selector switch near the receiver, as shown in Figure 2.

Each antenna should be positioned on the roof and oriented for least reflections. The antennas should be plain dipoles for 50- or 70-ohm receivers and folded dipoles for 300-ohm receivers.

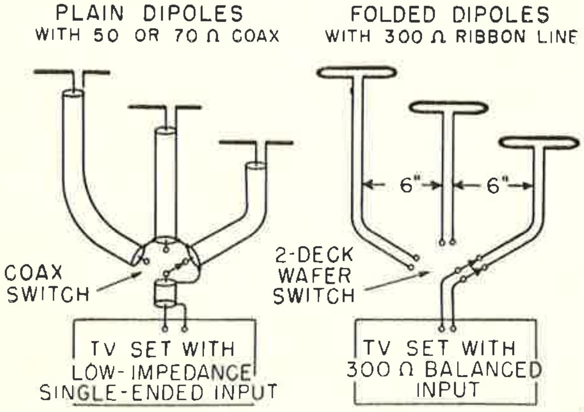

Fig. 2.

Consider a residential area in the city: Five stations, two on the high-band channels, all reasonably strong, all in the same general direction. What is the best antenna?

For receivers with 300-ohm input impedance, the best antenna in this location is an RCA-225A1 or 226A1 (the 225A1 has a reflector). This is a wide-band dipole. It was designed for these conditions. We will say more about it later.

Suppose we want to use a 50- or 70-ohm receiver in this same location: What antenna is required?

At least two antennas; one or more for the low channels and one for the high channels. Each to be plain dipole with reflector. The antennas may be connected together as specified by the manufacturer and fed into a 50- or 70-ohm line.

Let's move away from the city, say 30 miles: Three stations, one on the high band, all in the same general direction, no strong local reflections. What antenna is best?

Again, it is preferable to make a survey on the spot, but for receivers with 300-ohm input impedance, the RCA 225A1 dipole and reflector is recommended. For receivers with low-impedance input, use at least one antenna for the low-band, and one for the high-band. The antennas may be connected together as specified by the manufacturer, or separate lines can be run from each antenna to a switch near the receiver.

In very weak signal areas, what is the best antenna?

A separate "high-gain" narrow-band antenna for each of the low-band stations and one high-gain antenna for the high-band channels—(channels 5 and 6 are generally covered by a single high-gain antenna). Run a separate line for each antenna as shown previously.

It is preferable to use folded dipoles with reflectors and 300-ohm ribbon line for least loss of signal in the line.

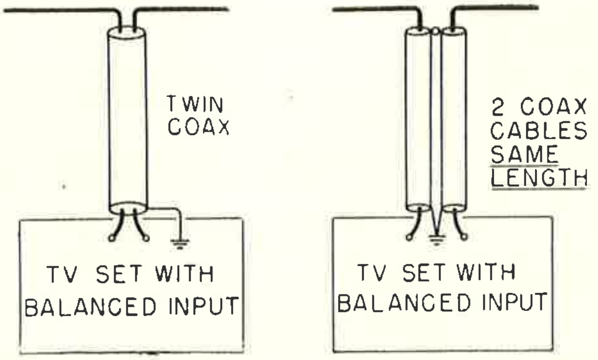

Fig. 3.

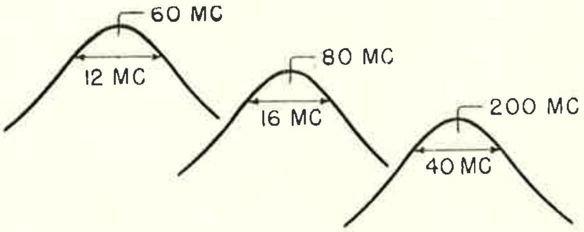

Fig. 4.

However, if noise pickup *on the line* is excessive, use plain dipoles and reflectors with 50- or 70-ohm coax. But decide this point carefully because coax has much higher loss of signal than 300-ohm ribbon line. When coax is used for a receiver having a balanced input, use twin coax as shown in Figure 3. The transmission line may be two-conductor shielded coax or two single coax cables *of the same length* run side by side with the braid outer conductor of each cable connected to the chassis close to the r-f input. The cables must be almost exactly the same length: If there is a difference of a half-wave length at some frequency, the signals from each cable arrive at the receiver in phase and cancel. A half-wave of coax on channel 13 is only about 16 inches long.

In extremely weak signal areas, a stacked array of two antennas may be used for any one station.

When you say a "separate" antenna for a low-band channel, do you mean an antenna that is cut to the correct length so it resonates at the frequency of the particular channel?

Yes. For optimum gain with a half-wave dipole on a particular low-band channel, the antenna should be cut to correct length so that it is *resonant* at the picture-carrier frequency; it should also be *matched* to the transmission line.

For satisfactory matching, use a folded dipole for 300-ohm line, and a plain dipole for 50- or 70-ohm line. The line impedance is, of course, determined by the receiver input impedance.

Such an antenna, with a correctly phased reflector, is classified here for simplicity as a high-gain narrow-band dipole.

How can such an antenna cover all the channels in the high band?

It is necessary to think in terms of percentage as shown in Figure 4. Assume an antenna with a bandwidth of 20%—

at 60 Mc/s, 20% bandwidth is 12 Mc/s
at 80 Mc/s, 20% bandwidth is 16 Mc/s
at 200 Mc/s, 20% bandwidth is 40 Mc/s.

Obviously for the same percentage bandwidth, an antenna covers more channels on the higher frequencies.

Other factors, such as the ratio of rod diameter to length, increase the bandwidth at the higher frequencies, so it is possible with one plain or folded dipole to cover all the highband channels.

For maximum gain on channels 2, 3, and 4, separate antennas are required, each being designed for its particular channel. Only one antenna is generally required to cover both channels 5 and 6.

A Double Conversion Ten- and Eleven-Metre Superhet*

By J. W. RICHARDT, Jr.

The Double-Conversion receiver utilizes intermediate frequencies of 1560 and 455 Kc/s. The tuned r-f stage and the first mixer employ Radiotron 6BH6 pentodes and the local high-frequency oscillator uses a 6C4 miniature triode. The rest of the valve lineup is as follows: 6BE6 second mixer and low-frequency oscillator; 6BJ6 455-Kc/s i-f amplifier; 6AL5 second detector, a.v.c. and automatic noise limiter; 6AQ6 first audio; and 6AK6 output.

