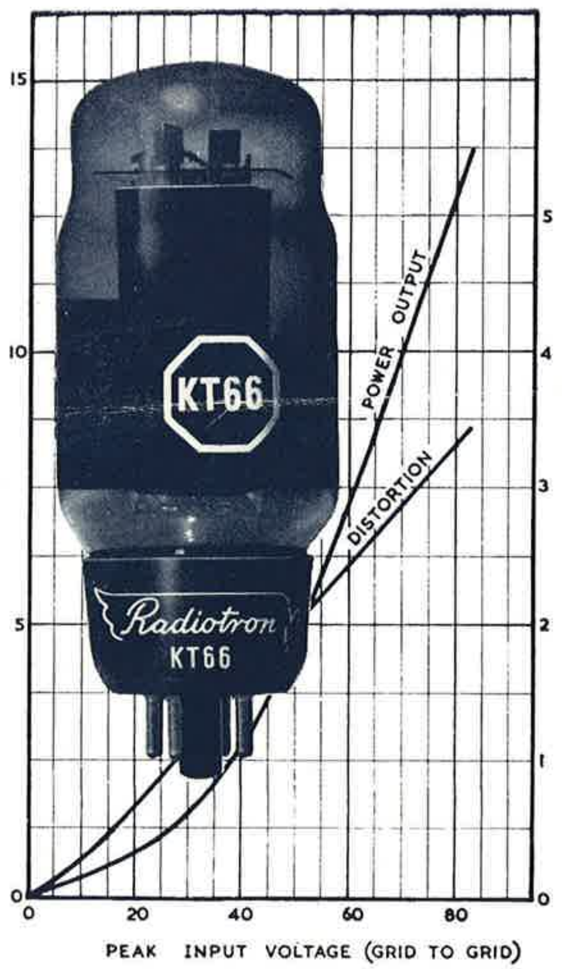
Radiotronics

AN A.W.V. TECHNICAL PUBLICATION DEVOTED TO RADIOTRONS AND THEIR APPLICATION

Number 142
April, 1950



Push-pull KT66's Triode-connected give high output with low distortion.

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RESEARCH • DESIGN • MANUFACTURE

NUMBER 142 APRIL 1950

RADIOTRONICS

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CIRCUIT LAB. REPORT **NUMBER 1**

Radiotron Receiver RC55*

FEATURING 7-PIN A.C. MINIATURE VALVES

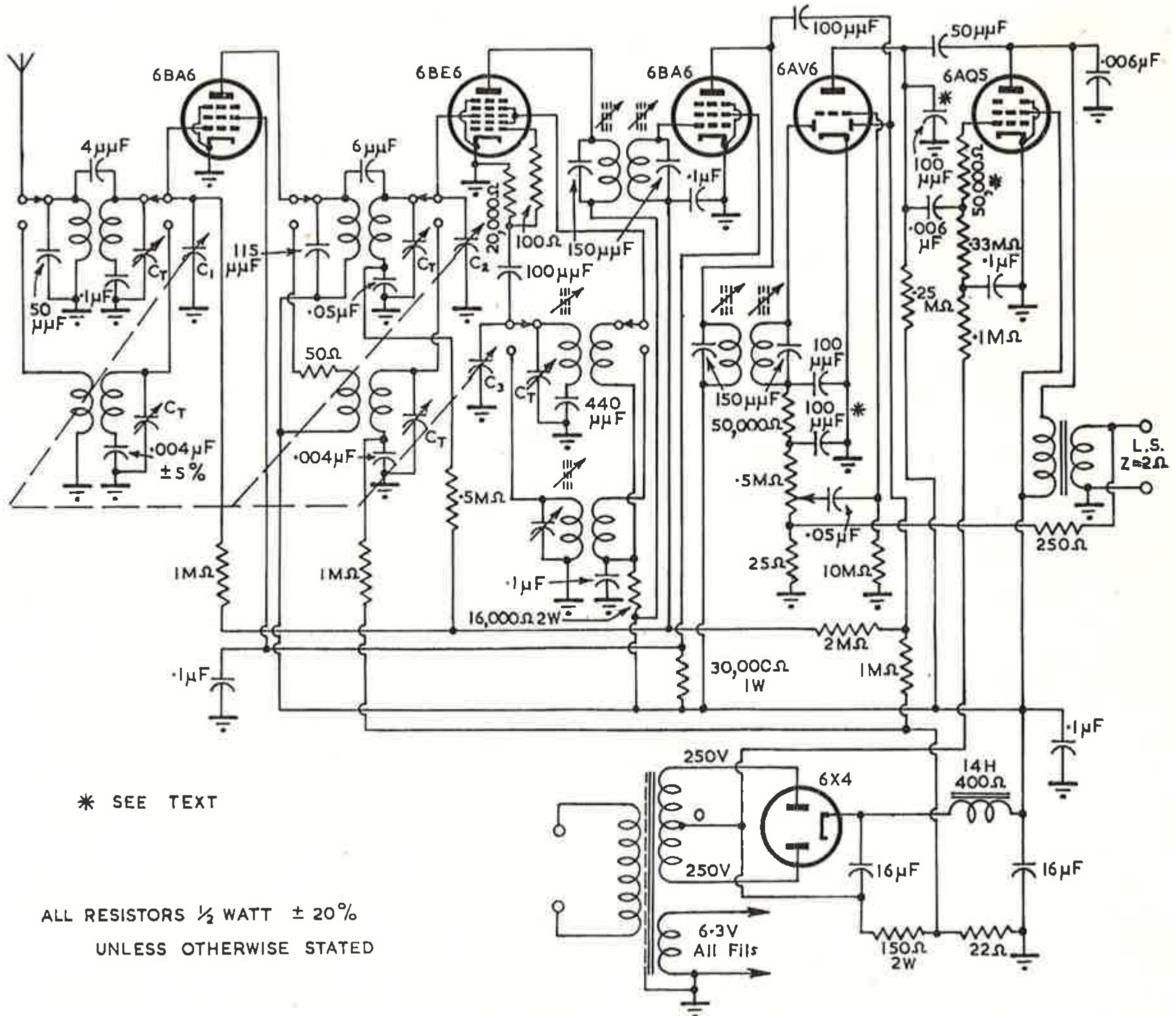
The Radiotron receiver RC55 is a high performance dual-wave receiver using the 7 pin a.c. miniature range and featuring harmonic mixing in the 6BE6 converter on the S.W. band.

The relatively low cost of this converter combined with the few components required is particularly attractive where economy in design is essential. As an indication of this, although two electrolytic condensers only are used in this receiver, no flutter occurs at the high frequency end of the S.W. band when the receiver is tuned slowly through an input signal of 0.1 volt with the volume control set at maximum.

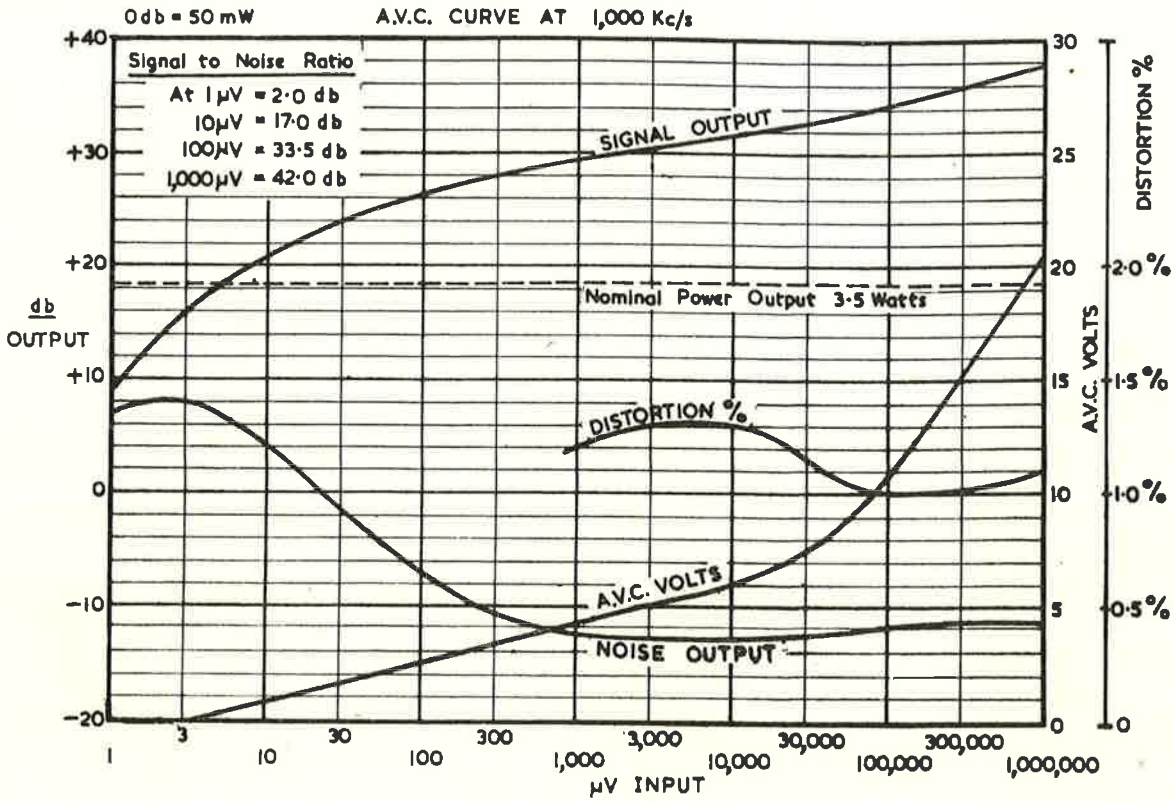
In order to simplify wiring, all cathodes are earthed and, with the exception of the 6AV6 triode section, the negative bias voltage of all valves is obtained from back bias resistors. The delay voltage for the a.v.c. diode circuit is taken from the −1.5 volt tap on the back bias network.

The screen voltage for the r-f and i-f amplifier valves is obtained from B+ through a common dropping resistor and decoupling condenser. In order to eliminate all traces of flutter under all normal conditions of operation the screen voltage of the converter is supplied through a separate dropping resistor. A common screen resistor may be used for the three stages, in which case performance on the broadcast band is unchanged, while at the high frequency end of the short-wave band flutter does not occur for any input signal until the power output exceeds 3 watts. At lower frequencies on the short-wave band flutter, of course, is better.

At the maximum available gain of the 6BA6 the sensitivity of the receiver on the broadcast band is excessive and, for this reason, the r-f stage gain is



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



Output, noise, a.v.c. and distortion vs. input voltage at 1000 Kc/s.

reduced by shunting the high impedance primary of the r-f transformer by a 115 μμF condenser.

On the S.W. band, the 50 ohm resistor in the primary of the low impedance r-f transformer equalises the gain over the band and prevents instability at the high frequency end.

Negative feedback is taken from the voice coil of the speaker and applied to the bottom of the volume control. This type of feedback is particularly desirable in that at low settings of the volume control a large amount of negative feedback is available, whereas with the control turned to maximum, the loss in gain is only 5 db.

A fixed amount of high frequency feedback is provided by the 50 μμF capacitor from the anode of the 6AQ5 to the anode of the 6AV6 which reduces the amount of noise on the side bands when tuning the receiver to a station. Although this feedback limits the high frequency response of the audio stages, this is still adequate when considering the high frequency attenuation brought about by the selectivity of the r-f and i-f circuits.

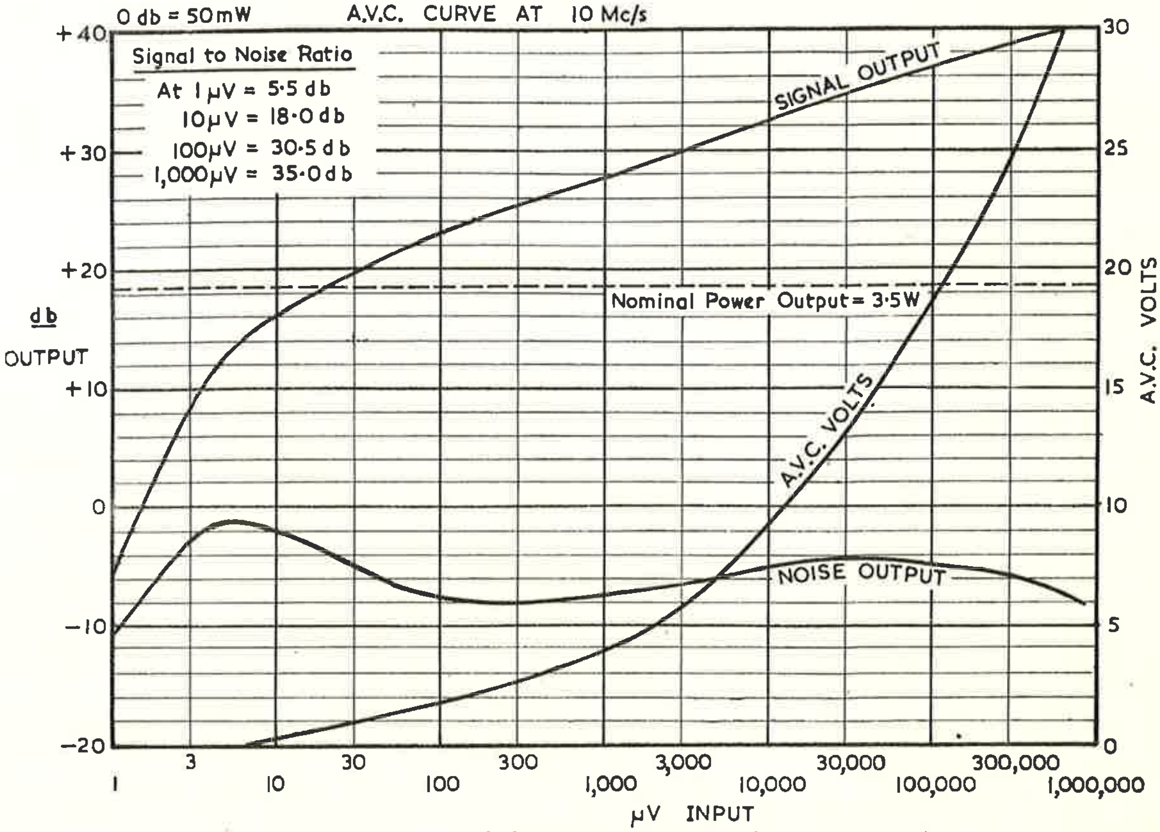
Converter operation.

To simplify oscillator coil manufacture the method of operation of the 6BE6 in which the cathode is taken to a tapping on the oscillator coil has not been used. Instead, the oscillator circuit is the conventional tuned-grid plate tickler type oscillator which allows the cathode to be earthed directly at the valve socket. With a grid leak of 20,000 ohms the grid current on the broadcast band is approximately 220 μA, with little variation over the band.

On the S.W. band harmonic mixing is used, i.e. the oscillator is operated at half the frequency required to give a difference in frequency of 455 Kc/s between signal and oscillator frequencies. In addition, the oscillator fundamental frequency is so chosen that the oscillator harmonic is on the low side of the signal frequency instead of on the high side and the padding capacitors therefore are in the signal tuned circuits.

