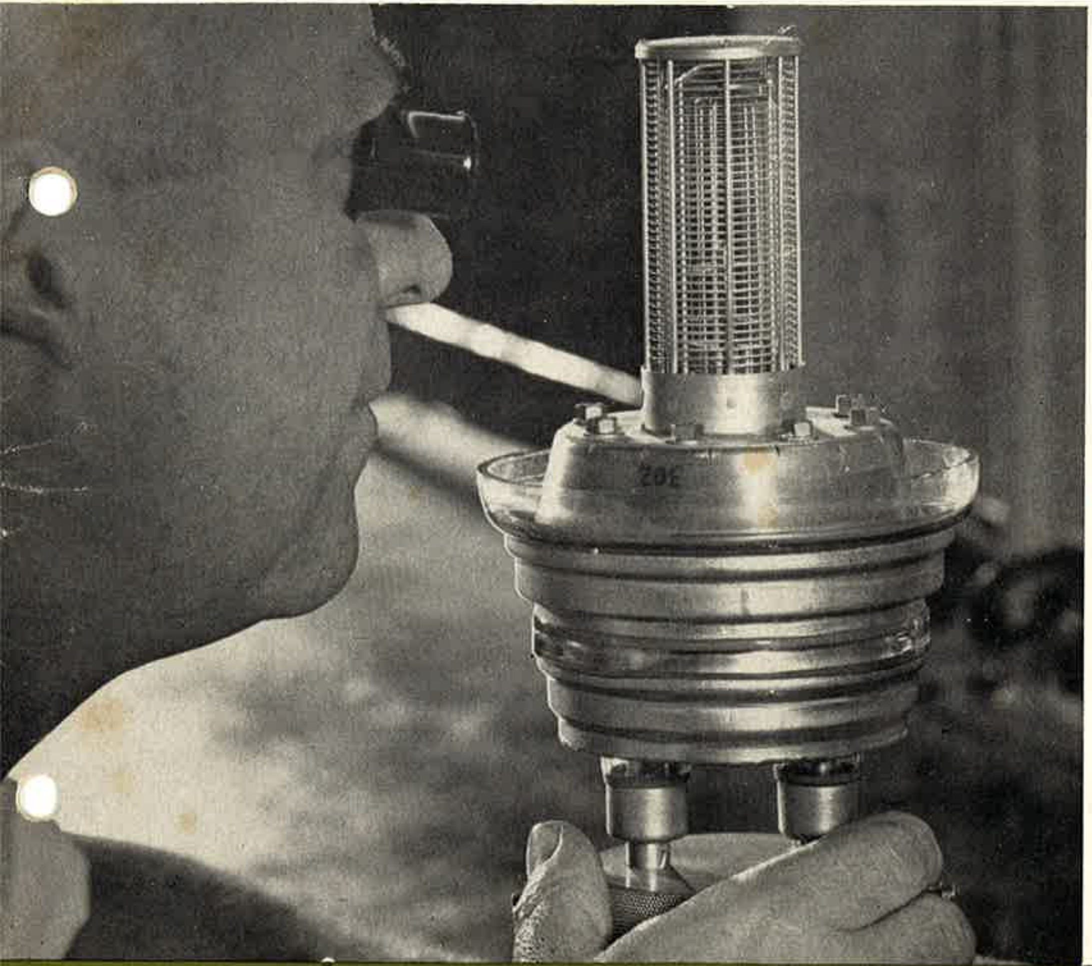
RADIOTRONICS



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COVER:

An operator at A.V.W. lining up the mounting assembly of a 6166 power valve.

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Automatic Black Level Control

This article is based on AWV Application Laboratory Report VR 102 by J. Van der Goot entitled "Investigation of Circuitry for the Automatic Control of Black Level in Television Receivers", dated February 28th, 1964. The object of the investigation which preceded the report was to design circuitry which keeps the black level at the picture tube constant regardless of different kinds of variations, such as setting of contrast control, changes in scene key, and so on. This presentation has been specially prepared for "Radiotronics".

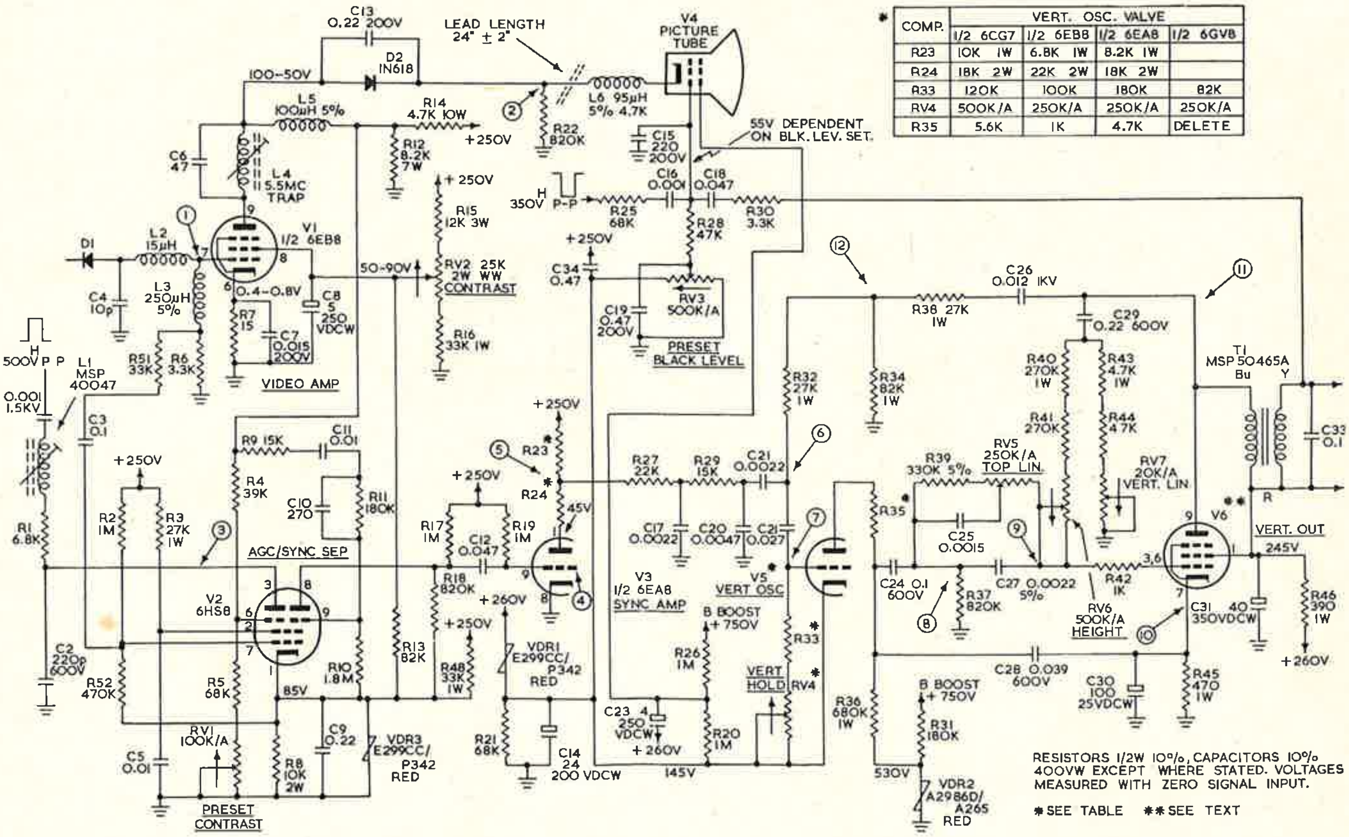
Requirements

Before outlining some existing and possible systems the requirements for a perfect system will have to be laid down. When a TV transmitter transmits black in accordance with the ABCB standards[1] the beam current in the picture tube of the TV receiver shall be zero, whereas luminance of any level transmitted shall cause a corresponding amount of beam current to flow.

Unfortunately there is no (simple) way in which black level can be permanently recovered in the receiver. However a constant level near black level that can be permanently recovered is the blanking level (level of front and back porches of horizontal synchronising pulses and the level during the vertical blanking period in between pulses). It is obvious that a near-to-ideal system could be provided if the blanking level were used as a reference level.

For the following discussion it is assumed that the difference between black level and blanking level may be ignored. Assuming that the blanking level at the picture tube "drive" electrode (most commonly the cathode) is held constant, all other picture tube electrode voltages particularly the grid No. 1 and grid No. 2 voltages must also be held constant regardless of variations such as the receiver's customer control adjustments and B+ voltage variations (due to video amplifier current drain variations with scene key

FIG. 1—Circuit diagrams of constant black-level and vertical deflection generator circuit.



COMP.	VERT. OSC. VALVE			
	1/2 6CG7	1/2 6EB8	1/2 6EA8	1/2 6GV8
R23	10K 1W	6.8K 1W	8.2K 1W	
R24	18K 2W	22K 2W	18K 2W	
R33	120K	100K	180K	82K
RV4	500K/A	250K/A	250K/A	250K/A
R35	5.6K	1K	4.7K	DELETE

RESISTORS 1/2W 10%, CAPACITORS 10% 400VW EXCEPT WHERE STATED. VOLTAGES MEASURED WITH ZERO SIGNAL INPUT.

* SEE TABLE ** SEE TEXT

and due to i.f. strip and tuner current drain variations with signal strength). If regulation of these voltages cannot be easily achieved, some kind of compensation may have to be introduced.

Assuming that this first part of the requirements has been fulfilled there remains the requirement of keeping the blanking level at the picture tube constant with changes in the blanking level transmitted (ABCB specifies[1] blanking level: 75 ± 2.5% of white-to-synchronising tip level).

A number of circuits have been devised to control the black level. The features important to the operation of these systems can be summarised as follows:

(a) Black level should be independent of:
  (i) setting of contrast control
  (ii) changes in d.c. component of the composite video signal (scene key)
  (iii) input signal strength
  (iv) variations of receiver tuning and difference in band-pass characteristics between channels
  (v) mains voltage fluctuations

(b) The shift in black level should be negligible when the transmitter's blanking level changes from one channel to another.

(c) The circuit should be compatible with synchronisation protection circuits.