The receiver uses a conventional a.v.c. circuit and a self-adjusting series-type noise-limiting diode (a.n.l.).

Layout considerations.

The cabinet of the receiver is a metal chassis 5 x 9½ x 3 inches with a perforated metal grille cut and fitted as a cover. Any convenient type of mounting arrangement can be used. The r-f head of the receiver is built on a small sub-chassis which is mounted to the left of the i-f strip. The other components of the receiver are built on a second sub-chassis. (See Fig. 4). The two sub-chassis units are drilled and fitted into the cabinet. When the mechanical work is completed, the units are removed for wiring. The front-panel layout and rear-terminal layout can be clearly seen in the photograph. Minor deviations in dimensions can be made.

The layout of the r-f head of the receiver is shown in the top view of the finished assembly (Fig. 4). The placement of all the parts as shown in the photographs should be closely followed.

Six holes have been drilled in the cabinet next to the r-f head to facilitate alignment of the stages with the unit in the cabinet. These holes may be covered with spring-clip type plugs if desired.

Wiring.

In wiring the r-f, i-f, and audio sections, all the leads are kept as short and direct as possible. Since the i-f amplifier has very high gain, it is advisable to shield all of the 455-Kc/s i-f transformer leads. The gain control, a.v.c. and a.n.l. on-off switches, and the other connectors should be permanently wired after the receiver has been tested and mounted in the cabinet.

Alignment.

The alignment procedure is conventional, but is briefly outlined for guidance. The first step is to check all operating voltages. These should be within ± 20% of the values specified on the chart. After the voltages have been checked, connect a signal

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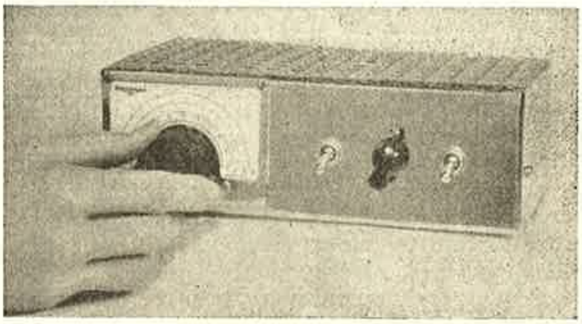


Fig. 1. Front view of the Tiny-Ceiver after final assembly.

generator to the control grid (grid No. 3) of the 6BE6 mixer and align the last two i-f transformers at 455 Kc/s with the a.v.c. in the off position.

After these stages have been carefully adjusted, reset the signal generator to 1560 Kc/s and adjust the low-frequency oscillator coil *L6* so that it is either 455 Kc/s above or below the signal from the generator. If the 8 μμF trimmer (*C21*) does not have enough range to facilitate this adjustment, the 62.5 μμF fixed capacitor (*C19*) will have to be replaced with one which will allow this adjustment to be made.

Next connect the signal generator to pin 7 of the first mixer valve, 6BH6, and align the first i-f transformer at 1560 Kc/s. When the alignment of this transformer has been completed, connect the signal generator, set at 26.8 Mc/s, to pin 1 of the 6BH6 mixer. With the variable capacitor *C10* set at full capacitance and the trimmer *C12* at about the halfway position, adjust the slug of oscillator coil *L5* so that the oscillator is 1560 Kc/s below the generator. After this adjustment has been completed, connect the generator set at 26.8 Mc/s to the antenna terminals and adjust the r-f and mixer slugs for maximum gain. Set the generator at 29 Mc/s and tune in this signal with the receiver tuning dial. At this point, the r-f mixer, and oscillator trimmers should be adjusted for tracking. It may be necessary to adjust the slugs and trimmers alternately to obtain a combination which will give the required bandwidth as well as proper tracking throughout the range of the receiver.

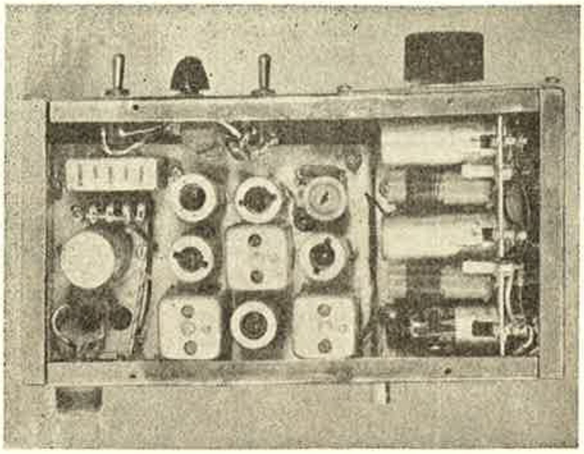


Fig. 3. Top view of the finished assembly showing placement of individual converter and i-f and a-f chassis within the compact case.

Power supply considerations.

Power requirements for this receiver are 250 volts at 58 milliamperes for the plate and screens and 6.0 volts at 1.5 amperes for the heaters. A power switch is not indicated because the author's installation provides a control box on the dash board for controlling the receiver and transmitter simultaneously. This switch, however, can be added to the receiver if desired. Most vibrator-type power supplies are filtered well enough so that no additional filtering is required. Should more be necessary, the filters