By using harmonic mixing, the necessity for neutralising is avoided, interaction of the signal and oscillator circuits is greatly reduced, and stability and flutter are consequently improved. The oscillator coil is a normal solenoid of somewhat greater number of turns than usual as the frequency range to be covered is approximately 2.7 – 8.95 Mc/s for the normal S.W. band coverage of 5.9 Mc/s to 18.4 Mc/s. A litz progressive universal winding may, however, be used for the oscillator coil if desired.

With harmonic mixing, somewhat higher oscillator grid currents are necessary in order to obtain the same conversion gain on the S.W. band as on the broadcast band. Typical values for type



Output, noise, a.v.c. and distortion vs. input voltage at 10 Mc/s.

6BE6 as used in this receiver are 500 µA at the low frequency end of the band to 750 µA at the high frequency end. The use of a 100 ohm resistor in series with the converter oscillator grid is desirable in order to limit the oscillator grid current at the high frequency end and avoid loss of conversion gain due to excessive oscillator voltage.

Two disadvantages normally experienced with short-wave harmonic mixing are that some conversion sensitivity is lost at the low frequency end of the band, and also that the receiver is able to receive signals 455 Kc/s either side of the oscillator fundamental frequency. The latter point is not serious in this receiver as a typical value for these responses with the oscillator set at 6 Mc/s is 65 db. down for the lower one and 60 db. down for the higher one. While these spurious responses do not occur in a receiver having an oscillator operated on the fundamental frequency other responses do occur at 455 Kc/s either side of the oscillator 2nd harmonic frequency.

For reasons of stability, a.v.c. was not applied to the converter on the S.W. band and, in practice, the steeper a.v.c. curve which results is to be desired at these frequencies because with a flat a.v.c. curve the gain of the receiver would increase greatly on badly fading signals and raise the noise level, whereas with a more sloping a.v.c. curve the fading will produce only quiet intervals in the received signal.

In the test report the figures given for selectivity are those measured with the a.v.c. operating. This method was used as it is the more usual one and is easy to carry out without modifications to the circuit for measurement purposes.

It may be possible to eliminate components marked with an asterisk on the circuit diagram. They have been included to minimise the possibility of instability with different receiver layouts.

TEST RESULTS RECEIVER RC55

Selectivity 1,000 Kc/s

Times Down	Total Bandwidth Kc/s
10 db.	8.5
20 db.	14.5
60 db.	28.0

Distortion for Nominal Power Output

9% for 3.5 watts in the primary.

Modulation Distortion

1 mV signal input at 1 Mc/s, 50 mW output at 400 c/s.

Modulation Depth %	% Distortion
20	0.9
40	1.2
60	1.8
80	3.3
100	3.9

Voltage and Current Analysis

	E_a	I_a	E_{sc}	I_{sc}	E_g
6BA6	225	7.6	78	3.2	-1.5
6BE6	225	3.9	96	7.5	-1.5
6BA6	225	7.6	78	3.1	-1.5
6AV6	83	0.3	—	—	—
6AQ5	225	33	213	3.1	12.3
6X4	—	—	—	—	—

Total B+ drain 69 mA.

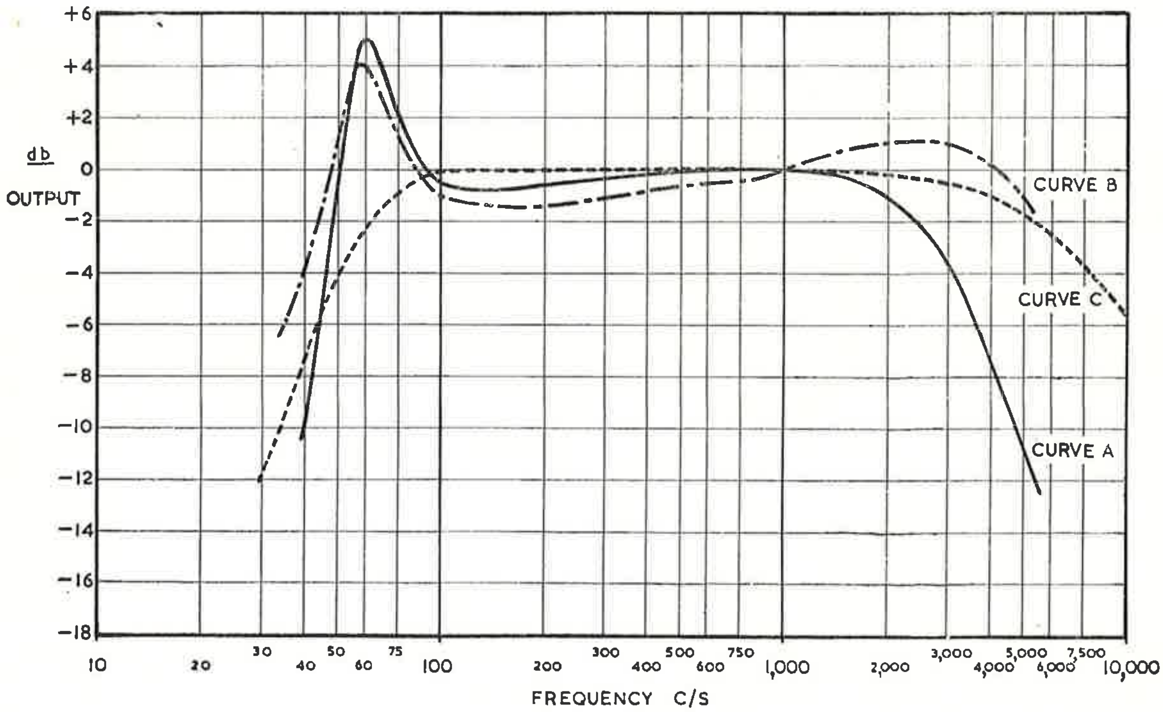
Sensitivities (for 50 mW Output).

	c/s.	I-F (Kc/s.)		BC (Kc/s)		S.W. (Mc/s)		
	400	455	600	1,000	1,400	6.0	10.0	17.8
6AQ5 grid (V)	1.1	(no a-f feedback)						
6AV6 grid (V)	0.014	(no a-f feedback)						
6AV6 diode plate (μV)		180,000						
6BA6 grid (μV)		1,200						
6BE6 grid (μV)		35	53	53	53	67	53	53
6BA6 grid (μV)			2.35	2.9	3.8	7.0	3.2	3.1
Aerial (μV)			<1.0	<1.0	<1.0	3.8	1.7	2.1
Image Ratio*			66,000	15,000	3,100	160	38	9.0
Noise sensitivity† (μV)			1.3	1.8	1.6	1.9	1.7	2.2
$I_{osc.}$grid (μA)			210	220	218	500	720	760

* Image ratio measured from 10 μV input.

† Input required to give equal signal and noise power output i.e. 3 db. signal-to-noise ratio.

AUDIO RESPONSE CURVES



(A) Output across voice coil, input through aerial terminal 1000 μV at 1000 Kc/s, 30% modulation, volume control adjusted to give 50 mW output. 0 db = 1.0 volt.

(B) Output across voice coil, input to top of volume control, volume control in same position as (A) 0 db = 1.0 volt.

(C) Output across 5000 ohm resistor across primary of transformer with speaker voice coil disconnected but feedback network left connected, volume control in same position as in (A) 0 db = 10.0 volts.

COIL DETAILS

AERIAL

Former:—$\frac{3}{4}''$ Diam. Paxolin.

	Primary Winding	Secondary Winding
Broadcast Band 535 Kc/s – 1620 Kc/s	250 turns, 9/41 Litz wound in a single $\frac{5}{32}''$ pie spaced $\frac{3}{16}''$ from secondary.	116 turns, 9/41 Litz wound in three $\frac{5}{32}''$ pies of 40, 40 and 36 turns spaced from each other $\frac{3}{32}''$.
S.W. Band 5.9 Mc/s – 18.4 Mc/s	20 turns, 36 B. & S., D.S.C. wound against cold end of secondary.	10.5 turns of 22 B. & S. enam. spaced 16 T.P.I.

R.F.

Former:—$\frac{3}{4}''$ Diam. Paxolin.

	Primary Winding	Secondary Winding
Broadcast Band 535 Kc/s – 1620 Kc/s	410 turns, 36 B. & S., S.S.E., wound in a single $\frac{5}{32}''$ pie spaced $\frac{3}{16}''$ from the secondary.	116 turns, 9/41 Litz wound in three $\frac{5}{32}''$ pies of 40, 40 and 36 turns spaced from each other $\frac{3}{32}''$.
S.W. Band 5.9 Mc/s – 18.4 Mc/s	7.5 turns, 36 B. & S., D.S.C., interwound at cold end of secondary.	10.5 turns of 22 B. & S. enam. spaced 16 T.P.I.

OSCILLATOR

Formers:—Broadcast Band $\frac{7}{16}''$ Diam. Paxolin with $\frac{3}{8}''$ diam. adjustable dust iron core.

S.W. Band $\frac{3}{4}''$ Diam. Paxolin with $\frac{3}{8}''$ Diam. adjustable dust iron core.

	Tuned Grid Winding	Feedback Winding
Broadcast Band 535 Kc/s – 1620 Kc/s	80 turns, 36 B. & S., D.C.C. wound in a single $\frac{1}{4}''$ pie.	20 turns 36 B. & S. D.C.C. wound in a single $\frac{1}{4}''$ pie spaced from grid winding $\frac{3}{32}''$.
S.W. Band 5.9 Mc/s – 18.4 Mc/s	19.5 turns of 22 B. & S. enam. close wound.	5.3 turns, 36 B. & S. D.S.C. interwound at cold end of grid winding.

Radiotron Type 12AT7 Twin Triode

(Reprinted by courtesy of Radio Corporation of America)

Radiotron type 12AT7 is a miniature type twin triode designed for use as a grounded-grid radio-frequency amplifier or as a frequency converter at frequencies below approximately 300 megacycles. It is also suitable for audio-frequency applications. A centre-tapped heater permits operation of the valve from either a 6.3 volt or a 12.6 volt heater supply.

General

Cathodes .... Coated Unipotential

	Series	Parallel	
Heater Voltage (a.c. or d.c.)	12.6	6.3	volts
Heater Current	0.150	0.300	ampere

Envelope .... T-6½ Glass
Base .... Small-Button Noval 9-Pin
Mounting Position .... Any
Direct Interelectrode Capacitances (approximate)*

(Grounded Cathode Operation)

Grid to Plate (Each Section)	1.45	μμF
Input (Each Section)	2.5	μμF
Output (Section Number 1)	0.45	μμF
Output (Section Number 2)	0.35	μμF
Grid to Grid (Max.)	0.005	μμF
Plate to Plate (Max.)	0.4	μμF
Heater to Cathode (Each Section) .	2.5	μμF

(Grounded Grid Operation)

Plate to Cathode (Each Section)	0.15	μμF
Input (Each Section)	5.0	μμF
Output (Section Number 1)	1.6	μμF
Output (Section Number 2)	1.5	μμF

Socket connections.

Pin 1 — Plate (Section Number 2).
Pin 2 — Grid (Section Number 2).
Pin 3 — Cathode (Section Number 2)
Pin 4 — Heater
Pin 5 — Heater
Pin 6 — Plate (Section Number 1).
Pin 7 — Grid (Section Number 1).
Pin 8 — Cathode (Section Number 1).
Pin 9 — Heater Centre-Tap.

Maximum ratings.

	Design Centre	Absolute
Plate Voltage	300	330 volts
Plate Dissipation (Each Section)	2.5	2.8 watts
D.C. Heater-Cathode Voltage	90	100 volts

Characteristics and typical operation.

Class A Amplifier *(Each Triode Section)*

Plate Voltage	100	180	250	volts
Grid Bias Voltage .	–1	–1	–2	volts
Amplification Factor.	54	62	55	
Transconductance ..	4000	6600	5500	micromhos
Plate Current	3.7	11	10	milliamperes
Grid Bias Voltage** .	–6	–8	–12	volts

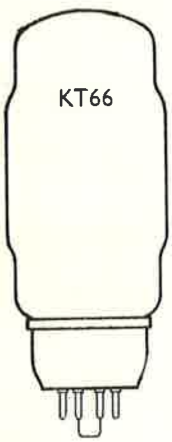
* Approximate values without external shield.
** Approximate values for 10 microamperes plate current.

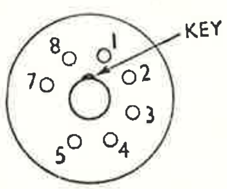
New Radiotron Type KT66 Power Tetrode

General.

Type KT66 is a versatile power tetrode with a number of useful applications. It may be used in the output stage of an audio-frequency power amplifier, either tetrode-connected for maximum sensitivity and power output or triode-connected for high quality working. In transmitting circuits using frequencies up to 30 Mc/s it may also be used as an oscillator or as a radio frequency amplifier.

The KT66 is designed as a beam tetrode with aligned grids. This alignment of the grids reduces losses in the screen and makes for the highest possible power conversion efficiency. With this system of construction high orders of power output may be obtained with a low screen dissipation, and the anode is designed to dissipate 25 watts continuously with a reliable life performance. It can be used as a substitute for the 6L6-G.





View looking on underside of base

Base connections and dimensions

7-Pin "Octal"

Pin 1	Not connected
Pin 2	Heater
Pin 3	Plate
Pin 4	Screen Grid, *g2*
Pin 5	Control Grid, *g1*
Pin 6	Omitted
Pin 7	Heater
Pin 8	Cathode

Maximum overall length $5\frac{5}{16}''$.
Maximum diameter $2\frac{1}{16}''$.

Maximum ratings

	Tetrode connected	Triode connected	
Heater Voltage	6.3	6.3	volts
Heater Current	1.27	1.27	approx. amps
Plate Voltage	525	400	max. volts
Screen Voltage	400	—	max. volts
Anode Dissipation	25	25	max. watts
Screen Dissipation	3.5	—	max. watts
Plate Impedance*	22500	1450	ohms
Mutual Conductance*	6300	5500	micromhos

Capacitances.

Grid to all other Electrodes	16.0	approx. $\mu\mu$F
Plate to all other Electrodes	11.5	,, ,,
Plate to Grid	1.1	,, ,,
Grid to Cathode	8.7	,, ,,
Plate to Cathode	15.8	,, ,,
Plate to Grid	7.2	,, ,,

General precautions in use, and interpretation of application data.

(1) For the prevention of parasitic oscillation always connect a resistance of 100-300 ohms close to the screen grid terminal of the valve socket. When the valve is used as a triode this resistance should be connected between the screen and plate. A control grid (stopper) resistance of 10,000-50,000 ohms is also recommended. Without these precautions to prevent parasitic short wave oscillation, "flash over" between the base pins of the valve may be experienced. The effect may also be minimised by separating the grid and anode leads, keeping them as short as possible and by avoidance of parallel paths.

(2) The maximum permissible d.c. resistance from control grid to cathode must not exceed 0.5 megohm for self bias and 0.1 megohm for fixed bias.

(3) Do not exceed the maximum plate and screen wattage dissipation.

(4) The use of a common self bias resistor is not recommended except in applications where the maximum plate dissipation is never reached.