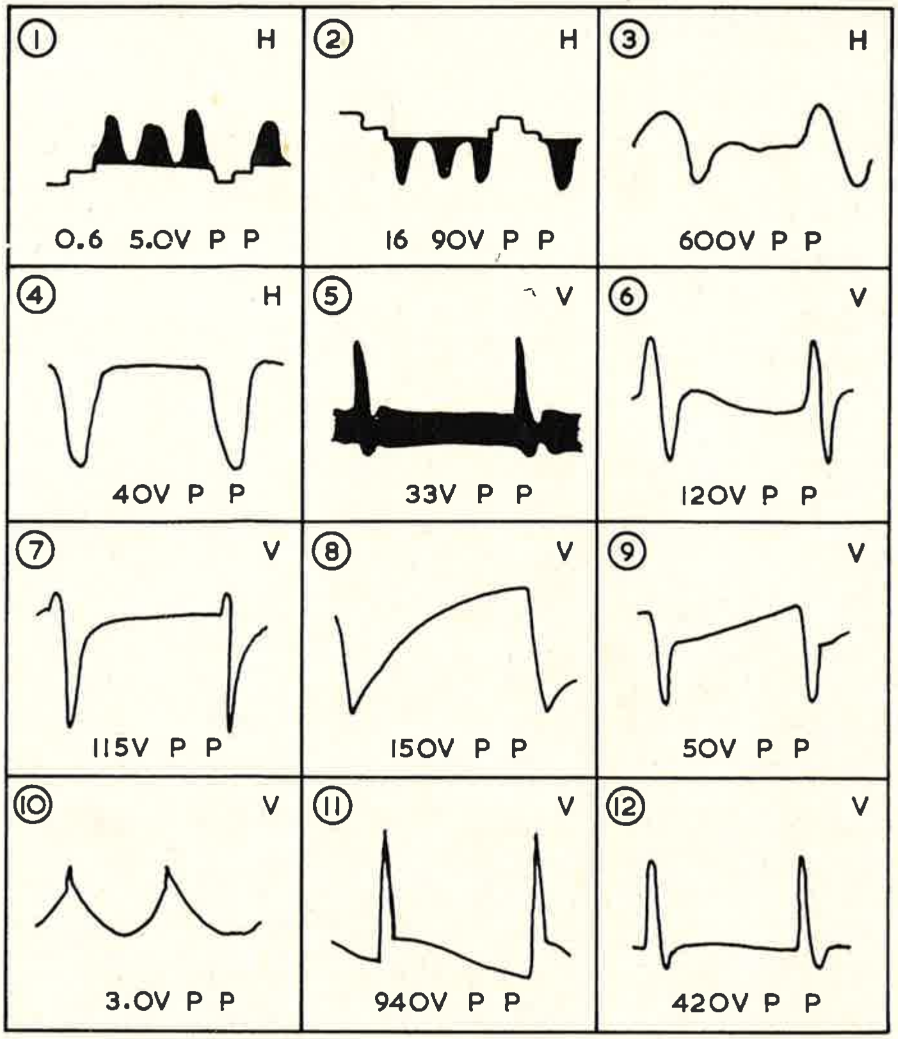


FIG. 2—Waveforms related to the reference points of the circuit diagram of Fig. 1.

EVALUATION OF DIFFERENT SYSTEMS

System A

In the first system to be evaluated, the keyed a.g.c. amplifier (constituting a high a.g.c. loop gain) virtually clamps the synchronising tips at a constant level. This level remains also virtually constant with variations of contrast control setting by the use of a potentiometer in the video amplifier's screen grid.

The effects with this system as related to the requirements specified in the foregoing sub-section on "Requirements" can be summarised as follows:

(a) Black level independence:
  (i) decreasing contrast by means of the control results in black level shifting toward "blacker than black" and vice versa
  (ii) change in d.c. component of the composite signal does not result in shift of black level
  (iii) change in input signal strength (above a.g.c. threshold) will not result in shift of black level
  (iv) variations in bandpass between channels will result in black-level shift
  (v) not known

(b) Black level shift relative to changes in transmitter blanking level. Change in the blanking level at the transmitter will result in a change of black level at the video amplifier output.

(c) Compatibility with synchronisation protection circuits. This system is compatible with synchronisation and a.g.c. protection circuits.

System B

In this system the a.g.c. amplifier clamps the synchronising tips, as under A, but the synchronising tip level at the picture tube "drive" electrode (most commonly the cathode) is made to vary with contrast setting to compensate for changes in blanking level and hence black level that would otherwise occur with changes in contrast. The AWA circuitry[2] is based on this principle.

The effects with system as related to the requirements previously specified can be summarised as follows:

(a) Black level independence:
  (i) black level is maintained
  (ii) d.c. component does not shift the black level
  (iii) black level does not shift with signal strength
  (iv) will result in shift of black level
  (v) special measures necessary to prevent the black level changing with mains voltage fluctuations

(b) Black level shift relative to changes in transmitter blanking level. The transmitter blanking level tolerance is ± 2.5%. Black level will shift due to this tolerance.

(c) Compatibility with synchronisation protection circuits. This system is compatible with synchronisation and a.g.c. protection circuits.

System C

In this system the a.g.c. amplifier clamps the back porch. The effects with this system as related to the requirements previously specified can be summarised as follows:

(a) Black level independence:
  (i) black level remains constant with setting of contrast control
  (ii) black level remains constant with change of d.c. component
  (iii) black level remains constant with change in input signal
  (iv) black level remains constant with variations in band-pass characteristics
  (v) special measures are necessary to prevent the black level changing with mains voltage fluctuations

(b) Black level shift relative to changes in transmitter blanking level. Here the black level is constant with changes in transmitter blanking level.

(c) Compatibility with synchronisation protection circuits. This system is compatible with synchronisation and a.g.c. protection circuits.

System D

A diode clamps the synchronising tip level. If used in combination with amplified a.g.c. the performance of this circuitry is equivalent to (A). However, its performance with noise is inferior to (A) due to the diode clamping at noise peak level. This circuitry is now seldom used.

The properties of systems A, B and C as described above are conveniently summarised in Table I. From this table it is evident that the back porch clamp circuitry offers the best performance. The operation of circuitry incorporating back-porch clamping has been adequately described[3].

Table I

PROPERTY	A.G.C. SYNC. TIP CLAMP (A)	AS (A) BUT WITH AUTOMATIC BLANKING LEVEL ADJUSTMENT (B)	A.G.C. BACK PORCH CLAMP (C)
Effect of setting of contrast control	No	No	No
Effect of change of scene key.	No	No	No
Effect of change of input signal strength (above threshold)	No	No	No
Effect of variations in band-pass characteristics between channels	Yes	Yes	No
Effect of mains voltage fluctuations	—	No	No
Effect of change of blanking level of Transmitter	Yes	Yes	No
Compatible with noise protection circuitry	Yes	Yes	Yes

AWV CIRCUIT FOR CONSTANT BLACK LEVEL WITH VALVE TYPE 6HS8

This circuit for providing constant black level is shown in Fig. 1, with the waveforms appropriate to the numbered points in the circuit shown in Fig. 2. This circuit was developed as an integral part of a television receiver. The different stages will now be described. It should be noted that four alternatives are available in the circuit diagram of Fig. 1 for V5, the vertical oscillator stage, the alternatives and component valves being tabulated in the drawing. No type number is specified in the drawing for V6, the vertical output stage. Three alternatives are available here, 6CZ5, 6EM5 and the pentode section of a 6GV8.

Video Amplifier

The pentode part of a 6EB8 is used as the video amplifier, V1. As the video amplifier is directly coupled to the picture tube (for low beam currents only) the plate voltage of the 6EB8 must not be permitted to rise above the EH-K(max.) of the picture tube. The EH-K(max.) (K pos.) of a typical picture tube, the 25QP4 for instance, is + 200V, therefore the plate load of the 6EB8 is in the form of a divider R14/R12 which reduces the effective E_{bb} to + 160V with a peak-white level at the "knee" which is approximately 45V and a synchronising tip level at approximately 145V. The max. output is approximately 100V p-p which corresponds to a maximum white-to-black level of approximately 65V p-p. This is sufficient to drive a (average drive characteristic) picture tube which has an E_{c2} relative to g_1 of 400V or slightly higher.

D2, C13 and R22 form a beam current limiter. The picture tube cathode current flows through R22. When the potential at the picture tube cathode caused by this current exceeds the voltage at the video amplifier plate the diode is biased in the reverse direction and the circuit then behaves as if it was capacitively coupled (C13 and R22). The advantage of capacitive coupling is that the average beam current which affects the E.H.T. voltage (to a degree determined by the E.H.T. regulation) remains constant with changes in scene key and/or contrast. Note that a certain percentage of d.c. coupling will be effected because of the back-resistance of the diode. The size of R22 determines the cross-over point from d.c. to a.c. coupling. The effect of the beam current limiter on the picture is as follows:

For low-key scenes (predominantly black) the picture tube is d.c. coupled and black level is maintained. For instance, a night scene appears as a night scene and does not appear "milky" as with capacitive coupling.

For high-key scenes (predominantly white) the picture tube is capacitively coupled. Blacks tend to appear "blacker-than-black" and dark greys will be rendered black. However, the latter is of little importance as detail in dark areas (which will be small because the scene is predominantly white) will be of little interest. The advantage is then that on the high key scene the picture will not exhibit "blooming" because of E.H.T. regulation limitations.

Contrast Control

The operation of the contrast control in the screen grid of the 6EB8 can be described as follows. The high loop gain of the a.g.c. amplifier tends to keep the a.g.c. reference level (in this case the back porch as described later) at the a.g.c. amplifier control grid (pin No. 6 of the 6HS8) very close to the cathode potential of the 6HS8. Since R4/(R5 + RV1) is a fixed ratio the back porch potential at the 6EB8 plate remains at a level which bears a constant relationship to the 6HS8 cathode potential. An increase of screen grid potential of the 6EB8 will result in an increase of its plate current. The latter in turn will result in a decrease of plate voltage and a decrease of potential at the a.g.c. amplifier control grid. This will result in a decrease of a.g.c. and an increase in gain of the receiver which means an increase of contrast. Conversely a decrease in screen grid potential of the 6EB8 will result in a decrease of contrast. The ratio of R4/(R5 + RV1) is made variable to allow for spread in component valves and valve characteristics. The control is obviously labelled "Pre-set Contrast".