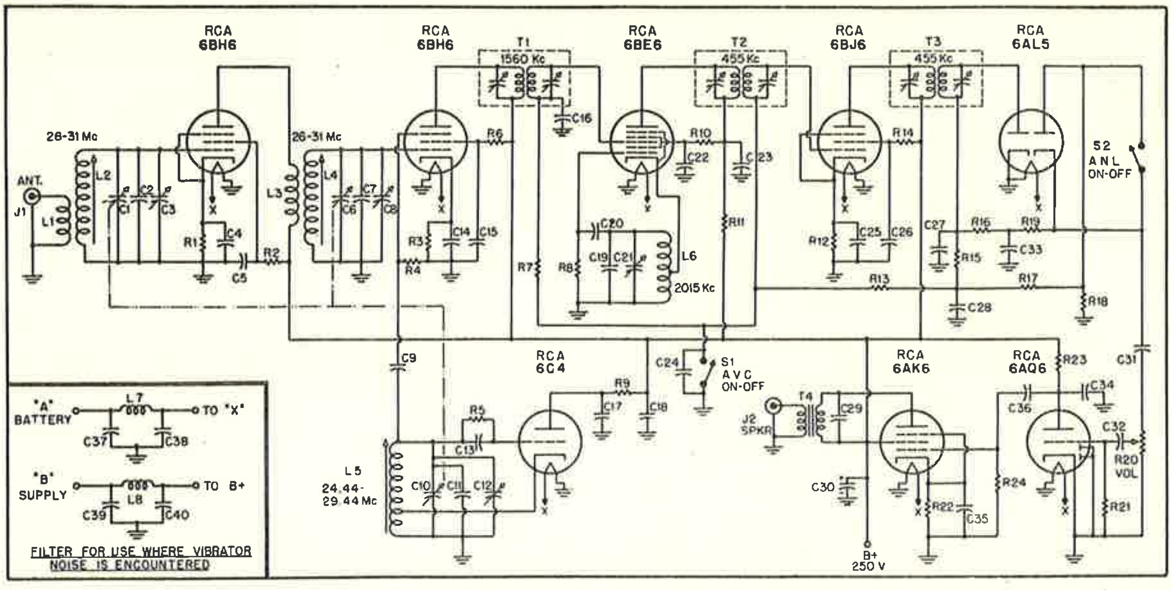


Fig. 2. Complete schematic of the Tiny-Ceiver 8 valve double-conversion 10-meter superhet.

shown in the sketch in Fig. 2 should be sufficient. The vacant space on top of the i-f chassis in front of the power connector was provided for this filter. In severe cases of interference, it may be necessary to enclose the filter components in a metal shield.

If a solid metal cover is used for the receiver, it is desirable to drill a number of holes in it, since even at the low power consumption of this receiver, its small size would make it run hot.

The sensitivity of the receiver is 1 microvolt for a power output of 0.050 watts into a 3.2-ohm load, or a gain of 118 db measured with a.v.c. and a.n.l. off. If the unit is constructed carefully and aligned properly, it should give many pleasant hours of mobile or home station operation.

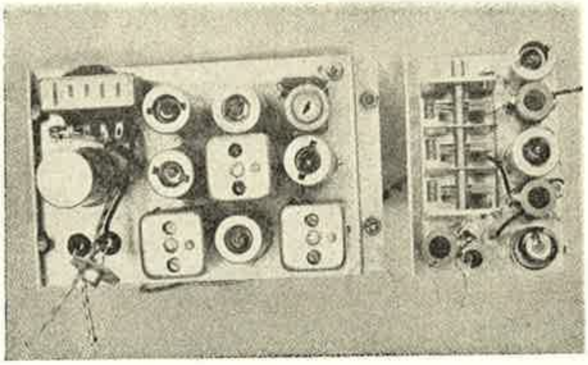


Fig. 4. Top view of the two units removed from the case shows placement of parts on both sub-chassis.

TINY-CEIVER VOLTAGE CHART

Valve	*Suitable Australian Types*	*Plate*	*Grid 2*	*Cathode*
6BH6 RF Amp	6AU6	250v	150v	1v
6BH6 First Mixer	6AU6	250v	50v	2v
6C4 H.F. Osc.	6C4	80-120v		
6BE6 2nd Mixer L.F. Osc.	6BE6	235v	100v	
6BJ6 I.F. Amp	6BA6	250v	125v	1v
6AQ6 1st Audio	6AV6	50v		
6AK6 2nd Audio	6AQ5	245v	250v	9v

All voltages measured with a vacuum tube voltmeter.

TINY-CEIVER PARTS LIST

C1, C6, C10	General Instrument Type 2801 3-gang variable, modified to have 2 rotor plates and 2 stator plates. (Note: Frame of this capacitor measures $1\frac{3}{8}''$ x $1\frac{3}{8}''$ x $2\frac{7}{16}''$ long.)
C2, C7, C11	10 μμF silver mica El Menco CM 20-100.
C3, C8, C12, C21	1-8 μμF variable tubular Erie 532.
C4, C5, C17, C29	5000 μμF Centralab Hi-Kap DA 048-001A.
C9	10 μμF Erie Ceramicon Type K.
C13, C20, C27, C28, 34	100 μμF silver mica El Menco CM 20-101.
C14, C15, C16, C18, C22, C23, C25, C26, C31, C32, C36	0.01 μF Centralab Hi-Kap.
C19	62.5 μμF Erie Ceramicon Type L.
C24, C33	0.01 μF 400v paper. Sprague 68P8.
C30	20 μF 450v (Part of Mallory FP332—2 10 μF 450-volt sections in parallel). See C35.
C35	10 μF 25v (Part of Mallory FP332). See C30.
C37	0.001 μF.
C38	0.25 μF 25v.
C39	0.001 μF.
C40	8 μF 350v.
L1	2 turns #20E close-wound over cold end of L2.
L2, L4	9 turns #20E, spaced to occupy winding area of National XR-50 coil form.
L3	5 turns #36DSC, inter-wound at cold end of L4.
L5	Same as L2 plus tap at 3 turns from ground end.
L6	Oscillator coil—Miller 5481-C.
L7	20 turns #14, $\frac{1}{2}''$ diameter-close-wound air core.
L8	2.5 mH RFC.
S1, S2	SPST Switch.
R1	100 ohms.
R2, R4, R14, R15	47000 ohms.
R3, R11	1000 ohms.
R5, R8, R9, R10	22000 ohms.
R6	150000 ohms.
R7	100000 ohms.
R12	82 ohms.
R13	2.2 megohms.
R16	1 megohm.
R17, R18, R23	270000 ohms.
R19	820000 ohms.
R20	500000-ohm potentiometer.
R21	10 megohms.
R22	560000 ohms.
R24	470000 ohms.
J1	Amphenol BNC 31-003.
J2	Bud., JP-248.
T1	1560-K/c i-f transformer—Miller Type 012W1.
T2	455-K/c i-f transformer—Miller Type 012C1.
T3	455-K/c i-f transformer—Miller Type 012C4.
T4	Output transformer for matching impedance of 6-8 ohm coil to 10,000 ohm load.
Dial Assembly	National MCN.