*measured at		Tetrode	Triode
	Plate Voltage	250	400 volts
	Screen Voltage	250	— volts
	Control Grid Voltage	−15	−38 volts

(5) In push-pull applications showing a large change in plate current between the "no-signal" and "full-signal" conditions, a choke input smoothing circuit having a good regulation should be used. A badly regulated supply will lead to a fall in power output and/or excessive quiescent anode dissipation. A suitable design is shown in Fig. 1.
(6) Cut off the B supply before any adjustments are made to any part of the circuit.
(7) Do not disconnect the anode load while the output is operating under full drive conditions.
(8) Allow adequate ventilation; considerable damage to valves and components will result if this is neglected.
(9) The maximum heater-cathode potential must not exceed 150 volts under any condition of operation.

In such applications where the potential might be greater, in cathode-coupled driven stages for example, the heater and cathode should be joined and a separate heater supply used for each valve.

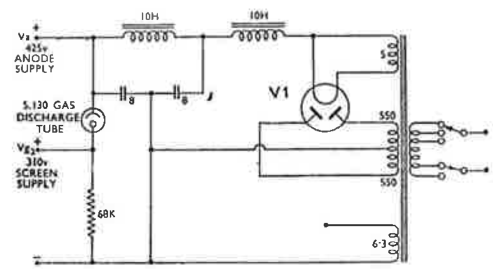


Fig. 1.—Power supply for KT66 valves in push-pull. V_1 = U52/5U4-G.

Interpretation of operating conditions.

The operating figures given for the KT66 are "design centre values" and have been chosen so that the valves will give satisfactory performance at these maximum ratings in equipment operated from power supplies where the normal voltage variation is within ± 7% of the specified centre value.

THE KT66 IN AUDIO-FREQUENCY POWER AMPLIFIER CIRCUITS

TETRODE-CONNECTED

(a) Single valve class A (self bias). Output 7.25 watts.

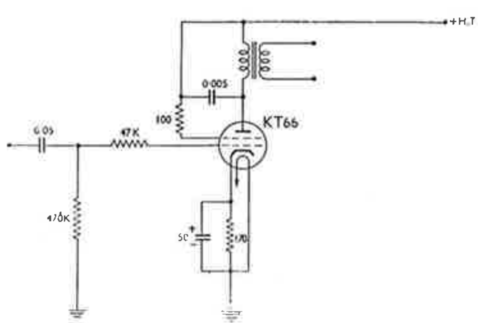


Fig. 2.—Single valve class A.

A typical circuit is shown in Fig. 2. The output transformer should be designed on generous lines because of the high plate current.

Operating conditions

Plate and Screen Voltage	250	volts
Control Grid Voltage	–15	approx. volts
Self Bias Resistance	170	ohms
Plate Current	85	mA
Screen Current	6.3	mA
Signal Input	15	peak volts
Load Resistance	2200	ohms
Plate Dissipation (no signal)	21.5	watts
Power Output	7.25	watts
Distortion	9	%

The performance is shown in Figs. 3 and 4.

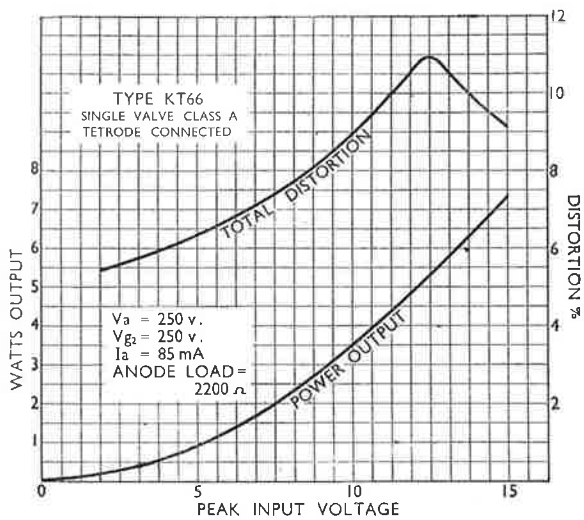


Fig. 3.

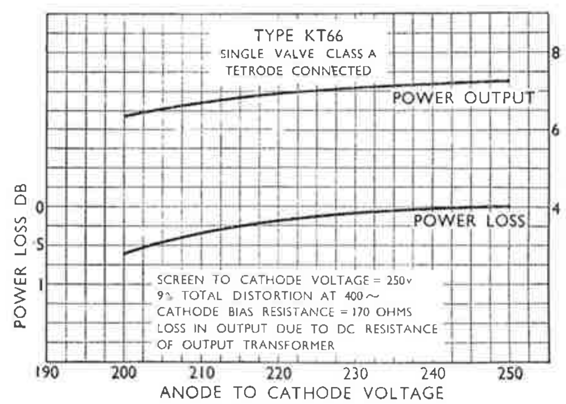


Fig. 4.

(b) Two valves push-pull class AB_1 (self bias). Output 30 watts.

A typical circuit is shown in Fig. 5. Under the 250-volt condition the screen grids and plates are both supplied from the same source; under the 400-volt condition the screen grids must have a supply at 300 volts of relatively low impedance, and a stabilised screen supply is recommended, as shown in Fig. 1 using a gas discharge tube.

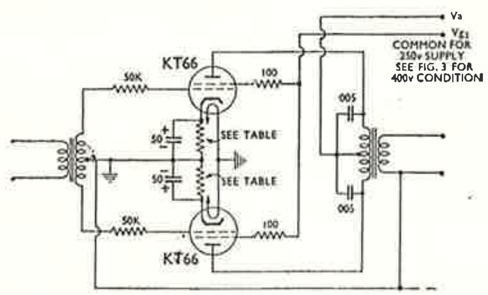


Fig. 5.—Two valves class AB_1.

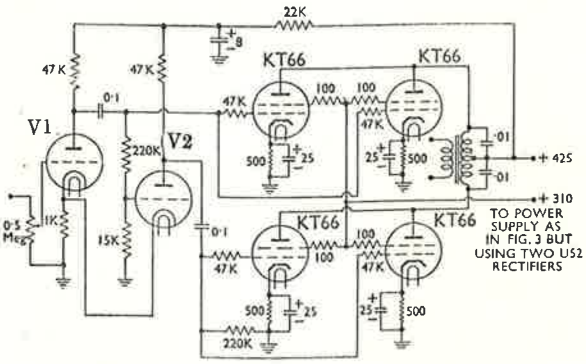


Fig. 6.—Four valves in push-pull class AB_1.
$V_1 = V_2 =$ *6J5-GT.*

Operating conditions

	No signal	Full signal	
250 VOLT SUPPLY			
Plate and Screen Voltage	250	250	volts
Control Grid Voltage	–17.5	—	approx. volts
Self Bias Resistance, per valve	200	—	ohms
Plate Current	162	165	mA
Screen Current	12	20	mA
Signal Input, grid-to-grid	—	36	peak volts
Load Resistance, plate-to-plate	—	4000	ohms
Power Output	—	17	watts
Distortion	—	4	%
450 VOLT SUPPLY			
Plate Voltage	415	390	volts
Screen Voltage	300	275	volts
Control Grid Voltage	–27	—	approx. volts
Self Bias Resistance, per valve	500	—	ohms
Plate Current, total	104	125	mA
Screen Current, total	5	18	mA
Signal Input, grid-to-grid	—	70	peak volts
Load resistance, plate-to-plate	—	8000	ohms
Plate Dissipation	21.5	9.5	watts
Power Output	—	30	watts
Distortion	—	6	%

(c) Four valves push-pull class AB_1 (self bias). Output 60 watts.

When outputs in excess of 30 watts are required, four valves may be used in parallel push-pull. It is essential to take full precautions against self oscillation; separate bias resistors should be used to minimise the effect of normal valve variations. The operating data may be obtained from the two valve condition shown in section (b) above, by doubling the plate and screen currents and halving the plate-to-plate load. The recommended circuit shown in Fig. 6 will give approximately 60 watts output with an input of 1.75 volts peak.

(d) Two valves push-pull class AB_1 (fixed bias). Output 50 watts.

An alternative method to (c) of obtaining an output in the neighbourhood of 50 watts, but employing only two KT66 valves is shown in Figs. 7 and 7A. The valves are operated with fixed bias but as no grid current is drawn a low impedance driver stage is not required. The circuit shown in Fig. 7 uses a push-pull transformer having a ratio of about 1:4 and each half-secondary is connected to a variable bias supply. An alternative circuit is shown in Fig. 7A using a phase-splitting valve.

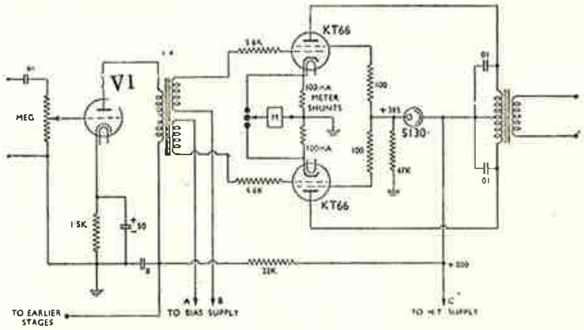


Fig. 7.—Two valve push-pull amplifier with fixed bias tetrode connection transformer coupled.
$V_1 =$ *6J5-GT.*

A convenient method of obtaining the required negative bias is shown in Fig. 7B. It is somewhat unorthodox but avoids the need for a special bias transformer. The rectifier is fed from the anode supply transformer via a high voltage 0.02 μF capacitor.

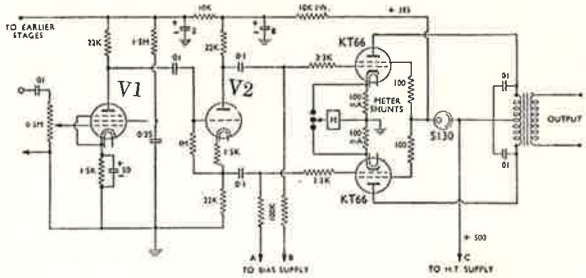


Fig. 7A.—Two valve push-pull amplifier with fixed bias tetrode connection resistance capacity coupling.
$V_1 =$ *6J7G*, $V_2 =$ *6J5-G.*

Because of the large variation in screen current between quiescent and full output, a gas-filled cold-cathode diode, type S130, is used to supply the screen voltage. As the voltage drop across the diode does not change materially with varying current, the screens are maintained at approximately 115 volts below the anode supply voltage under all conditions of operation. This reduced voltage may be used to supply the earlier valves. The screen grids must not be connected directly to the 500 volt supply.

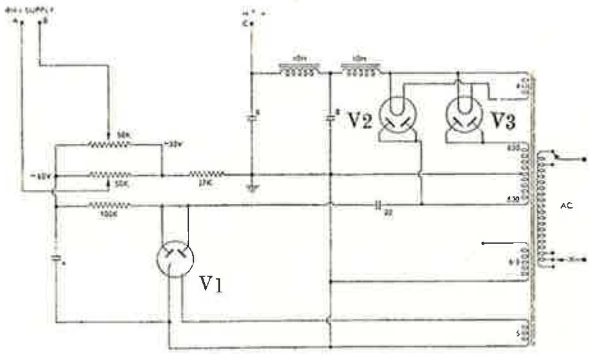


Fig. 7B.—Power supply for figures 7 and 7A.

V_1 = 5Y3-GT, V_2 = V_3 = 5R4-GY.

Operating conditions for low distortion and continuous full load operation.

	No signal	Full signal	
Plate Voltage	525	480	volts
Screen Voltage	420	385	volts
Plate Current, per pair	80	175	mA
Screen Current, per pair	3	21	mA
Control Grid Voltage	–45	— approx.	volts
Input Voltage, grid-to-grid	—	90	peak volts
Load Resistance, plate-to-plate	—	6000	ohms
Power output	—	50	watts
Distortion	—	5	%

Where greater distortion is not objectionable the following operating conditions may be used. In (1) the screen is supplied through a 250-volt 25-watt lamp to limit the screen dissipation with a sustained input signal. In (2) this lamp is omitted and continuous operation at full output is not permissible.

(1) For medium quality and inputs above 30% modulation

	No signal	Full signal	
Plate Voltage	525	480	volts
Screen Voltage	475	385	volts
Plate Current, per pair	80	175	mA
Screen Current, per pair	3	21	mA
Control Grid Voltage	–50	— approx.	volts
Input Voltage, grid-to-grid	—	100	peak volts
Load Resistance, plate-to-plate	—	5000	ohms
Power Output	—	50	watts
Distortion	—	8	%

(2) For medium quality and inputs up to 30% modulation

	No signal	Full signal	
Plate Voltage	450	420	volts
Screen Voltage	450	420	volts
Plate Current, per pair	80	200	mA
Screen Current, per pair	7	36	mA
Control Grid Voltage	–48	— approx.	volts
Input Voltage, grid-to-grid	—	96	peak volts
Load Resistance, plate-to-plate	—	5000	ohms
Power Output	—	50	watts
Distortion	—	8	%

TRIODE-CONNECTED

(e) Single valve class A (self bias). Output 5.8 watts.

A typical circuit is shown in Fig. 8. When a plate voltage of 250 is used a KT66 valve is not operating at its maximum dissipation due to the low plate current of 60 mA. There is, however, no advantage to be gained by increasing this current.

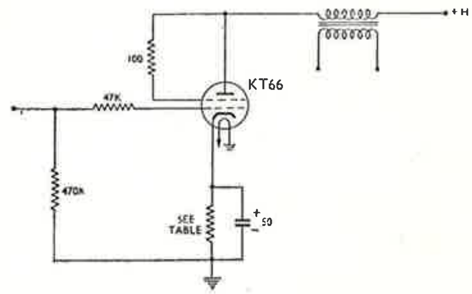


Fig. 8.—Single valve class A triode connected.

Operating conditions

Plate and Screen Voltage	250	400	volts
Control Grid Voltage	–19	–38 approx.	volts
Self bias Resistance	330	620	ohms
Plate Current	60	62	mA
Input Voltage	19	38	peak volts
Load Resistance	2750	4500	ohms
Plate and Screen Dissipation	15	25	watts
Power Output	2.2	5.8	watts
Distortion	6	7	%

(f) Two valves in push-pull class AB_1 (self bias). Output 14.5 watts.

The circuit is shown in Fig. 9. This circuit is popularly employed for an amplifier intended for high fidelity sound reproduction. Negative feedback may be included over the whole amplifier to reduce overall distortion still further. It is important in such cases that a correctly designed output transformer should be employed to reduce phase shift to a minimum.

Operating conditions

Plate and Screen Voltage	250	400	volts
Control Grid Voltage	–20	–38 approx.	volts
Self bias Resistance, per valve	390	620	ohms
Plate and Screen Current, total	104	124	mA
Input Voltage, grid-to-grid	40	80	peak volts
Load Resistance, plate-to-plate	2500	4000*	ohms
Power Output	4.5	14.5	watts
Distortion	2	3.5	%

* May be 10,000 ohms in class A, giving slightly reduced output and lower distortion.

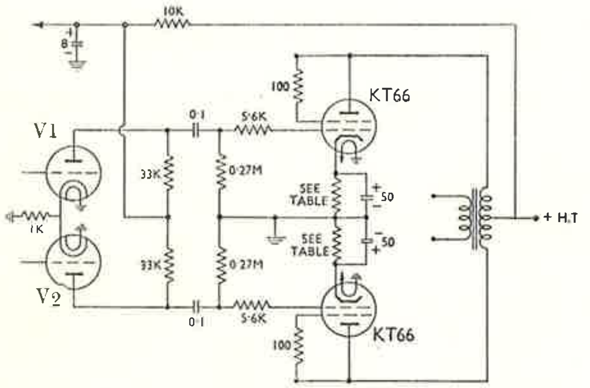


Fig. 9.—Two valves push-pull, triode connected, class AB_1. $V_1 = V_2 = $ 6J5-GT.