Ideally the 6EB8 screen grid supply should have zero impedance. In the AWA circuitry mentioned above the screen grid is actually supplied via a cathode follower. In the AWA circuitry this is essential as the black level is derived as a ratio and not derived from the blanking level (back-porch) as in the circuitry described in this report. The effect of a finite screen grid supply impedance in this circuitry is as follows. A low-key scene results in a less negative average grid voltage at the video amplifier control grid and consequently the average screen current will fall and the average screen voltage will rise. As discussed before, an increase in screen potential results in an increase of contrast. Conversely a high-key scene will result in a decreased contrast. The overall effect then is not undesirable, as a low-key scene will have a better distinction between blacks and the various shades of grey of which the low-key picture is mainly composed.

To compensate for the difference between black level and blanking level at the video amplifier plate, R13 is introduced. The difference between black level and blanking level in the transmitted signal is termed "set-up" (see reference[1], Drawing ZC-9, Sheet 1, Issue 3).

Without R13 the black level shifts with contrast control adjustment. This can be explained as follows. With increase of contrast the blanking level remains constant. The "set-up" (measured in volts), however, increases and therefore the black level shifts in the "white" direction. Conversely a decrease in contrast results in a black level shift in the "blacker-than-black" direction.

The effect of R13 is as follows. With increase of contrast the video amplifier screen grid voltage increases. Therefore the cathode voltage of the 6HS8 rises to an extent determined by the voltage divider action of R13, the 6HS8, R8 and VDR3. Since the blanking level at the video amplifier plate bears a constant relationship to the 6HS8 cathode potential (as mentioned before) the blanking level and also the black level shifts in the "blacker-than-black" direction. Conversely, a decrease in contrast results in a shift in the "white" direction. By judicious choice of R13 the black level can be made to remain constant with changes in contrast. The value of R13 is determined empirically.

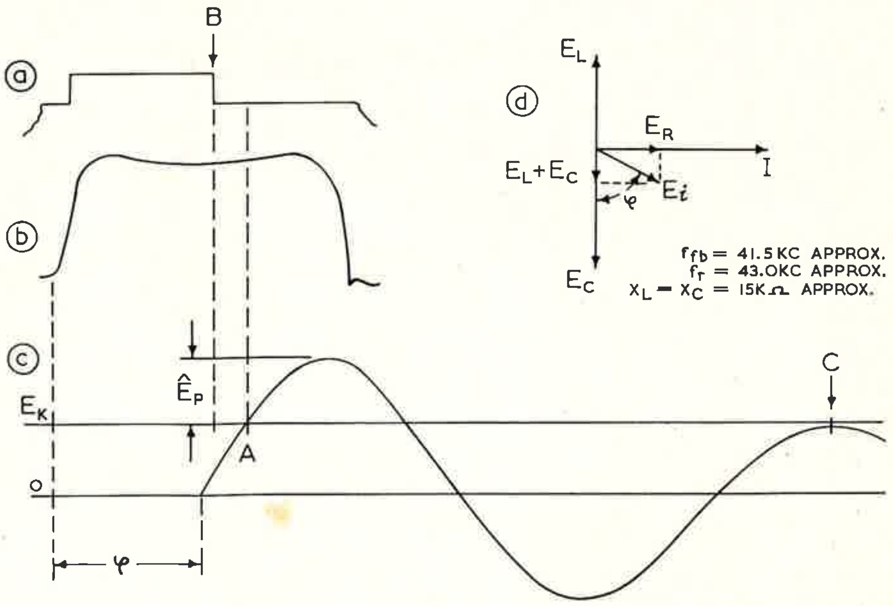


FIG. 3—AGC amplifier waveforms.
(a) Horizontal sync. pulse
(b) Horizontal flyback pulse
(c) Ringing voltage at plate of a.g.c. amplifier

6HS8 Sync. Separator and AGC Amplifier

The values of components surrounding the 6HS8 as presented in Fig. 1 are not optimum. Investigations carried out after the original preparation of the figure have shown that it is of advantage to operate the 6HS8 with as high a screen voltage as permitted by the ratings for maximum screen dissipation and cathode current. However, as far as the black level circuitry is concerned there is no fundamental difference in operation.

AGC Amplifier

If the a.g.c. amplifier is to clamp the back porch, C1 must be charged during the back porch. For this purpose L1, R1 and C2 are inserted between the fly-back pulse source and the plate of the a.g.c. amplifier. They form a series-tuned circuit. At resonance the current in the tuned circuit will be in phase with the applied voltage (the fly-back pulse). The voltage across L1 will then lead the applied voltage by 90° and the voltage across C2 will lag the applied voltage by 90°. By altering the tuning the phase shift (in Fig. 3c) can be conveniently controlled. The size of R1 mainly affects the damping (decay of amplitude between fly-back pulses). When tuned away from resonance the size of R1 also affects the phase angle. This arrangement permits the generation of a positive half-sine wave which occurs during the back porch and extends past the porch as shown in Fig. 3c. It is theoretically possible that a.g.c. is developed by video immediately following the back porch. However, the maximum video level that can occur at the control grid of the a.g.c. amplifier is lower than the back-porch or blanking level. (The difference between the two levels is the "set-up" as described above). In practice the effect on the performance of the circuitry is so small as to be ignored.

The values of L1, R1 and C2 are determined empirically. The following factors must be taken into account, however. To obtain a maximum ratio between the plate voltage of the a.g.c. amplifier (i.e. the voltage across C2, E_c in Fig. 3d) and the input voltage (i.e. the amplitude of the fly-back voltage, E_i in Fig. 3d) the ratio of L1 and C2 should be large. The size of L1 is limited by economy reasons and that of C2 by the fact that it should be large compared with the plate capacitance of the valve. An inductance of 60 mH can be obtained with the MSP 40047 horizontal blocking oscillator transformer. C2 and R1 are then adjusted so that the following conditions hold:

(i) the plate voltage should equal the cathode potential, E_k, at the instant indicated by A, Fig. 3c, which occurs later than the trailing edge of the horizontal synchronising pulse, point B, Fig. 3d.

(ii) the peak of the second positive half sine-wave, at point C should not exceed the cathode potential of the 6HS8 (so that no a.g.c. can be developed at this point). Note that before carrying out any of these adjustments it is essential that the horizontal phasing is correct (i.e. the positioning of the fly-back pulse, Fig. 3b, relative to the horizontal synchronising pulse, Fig. 3a) as shown in Fig. 3a and 3b.

To determine the optimum value of E_p consider the average plate characteristics for the 6HS8 (see valve data book). The curves tend to converge to a straight line at maximum attainable plate current (top right). To get optimum performance (i.e. a high plate current at zero bias) it is of no advantage to take the plate voltage much higher than that necessary to develop say 90% of the maximum attainable plate current. Consider a maximum a.g.c. voltage (a.g.c. voltage for minimum gain of i.f. strip and tuner) of —35V. With a screen grid voltage of 67.5V* and a maximum positive peak plate voltage of between 200V* and 250V* a peak plate current of approximately 2.25 mA can be obtained. For the sake of simplicity it is assumed that a constant plate current of 2 mA flows for 4μ sec. during the back-porch. By definition

$$C \times E = 1 \times t$$

Where $C = 10^{-9}$ farads, $1 = 2 \times 10^{-3}$ amperes, and $t = 4 \times 10^{-6}$ seconds, this yields $E = 8$ volts.

Considering that the voltage across the capacitor, C1 consists of a d.c. component with a non-linear sawtooth superimposed, it may be assumed that the voltage swings from —32V to —40V (—35V, +3, –5). If the capacitor is charged for 4 μsec. it discharges for 64—4 = 60 μsec. During discharge

$$\frac{c}{E} = 1 - E^{-\frac{t}{RC}}$$

Where $e = 32$ volts, $E = 40$ volts, $t = 4 \times 10^{-6}$ seconds and $C = 10^{-9}$ farads, this yields $R = 268 \times 10^3$ ohms. Therefore the resistive load presented by the i.f. and tuner a.g.c. circuitry should not be less than 268k ohms.

FOOTNOTE * Relative to cathode.

BLACK LEVEL STABILISATION WITH MAINS VOLTAGE CHANGES

During the course of investigations it was found that the circuitry was very susceptible to mains voltage changes. A transient as well as a permanent shift in black level was observed with a change in mains voltage. The causes are as follows:

Permanent Black-Level Shift

When the mains voltage changes, the No. 1 grid and cathode potentials of the picture tube follow the change. The E_{c2}, however, since it is partly derived from the stabilised B_{boost} does not vary in the same proportion. Therefore, with an increase in mains voltage the g_2–k voltage is reduced resulting in a shorter grid base and blacks go "blacker-than-black". Conversely, a decrease in mains voltage results in blacks going grey. Three modifications were carried out to offset this effect.

(a) R20 was returned to the vertical oscillator cathode supply instead of the B + supply. The voltage at this point changes in the same direction as the mains voltage but (due to VDR1) at a higher rate. This tends to offset the stabilising effect produced by the B_{boost} on the E_{c2} of the picture tube (R26).

(b) The E_{c1} of the picture tube was fed from the same point as in (a). When the mains voltage changes the E_{c1} changes at a higher rate, thereby tending to offset the change in grid base as mentioned above.

(c) A voltage dependent resistor (VDR3) was incorporated in the cathode circuit of the a.g.c. amplifier. When the mains voltage changes the cathode potential changes in the same direction but at a lower rate. An increase in mains voltage results in a smaller increase in cathode potential in turn resulting in a steeper increase of a.g.c. voltage. Conversely a decrease in mains voltage results in a lesser decrease in a.g.c. In fact, VDR3 tends to stabilise the amplitude of the composite video signal at the output of the video detector. When the mains voltage increases the grid base of the video amplifier increases and since the black level of the composite video signal at the video amplifier control grid is maintained, the plate current corresponding to the black level of the composite video signal is increased with the result that at the picture tube the blacks tend to shift towards grey. Conversely, a decrease in mains voltage results in a shift of black towards "blacker-than-black".