All capacitors are 400-volts unless otherwise specified.
All resistors are ½ watt unless otherwise specified.

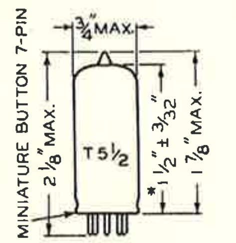
The make and type numbers of the above parts are the ones used in the original construction. Substitutions may be made if physical and electrical characteristics are similar.

Radiotron Type 2D21 Thyratron

HOT-CATHODE GAS-TETRODE, MINIATURE TYPE

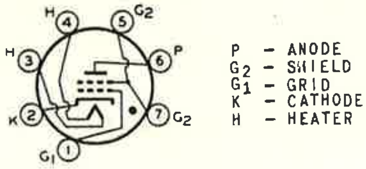
(Reprinted by courtesy of Radio Corporation of America)

The 2D21 is a sensitive, four-electrode thyratron of the indirectly heated cathode type designed for use in relay applications. It has a steep control characteristic (high control ratio) which is essentially independent of ambient temperature over a wide range, extremely small preconduction of gas-leakage currents right up to the beginning of conduction, very low grid-anode capacitance, and very low grid current. Having very low grid-anode capacitance, the 2D21 is not appreciably affected by line-voltage surges; and having very low grid current, it can be used with a high value of resistance in the grid circuit to give high circuit sensitivity. In a high sensitivity circuit, the 2D21 can be operated directly from a high-vacuum phototube.



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

Bottom View of Socket Connections



GENERAL

Heater Voltage (a.c. or d.c.)*	6.3	volts
Heater Current	0.6	ampere
Direct Interelectrode Capacitances (with no external shielding):		
Control Grid — Anode	0.02	μμF
Input	2.4	μμF
Output	1.6	μμF
Valve Voltage Drop (Approx.)	8	volts
Control Ratio at Breakdown (Approx.):		
Control Grid — Anode (Shield-grid volts = 0)	250	
Shield Grid — Anode (Control-grid volts = 0)	1000	
Maximum Overall Length	2-⅛"	
Maximum Seated Height	1-⅞"	
Maximum Diameter	¾"	
Bulb	T-5-½	
Base	Miniature Button 7-Pin	
Mounting Position	Any	

MAXIMUM RATINGS AND TYPICAL OPERATING CONDITIONS

Maximum Ratings Are Absolute Values.

Peak Forward Anode Voltage	650 max.	volts
Peak Inverse Anode Voltage	1300 max.	volts
Shield-Grid (Grid No. 2) Voltage	−100 max.	volts
Control-Grid (Grid No. 1) Voltage	−100 max.	volts
Peak Heater-Cathode Potential:		
Heater negative with respect to cathode	100 max.	volts
Heater positive with respect to cathode	25 max.	volts
Peak Cathode Current	500 max.	mA
Average Cathode Current**	100 max.	mA
Control-Grid Circuit Resistance	10 max.	megohms
Ambient Temperature Range	−55 to +90	°C
Typical Operation in Relay Service:		
Anode Voltage (r.m.s)	400	volts
Shield-Grid Voltage	0	volts
Control-Grid Bias Voltage (r.m.s)♯	5	volts
Control-Grid Signal Voltage (Peak)	5	volts
Control-Grid Circuit Resistance	1	megohm
Anode Circuit Resistance♯♯	2000	ohms

* Heater voltage must not deviate more than 10% from the rated value, and must be applied at least 10 seconds before application of anode voltage.

** For an averaging period of 30 seconds.

♯ Approximately 180° out of phase with the anode voltage.

♯♯ Sufficient resistance, including the valve load, must be used under any conditions of operation to prevent exceeding the current ratings.

Installation.

The base of the 2D21 fits a button-base socket which may be installed to hold the valve in any position. The centre hole of the socket designed for the button base provides for the possibility that this valve may be manufactured with the exhaust tube tip at the base end. For this reason, it is recommended that in equipment employing this valve, no material be permitted to obstruct the centre hole of the socket. The socket should be made of an insulating material having leakage current low in comparison with the pre-conduction grid current of the 2D21.

The heater is designed to operate on either a.c. or d.c. at 6.3 volts. When it is operated on a.c. with transformer, the heater winding should operate the heater at its recommended voltage value for full-load operating conditions at average line voltage. When it is operated from a battery with charger attached, provision must be made through the use of regulation-control devices to hold the voltage across the heater to the recommended value. Regardless of the heater-voltage supply used, *the heater voltage must never be allowed to deviate more than 10% from its rated value.* Heater operation outside of this voltage range will impair valve performance and may cause valve failure. Low heater voltage causes low cathode temperature with resultant cathode sputtering and consequent destruction of the cathode; high heater voltage causes high cathode temperature with resultant heating of the grid and consequent grid emission which produces unpredictable shifts in the critical grid voltage for conduction.

The cathode should be allowed to reach normal operating temperature before anode voltage is applied to the valve. *The delay period should not be less than 10 seconds after application of heater voltage.* Unless this recommendation is followed, the cathode will be damaged.

When the 2D21 is operated from a transformer, the cathode should preferably be connected directly to the mid-point or to one side of the heater circuit. When the 2D21 is used in equipment employing a storage battery for the heater supply, the cathode circuit is usually tied directly to the negative side of the heater supply. In circuits where the cathode is not connected directly to the heater, the potential difference between them should never exceed the recommended peak potentials shown under MAXIMUM RATINGS.

The shield grid is normally connected to the cathode. It may, however, be used as a control electrode because the control characteristic of grid No. 1 may be shifted by varying the potential of grid No. 2 (shield). As the shield is made negative, the grid No. 1 characteristic is shifted in the positive direction. With –2 volts on grid No. 2, the grid No. 1 characteristic lies completely in the positive region. The use of grid No. 2 as the control electrode (with grid No. 1 tied to cathode) has the advantage of increased sensitivity but consideration must be given to the higher preconduction current, higher capacitance to anode, and less stability of operation.