(g) Operating as cathode follower.

A typical circuit which is useful for driving a class B output stage is shown in fig. 10. The impedance between cathode and earth provided by a triode connected in this way is given approximately by:—

$$\frac{m}{m+1} \times \frac{1}{g}$$

Where m = amplification factor
g = mutual conductance

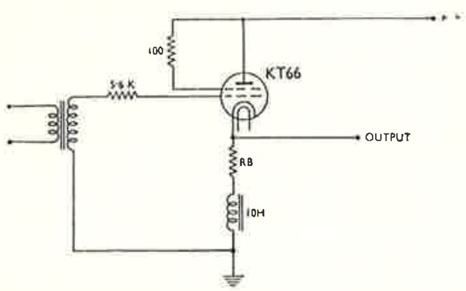


Fig. 10.—Operation as cathode follower.

In the case of KT66 valves this is about 200 ohms. Self bias is provided by the total d.c. resistance in the cathode circuit and this should amount to 315 ohms for 250 volt or 600 ohms for 400 volt operation. A small resistance *RB* is added to the d.c. resistance of the choke or transformer to make up this value.

"Flash-over" on KT66 valves.

Some users of KT66 valves, particularly in the 400 volt push-pull condition, have experienced trouble due to "flash-over" between the base pins of the KT66. It has been alleged that the trouble is due to faulty bakelite either in the valve base, or socket, and in fact some users have attempted to cure the trouble by the use of ceramic sockets.

This effect has not been encountered in the laboratory due to the care taken with the mechanical layout even though overload conditions are resorted to. An examination of actual cases where the "flash-over" has occurred, has shown that it was due to parasitic short wave oscillation or "squegger".

In the particular cases under review, the grid and plate leads to the KT66 valves followed parallel paths over an appreciable length, and in fact the conditions were closely similar to those of an ultra-short wave self-oscillator. It has been found possible to obviate the effect by separating these leads and keeping them as short as possible, but even with the original arrangement, the self-oscillation was completely stopped by the use of a grid stopper resistance of 50,000 ohms connected close to the control grid base pin of each valve.

From this experience, taken in conjunction with the earlier tests, it can now be safely assumed that any case of flash-over with the KT66 valve, either with or without signal drive, is definitely due to short wave self-oscillation, and the following precautions should be adopted.

1. In laying out a new receiver or amplifier, keep grid and anode leads as short as possible, and avoid parallel paths.
2. In addition, fit a 50,000 ohm grid stopping resistance in each grid lead, close to the base pin.
3. Fit screen grid stopping resistances of 100 ohms in each screen lead close to the screen grid base pin.

In cases where existing apparatus is showing the flash-over effect, and where it is not possible to modify the wiring, then it should be possible to prevent the effect by adopting precautions two and three above.

THE KT66 IN A TRANSMITTER

The KT66 valve is suitable for use in the r-f stages of a transmitter, especially at frequencies below 15 Mc/s, see Fig. 11. Suitable operating conditions are given as follows:—

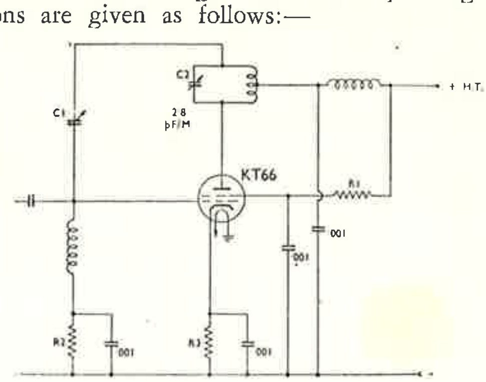


Fig. 11.—Circuit suitable for class "C" telegraphy.

Radio-Frequency Amplifier-Class C. Telegraphy.

Maximum permissible ratings

Key-down conditions per valve without modulation.

Plate Voltage	400 max. volts
Screen Voltage	300 max. volts
Grid Voltage	–200 max. volts
Plate Current	90 max. mA
Screen Current	14 max. mA
Grid Current	30 max. mA
Plate Input	36 max. watts
Screen Input	3.5 max. watts
Plate Dissipation	25 max. watts

Typical operating conditions

Frequency	10.5	10.5	10.5	Mc/s
Plate Voltage	400	300	200	volts
Screen Voltage	300	300	200	volts
Grid Voltage	–40	–40	–40	volts
Plate Current	80	78	57	mA
Screen Current	9.5	10	14	mA
Grid Current*	0.7	1.1	3.2	mA
Peak R-F Grid Volts	48	50	58	volts
Driving Power*	0.03	0.05	0.17	watts
Power Output	21	13.5	6	watts
Plate Efficiency	66	58.5	54	%

*Subject to wide variations depending on the impedance of the load circuit.

Values of screen, control grid and cathode resistors for various plate voltages at full drive.

Plate Voltage	400	300	200	volts
Series Screen Resistor (*R1*)	10,000	—	—	ohms
Grid Leak (*R2*)	60,000	35,000	12,500	ohms
Cathode Resistor for safe dissipation with drive off (*R3*)	450	250	50	ohms

The screen supply for this class of service may be obtained from the plate supply through a series resistor or from a potentiometer; the series resistor method will normally be the most convenient and the table gives typical resistance values. Grid bias may be obtained from batteries, from a grid leak, from a cathode resistor or from a combination of these methods. It is strongly recommended that grid leak bias should only be used in conjunction with some cathode resistance or fixed bias in order to limit the plate current to a safe value in the event of failure of the drive. If the transmitter is to be keyed by interrupting the drive it is essential to use sufficient cathode or fixed bias to limit the anode dissipation to 25 watts in the "key-up" condition, taking into account the regulation of the power supply. If circumstances arise in which it is necessary to use grid leak bias alone, the provision of adequate fuses in each B supply lead is most important. The values of grid leaks shown in the table apply when neither cathode nor fixed bias are used; when additional sources of bias are employed the resistance of the grid leak may be reduced, but the value is not critical. The values of cathode resistors given in the table designed to keep the plate dissipation within its rated maximum when the drive is removed.

When cathode bias is employed and the maximum B supply voltage available is limited, the series resistor method of obtaining the screen supply should be avoided, since the rise in screen voltage in the key-up condition makes necessary a higher value of cathode resistor than would otherwise be required. This results in an increased voltage drop across the cathode resistor in the "key-down" condition. The cathode resistance values shown in the table assume a constant screen voltage. Removal of the screen voltage normally reduces the output to zero so that keying in the screen supply lead is satisfactory.

It is not usually necessary to screen the input and output circuits from one another since the plate-grid capacitance of a KT66 valve is large enough to make neutralising necessary. However, the neutralising capacitance required is only of the order of 1 $\mu\mu$F, and sufficient capacitance can often be obtained by suitable location of the components and wiring. This capacitance is represented by *C1* in the circuit diagram of Fig. 11. This figure shows a suitable circuit using KT66 valves in a Class C telegraphy amplifier. The inductance *L1* should be designed to resonate at the desired frequency with its tuning capacitor *C2*.

The results described in this publication may be obtained at frequencies up to 15 Mc/s: at higher frequencies the efficiency will fall off somewhat and considerable variations may be found between individual valves, owing to variations in output impedance caused by the carbonising of the bulb.

Crystal oscillator.

The maximum conditions for this class of service are the same as those given for Class C telegraphy. Grid bias will normally be obtained from a grid leak, but the use of some cathode bias is strongly recommended as a safety measure. The cathode resistor should be large enough to limit the anode dissipation to 25 watts in the event of the valve falling out of oscillation; its value will depend to some extent on the regulation of the B supply and on the load drawn by other parts of the transmitter when the drive is removed. The screen supply may be obtained by any convenient method; a series resistor from the plate supply will normally be employed. The value of this resistance for a given plate voltage will be the same as that given under Class C telegraphy.

Although the valve may be operated at its full telegraphy ratings, it is recommended that the plate voltage should be kept as low as possible in order to limit crystal heating and so reduce frequency drift due to changes in crystal temperature; this is more important for X and Y-cut crystals than for the low temperature co-efficient crystals such as the AT-cut. In crystal oscillator circuits it may be found that the plate-grid capacitance is sufficiently low to render oscillation difficult; this may be overcome by connecting a small external capacitance between plate and grid. In order to avoid excessive feedback which might damage the crystal, the value of this capacitor should be kept as low as possible and should rarely exceed 1 $\mu\mu$F. It is shown as *C3* in Fig. 12, which is a circuit diagram showing a type KT66 valve as a conventional crystal oscillator with

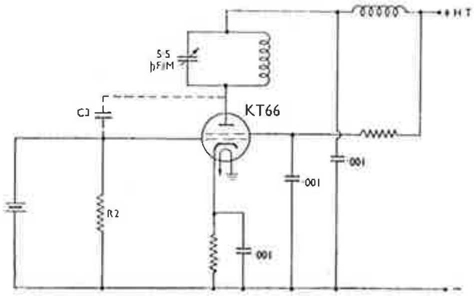


Fig. 12.—Conventional crystal oscillator circuit.

the output circuit tuned to the crystal frequency. An output of about 15 watts can be obtained at frequencies up to 10 Mc/s without damaging the crystals.

Fig. 13 shows the valve in a Dow crystal oscillator circuit, commonly known as a tri-tet circuit. The screen grid is earthed to radio frequency and acts as the plate of a triode oscillator, of which the cathode assumes a r-f potential above earth; the load circuit is connected to the valve plate and is thus electron coupled to the oscillatory system. The cathode tuned circuit *L1 C4* is not tuned to the crystal frequency, but to a considerably higher value, and the *L/C* ratio should be kept as high as possible in order to reduce the voltage across the

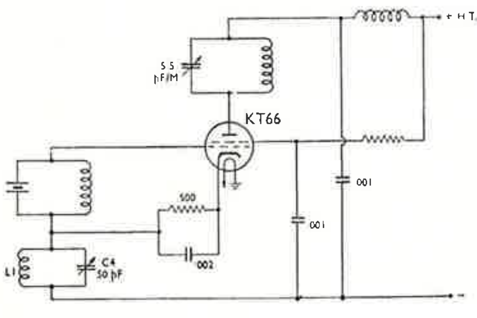


Fig. 13.—Dow crystal oscillator for harmonic output. Component values in the three circuits (Fig. 21, 22, 23) are suitable for wavelengths between 15 and 50 metres.

crystal. Grid bias may be obtained from a grid leak, but the use of cathode bias only, as shown in Fig. 13, is recommended, as the r-f crystal current is substantially lower with this arrangement. Since the KT66 is not a fully screened valve, the load circuit of the Dow oscillator should be tuned to some harmonic of the crystal frequency, as operation at the fundamental frequency is likely to lead to excessive feedback and crystal fracture. With the output tuned to the 2nd harmonic of the crystal frequency an output of 12 watts can be obtained at frequencies up to 10 Mc/s.

Frequency multiplier.

The maximum conditions for this class of service are the same as those given for Class C telegraphy. Generally the same circuit arrangement can be used and if a neutralising capacitor is employed it may be left in position; it will provide slight regeneration in doubler service and give some increase in output. Grid bias can be obtained from a grid leak, but some cathode or fixed bias should be used as a safety measure. For doubler service a total d.c. bias of 100 volts negative is sufficient, and at maximum rating this may be obtained from a cathode resistor of 450 ohms and a grid leak of 30,000 ohms. A d.c. grid current of about 3 mA will be required at 140 volts peak corresponding to about 0.4 watt of driving power. Very high values of grid bias should be avoided as they tend to limit the maximum available anode current consistent with good efficiency and their use results in reduced output. This is due to the shape of the valve characteristic in the positive grid region.

The screen supply will normally be obtained from a series resistor, and the values given under Class C telegraphy are suitable. When the valve is used as a frequency doubler an output of 16 watts can be obtained at frequencies up to 15 Mc/s. Useful amounts of higher order harmonic output can also be obtained, but accurate figures cannot be quoted since outputs may vary between valves.

Driver stages.

Since the driving power required by a KT66 is very small, a receiving valve such as the type 6J5-GT may be used as a driver. Any convenient method may be employed to couple the amplifier to the driver; capacitive coupling is very simple, but in some cases link coupling may be more convenient.

It is often convenient to build a transmitter employing the same type of valve in a number of stages, and the KT66 is particularly applicable to such a design. A three-valve transmitter employing a KT66 crystal oscillator followed by a buffer amplifier or doubler driving a KT66 final amplifier is a typical arrangement; it would be capable of delivering an output of 15 watts or more on a number of frequencies, since the crystal and buffer stages could be designed to work both at their driver frequency and as doublers.

THE KT66 IN A VOLTAGE STABILISER.

The KT66 is very suitable for use as the series regulator valve in a voltage stabiliser, a typical circuit being shown in Fig. 14. The KT66 operates in conjunction with the 6J7-G valve and the gas-filled cold-cathode diode, QS95/10.

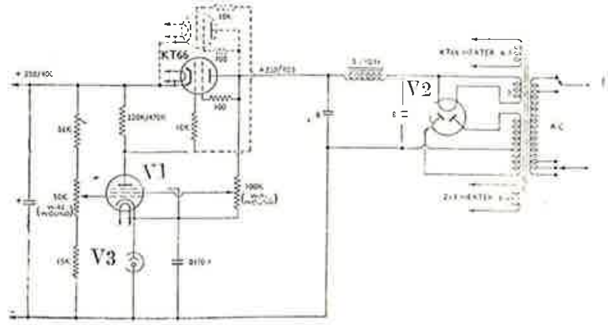


Fig. 14.—Stabilised power supply using KT66. V_1 = 6J7-G, V_2 = U52/5U4-G. V_3 = QS95/10.

The plate dissipation of the KT66, which represents the power handling capacity of the stabiliser, is 25 watts, and for multiples of this output power two or more valves in parallel, as shown in the circuit, will be required. When more than one KT66 is used, for outputs greater than 25 watts, the requisite number of U52/5U4-G operating in must also be used. It must be appreciated that the dissipation is at a maximum at low output voltage, when the valve is absorbing the difference between the supply and the output voltages.

Radiotron Type 12AU7 Twin-Triode Amplifier

(Reprinted by courtesy of Radio Corporation of America)

Radiotron type 12AU7 is a heater-cathode type of medium-mu, twin-triode amplifier featuring a small glass envelope with integral button 9-pin base, separate terminals for each cathode, and a mid-tapped heater to permit operation from either a 6.3- or 12.6-volt supply.

Having characteristics which are very similar to those of the larger types 6SN7-GT and 12SN7-GT, the 12AU7 like these types is useful in many diversified applications including multivibrators, synchronising amplifiers, oscillators, mixers, and numerous industrial control devices. In such equipment, the 12AU7 can be used to advantage because of its compact size, its separate cathode terminals, and its economical consumption of heater power at either of the two voltages.