The combination of the three modifications described above gives 100% compensation of permanent shift. By altering the size of R8 and/or the type of VDR3, under or overcompensation can be obtained.

Transient Black-Level Shift

A transient black level shift with a change of mains voltage is experienced because of the relatively large time constant of C8 and the contrast voltage divider consisting of R15, RV2, R16 and the equivalent resistance presented by the screen grid current of V1 (RC = approximately 35 msec. C8 requires approximately 0.1 sec. to "recover").

This effect may be substantially minimised by connecting a capacitor, C34, between B + and the picture tube control grid (brightness) circuit, the junction of R47 and RV3. The values of R47, RV3 and C34 are chosen so as to produce maximum reduction of the transient effect.

Conclusion

The requirements for a "perfect" black-level control have been outlined.

Four different systems which provide varying degrees of black-level control have been described and their relative merits discussed.

A black-level control circuit with back-porch clamping incorporated in an integrated TV receiver circuit has been described including methods by which black-level stabilisation with mains voltage changes has been realised.

References

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3 "A New Approach to the DC-Restoration Problem", by Stephen P. Ronzheimer (Hazeltine Research Inc.), I.R.E. Transactions (U.S.A.) on Broadcast and T.V. Receivers. November, 1962.

4 "An Improved Single Picture-Control for Television Receivers (Automatic Black Level)", Hazeltine Electronics Corporation, Report No. 7034, December 15th, 1947.

5 "Black-Level Performances of T.V. Receivers: Measurement Techniques and Receiver Design Consideration", Hazeltine Technical Report No. 8030, October 13th, 1961.

NEWS & New Releases

Gemini 7 Astronauts To Contact Earth With Light Beam Using Laser Transmitter Developed By RCA For NASA

The National Aeronautics and Space Administration has scheduled a unique test of light beam communications between the Gemini 7 two-man spacecraft and the earth using a laser device built by the Radio Corporation of America, RCA announced recently.

The experiment will mark the first attempt to establish contact with the earth from an orbiting spacecraft using the narrow, intense light beam of a laser, according to Dr. James Vollmer, RCA project manager for the programme.

The laser transmitter, about the size of a home movie camera, was developed by RCA's Applied Research Organization in Camden, N.J., for NASA's Manned Spacecraft Center, Houston, Texas. RCA is building three of the self-contained, hand-held laser devices for NASA under an $88,657 contract awarded in May of this year.

Dr. Vollmer said two transmitters already have been delivered — one to NASA and the other to McDonnell Aircraft Corp., St. Louis, Mo., Gemini prime contractor — and the third will be delivered early in September.

Dr. Vollmer said either Astronaut Frank Borman or James Lovell will aim the laser device at another laser light beamed at their spacecraft from the White Sands Missile Range in New Mexico as Gemini 7 passes in the vicinity of this location during its 14-day flight scheduled for early 1966.

Dr. Vollmer said that when the light beam receiver, mounted on an RCA built AN/FPS-16 tracking radar at White Sands, picks up the pulses of light from the spacecraft, the ground laser beacon will flash indicating to the astronauts that contact has been established.

Dr. Vollmer said the astronaut will aim the six-pound transmitter using an optical telescopic sight. He will attempt to contact the ground with the laser pulsing 100 times a second, and will change the pulse rate to 7,000 pulses per second using a button on the side of the three by six by eight-and-one-half inch unit. Dr. Vollmer explained that this higher pulse rate will be modulated to carry the astronaut's voice over the light beam to the ground.

The heart of the device is a new type of laser, called a gallium arsenide injection diode laser, which operates at room temperature, eliminating the need for cryogenic fluids such as liquid nitrogen in the spacecraft. Four of the lasers are used in the transmitter.

The new laser was developed by RCA's David Sarnoff Research Laboratories, Princeton, N.J.

The first transmitter was developed and delivered to NASA June 15, less than two months after receipt of the contract. It is now undergoing qualification tests at the Manned Spacecraft Center in Houston. Dr. Vollmer said that original specifications required a ten-pound unit emitting eight watts, and the device delivered to NASA by RCA weighs six pounds and transmits 16 watts of light power.

Transistors and nuvistors in a two-meter transceiver

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Dramatic strides in electronics attending the introduction and full growth of the transistor have been accompanied by far-reaching effects in the field of ham equipment, where almost daily radio amateurs are being advised of new, ingenious designs proclaiming the versatility, compactness, and high quality of advanced solid-state gear.

In this article the author offers readers a novel departure from the more conventional, transistorised apparatus — a two-meter transceiver that utilizes both transistors and RCA nuvistors to achieve an effective compromise in all-around economy and operating efficiency.

Constructed more than seven months ago, this unique rig already has been used by the author in hundreds of successful QSO's at ranges up to, and exceeding, 100 miles.

Although a sufficient number of high-frequency transistor types are available to construct an all-transistor transceiver, the high cost of these types makes their use impractical if not prohibitive. An investigation of RCA nuvistors by the author showed these tiny metal-ceramic tubes to be far more economical — even after due consideration of the high-voltage supply they require. As a result, nuvistors were employed in both the receiver and transmitter sections of the unit.

The block diagram and schematic drawing (Figures 2 and 3, respectively) reveal in detail the several features that are incorporated in the design of the transceiver. In addition to this, of course, a versatile power transformer is required to permit operation from either a 12-volt automobile or 117-volt residential source. Transformers meeting such requirements are readily available, and can serve very well in operating either directly from line voltage or from a 12-volt-DC source as part of a DC-to-DC converter. The proper circuitry is automatically chosen by the power plug that is used.

If a reasonably good antenna is employed, the transmitter power level is more than adequate for mobile operation, local net contacts, and field-day work.

Basic Design Concepts

The 144-Mc receiver front-end is a well-proven unit.

The first intermediate frequency selected was 11.7 megacycles. This is sufficiently high to give good front-end image rejection and allow use of commercial transformers. It is also possible to broad-tune this stage for a 4-megacycle bandwidth from 9.7Mc to 13.7Mc.

Because the bandwidth of the first IF stage is broad enough to afford coverage of the entire 2-meter band, the front-end of the receiver is fixed-tuned. As a result, no tracking problems exist. Station selection is accomplished by tuning the second-converter variable-frequency oscillator.

The second intermediate frequency selected was one megacycle. Consequently, when the oscillator is set one megacycle above the 9.7Mc signal (144Mc), it is also possible to convert the signal at 11.7Mc (146Mc) down to one megacycle. In this way, use is made of the customarily rejected image to provide simultaneous reception of 144-146 megacycles and 146-148 megacycles. On most bands, this arrangement would be chaotic. On two meters, however, where most of the stations operate below 146 megacycles, it provides a simple way of tuning the entire band without tracking problems.

The transistor oscillator and buffer are extremely stable in this transceiver; the RCA-2N371 was originally designed and tested especially for stable oscillator operation at even higher frequencies. The 2N384 buffer prevents loading of the oscillator.

The six tuned circuits in the IF stages, together with the 1-Mc crystal filter (X$_5$), provide adequate selectivity. If a high degree of selectivity is not deemed necessary, the builder may omit the crystal unit from the circuit.

A simple, yet important, innovation in the transceiver is the use of the receiver speaker as a microphone in the transmit position. No longer need the amateur be concerned over the fact that he forgot his microphone at home! The transceiver, however, also incorporates a jack for a conventional carbon microphone.

Use of the same nuvistor crystal oscillator for receiving and transmitting eliminates the need for a socket and tube. The proper crystal is chosen by the send-receive switch, S$_2$. Extra crystal sockets and a switch (S$_7$) are provided to allow a choice of three transmitting frequencies.

Initial operation of the transceiver from AC power lines led to the presence of excessive ripple in the 12 volts supplied to the transistors. The additional filtering capacitance required to eliminate the ripple was subsequently provided by the regulator circuit utilizing transistor, Q$_{11}$. This circuit acts as a capacitance multiplier and adds approximately 10,000 microfarads of capacitance to the filter circuit.

The multi-scale-type meter is used both for transmitter tuning and for measuring signal strengths. In the transmit position of S$_3$, the meter can be switched to measure the grid or plate current or the power output from the final amplifier. In the receive position, the meter functions as an "S" meter.

Construction

As in all VHF circuitry, the layout of this transceiver requires proper parts-orientation and short leads. If the general layout shown in Figs. 4 and 5 is followed, no trouble should be encountered in either the transmitter or the receiver units. Special care should be taken to make certain that the parts are located so as not to interfere with the send-receive switch mounted on the main chassis.

A good solder joint between the nuvistor sockets and the brass plate will ensure solid grounding of the nuvistor shell to provide noise-and-oscillation-free operation.

FIGURE 1: W2OKO's two-meter transceiver features large dial with vernier for easy tuning. All operating controls of the unit are located on the front panel.

The author took full advantage of the small size and low-voltage requirements of transistors by mounting all audio circuits — except the power amplifier — on a phenolic board, thus utilizing space under the chassis that otherwise might have been wasted. Detailed in Figure 6, this terminal board was hand drilled and fitted with Alden No. 651T lugs. It can be duplicated by use of pre-drilled boards and lugs made by Vector (85F24EP board and T 9.4 lugs) or any other such manufacturer. To prevent local motorboating or feedback howls. it is recommended that no appreciable deviation in layout be made. For example, audio transformer T_5 — originally mounted on the audio-AVC-ANL terminal — was relocated near the front of the chassis to eliminate a hum caused by magnetic pickup from the power transformer above the chassis. Relocation of T_5 greatly reduced the hum; but the placement of a shortened loop of copper around the outside of the power transformer core virtually eliminated the disturbance through dissipation of the radiated energy.

To minimize the magnetic coupling between the two transformers (T_5 and T_9), an aluminium chassis and cabinet are recommended to all builders of the transceiver.

The 12-volt rectifier and the regulator transistor, Q_{11}, are mounted on a 2-inch-by-3-inch-by-$\frac{1}{8}$-inch aluminium plate that is insulated (electrically) from the main chassis, and is an adequate heat sink for those components.