A control-grid resistor having a value as high as 10 megohms to give circuit sensitivity can be used with the 2D21 because its control-grid current is very low. However, when a high value of grid resistor is used, care should be taken to keep the valve base and socket clean and dry in order to make the effect of leakage currents between the control-grid base pin and anode base pin very small.

When the 2D21 is operated with a high value of grid resistor and a.c. voltage on the anode, the circuit capacitance between control grid and anode should be kept low by placing the grid resistor directly at the socket grid terminal, by connecting both shield-grid terminals (pins 5 and 7) to the cathode (pin 2) at the socket, and by using a close-fitting shield connected to the cathode terminal.

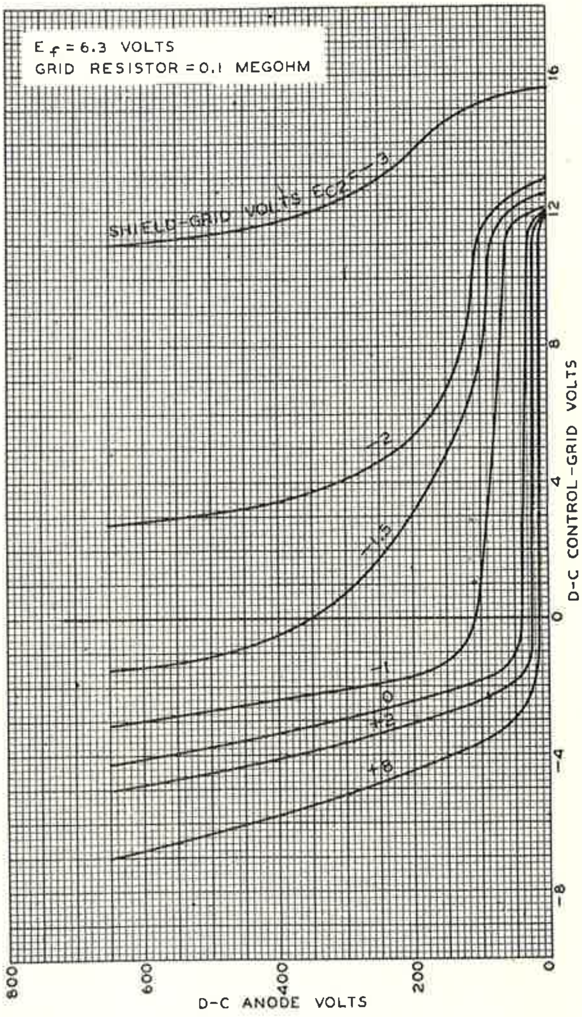


Fig. 1.—Average control characteristics.

Application.

The 2D21 operates by virtue of the fact that, for any specific shield-grid potential and positive anode potential, there is a critical value of control-grid voltage. If the control grid is maintained more negative than this critical value the valve does not conduct and the anode current remains zero. If the control grid is made less negative than the critical value, the valve will conduct and the anode current assumes a value determined by the applied anode potential and the impedance in the anode circuit. In the conducting condition, the 2D21 has a voltage drop which is quite low and substantially independent of both anode current and control-grid bias. Conduction may be stopped and grid No. 1 allowed to regain control by reducing the anode voltage to zero or making it negative. A family of curves showing the relations between control-grid voltage, shield-grid voltage, and anode voltage is given in Fig. 1.

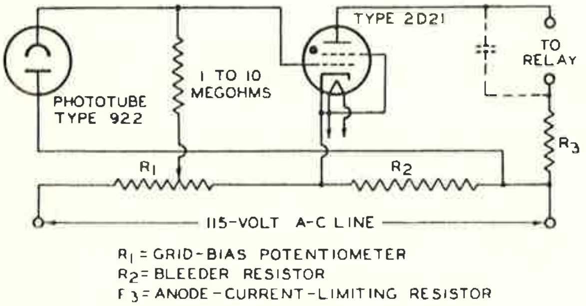


Fig. 2.—Typical light-operated relay circuit.

The 2D21 is designed primarily for use as a relay valve in applications where the operating frequency is relatively low. Because of its small size and high sensitivity capabilities, the 2D21 is especially suitable for very compact equipment.

In the design of equipment to utilize the 2D21, consideration should be given to its maximum ratings which are on an *absolute maximum* basis. In order not to exceed these absolute ratings, the equipment designer has the responsibility of determining an average design value for each rating below the absolute value of that rating by an amount such that the absolute values will never be exceeded under any usual conditions of supply-voltage variation, or manufacturing variation in the equipment itself.

The 2D21 can be operated with an anode voltage obtained from either an a.c. or a d.c. source. When a d.c. supply is used, the circuit has a lock-in feature because the grid loses control when conduction starts. In order for the grid to regain control and to restore the valve to the non-conducting condition, the anode voltage must be removed momentarily. When an a.c. supply is used, the circuit has no lock-in feature because the anode becomes negative during the negative half of the a.c. cycle and allows the grid to resume control. In a.c. operation, control of the average anode current may be accomplished by relative phasing of the control-grid, shield-grid, and anode potentials. A typical light-operated relay circuit is shown in Fig. 2.

The electrode structure of the 2D21 provides a very low grid-anode capacitance with the result that the valve is insensitive to line-voltage surges. This characteristic also insures that only very low values of capacitive current will flow through the grid resistor. Such a characteristic is very desirable because the control-grid bias required to prevent conduction is increased by an amount equal to the peak *IR* drop across the grid resistor. In some cases, a neutralizing circuit may be used to cancel the effect of the *IR* drop.

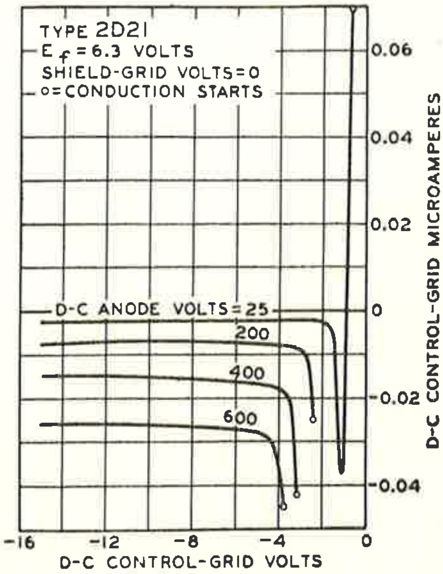


Fig. 3.—Average grid characteristics before anode conduction.