GENERAL DATA

Electrical:

Heater, for Unipotential Cathodes:

Heater Arrangement	Series	Parallel	
Voltage (a.c. or d.c.) ..	12.6	6.3	volts
Current	0.15	0.3	ampere

Direct Interelectrode Capacitances:°

	Triode Unit T_1	Triode Unit T_2	
Grid to Plate	1.5	1.5	$\mu\mu$F
Grid to Cathode	1.6	1.6	$\mu\mu$F
Plate to Cathode	0.50	0.35	$\mu\mu$F

Mechanical:

Mounting Position	Any
Maximum Overall Length	2-$\frac{3}{16}$″
Maximum Seated Length	1-$\frac{15}{16}$″
Length from Base Seat to Bulb Top (excluding tip)	1-$\frac{9}{16}$″ ± $\frac{3}{32}$″
Maximum Diameter	$\frac{7}{8}$″
Bulb	T-6-$\frac{1}{2}$
Base	Small Button Noval 9-Pin

°With no external shield.

Socket Connections:

Pin 1—Plate (Triode No. 2).
Pin 2—Grid (Triode No. 2).
Pin 3—Cathode (Triode No. 2).
Pin 4—Heater.
Pin 5—Heater.
Pin 6—Plate (Triode No. 1).
Pin 7—Grid (Triode No. 1)
Pin 8—Cathode (Triode No. 1).
Pin 9—Heater Mid-tap.

CLASS A_1 AMPLIFIER

Values are for each unit

Maximum ratings, *Design-Centre Values:*

PLATE VOLTAGE	300 max. volts
PLATE DISSIPATION	2.75 max. watts
CATHODE CURRENT	20 max. mA

Peak Heater-Cathode Voltage:

Heater negative with respect to cathode	180 max. volts
Heater positive with respect to cathode	180 max. volts

Characteristics:

Plate Voltage	100	250	volts
Grid Voltage	0	−8.5	volts
Amplification Factor ...	19.5	17	
Plate Resistance	6250	7700	ohms
Transconductance	3100	2200	micromhos
Plate Current	11.8	10.5	mA

Maximum Circuit Values

(for maximum rated conditions):

Grid-Circuit Resistance:

For cathode-bias operation	1.0 max. megohm
For fixed-bias operation ..	0.25 max. megohm

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 Department

Note.—Part 1 of this article appeared in Radiotronics No. 141, February, 1950.

PART II

The D-C component of the picture signal

The visual and aural senses differ in one important respect which places a requirement on the television transmission system which has no counterpart in the sound transmission system. The response of the ear to sound is actually a response to variations in air pressure. While the ear is very sensitive to rapid variations in pressure, it is completely unconscious of absolute values of air pressure, or of slow variations in pressure, as sound. In other words, there is a definite low limit to the frequency of pressure variations which the ear accepts as sound. Therefore there is no need for a sound transmission system to pass frequencies below the aural limit which is somewhere in the neighbourhood of 15 cycles per second. The circuits may be a-c coupled without loss of essential information. Even the best of practical systems have a low frequency cutoff at about 30 cycles, and most others cut off somewhere between 50 and 100 cycles.

The eye, on the other hand, is sensitive to absolute intensities of light and to slow variations of intensity. As the frequency of variation increases, the eye rapidly loses its ability to follow the changes and tends to produce a sensation which is an average of the variations. It is this averaging ability that enables the eye to interpret a rapid succession of still pictures as a portrayal of smooth motion. This phenomenon is the basis of both motion picture and television systems.

The important point in the present discussion is that the eye recognizes a slow change in light intensity. The period of the change may be a fraction of a second or it may be a minute, an hour, or a half day in length. A television system must be capable of conveying these slow changes, no matter how long the period, to the receiver. The rapid scanning of the image of the scene in the camera produces a signal containing these slow changes as well as very rapid variations caused by the passage of the scanning beam over small light and dark areas of the image. The slow changes often have periods so long that they may be considered as d-c levels

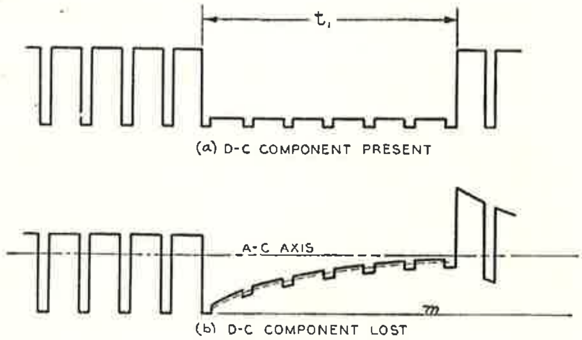


Fig. 6.

Fig. 6.—When a signal containing a d.c. component, as in (a), is passed through an a.c. coupled system the d.c. component is lost, as shown in (b).

which simply change value occasionally. Hence, the signal is said to contain a *d-c component*. The television system must either pass the entire spectrum, including the d-c component, in each of its stages, or the signal must contain such information that it will be possible to restore the d-c component, which would be lost in an a-c coupled system, when it finally arrives at the reproducer. Because of the well-known difficulties in constructing multistage d-c coupled amplifiers, it is obviously desirable to use an a-c coupled system. It is fortunate that relatively simple means are known for d-c restoration thus making possible the use of an a-c coupled system.

Fig. 6(a) illustrates a signal which contains a d-c component in the form of a temporary change in the amplitude of the pulses. The period t_1 embracing the low-amplitude pulses may be of any arbitrary length. The original signal is characterized by the constant level of the negative peaks of all the pulses regardless of amplitude. After passing through an a-c coupled system (in which the time constants of the coupling networks are short compared to the period t_1) the signal becomes distorted approximately as shown in Fig. 6(b). Here the

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negative pulse peaks no longer fall on a constant level, but the signal tends to adjust itself in a consistant manner about an axis called an a-c axis.

The a-c axis of a wave is a straight line through the wave positioned so that the area enclosed by the wave above the axis is equal to the area enclosed by the wave below the axis. The broken line marked a-c axis in Figure 6(b) is actually the correct axis only for a wave composed of large pulses like the first four at the left. During the transient condition following the first short pulse, the line shown is not the true a-c axis, but represents the operating point of the amplifier in the a-c coupled system. The actual a-c axis of the short pulses (shown by the dotted line) gradually adjusts itself to coincide with the operating point of the amplifier. This adjustment is shown by the exponential rise of the signal during the interval t_1, but it is interrupted before completion by the resumption of the large pulses. Thence a second transient condition takes place leading to a gradual restoration of the signal to its original form.

The departure of the pulse peaks from the original constant level indicated by the line *m*, is called *loss of the d-c component* or loss of "lows". It is interesting to note that this loss causes an increase in the peak-to-peak amplitude of the signal, a condition which is undesirable, especially in high-level amplifiers.

Black level

An absolute system of measurement must have a fixed standard reference unit or level. This rule applies to the problem of reproducing absolute light intensities. The simplest and most obvious reference for such a system is zero light, or *black level* as it is often called. This is a reference level which can be reproduced arbitrarily at any point in the system. Now if the television signal can be synthesized in such a way that frequent short intervals have some fixed relationship to actual black in the scene, then it becomes possible to restore the d-c component by forcibly drawing the signal to a fixed arbitrary level during these intervals.

D-C insertion and D-C restoration

Because the blanking or retrace periods are not useful for transmitting actual picture information, they offer convenient intervals for performing special control functions such as d-c restoration as mentioned in the previous paragraph. If the peaks of the blanking pulses are coincident with black level, or differ from black level by a constant amount, then d-c restoration can be accomplished simply by restoring these peaks to an arbitrary reference level. Thus, in Fig. 6(b), if the peak of each pulse can be restored to the line *m*, then the signal will appear as in (a) and the d-c component will have been restored. Small errors will remain corresponding to the displacements in level between pulses, but these are usually negligible an in any case do not become cumulative. Hence the restoration is essentially complete.

It now becomes apparent that an extremely important step in the synthesis of the television signal is that of making the peaks of the added blanking pulses bear some fixed relationship to actual black level in the scene. It was pointed out previously that the peaks of these pulses are produced by clipping off unwanted portions of the signal as illustrated in Fig. 4, C and D. A second, and most important, function is performed when the clipping is controlled in such a way that the resultant peaks have the required fixed relationship to black level. This process of relating the blanking peaks to actual black level is called *d-c insertion,* or insertion of the d-c component. A subsequent process, later in the system, of bringing these peaks back to an arbitrary reference level is called *d-c restoration.* D-c restoration must be accomplished at the input of the final reproducing device (the kinescope) in order to reproduce the scene faithfully if an a-c amplifier is used. It is desirable to restore the d-c component at other points in the system also, because the process reduces the peak-to-peak excursions of the signal to a minimum by removing increases in amplitude caused by loss of the d-c component. In a similar way, it is possible to remove switching surges, hum, and other spurious signal components which have been introduced by pure addition to the signal. Maintaining minimum excursion of the signal is important, especially at high level points in the system, in order to avoid saturation in amplifiers and consequent distortion of the half tones in the scene. For a specific example, d-c restoration helps to maintain constant sync. amplitude in high-level amplifiers. In other words, it makes possible economies in the power capabilities of amplifiers such as the final stage in the picture transmitter.

Diagram 3 in Fig. 5 illustrates part of a typical picture signal including two horizontal blanking pulses. It may be seen that there is a distinct difference between actual black level and blanking level which is prescribed as 5% of maximum blanking pulse amplitude. This difference is usually called *setup* and its magnitude was set as a reasonable compromise between loss of signal amplitude range and the need for a tolerance in operating adjustment. Setup is desirable as an operating tolerance in the initial manual adjustment of the clipper in that part of the system where the d-c is *inserted.* It simply insures that no black peaks in the actual picture signal are clipped off.

The accuracy with which setup is maintained depends on characteristics of the pickup or camera tube. Some types of pickup tubes produce signals during blanked retrace periods which are the same as, or are constantly related to, black level. In systems where such tubes are used, the magnitude of setup may be held constant automatically at whatever value is determined in the initial manual adjustment of the clipper circuit. In general, pickup tubes employing low velocity scanning, such as the image orthicon, provide this kind of basic black level information.

The iconoscope is different from orthicons in this respect because the secondary emission resulting from the high velocity scanning, produces a potential distribution on the mosaic in which black level is far from the level existing during the retrace periods when the beam is cut off. In fact, the difference between black level and blanking level varies continuously as the scene brightness changes because the potential distribution caused by resettling of the secondaries likewise changes. Automatic maintenance of setup, or pedestal height, cannot therefore be obtained by reference to the signal during blanked retrace periods in the iconoscope, but may be obtained by reference to actual black peaks in the picture signal. Where such reference is not practical, a manual control may be readjusted from time to time to keep the setup at the required value.

Synchronizing

The horizontal and vertical scanning circuits in a receiver are two entirely independent systems both of which require extremely accurate information to keep them in step with the corresponding scanning systems in the camera where the signal originates. Because the duration of sync. pulses may be rather short they may be added to the picture signal in such a way as to increase the overall amplitude of the final signal without increasing the average transmitted power level very much. Thus, simple amplitude discrimination can be used to separate the synchronizing information from the incoming composite signal in the receiver. It is, however, desirable that a second increase in amplitude should not be used to distinguish between horizontal and vertical sync. One reason for this is that a further increase in signal amplitude would make necessary an increase in the peak power rating of the transmitter or else would unnecessarily restrict the power available for signal.

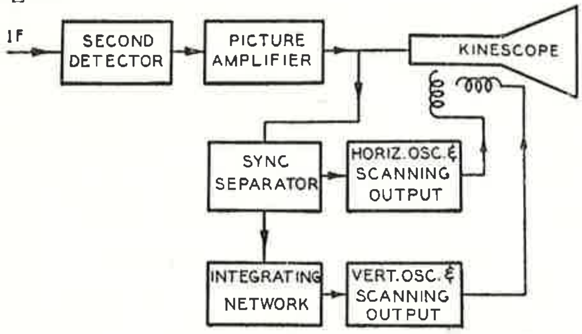


Fig. 7.—Block diagram of picture signal amplifier and scanning circuits in typical receiver.

A synchronizing system has therefore been chosen in which both vertical and horizontal pulses have the same amplitude, but different waveshapes. Frequency discrimination may then be used to separate them in the receiver. The shapes of these pulses and their relation to the blanking pulses are illustrated in detail in Fig. 5. Fig. 7 is a functional block diagram showing the steps necessary to utilize the sync. signals.

Diagrams 1 and 2 of Fig. 5 illustrate a typical complete composite picture signal in the neighborhood of the vertical blanking pulse in each of two successive fields. Interlacing of the scanning lines is shown by the time-displacement of the horizontal blanking pulses in one diagram with respect to those in the other diagram. This displacement is one half of the interval of a scanning line (H/2).

All sync. pulses appear below black level in an amplitude region which is sometimes called *blacker-than-black;* hence they can have no effect on the tonal gradation of the picture. Horizontal sync. pulses are (except during the first portion of the vertical blanking interval) simple rectangular pulses such as those appearing at the negative peaks or bases of the horizontal blanking pulses and during the last portion of the vertical blanking pulses. The duration of a horizontal sync. pulse is considerably less than that of the blanking pulse, and the leading edge of the sync. pulse is delayed with respect to the leading edge of blanking, forming a step in the composite pulse which is called the *front porch.* Correspondingly, the step formed by the difference between the trailing edges of sync. and blanking is called the *back porch.* The purpose in forming the front porch is to insure that the horizontal retrace in the receiver (initiated by the sync. pulse) does not start until after the blanking pulse has cut off the scanning beam. It also insures that any discrepancies which may exist in the leading edge of blanking do not affect either the timing or the amplitude of sync.

The choice of the nominal width of horizontal sync. (0.08 H, see diagram 5 in Fig. 5) was influenced by three factors. First, the width should be as great as possible so that the energy content of the pulses will be large compared to the worse type of noise pulses which may be encountered in the transmission process, thus providing maximum immunity to noise. Second, the width should not be greater than is necessary to meet the first condition because average power requirements of the transmitter may thereby be minimized. Modulation of the picture transmitter is such that sync. pulses represent maximum carrier power; hence it is desirable to keep the duty cycle as small as possible. Third, the horizontal sync. pulses should be kept as narrow as possible so as to maintain a large difference between these pulses and the segments of the vertical sync. pulses described in the following paragraph. Such a large difference makes it easier to separate the vertical sync. from the composite sync. signal. It has also been recognized that the back porch is useful for a special type of clamping for d-c restoration. Hence it should be as wide as possible.

Vertical sync. pulses are also basically rectangular in shape, but are of much greater duration than the horizontal pulses thus providing the necessary means for frequency discrimination to distinguish between them. However, each vertical sync. pulse has six *slots* cut in it which make it appear to be a series of six wide pulses at twice horizontal frequency, i.e., wide compared to horizontal sync. pulses. The slots contribute nothing to its value as a vertical sync. pulse but do provide means for uninterrupted information to the horizontal scanning circuit.

Before and after each vertical pulse interval are groups of six narrow pulses called *equalizing pulses*. These also are for the purpose of maintaining continuous horizontal sync. information throughout the vertical sync. and blanking interval. The repetition frequency of the equalizing pulses and the slots in the vertical pulses is twice the frequency of the horizontal sync. pulses. This doubling of the frequency does two things. First, it provides an arrangement in which the choice of the proper alternate pulses makes available some kind of a horizontal sync. pulse at the end of each scanning line in either even or odd fields. Second, it makes the vertical sync. interval and both equalizing pulse intervals exactly alike in both even and odd fields. The importance of this latter result will become evident in following paragraphs. It is important to point out that the leading edge (downward stroke) of each horizontal sync. pulse and of each equalizing pulse, and the trailing edge (again the downward stroke) of each slot in the vertical pulses are responsible for triggering the horizontal scanning circuit in the receiver; hence the intervals of H or H/2 apply to these edges.