To dissipate the heat generated in the audio power stage, Q_{10} — especially during its operation as a modulator in the transmit mode — a much larger heat sink is required. This transistor, in which the collector is internally connected to the case, is thermally grounded but electrically insulated from the chassis through the use of mica or anodized aluminium washers between the case and the chassis.

Because the collectors of Q_{12} and Q_{13} operate at circuit ground, these transistors are mounted directly on the main chassis.

Receiver Alignment and Adjustment

Alignment of the receiver section is accomplished by using the "S" meter as the alignment indicator. An up-scale movement of approximately one division (when the AC power is applied to the transceiver) is an indication that the meter circuit is properly balanced. If this movement is more than one division, the value of R_{43} should be changed. The alignment procedure is as follows:

1 — Apply a 1-Mc signal from a signal generator to the base of Q_1 (2N372 mixer stage) and successively tune each of the six IF-transformer windings for a maximum "S"-meter reading. As the tuning progresses, reduce the input signal strength.

2 — Apply an audio frequency to the 1-Mc signal. A good clean tone from the speaker indicates that the audio system is operating properly.

3 — Adjust the tunable receiver oscil-

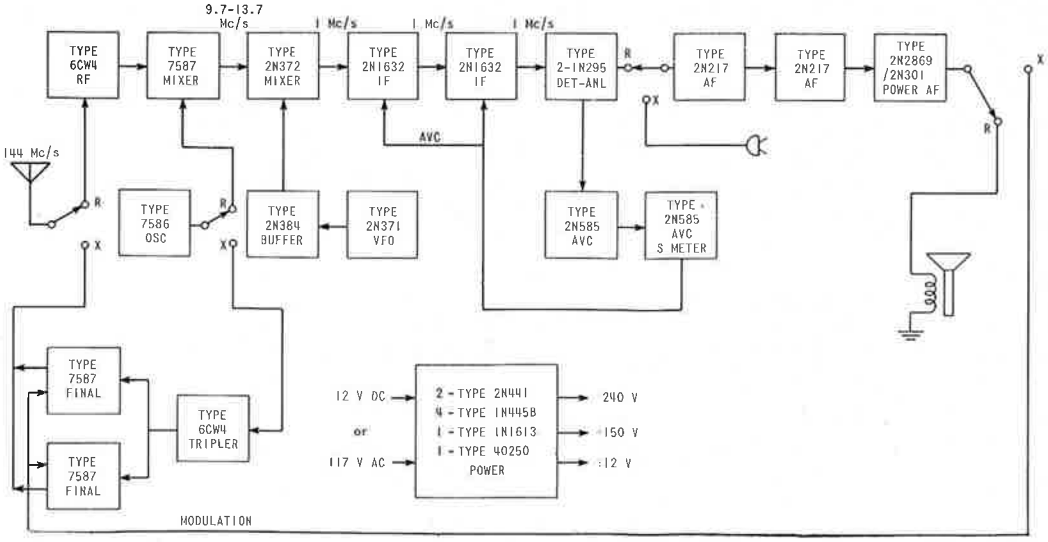


FIGURE 2: Block diagram of W2OKO's two-meter tratnsceiver shows all RCA nuvistor and transistor types employed in the unit, together with their individual circuit assignments.

lator (VFO) as follows:

(a) With trimmer C_{32} set to mid-range and the receiver dial of the transceiver at close to full scale, pick up the oscillator signal on a communications receiver that is tuned to 12.7 Mc.

(b) Adjust the trimmer so that 12.7 Mc appears at about 90 on the transceiver dial.

(c) Search the bottom end of the transceiver dial for a signal of 10.7 megacycles.

(d) C_{32} should be adjusted to centre the oscillator range of 10.7 to 12.7 megacycles across 80% of the tuning dial.

4 — Apply an 11.7-Mc signal to the grid of V_2 (7587 mixer stage) and adjust the top slug of T_1 for a maximum "S"-meter reading.

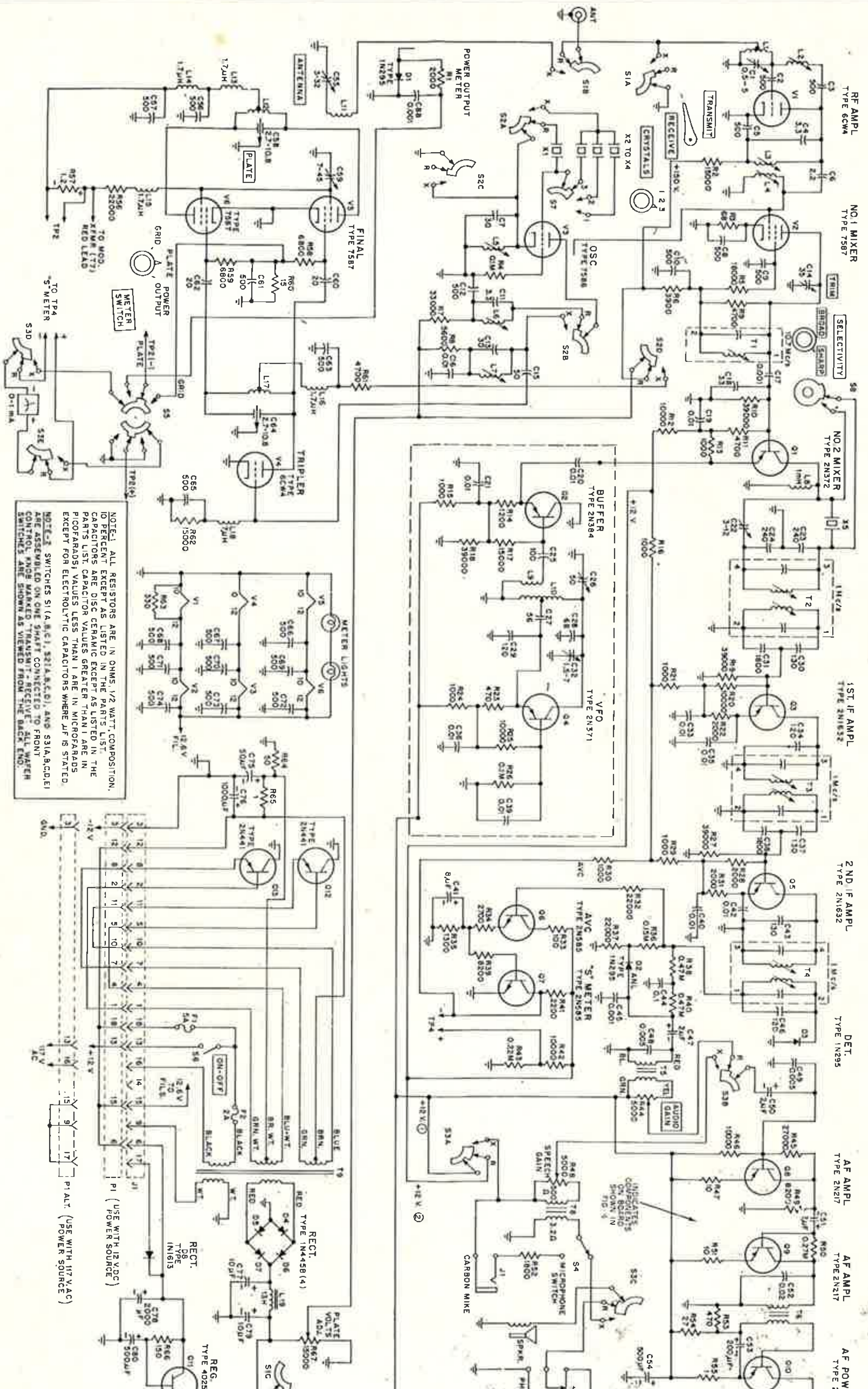
5 — The final step is the tuning of the front-end for 144-to-148 Mc operation. Using a grid dip meter, set L_1, L_3, and L_4 to 146 megacycles; L_5 to 45 megacycles; and L_6 to 134 megacycles. Connect the antenna. If all wiring is correct, 2-meter signals should be heard. If no signals are heard, verify operation of the crystal oscillator by removing the 44.76-Mc crystal from its socket. The background noise should fall off. A slight readjustment of L_5 may be necessary to start the oscillation. L_6 should be peaked for maximum oscillator output. Tune in a signal at approximately 145 megacycles and adjust L_3 for a maximum "S"-meter reading. Repeat with a signal at 147 megacycles and tune L_4 for a maximum "S"-meter reading. Remember that the receiver is simultaneously tuning both 145 and 147 megacycles; be sure the signal frequency corresponds to the coil that is being tuned.

The antenna coil, L_1-C_1, should be tuned to approximately 146 megacycles. The top slug of T_1 may be adjusted slightly for equal reception over the whole band.

Neutralization is easily obtained by adjusting L_2 for minimum feedthrough of a strong signal when the plate voltage of V_1 is zero. This adjustment is not too critical. The receiver alignment is then complete except that touch-up of the tuning might be necessary to provide the whole band with equal sensitivity.

ACKNOWLEDGMENT. The author wishes to thank Harry Thanos, Entertainment Applications, Commercial Receiving Tube and Semiconductor Division, RCA Electronic Components and Devices, Somerville, New Jersey, for his mathematical assistance in transistor applications.