The effective control-grid bias obtained when a high value of grid-circuit resistor is used is influenced by the grid currents flowing before and during conduction. The magnitude of these currents before conduction is shown in microamperes in Fig. 3 and during conduction in milliamperes in Fig. 4.

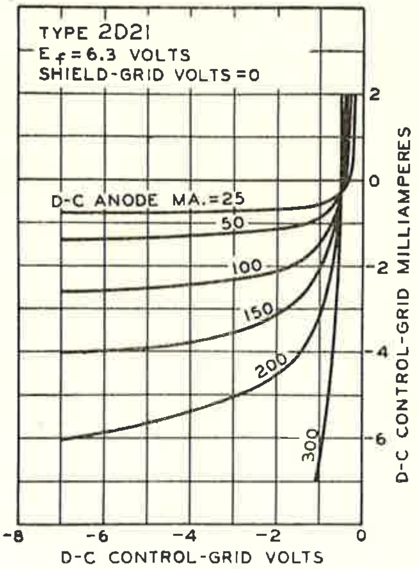


Fig. 4.—Average grid characteristics during anode conduction.

The critical grid voltage for any given anode voltage is that value of grid voltage which will permit conduction to start. The value of critical grid voltage is affected by several factors including: the operating anode voltage, shield-grid voltage, variation of heater-supply voltage, value of grid resistor, and individual valve variations both initially and during life. In Fig. 5 is shown the range of critical grid voltage based on the combined effect of individual valve variation, and variation throughout valve life for a grid resistor value of 0.1 megohm and also for a value of 10 megohms. About 10% of the total variation range for any particular operating condition is attributable to a heater-voltage variation of ± 10% from the rated value.

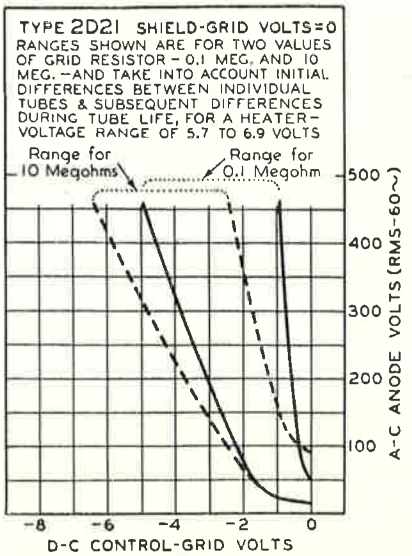


Fig. 5.—Operational range of critical grid voltage.

The equipment designer should give careful consideration to the range values shown in Fig. 5. From them he can determine for specified operating conditions not only the proper value of control-grid bias to prevent conduction until it is desired but also the magnitude of the signal (triggering) voltage to cause conduction. Ample triggering voltage should always be provided to insure anode conduction even under the worst operating conditions to which the equipment will usually be subjected.

In addition to the factors contributing to the range values shown in Fig. 5, the effect of high ambient temperature may become important. When the 2D21 is used with a low value of grid resistor, the effect on critical grid voltage of ambient temperature changes between −55°C and +90°C is negligible. However, when the grid resistor has a high value, temperature changes within the operating range of −55°C to +90°C may cause a change of 20% in the critical grid voltage. Most of this change occurs at the higher temperatures and is caused by leakage outside the valve. Therefore, under operating conditions involving a high value of grid resistor and high ambient temperature, it is essential that precautions be taken to minimize leakage.

The maximum average cathode current is the highest current which can be drawn continuously through the valve and is based on the allowable heating of the valve. The average current should be determined on the basis of 30-second operation of the valve. If the cycle of operation during the 30-second period is rapid, the average current can be read on a d.c. meter. If the cycle is long, it is necessary to calculate the average current from readings taken during the 30-second period.

NEW RCA RELEASES

Radiotron type 6AX5-GT — is a full-wave, vacuum rectifier especially designed to provide for the economical design of a.c. receivers and to facilitate the design of automobile receivers having high power output.

The 6AX5-GT features a unipotential cathode having a 6.3-volt heater and a relatively wide plate-cathode spacing chosen to minimize sputter and yet provide good regulation. The heater provides for economical design of a.c. receivers because it can be operated from the same transformer winding that supplies other 6.3-volt heater types in the receiver Furthermore, use of the 6AX5-GT which has the same heating time as other heater-cathode types in the receiver, limits the voltage appearing across the filter capacitors during the warm-up period. Consequently, electrolytic filter capacitors having lower peak voltage ratings than required for a filament-type rectifier tube can be used with the 6AX5-GT. *

The 6AX5-GT is also especially useful in supplying the d.c. requirements of automobile receivers having high power output, because it is capable of delivering about 75 per cent. more d.c. current than other commonly used automobile rectifier tubes.

Radiotron type 6AS6 — is a sharp-cutoff pentode of the 7-pin miniature type. Designed so that grid No. 1 and grid No. 3 can each be used as independent control electrodes, the 6AS6 is especially useful in gated amplifier circuits, delay circuits, gain-controlled amplifiers, and mixer circuits. It is also suitable for use as an r-f amplifier at frequencies up to about 400 megacycles per second.

Refer to 'Editors' & Engrs.' Radio Handbook, 13th Edition, P. 588.

Versatile Grid Controlled Power Supply*

By J. H. OWENS *and* G. D. HANCHETT

A power supply that will deliver up to 200 mA at any voltage from about 50 to 400 volts! Does this appeal to you? If it does, and if you want this convenience at low cost without the losses of tapped bleeder resistors or expensive variable transformers, but with good voltage regulation, just by setting a small potentiometer—here's how!

It's done with grid-controlled rectifiers, commonly known as thyratrons. And what are they? They are simply rectifiers containing gas to reduce the voltage drop and to improve the efficiency, and having one or more grids interposed between the plates and cathodes to control the start of plate current flow.