Perhaps the most difficult problem in synchronizing, and the one in which there is the largest number of failures, is that of maintaining accurate interlacing. Discrepancies in either timing or amplitude of the vertical scanning of alternate fields will cause displacement, in spaces of the interlaced field. The result is non-uniform spacing of the scanning lines which reduces the vertical resolution and makes the line structure of the picture visible at normal viewing distance. The effect is usually called *pairing*. The maximum allowable error in line spacing in the kinescope to avoid the appearance of pairing is probably 10% or less. This means that the allowable error in timing of the vertical scanning is less than one part in 5000. This small tolerance explains why so much emphasis is placed on the accuracy of vertical synchronizing.

The presence of a very minute 30-cycle component in the *vertical* scanning invariably causes pairing. The fact that the rasters produced in alternate fields are displaced with respect to each other by half a line means that the *horizontal* sync. signal has an inherent 30 cycle component. It is this situation and the need to prevent any transfer of the 30-cycle component into the vertical deflection which account for the introduction of the double-frequency equalizing pulses before and after the vertical sync. pulses. The vertical sync. pulses are separated from the composite sync. signal, before being applied to the vertical scanning oscillator, by suppressing the horizontal sync. pulses in an integrating network similar to that illustrated in Fig. 8(a).

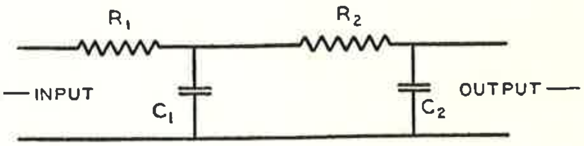


Fig. 8 (a).—Two-stage integrating network used to separate vertical sync. pulses from the composite picture. Circuit values are as follows:—

$R_1 = 10{,}000$ *ohms*	$R_2 = 200{,}000$ *ohms.*
$C_1 = .012\ \mu F$	$C_2 = 150\ \mu\mu F$
$T_1 = 120$ *ms.*	$T_2 = 30$ *ms.*

Most receivers employ integrating networks of three stages instead of the two illustrated. However, the general character of the circuit action is clearly shown by the wave form diagrams in Fig. 8(b). In simple terms, the equalizing pulses before the vertical sync. pulses cause the integrating network to "forget" the difference between alternate fields by the time the vertical sync. pulses begin. This is illustrated by the gradual convergence of curves *f* and *g* during the equalizing pulse interval, as the result of integration in the first stage alone. The effect of further integration in the second stage is shown by curve *h*, which is typical of the pulses applied to the vertical oscillator in a receiver. Thus, the 30-cycle component is effectively eliminated, from the standpoint of accurate timing of the start of vertical retrace, by the addition of the first set of equalizing pulses and the slots in the vertical pulse itself. The second set of equalizing pulses which follow the vertical pulse affect to some extent the impedance of the circuit to which the vertical scanning oscillator is coupled, and thus affect the amplitude of its output; hence these pulses help to provide more nearly constant output of the oscillator. Both sets of equalizing pulses contribute materially to the necessary accuracy of vertical synchronizing.

The width of an equalizing pulse is half the width of a horizontal sync. pulse (see diagram 4 of Fig. 5, and Fig. 8). This width is chosen so that the a-c axis of the sync. signal does not change at the transition from the line-frequency horizontal sync. pulses to the double-frequency equalizing pulses. The curves f_2 and g_2 in Fig. 8 illustrate the undesirable effect of making the equalizing pulses the same width as the horizontal sync. pulses. There is a slight rise in the integrated wave during the equalizing pulse interval which could cause premature triggering of the vertical oscillator in the receiver if the hold control were adjusted near one end of its range. This rise in the integrated wave results from the change in the a-c axis.

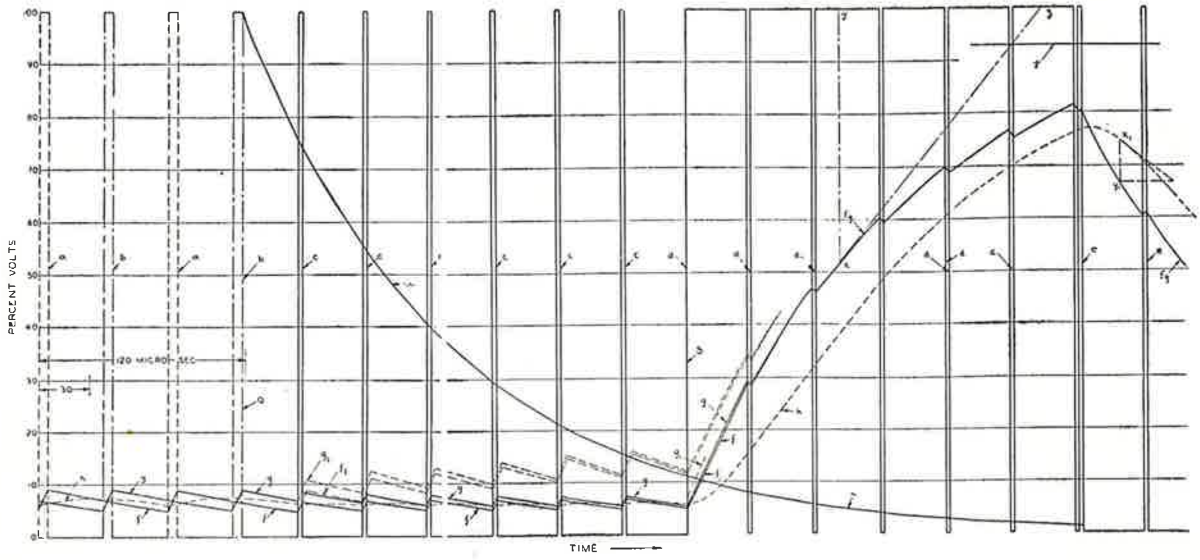


Fig. 8 (b) General action of the integrating circuits in the region of the equalizing pulses and beginning of the vertical sync. pulse. Curves are identified as follows:—Curve acde is Sync. Signal in region of even field pulse. Curve bcde is Sync. Signal in region of odd field pulse. Curve f is curve acde "integrated" by circuit having T = 120 micro-sec.*
Curve g is curve bcde "integrated" by circuit having T = 120.
Curve h is curve acde "integrated" by two stages, having T_1 = 120 and T_2 = 30.
Curve i shows rate at which equalization occurs for T = 120.
NOTE: In this diagram the slots in the vertical pulse are shown the same width as the equalizing pulses, whereas they are actually the same width as the horizontal sync. pulses (see Fig. 5).

The width of the slots in the vertical sync. pulses is approximately equal to the width of the horizontal sync. pulses. The slots are made as wide as possible so that noise pulses or other discrepancies occurring just prior to the leading edge of a slot (i.e., near the end of the preceding segment of a vertical pulse) do not trigger the horizontal oscillator. Premature triggering can happen if the noise pulse is high enough and if it occurs very close in time to the normal triggering time. Increased time-separation (a wider slot) reduces likelihood of such premature action. Here again, the requirements of special clamping also are met more easily if the slots are made as wide as possible.

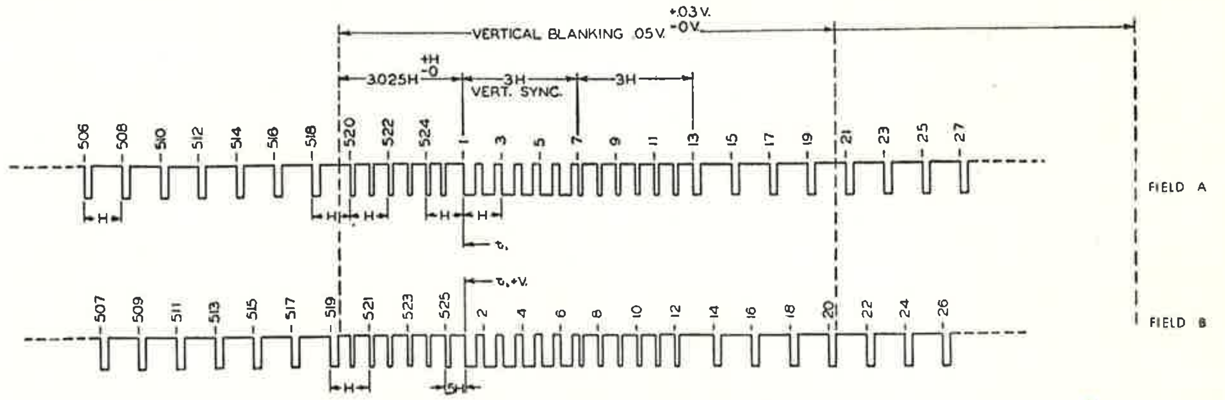
A further important advantage of the RMA system of separating the vertical sync. by means of frequency discrimination is that the integrating network is a potent factor in reducing the effect of noise on vertical synchronizing. Noise signals contain mostly high-frequency components; hence they are almost completely suppressed by the integrating circuit.

Differentiation, or suppression of the low-frequency components, of the sync. signal before it is applied to the horizontal scanning oscillator is done sometimes, but is not necessary, and has not been indicated in Fig. 7.

The methods just described for synchronizing the scanning circuits in a television receiver are complicated by the need for transmitting the complete information over a single channel. In the case of the scanning circuits in the cameras, however, the situation is very different. The cameras and the synchronizing generators are so close to each other that there is no problem in providing as many wire circuits as may be desired. Therefore it is customary to use what are called *driven* scanning circuits in cameras and sometimes in picture monitors used with the cameras. Separate pulse signals, called *driving signals*, are produced in the synchronizing generator for exclusive use in the terminal equipment. Horizontal and vertical driving signals are completely independent of each other in the RCA system and are carried on separate transmission lines to the points of application. The driving signal pulses trigger directly the sawtooth generators which produce the scanning waveforms. This method reduces interlacing errors in the terminal equipment to the errors inherent in the driving signals.

Fig. 9 illustrates a portion of the scanning lines appearing on a kinescope as a result of the application of a television signal composed of RMA sync. and blanking pulses. The group of lines shown are those occurring in the neighbourhood of a vertical retrace period including a few before and a few after the vertical blanking pulse. As noted on the diagram, the triggering of the lines has been displaced both vertically and horizontally so that the shadows produced by the sync. and blanking pulses appear near the centre of the raster rather than in the normal positions at the edges of the raster. This displacement is brought about simply to clarify the illustration of the effect of the pulses on the raster.

*Diagram prepared by A. V. Bedford, RCA Laboratories, for presentation to the N.T.S.C.



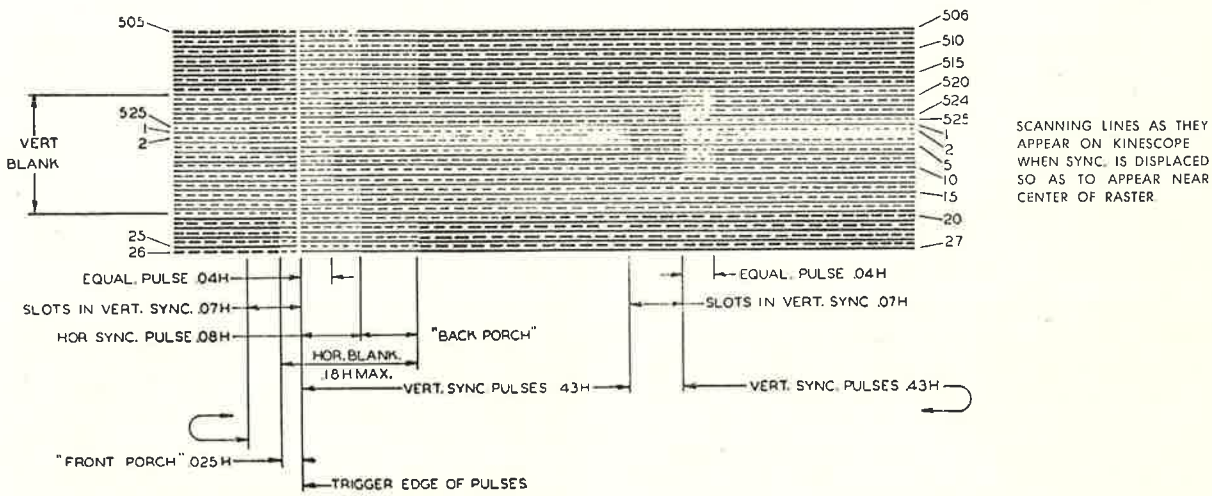


Fig. 9.—A portion of the scanning lines appearing on a kinescope as a result of the application of a television signal composed of RMA sync. and blanking pulses. The group of lines shown are those occurring in the neighbourhood of the vertical retrace period, including a few before and a few after the vertical blanking pulse.

The shadows produced thus are called a *pulse cross.* When expanded vertically so that individual scanning lines become easily apparent, the pulse cross becomes a ready means of checking the performance of the sync. generator. The shadows produced by each different kind of pulses are indicated clearly on the diagram. With linear scanning, the horizontal dimensions of the shadows are measures of time or pulse width, and, because of the expanded scale, they provide a relatively accurate means of measuring pulse width. Furthermore, by counting appropriate lines, the numbers of equalizing pulses, slots, vertical sync. pulses, etc., can be checked easily.

A useful piece of station test equipment can be made by modifying the deflection circuits in a picture monitor to provide the displacement of the lines and the extra large vertical expansion described.

(To be concluded in the next issue of Radiotronics.)

CORRECTION RADIOTRONICS 141

Page 16, column 2, line 4 should read:

Line impedance $= \sqrt{170 \times 600}$.

Noted.

AMENDMENT RADIOTRONICS 138

On page 66 it was stated that the maximum-signal power output was approx. 30 watts for triode connected 807's in Class AB_1 under the new operating conditions as listed. RCA now advise that this value should be 15 watts.

Noted.

Television Antennas And Transmission Lines*

By JOHN R. MEAGHER

Television Specialist, RCA Renewal Sales

Part 1 of this article was published in Radiotronics, No. 141, February, 1950.

PART II—GHOSTS

We did not originally intend to devote much time to ghosts because the subject is generally well understood, at least in regard to the common or garden variety of ghosts. But on analyzing the subject, we were surprised to realize the great variety of ghosts that may, under unfavourable circumstances, unhappily haunt the kinescope. So we decided to make you better acquainted with them.

For instance, have you met all of the following members of the ghost family?—

Leading ghosts,
Trailing ghosts,
Positive ghosts,
Negative ghosts,
Multiple ghosts,
Fluttering ghosts,
Transmission Line ghosts,
Tunable ghosts.

If you do not recognize all of them, you may have wasted time trying to eliminate some varieties by orienting the antenna, which doesn't phase them a bit.

Trailing ghosts.

The usual type of ghost, or echo, or secondary image, is caused by reflection of the transmitted signal from a building or other structure or from a hill or cliff. The reflected signal, which is usually weaker than the direct signal, arrives at the receiving antenna later than the direct signal, and the ghost, therefore, appears on the right-hand or trailing side of the original picture.

The building or other object from which the signal is reflected may be situated in various locations with respect to the TV station and to the receiver, as shown in Fig. 1.

In Fig. 1A where the direct and reflected signals arrive at the receiver from the same general direction, there is no practical remedy at present to eliminate the ghost.