C_1—0.5 to 5.0 pf tubular trimmer
C_4, C_{11}—3.3 pf ceramic tubular Centralab
C_5, C_8, C_9, C_{10}—500 pf silver button
C_6—2.2 pf ceramic tubular
C_{13}—30pf tubular ceramic
C_{14}—35 pf air variable
C_{15}—50 pf tubular ceramic
C_{22}—3 to 12 pf ceramic trimmer
C_{24}, C_{25}—240 pf silver mica
C_{26}—50 pf air tuning capacitor
C_{27}—56 pf silver mica
C_{28}—68 pf silver mica
C_{29}, C_{34}, C_{46}—120 pf silver mica
C_{30}, C_{37},C_{43}—130 pf silver mica
C_{32}—1.5-7.0 pf ceramic trimmer NPO
C_{41}—8 μf electrolytic, 15 volt
C_{44}—0.1 μf paper
C_{47}, C_{50}, C_{51}—2 μf electrolytic 15 volt
C_{53}—200 μf electrolytic, 3 volt
C_{55}—3 to 32 pf air capacitor
C_{58}, C_{64}—2.7 to 10.8 pf butterfly air capacitor
C_{59}—7 to 45 pf ceramic trimmer
C_{60}, C_{62}—20 pf ceramic tubular
C_{75}—50 μf electrolytic, 15 volt
C_{76}—1,000 μf electrolytic, 15 volt
C_{77}, C_{79}—10 μf, 450 volt
C_{78}—2,000 μf electrolytic, 15 volt
C_{80}—500 μf electrolytic, 15 volt
L_1—5 turns, No. 16 bare wire, ¼-inch (spaced wire) diameter, tap 2 turns up from bottom
L_2—12 turns, No. 26 enamelled wire on ¼-inch-diameter slug form
L_3, L_4—4 turns, No. 26 enamelled wire closewound on ¼-inch-diameter slug form
L_5—4 turns, No. 26 enamelled wire closewound on ⅜-inch-diameter phenolic slug form
L_6—5 turns, No. 26 enamelled wire closewound on ¼-inch-diameter slug form
L_7—4 turns, No. 26 enamelled wire, ⅜-inch (spaced wire) diameter phenolic slug form
L_8—1mh
L_9—2 turns, air wound ½-inch diameter from same miniductor as L_{10}
L_{10}—21 turns, air wound, -inch diameter, tap 2 turns up from ground
L_{11}—1 turn, No. 14 bare wire, ¾-inch diameter and "spaghetti" insulated
L_{12}—5 turns, No. 14 bare wire, ½-inch diameter, ⅝-inch long and tapped at centre
L_{13}, L_{14}, L_{15}, L_{16}—1.7 μh choke
L_{17}—4 turns, No. 16 bare wire, ½-inch diameter ⅝-inch long and tapped at centre
L_{18}—7 μh choke
L_{19}—13h 65 ma choke
Meter—0 to 1.0 ma, with 0-5 and 0-10 scales and "S"-meter scale
R_2—Composition, 15,000 ohm, 1 watt
R_7—Composition, 33,000 ohm, 1 watt
R_8—Composition, 5,600 ohm, 1 watt
R_{44}, R_{48}—5,000 ohm, carbon potentiometer
R_{53}—Composition, 470 ohm, 1 watt
R_{54}—Composition, 27,000 ohm, 1 watt
R_{55}—1 ohm, 1 watt
R_{56}—Composition, 22,000 ohm, 1 watt
R_{57}—Composition (5%), 1.2 ohm, ½
R_{63}—Composition, 330 ohm, 1 watt
R_{64}—50 ohm, 10 watt, wirewound
R_{65}—1 ohm, 5 watt
R_{66}—Composition, 150 ohm, 1 watt
R_{67}—15,000 ohm, slide wire, 25 watt
S_1—3 pole, 5 position, non-shorting, ceramic
S_2—5 pole, 3 position, non-shorting, ceramic
S_3—5 pole, 3 position, shorting, ceramic
(Note: S_1, S_2, and S_3 joined on same 6-inch shaft assembly)
S_4, S_6—SPST Toggle
S_5—2 pole, 3 position, miniature
S_7—1 pole, 3 position (Centralab)
S_8—1 pole, 2 position
T_1—10.7-Mc transformer (Miller 1601 or equiv.)
T_2, T_3, T_4—1.5-Mc transformer (tuned to 1.0 Mc by extra capacitors in circuit)
T_5—Audio transformer, 20,000 ohms to 800 ohms
T_6—Audio transformer, 10,000 ohms to 15 ohms
T_7—Audio universal transformer, used in reverse, 8 watt
T_8—Audio transformer, 500 to 3.2 ohms
T_9—Power transformer: primaries 12-volt DC or 117-volt AC; secondaries 280-volt DC at 150 ma and 12.6-volt AC at 3 amps
X_1—Crystal, third overtone 44.766 Mc, style FA5 holder for receivers
X_2, X_3, X_4—Crystals for transmitter, 48.0000 to 49.3333 Mc, third overtone, style FA5 holders
X_5—1-Mc fundamental for filter
Miscellaneous—Sockets for nuvistors and transistors; terminal board prepunched with lugs (see text); chassis 12 inches by 8 inches by 3 inches; cabinet; heat sink for 2N2869; speaker (3½-inch diameter, 3.2 ohm); 18-pin male socket and two 18-pin female plugs



NOTE-1 ALL RESISTORS ARE IN OHMS, 1/2 WATT, COMPOSITION, 10 PERCENT EXCEPT AS LISTED IN THE PARTS LIST. CAPACITORS ARE DISC CERAMIC EXCEPT AS LISTED IN THE PARTS LIST. CAPACITOR VALUES GREATER THAN 1 ARE IN PICOFARADS; VALUES LESS THAN 1 ARE IN MICROFARADS EXCEPT FOR ELECTROLYTIC CAPACITORS WHERE µF IS STATED.

NOTE-2 SWITCHES S1(A,B,C), S2(A,B,C,D), AND S3(A,B,C,D,E) ARE ASSEMBLED ON ONE SHAFT CONNECTED TO FRONT CONTROL KNOB MARKED "TRANSMIT - RECEIVE". ALL WAFER SWITCHES ARE SHOWN AS VIEWED FROM THE BACK END.

FIGURE 3: Schematic diagram and parts list of W2OKO's two-meter transceiver.

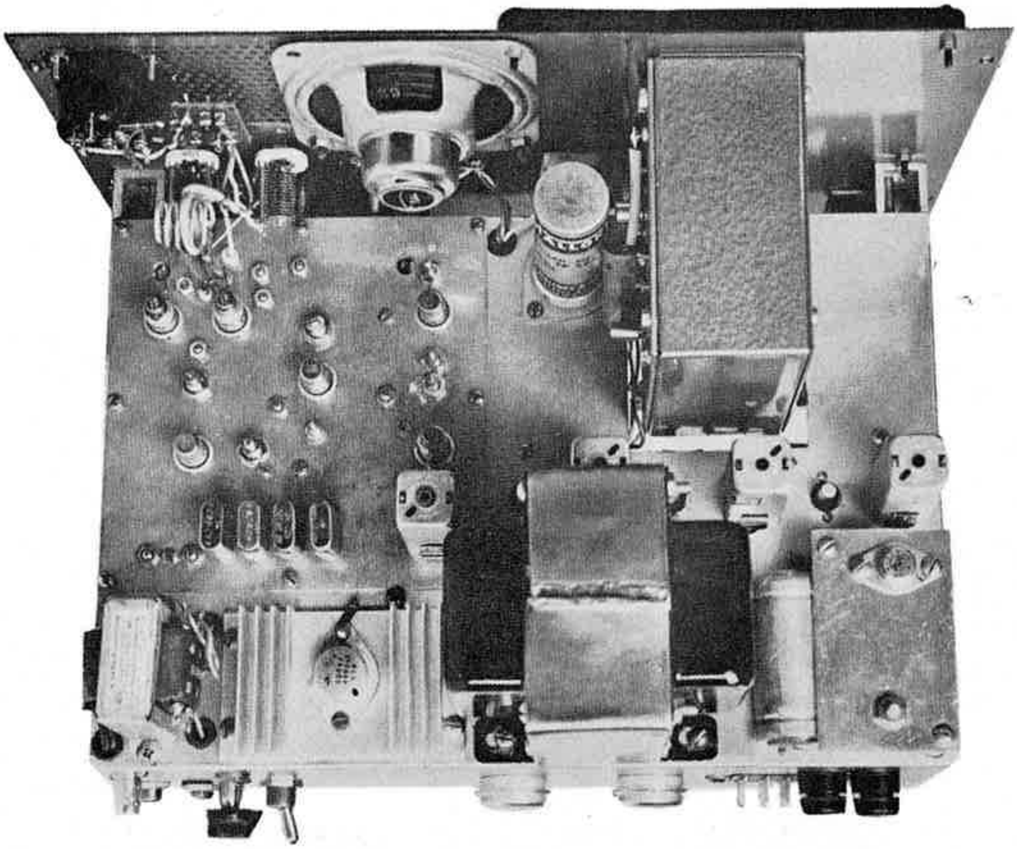


FIGURE 4: Top view of transceiver shows locations of nuvistors, transistors, and other major components. Incorporated on back of chassis are the microphone jack, speech-gain control, microphone switch, earphone jack, antenna connection and the transceiver's power plug and fuses.

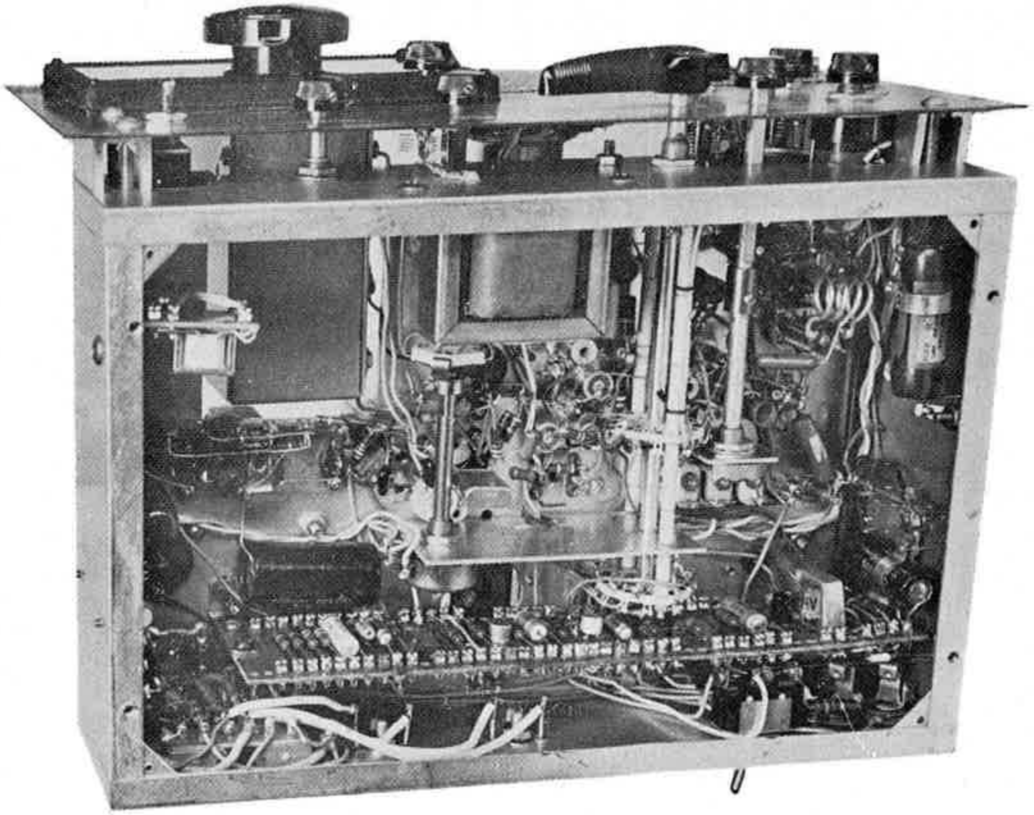


FIGURE 5: Bottom view of transceiver highlights terminal strip and its mounted components. Also visible in photo are the transmit-receive switch, transmitter crystal switch, crystal filter, and the speech-gain control. Note new location of audio transformer (upper right), which was removed from original position to minimize a hum that was caused by magnetic pickup from the power transformer located on the top side of the chassis.