In the power supply to be described, a pair of Radiotron-2050's are used to deliver the current at the desired voltage. Within its capabilities a unit like this permits the convenient reduction of power during tune-up of that new rig, and a moment later, its operation at full input. For experimental work, such a unit is an invaluable laboratory tool.

Theory of operation.

Refer to Figure 1 which illustrates the critical control characteristics of a thyratron valve. The heavy solid line represents the a.c. voltage impressed on the plate of one of the rectifiers in a full-wave circuit; and the dashed line represents the critical instantaneous grid voltage that must simultaneously be put on the control grid of this valve to prevent it from ionizing or "firing". In this condition, neither valve will pass plate current, and the output of the rectifier will be zero.

The dotted line represents an inphase voltage which, if impressed upon the grid of the thyratron, will cause it to fire at the start of the cycle and conduct throughout its duration, at which time the plate voltage drops to zero and the valve deionizes, thereby restoring grid control. In this condition, both of the valves act like regular diode rectifiers and deliver maximum power to the load.

Figure 2 shows the relationship of plate voltage versus critical-grid-voltage when a voltage of 90° displacement is impressed on the grid. The arrows indicate the instant where the actual negative grid voltage becomes more positive than the critical voltage for the applied plate voltage. At this point, ionization occurs, and current flows during the remaining part of the cycle as indicated by the shaded area. The d.c. output voltage delivered by the filter will be about three-quarters of the maximum obtainable. From this, it can be seen that variations in phase between applied anode voltage and grid voltage will produce more or less rectifier output. Carried to extremes, this means either full-voltage at full conduction or zero-voltage at zero conduction.

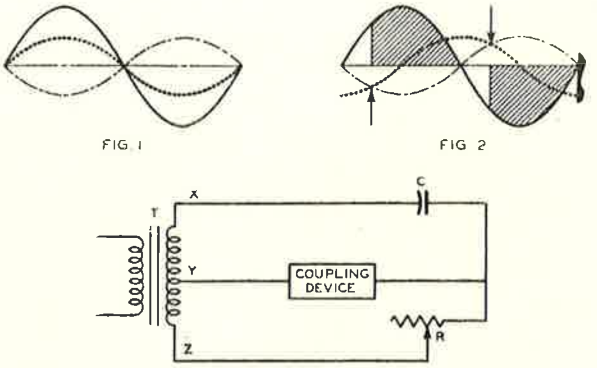


FIG 1 FIG 2

FIG. 3

Control characteristics of thyratron tubes and a basic phase controlling network.

Phasing circuit.

Figure 3 shows the basic phase-controlling network. A transformer (T) has a centre-tapped secondary winding connected to the coupling device. If the centre-tap (Y) is used as a zero point, the voltage on one side (X) is, of course, 180° out of phase with the voltage on the other side (Z). Then, if the resistance (*R*) is high compared with the reactance of the capacitor (*C*), the coupling device is effectively connected across the upper half of the secondary (XY), and the voltage across it is in equal phase. But if the resistance (*R*) is low compared with the reactance of the capacitor (*C*), the coupling device is effectively connected across the lower half of the transformer secondary (YZ), and the voltage across it is now of reversed phase. In this position, the capacitor (*C*) is connected across the entire winding (XZ), but its reactance is high compared with the reactance of the transformer secondary, and no ill-effects are produced. Intermediate values of resistance (*R*) will cause intermediate phase differences across the coupling device, and will provide the control that is so desirable.

Construction details.

Figure 4 shows the complete circuit of the unit. A separate filament transformer is used to heat the filaments of the Radiotron-2050's, light the pilot lamps, and supply the phasing voltage. A low-cost, unmounted transformer is used, and is located

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* Reprinted from Ham Tips (Nov.-Dec., 1946) by courtesy of Radio Corporation of America.

underneath the chassis. The 6.3- and 5-volt windings on the power transformer are left free and available for heating the filaments of a wide variety of valves operated from the power supply.

Since a capacitance-input filter is employed, a resistor is used in series with the input capacitor to limit the peak current to the maximum rating. The value of this series resistor is approximately equal to 0.9 ohm per r.m.s. volt of ½ the total secondary voltage of the supply transformer. For an 800-volt centre-tapped secondary, the value of the resistor is approximately 800/2 x 0.9, or about 360 ohms.

The 100,000-ohm grid resistors are used to prevent excessive 2050 grid current and consequent loading of the phasing transformer. It may be necessary to reverse the transformer grid connections to get proper phase relation so that firing is prevented when the potentiometer is in a maximum-resistance position.

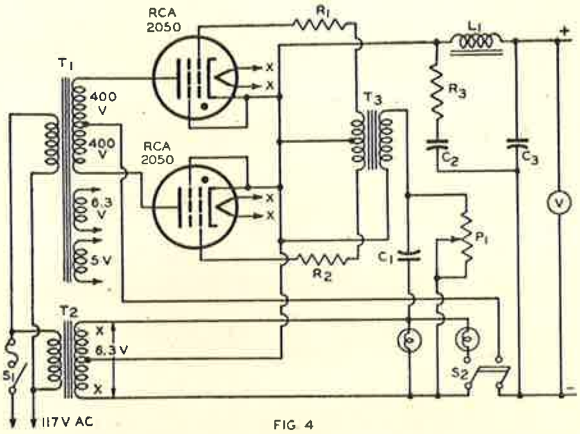


FIG 4

Fig. 4.—Power supply schematic.

Don't worry about the 10-μF electrolytic capacitor being used in an a.c. circuit. Its reactance, or capacitance is practically the same in both directions, and the peak voltage of less than 10 is not high enough to cause it to be damaged.

The phasing transformer is a small-size audio unit, single plate to push-pull grids. It is mounted underneath the chassis in a convenient position.

Two switches are used to cut the unit on and off. *S1* puts voltage on all tube heaters, and *S2* delivers high voltage to the rectifiers. *S2* should never be closed until the 2050 heaters have had a warm-up of at least 10 seconds, and preferably 30 seconds.

Operating precautions.

Because a capacitance-input filter is used, the voltage regulation will compare favourably with regular high-vacuum rectifiers. Therefore, the output voltage will rise considerably if the load is removed. The use of a swinging choke at the input to the filter will provide equivalent voltage regulation to standard circuits, but it will also limit the d.c. output voltage to approximately 90% of the r.m.s. voltage of one-half the high-voltage transformer winding.