Occasionally, in the hope that the plane of polarization of the reflected signal has been changed from horizontal, experimenters try tilting the antenna in various planes to get it at right angles to the plane of the reflected signal. Unfortunately, such trials have usually proved inconclusive or futile.

In Fig. 1B, where the reflected signal is arriving at the receiver from one side, it can be minimized by orienting the antenna for least pick-up in this direction.

*Reprinted from RCA Radio Service News (Jan.-Feb., 1949) by courtesy of Radio Corporation of America.

In Fig. 1C, where the reflected signal is arriving at the receiver from the rear, a reflector on the antenna is helpful because it reduces rear pick-up to some extent. As mentioned previously, it is not a cure-all for this condition. A reflector of large dimensions, such as a metal billboard, or a large screen of chicken wire, is generally helpful in reducing rear pick-up. In a few cases, it is possible to position the antenna so it is "shielded" from the rear by a closely adjacent steel building.

In mid-city locations, it is sometimes advantageous to orient the antenna for maximum pick-up of a strong reflected signal, and minimum pick-up of the direct signal. This expedient may be necessary in locations where the direct signal is blocked by an intervening building, as in Fig. 1D.

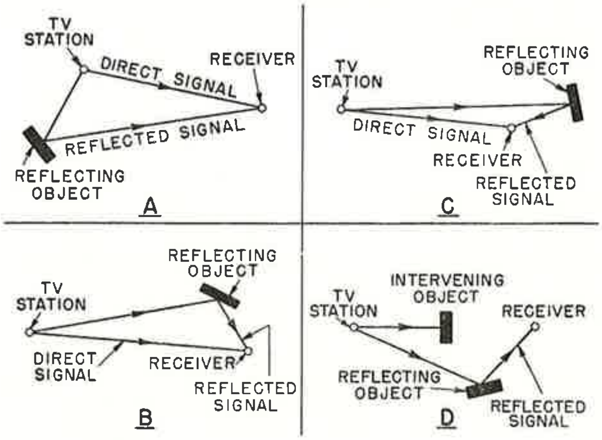


FIG 1

In locations where several reflected signals from different buildings reach the receiver, there are several ghost images in the picture. These images are referred to as multiple ghosts, or multiple reflections. A typical condition in which multiple ghosts are produced is shown in Fig. 2.

A ghost may be either positive or negative, using these terms in their photographic sense, where a negative is a reversed image; that is, the black portions are white and the white portions are black.

Whether a ghost is positive or negative depends on the relative r-f phase of the direct and reflected signals. See Fig. 3. The relative phase depends on the position of the antenna. If the antenna is moved some distance toward or away from the transmitter, the relative phase changes, and the direct and reflected signals either aid or oppose, producing a positive or a negative ghost respectively.

Fluttering ghosts.

When an aircraft is in the vicinity of the receiver, it reflects signals from the TV transmitter to the receiver. The receiver also gets direct signal from the transmitter. The relative phase of the direct and reflected signals arriving at the receiver changes as the plane travels along. The two signals alternately aid and oppose each other, producing a flutter in picture brightness and also a flutter in the ghost image. In TV receivers with automatic gain control on the picture i-f amplifier, the fluctuation in brightness is largely eliminated. Refer to Fig. 4.

The rate of flutter depends on the position, height, speed, and direction of the plane. The rate of flutter changes as the plane moves along.

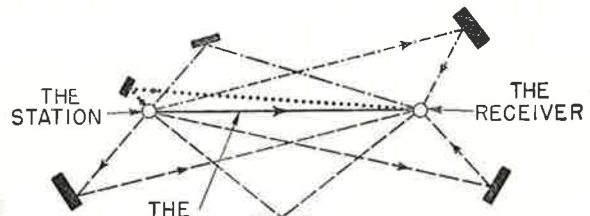


FIG. 2

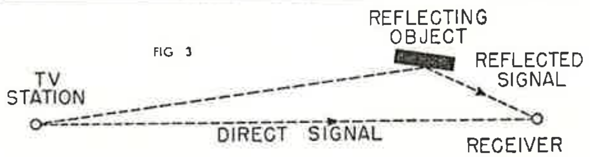


FIG 3

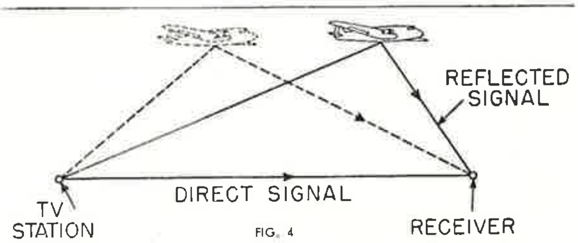


FIG. 4

Occasionally signals from a distant TV station that is beyond normal receiving range may be seen for short periods due to reflection from a plane as shown in Fig. 5. This occurrence demonstrates that the signals are passing overhead and could be intercepted if it were possible to place an antenna high enough in the air.

The usual type of ghost appears on the *right-hand* side of the picture. It is termed a "trailing" ghost because the reflected signal travels a longer path than the direct signal and arrives later than the direct signal.

There is a condition where one or more images may appear on the left-hand side of the picture. We have termed these "leading" ghosts for lack of a better description.

This type of ghost appears in locations where the following conditions exist, as shown in Fig. 6;

1. Location relatively close to the transmitter.
2. Considerable signal pick-up in the r-f or 1st detector circuits of the receiver, with the antenna disconnected.

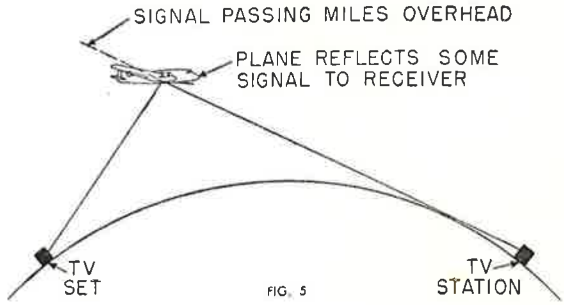


FIG. 5

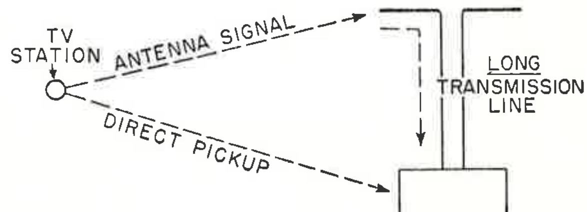


FIG 6

3. Long run of transmission line to the antenna.

Because the antenna signal is delayed travelling down the transmission line, the direct signal picked up in the r-f or 1st detector circuits appear ahead, or on the left-hand side of the antenna-signal picture.

The remedy in this case is to—

1. Reduce the direct pick-up in the receiver by shielding the r-f and detector circuits, or the entire chassis.
2. Increase the signal from the antenna. (If there is some type of antenna distribution system, it may be defective, or attenuating the signal too much.)

In any case where direct signal pick-up by the receiver is evident in the picture, this signal will be altered by persons moving around the room, or near the TV receiver. (Remember that movement close to an unshielded transmission line will alter the antenna signal, particularly when the receiver does not terminate the line correctly.)

Under the conditions mentioned above, but where the transmission line is shorter than a few hundred feet, the direct signal will not appear as a separate image but will blend with the antenna signal to produce a picture of poor quality, which will change in quality when someone moves around the room or near the receiver.

In one actual case where leading ghosts were encountered, the following checks were made:

1. With the antenna connected to the receiver and contrast and brightness correctly adjusted, there were about 12 distinct images on the kinescope.
2. Disconnecting the antenna from the receiver, without disturbing the contrast control, it

was found that there were about 10 images.

3. From this it was assumed that the antenna was contributing very little to the signal, and that most of the pick-up was from signals "coming in the window" and being picked up in the r-f or detector circuits. The numerous images were due to reflected signals from different tall buildings in the vicinity.
4. Shielding of the r-f and detector circuits did not help in this case.
5. It was then assumed that the r-f valve had no gain, to account for the fact that connecting the antenna to the receiver produced very little difference in the picture on the kinescope. This proved to be the trouble: A resistor in the r-f bias circuit had dropped to a very low value, which resulted in the r-f valve being biased off at all times.
6. When this trouble was corrected and the antenna was connected to the receiver, the picture contrast control had to be turned back considerably because the antenna signal was then being amplified in the r-f stage.
7. Leaving the contrast control in this new setting, the antenna was disconnected, and it was observed that no images were visible on the kinescope, indicating that at this low-gain setting of the contrast control, the direct signal pick-up in the r-f and detector circuits was not strong enough, compared to the antenna signal, to cause trouble.
8. The antenna was then reconnected, and after some time spent in finding the correct position and orientation, the final picture was excellent with only a few very faint reflections or trailing ghosts.

Another case in a similar location was traced to a defective component in the antenna distribution system.

These two actual cases are mentioned here because usually the presence of multiple ghosts is blamed on the particular location, and on the position and orientation of the antenna. It is worthwhile, at least in strong-signal areas, to check other factors, as proved in these two instances.

In working on reflection problems, it is sometimes helpful to know exactly which building or structure is acting as the reflecting object in producing a particular ghost.

To locate the reflecting object, it is necessary first to determine the "additional air-path distance" that the reflected signal must travel. (Any reflected signal travels a longer distance than the direct signal.)

The additional air-path distance is determined from knowledge of two facts:

1. The scanning spot in the kinescope requires approximately 53 millionths of a second, or 53 microseconds, to travel from the left- to the right-hand side of the picture. (Unblanked portion of picture.)
2. Radio signals travel approximately 1000 feet in one microsecond in air. In 53 microseconds a radio signal travels approximately 53,000 feet or 10 miles. Therefore, during the time it takes for the spot to travel from the left- to the right-hand side of the picture, a radio signal travels about 10 miles. The horizontal width of the picture provides a distance scale, somewhat like the range scale on the radar "A" scope.

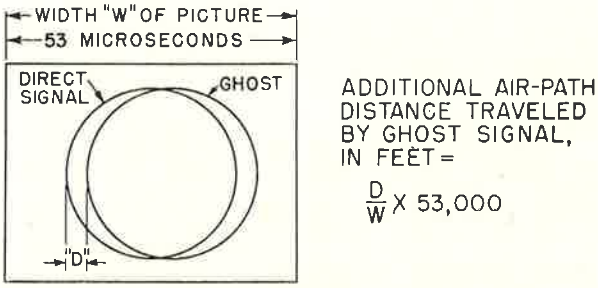
The procedure in determining the additional air-path distance of the reflected signal is as follows:

1. Adjust the picture width so it is the same size or slightly smaller than the mask, and adjust for the best possible horizontal linearity.
2. Measure the horizontal distance in inches between a point in the original picture, and the corresponding point in the ghost.
3. Measure the width of the picture in inches.

The additional air-path distance in feet is

$$\frac{\textit{distance between corresponding points}}{\textit{width of picture}} \times 53{,}000$$

As an example, if the distance between corresponding points in the original and ghost pictures is one inch, and the width of the picture is 8 inches, the additional air-path distance is $\frac{1}{8} \times 53{,}000$ or approximately 6,600 feet.



(RADIO WAVES TRAVEL APPROX. 1000 FEET IN ONE MICROSECOND.)

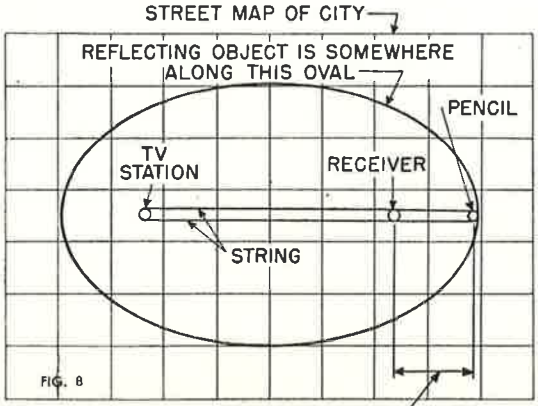
FIG 7

Note that this is the *additional* air-path distance that the reflected signal travels. It is NOT the distance from the reflecting object to the receiver or to the transmitter. For instance, if the distance from the transmitter to receiver is 50 miles, the direct signal travels 50 miles, and in the above example the reflected signal travels 50 miles plus 6,600 feet.

In this particular example, if the reflecting object were directly in back of the receiver, it would be one-half of 6,600 feet or 3,300 feet in back of the receiver.

To find the buildings or other objects that could produce a ghost with a specific additional air-path distance, it is possible to draw an oval line on a map as shown in Fig. 8. The additional air-path distance of a reflected signal is the same for all points along this line. Any large building or structure along this line can be the reflecting point.

This method is useful only when the distance between transmitter and receiver is relatively short.



Transmission-line ghosts

When the transmission line is not correctly terminated by the receiver, a portion of the signal is reflected at the receiver and travels back up the line to the antenna. If the antenna does not correctly terminate the line, a portion of this signal is reflected and travels down the line to the receiver, where it produces a trailing ghost.

With a normal length of transmission line, the reflected signal takes very little time in travelling up and down the line, so it is only slightly delayed and does not appear as a separate ghost. It merges with the original picture signal and affects the picture quality.

With a sufficiently long run of transmission line, the reflected signal appears as a separate trailing ghost.

Occasionally it is necessary to determine whether a particular ghost is due to incorrect termination of the transmission line, or to an external reflected signal. This determinatiton will show whether it is necessary to improve the line termination or to reorient the antenna.

First determine the additional air-path distance of the ghost, as described previously.

Then determine the equivalent air-path length of the particular transmission line, which is equal to:

$$\frac{\textit{length of line in feet} \times 2}{k}$$

where k is the velocity constant of the particular line and is approximately 0.83 for some types of 300 ohm ribbon line.

As an example, let us assume that the additional air-path distance of a ghost is 4,000 feet and that the 300-ohm transmission line is 500 feet long. Then the equivalent air-path length for 500 feet of 300-ohm ribbon line, for a single reflection is

$$\frac{500 \times 2}{.83} \quad 1{,}200 \text{ feet (approx.)}$$

Because the ghost signal in this example has an additional air-path distance of 4,000 feet, it cannot be caused by reflection in the transmission line which has an equivalent air-path length of 1,200 feet.

Tunable ghosts

Echoes that vary in number and intensity with adjustment of the tuning control on the TV receiver are referred to as "tunable ghosts" or tunable echoes, and may be caused by incorrect alignment of the r-f picture i-f amplifiers, or by regeneration.

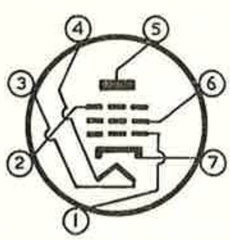
(*Continued in next issue of Radiotronics*)

CIRCUIT LAB. REPORT **NUMBER 2**

Grounded-Plate Type 6AU6 Triode Connection For Pre-Amplifier Use*

Difficulty is usually experienced in reducing to satisfactorily low levels the hum and noise from the input stage of a high gain a-f amplifier. If for example noise is required to be 60 db. below full output in an amplifier which is fully driven by a 5 millivolt input signal, then the noise and hum voltage at the grid of the first valve must be 5 microvolts or less.

In valves having the grid brought out to a top cap, the hum voltage induced electromagnetically from stray magnetic fields into the loop of wiring between grid and cathode of the valve may be excessive, and with double-ended valves it is not easy to reduce the area of this loop sufficiently. Further, electrostatic shielding of the grid circuit components and leads, and of the grid cap itself, is



Pin 1	Grid No. 1.
Pin 2	Grid No. 3.
Pin 3	Heater.
Pin 4	Heater.
Pin 5	Plate.
Pin 6	Grid No. 2.
Pin 7	Cathode.