Transmitter Turning and Adjustment

Tuning the transmitter is easily accomplished as follows:

1 — With power off and all nuvistors in place, tune L_7 to 49 megacycles and L_{12} and L17 to 144 megacycles, using the grid dip meter. After this, remove the two 7587 final nuvistors and turn on the power.

2 —Throw the send-receive switch to the transmit position. Connect a high-impedance voltmeter across R_{62} and adjust L_7 for a maximum reading (approximately 10 volts) on the voltmeter for indication that V_3 is oscillating. To insure positive starting of the oscillator, back the slug out to give a slightly higher tuned frequency.

3 — After turning off the main power, disconnect the 240 volts from the plates and screen grids of the final amplifier. Plug in the 7587 final nuvistors, turn on the power, and set the meter switch to TP_1 (final grid current). Tune C_{64} for a maximum meter reading (usually between one and two milli-amperes). Rotation of the plate-tuning capacitor, C_{58}, through its entire range should have very little effect on the grid current. Minimize this effect by adjusting C_{59}, the screen-grid bypass capacitor.

4 — Again, disconnect the AC power and reconnect the 240 volts to the final amplifier plates and screen grids. Switch the meter to TP_2 (plate current) and attach the antenna or a dummy load to the transceiver. Turn on the power. After the nuvistors warm up, tune C_{58} (in the final-amplifier plate circuit) for a dip.

5 — Turn the meter switch to TP_3 (power output) and adjust C_{58} and C_{55} for a maximum power-output reading. The capacitor, C_{55}, tunes out feed-line reactance.

Modulation can be introduced through either an external carbon mike or the built-in speaker. Adjust the gain control on the rear of the chassis for 100% modulation with no distortion.

The receiver is now completely tuned. When transmitter frequencies are changed, only the final-amplifier capacitors on the front panel (C_{55} and C_{58}) need be readjusted.

If the DC-to-DC converter and power plug have been wired correctly, the unit can also be portably operated from any negative grounded 12-volt DC supply.

The transceiver will perform very well for long periods with little maintenance and will provide many hours of pleasurable operation.

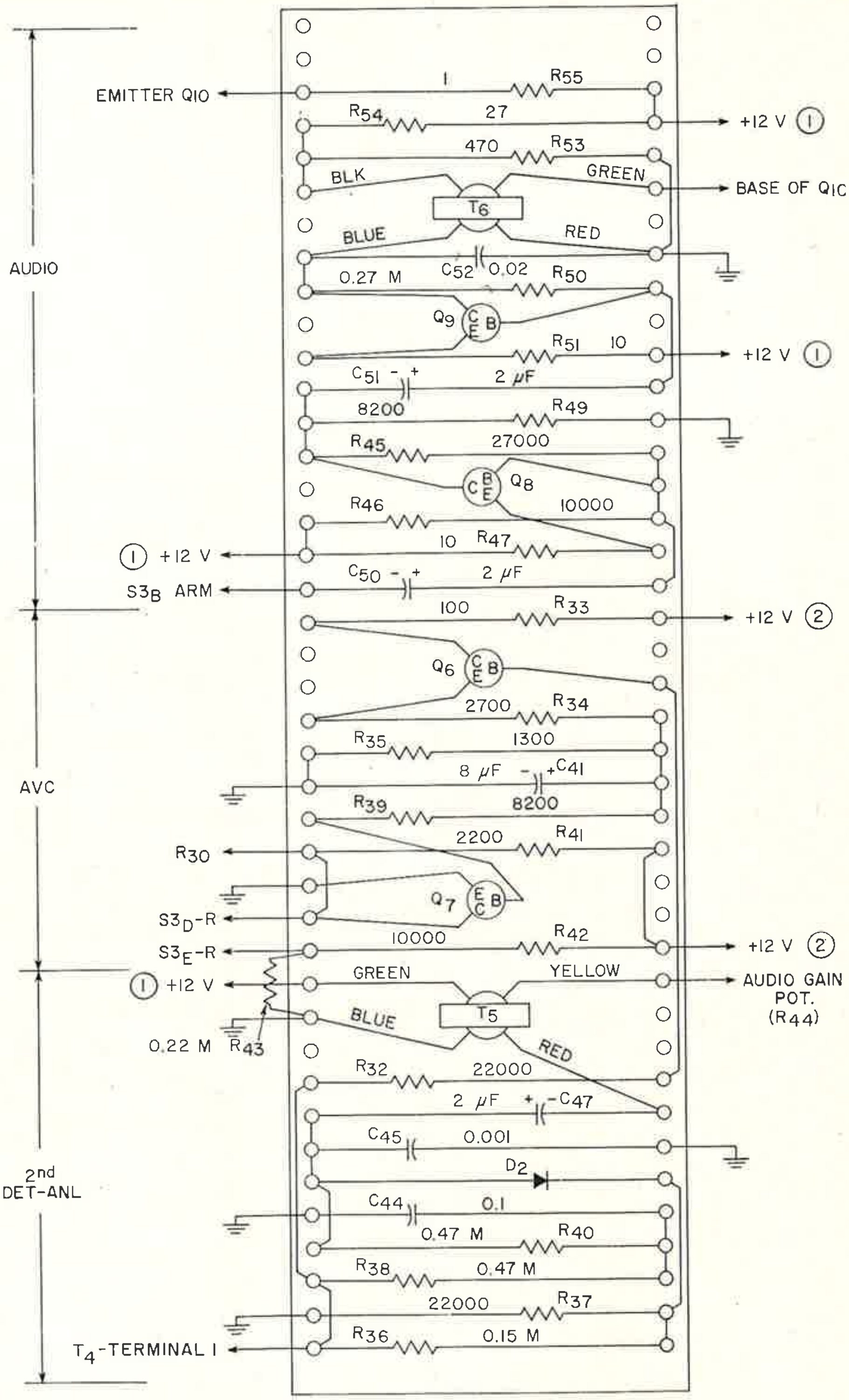


FIGURE 6: Detail on transceiver's Audio-AVC-ANL terminal board, including pertinent footnotes.

NOTES

(1) If hum is objectionable, T5 may have to be moved.

(2) All + 12V (1) lines may be joined and connected to arm of $S3_A$.

(3) All + 12V (2) lines may be joined and connected to $S3_A$ — receive.

(4) All grounds may be joined to form a common ground.

(5) Detector D3 (Type 1N295) and C46 (120 pf) are mounted at T4 (2nd IF).

SUPER RADIOTRON
23GSP4
PICTURE
TUBE

The 23GSP4 is a directly viewed glass picture tube having an aluminised screen 19$\frac{1}{4}$″ x 15$\frac{1}{8}$″ with a minimum projected area of 282 square inches. It employs 110° magnetic deflection and low voltage electrostatic focus. Integral implosion protection is provided by a formed rim band and tension band around the periphery of the tube panel. Mounting lugs have also been included in the tube design, thus eliminating the need for complicated mounting and implosion protection equipment in the receiver.

GENERAL

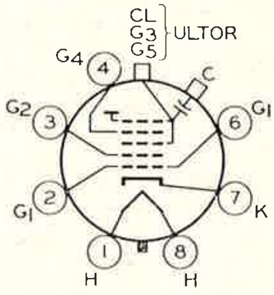
Heater Voltage	6.3 volts
Heater Current	0.6 amp
Direct Interelectrode Capacitances:	
Cathode to all other electrodes	5 pf
Grid 1 to all other electrodes	6 pf
External conductive coating to anode:	
Maximum	2500 pf
Minimum	1700 pf
Faceplate	Filterglass
Light Transmission	42%
Phosphor	Aluminised P4 Sulphide
Fluorescence	White
Phosphorescence	White
Focusing Method	Electrostatic
Deflection Method	Magnetic
Deflection Angles (approx.):	
Diagonal	110°
Horizontal	99°
Vertical	82°

Tube Dimensions:			
Overall Length	14.875	±0.281	inches
Greatest Width	20.650	±0.125	inches
Greatest Height	16.650	±0.125	inches
Diagonal	23.625	±0.125	inches
Neck Length	5.125	±0.125	inches

Screen Dimensions (min.):	
Horizontal	19.250 inches
Vertical	15.125 inches
Diagonal	22.312 inches
Area	282 sq. in.
Electron Gun	Unipotential
Bulb	J187-K1
Bulb Contact	JEDEC J1-21
Base	JEDEC B7-208