Benefits.

All we can say here is that once you have built and used one of these grid-controlled thyratron power-supplies, you will wonder how you ever managed to do without it in the past.

PARTS LIST

T1	Power transformer, 800 V., centre-tapped secondary, 200 mA capacity.
T2	Filament transformer, 6.3 V., 1.2 amps.
T3	Interstate audio transformer, single-plate to P-P grids.
C1	10 μF, 150 V., electrolytic.
C2, C3	8 μF each, dual electrolytic, 450 V. working.
R1, R2	100,000 ohms, ½ watt, carbon.
R3	360 ohms (approx.), 25 watt, wire-wound (see text).
P1	10,000 ohm wire-wound potentiometer.
L1	Choke, 10 henries (approx.), 200 mA.

The Radiotron 7-pin miniature 2D21 can also be used in this circuit, provided the peak plate current is limited to half that of the 2050. This is achieved by suitably increasing the current-limiting resistance in series with the input condenser. The d.c. current remains 200 mA.

See data on P66

APPARENT DUPLICATION OF TYPE NUMBERS

TYPE 6AR7-GT.

Questions have been asked by those who have noticed type 6AR7-GT listed as a double diode in the A.R.R.L. Handbook and in other overseas publications. The answer is that this type number was reserved by R.M.A. for the General Electric Company (U.S.A.) in 1945, but registration was not carried out and the request for reservation was subsequently cancelled. This type number, 6AR7-GT, was subsequently registered by the Radio Manufacturers Association (U.S.A.) on application by Amalgamated Wireless Valve Company, for a duo-diode-pentode manufactured in Australia. The use of this type number for any other valve is erroneous. *

RADIOTRON

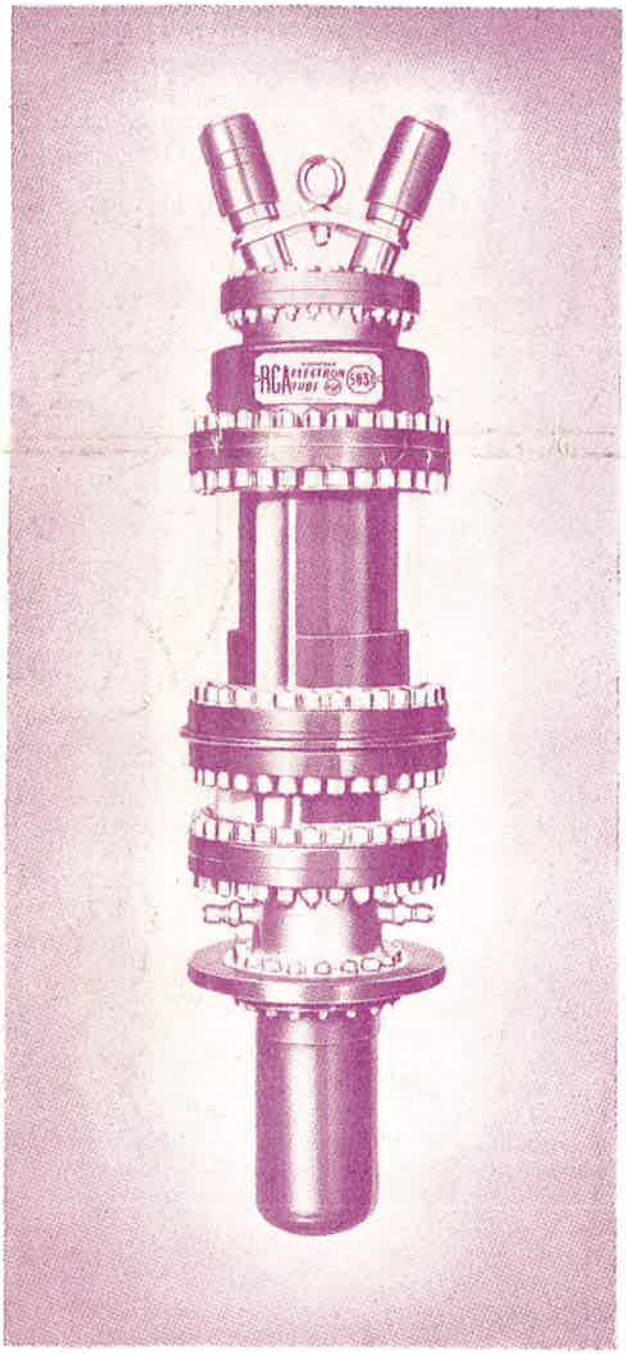
5831

SUPER-POWER BEAM TRIODE

650-Kw Plate Input
Very High Power Sensitivity

Water Cooled
TENTATIVE DATA

150-Kw Plate Dissipation
Thoriated-Tungsten Filament



RCA-5831 is a water-cooled beam triode of unique design capable of generating several hundred kilowatts of power at high efficiency and with exceptionally low driving power. It is intended primarily for use as a class C rf power amplifier, either modulated or unmodulated, but is also useful as a class B af power amplifier and modulator. In unmodulated class C service, the 5831 has a maximum plate voltage rating of 16000 volts, a maximum plate input of 650 kilowatts, and a maximum plate dissipation of 150 kilowatts. It can be operated with maximum rated plate voltage and plate input at frequencies up through the "Standard Broadcast Band" and much higher. The limitations for operation at the higher frequencies and at higher power have not yet been determined but requests for information on specific applications will be welcomed.

The 5831 is unique in that it features a symmetrical array of unit electron-optical systems embodying a mechanical structure which permits close spacing and accurate alignment of the electrodes to a degree unusual in high-power tubes. Ducts for water cooling the plate and the beam-forming cylinder are built in and have simplified hose connections. The grid-terminal flange requires a water-cooled connector. Because of the electron-optical principles incorporated in its design, the 5831 has low grid current and hence requires less than 2 kilowatts of driving power.

Other features of the 5831 include a multi-strand, thoriated-tungsten filament for economical operation as well as high emission capability, low-inductance rf leads and flange terminals, and compactness—features contributing to the overall suitability of the 5831 in high-efficiency, high-power applications.

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