Fig. 1.—Type 6AU6 base connections.

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

also necessary, but is difficult to achieve satisfactorily with this type of valve construction.

Single-ended valves minimize these troubles, but introduce a new one. The proximity of the grid, heater and plate contacts on the valve socket make for severe leakage and capacitance requirements of the valve base and socket. Thus, if leakage were the only consideration the insulation resistance between control grid and heater pins, in order to realise the above noise level, would need to be of the order of 50,000 megohms if the grid circuit impedance were 0.1 megohm.

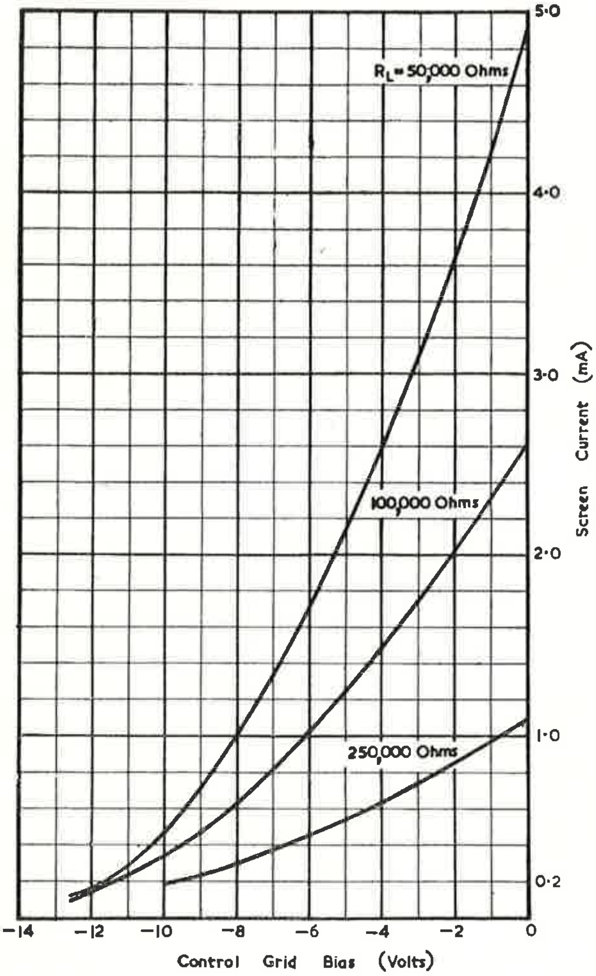


Fig. 2.—Curves of 6AU6 screen current vs. control grid bias ($E_p = E_{cs} = 0$) with different load resistors.

However, by using the guard ring principle satisfactory attenuation can be achieved more readily, and if, for instance, earthed contacts and the earthed centre spigot can be arranged to isolate the heater contacts from the grid and plate contacts, then hum problems due to leakage are greatly minimized. It is assumed in such cases that neither side of the heater winding is earthed, as centre tapping is usually necessary to reduce the hum level to a minimum.

Figure 1 shows, that with the triode connection mentioned in Radiotronics No. 139 (grid Nos. 2 and 3, plate and shield connected together), there is no low impedance connection to ground between grid 1 and heater, so that hum trouble in the grid circuit is possible. By earthing pin 2 (the suppressor) this trouble may be overcome, but under the recommended triode operating conditions, this leads to the maximum rating for the screen dissipation being exceeded in some valves. In addition, with pin 2 earthed, the leakage from heater to pins

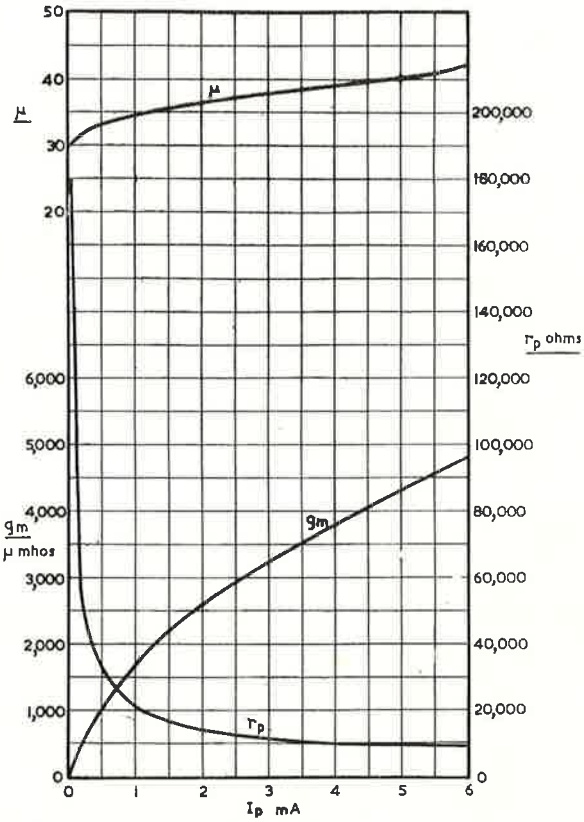


Fig. 4.—Curves of 6AU6 μ, g_m and r_p vs. I_p connected with grid No. 2 used as plate and with (E_p = 100V, E_{c1} varied). The valve is triode grid No. 3 and the pentode plate earthed.

5 and 6 (triode plate) is unbalanced and will be troublesome at low signal levels. Leakage in the 6AU6 itself will not be serious because its base material is glass, and in this respect it is an improvement over previous types with moulded bakelite bases (particularly single-ended types), but the insulation requirements for the socket will still be severe.

A recommended method of operating the 6AU6 as a pre-amplifier to overcome these troubles is to use grid 2 as the triode plate, with grid 3, the pentode plate and the external shield grounded. This gives an earthed contact on either side of the

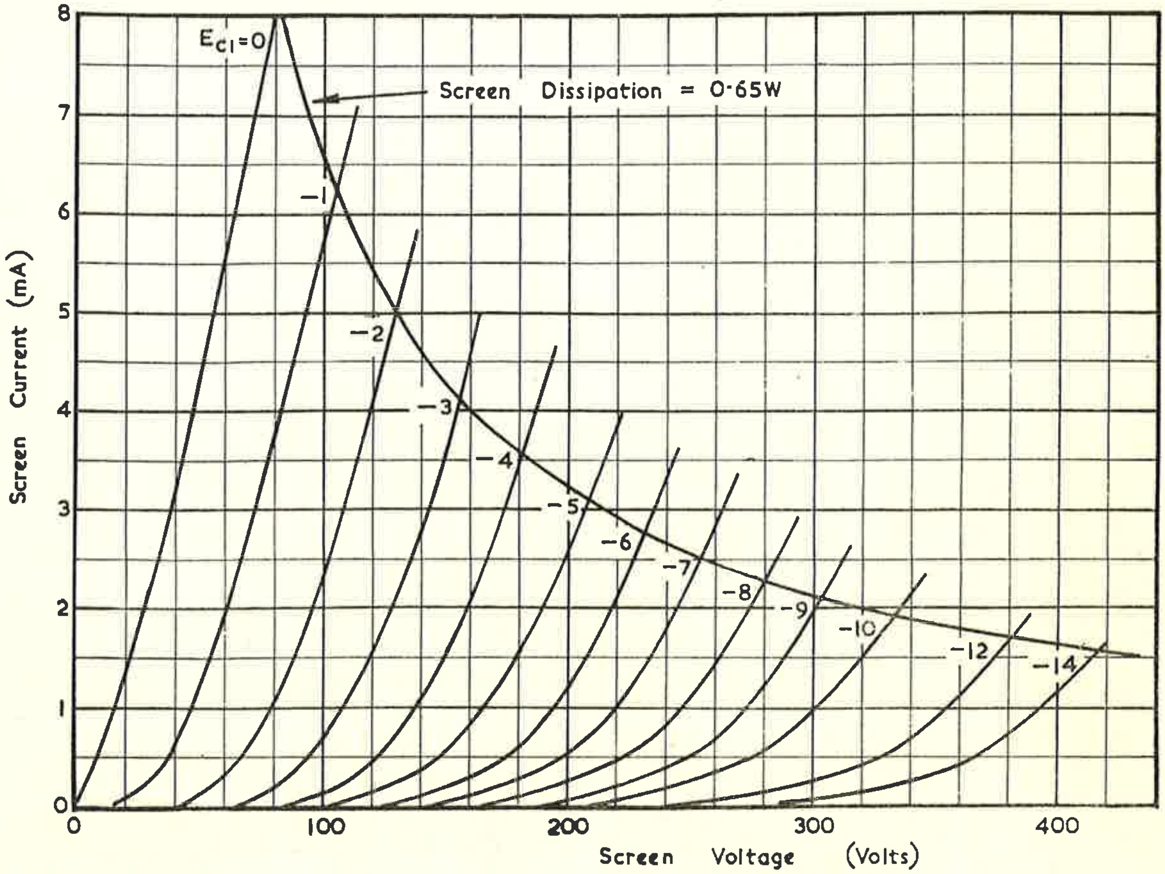


Fig. 3.—Curves of 6AU6 screen current vs. screen voltage for fixed values of grid No. 1 bias ($E_p = E_{c3} = 0$).

heater pins and the plate contact, while the grid has an earthed contact on one side and the cathode on the other.

Triode connection is desirable for a pre-amplifier because triodes are inherently less noisy than pentodes, firstly because of their higher $g_m/\sqrt{I_b}$ ratio and secondly because there is no noise due to random division of current between plate and screen.

The linearity of the mutual characteristic of grid 2 used as the plate is good, as indicated in Fig. 2, in which the dynamic characteristic for three different resistive plate loads is shown.

There are, however, two disadvantages of this method of connection. Firstly, the screen grid dissipation must not exceed 0.65 watt, which may restrict a transformer or choke-coupled design, and secondly, at high output levels distortion is higher than with the standard triode connection. This grounded-plate connection is accordingly recommended only for low level pre-amplifier use.

Figure 3 gives the plate characteristic of the 6AU6 operated in this manner, and Figure 4 shows μ, g_m and r_p as a function of plate current.

Some typical operating conditions are set out below.

Typical Conditions

E_{bb} = 300V.

R_L (Megohms)	0.05	0.1	0.25
R_K (Ohms)	230	450	1,000
I_b (mA)	4.3	2.35	1.0
Stage gain*	25	25	23

E_{bb} = 180V.

R_L (Megohms)	0.05	0.1	0.25
R_K (Ohms)	450	750	1,600
I_b (mA)	2.4	1.3	0.59
Stage Gain*	21	23	21

E_{bb} = 90V.

R_L (Megohms)	0.05	0.1	0.25
R_K (Ohms)	1,100	2,000	4,000
I_b (mA)	0.93	0.54	0.25
Stage Gain*	14	14	13

* Resistance of following grid leak = 0.5 megohm.

Radiotron Type 12AX7 High-Mu Twin-Triode Amplifier

(Reprinted by courtesy of Radio Corporation of America)

Radiotron type 12AX7 is a heater-cathode type of high-mu, twin-triode amplifier featuring a small glass envelope with integral button 9-pin base, separate terminals for each cathode, and a mid-tapped heater to permit operation from either a 6.3- or 12.6-volt supply.

Having characteristics which are similar to those of the larger types 6SL7-GT and 12SL7-GT, the 12AX7 like these types is useful in many diversified applications including phase inverters, multivibrators, and numerous industrial control devices where high voltage gain and low heater power are important design factors. In such equipment, the 12AX7 can be used to advantage because of its compact size, its separate cathode terminals, and its economical consumption of heater power at either of the two voltages.

GENERAL DATA

Electrical:

Heater, for Unipotential Cathodes:

Heater Arrangement	*Series*	*Parallel*	
Voltage (a.c. or d.c.)	12.6	6.3	volts
Current	0.15	0.3	ampere

Direct Interelectrode Capacitances:°

	Triode Unit T_1	*Triode Unit T_2*	
Grid to Plate	1.7	1.7	μμF
Grid to Cathode	1.6	1.6	μμF
Plate to Cathode	0.46	0.34	μμF

Mechanical:

Mounting Position	Any
Maximum Overall Length	$2\frac{3}{16}''$
Maximum Seated Length	$1\frac{15}{16}''$
Length from Base Seat to Bulb Top (excluding tip)	$1\frac{9}{16}'' \pm \frac{3}{32}''$
Maximum Diameter	$\frac{7}{8}''$
Bulb	T-$6\frac{1}{2}$
Base	Small-Button Noval 9-Pin

Socket connections:

Pin 1 — Plate (Triode No. 2)
Pin 2 — Grid (Triode No. 2)
Pin 3 — Cathode (Triode No. 2)
Pin 4 — Heater
Pin 5 — Heater
Pin 6 — Plate (Triode No. 1)
Pin 7 — Grid (Triode No. 1)
Pin 8 — Cathode (Triode No. 1)
Pin 9 — Heater Mid-Tap

CLASS A_1 AMPLIFIER

Values are for each unit

Maximum Ratings, *Design-Centre Values:*

Plate Voltage	300 max. volts
Plate Dissipation	1 max. watt
Grid Voltage:	
Negative bias value	50 max. volts
Positive bias value	0 max. volts
Peak Heater-Cathode Voltage:	
Heater negative with respect to cathode	180 max. volts
Heater positive with respect to cathode	180 max. volts

Characteristics:

Plate Voltage	100	250	volts
Grid Voltage	−1	−2	volts
Amplification Factor	100	100	
Plate Resistance	80000	62500	ohms
Transconductance	1250	1600	micromhos
Plate Current	0.5	1.2	mA

Typical Operation—Resistance-Coupled Amplifier:

Plate-Supply Voltage	90			180			300			volts
Plate Load Resistor	0.1	0.22	0.47	0.1	0.22	0.47	0.1	0.22	0.47	megohm
Grid Resistor (of following stage)	0.22	0.47	1.0	0.22	0.47	1.0	0.22	0.47	1.0	megohm
Cathode Resistor	4700	7400	13000	2000	3500	6700	1500	2800	5200	ohms
Cathode Bypass Capacitor●	2.4	1.4	0.8	3.5	2.1	1.1	4.0	2.3	1.3	μF
Blocking Capacitor●	0.013	0.006	0.003	0.013	0.006	0.003	0.013	0.006	0.003	μF
Peak Output Voltage■	6	9	11	25	34	39	57	69	77	volts
Voltage Gain	35#	45##	52‡	47▲	59▲	66▲	52▲	65▲	73▲	

° With no external shield.
\# At an output voltage of 2 volts r.m.s.
‡ At an output voltage of 4 volts r.m.s.
● The cathode by-pass capacitors and blocking capacitors have been chosen to give output voltages at 100 c/s (f_1) which are equal to 0.8 of the mid-frequency value. For any other value of (f_1), multiply the values of cathode by-pass and blocking capacitors by $100/f_1$.
\## At an output voltage of 3 volts r.m.s.
▲ At an output voltage of 5 volts r.m.s.
■ This peak output voltage is obtained across the grid resistor of the following stage at any frequency within the flat region of the output vs. frequency curve, and is for the condition where the signal level is adequate to swing the grid of the resistance-coupled amplifier valve itself to the point where its grid starts to draw current.

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