SOCKET CONNECTIONS

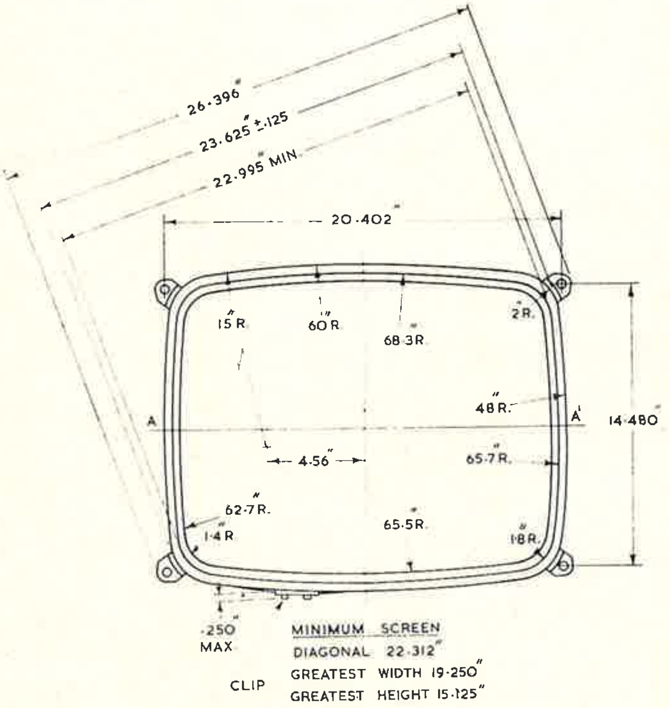
Pin 1—Heater
Pin 2—Grid No. 1
Pin 3—Grid No. 2
Pin 4—Grid No. 4
Pin 5—Blank
Pin 6—Grid No. 1
Pin 7—Cathode
Pin 8—Heater
Bulb Contact—Anode

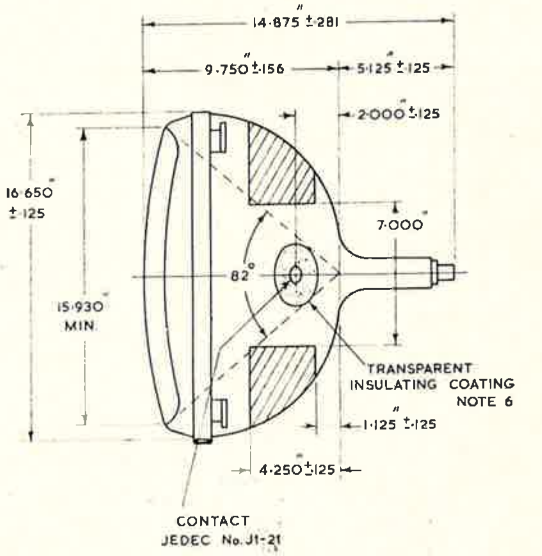


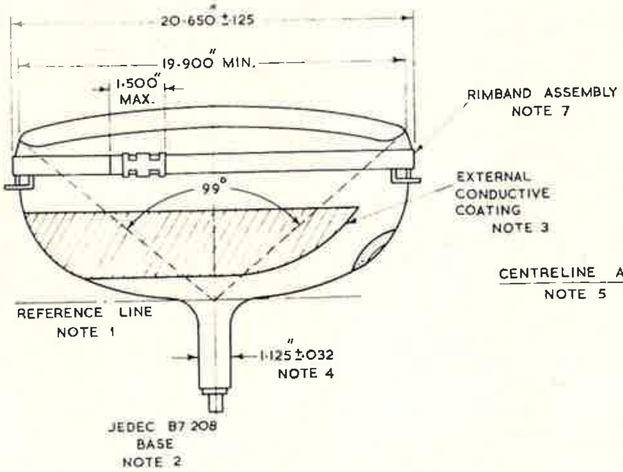
RATINGS, DESIGN MAXIMUM SYSTEM

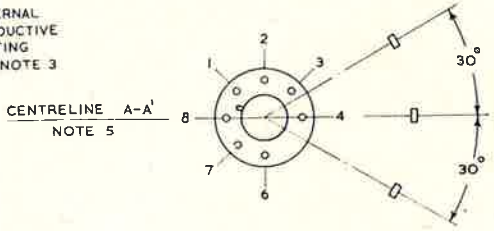
(Unless otherwise specified, voltage values are positive, and measured with respect to cathode)

Maximum Anode Voltage	22,000 volts
Minimum Anode Voltage	12,000 volts
Maximum Grid No. 4 Voltage	+1100, —550 volts
Maximum Grid No. 2 Voltage	550 volts
Minimum Grid No. 2 Voltage	200 volts
Grid No. 1 Voltage:	
Maximum Negative Value	—154 volts
Maximum Negative Peak Value	—220 volts
Maximum Positive Value	0 volts
Maximum Positive Peak Value	2 volts









Maximum Heater-Cathode Voltage, Heater Negative with respect to Cathode:	
During Warm-up, 15 secs	450 volts
After Warm-up Period	200 volts
Maximum Heater-Cathode Voltage, Heater Positive with respect to Cathode:	200 volts

TYPICAL OPERATION, GRID DRIVE SERVICE

Unless otherwise specified, all voltage values are positive with respect to cathode)

Anode Voltage	18,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Grid No. 1 Voltage	—36 to —94 volts dc

TYPICAL OPERATION, CATHODE DRIVE SERVICE

(Unless otherwise specified, all voltage values are positive with respect to Grid No. 1)

Anode Voltage	18,000 volts dc
Grid No. 4 Voltage*	0-400 volts dc
Grid No. 2 Voltage	400 volts dc
Cathode Voltage	36 to 78 volts dc

MAXIMUM CIRCUIT VALUE

Grid No. 1 Circuit Resistance	1.5 megohms

* The grid No. 4 (or grid No. 4 to grid No. 1) voltage required for optimum focus of any individual tube will be a value between 0 and 400 volts independent of anode current. It will remain essentially constant for values of anode (or anode to grid No. 1) voltage and grid No. 2 (or grid No. 2 to grid No. 1) voltage within the ranges shown for these items.

NOTES

NOTE 1. Yoke Reference Line is determined by plane surface of flared end of JEDEC Reference Line Gauge No. 126 when seated on funnel of tube. With minimum neck length tube, the PM centring magnet should extend no more than $2\frac{1}{4}''$ from Yoke Reference Line.

NOTE 2. Lateral strains on the base pins must be avoided. The socket should have flexible leads permitting movement. The perimeter of the base wafer will be inside a $1\frac{3}{4}''$ diameter circle concentric with the tube axis.

NOTE 3. External conductive coating forms supplementary filter capacitor and must be grounded

NOTE 4. Neck diameter may be a maximum of 1.168″ at the splice.

NOTE 5. Base pin No. 4 aligns with centreline A-A′ within 30° and is on the same side as anode contact J1-21.

NOTE 6. To clean this area, wipe only with a soft, dry lintless cloth.

NOTE 7. The Rimband assembly must be grounded.

PROTECTION OF TRANSISTORS IN CLASS B AUDIO OUTPUT STAGES

Failure of transistor class B output stages can occur when adverse conditions cause the output transistors to operate beyond their intended limits. This note describes a practical protection circuit which limits current flow from the power supply to the output transistors and thus protects the class B stage against the damaging effects of voltage surges, shorts, or other adverse conditions.

DESTRUCTIVE CONDITIONS AND PROTECTION REQUIREMENTS

In class B output stages, transistor failure is usually initiated by one or more of the following conditions:

(1) exceeding the collector-to-emitter breakdown voltage,
(2) exceeding the transient breakdown capabilities of the transistor,
(3) thermal runaway,
(4) shorted load, or load impedance substantially below normal,
(5) wiring errors or accidental shorts in the output circuit.

In each of these cases, transistor failure is caused by excessive current flow from the power supply to the output transistor. Destruction of the transistor may occur in a few microseconds or less after the failure mode is initiated.

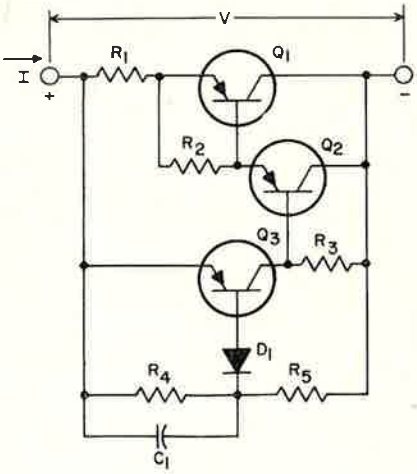
Although simple circuits can be devised to limit drive signal, or to sense load impedance for protection against a shorted load, such circuits do not provide protection against all the causes listed above. Because the destruction is caused by excessive current in all cases, the ideal protection circuit should limit maximum transistor current and disconnect the power supply whenever excessive current tends to flow. To be reliable, the circuit should react more quickly than the fastest rise time of current expected. Current limiting should begin within a few microseconds after a failure mode is initiated, and the power supply should be disconnected within 100 micro-seconds after limiting occurs.

PRACTICAL PROTECTION CIRCUIT

A simple two-terminal electronic switch which meets the requirements described above is shown in Fig 1.

by
M. S. FISHER
Radio Corporation of America
Electronic Components and Devices
Somerville, N.J.

This "electronic circuit breaker", which is placed in series with the power supply, is designed for use at power-supply voltages up to 30 volts and has a tripping current level of 3 amperes.



C_1 = 0.008 μf
D_1 = 1N3754
Q_1 = 40022
Q_2, Q_3 = 2N2614
R_1 = 0.5 ohm, 5 watts
R_2 = 100 ohms, 0.5 watt
R_3 = 300 ohms, 5 watts
R_4 = 680 ohms, 0.5 watt, 5%
R_5 = 3300 ohms, 0.5 watt, 5%

FIG. 1.—Schematic diagram for electronic circuit breaker.

The dc voltage-current characteristic for this protection circuit is shown in Fig. 2. In the region from A to B, operating conditions are normal and the breaker acts as a low-value resistance. In this mode, transistors Q_1 and Q_2 are held in saturation by the bias voltage developed across R_3. Transistor Q_3 is off (except for some small leakage current in the collector circuit) because the voltage across R_4 is not sufficient to overcome the offset voltage of Q_3 and the diode D_1.

When the current (I) in Fig. 1 reaches the predetermined level shown at point B in Fig. 2, the voltage drop (V) across the circuit produces sufficient voltage across R_4 to turn on Q_3. The drop across R_3 then increases, reducing the bias on Q_1 and Q_2 so that they start to cut off. As the voltage V increases to the value shown at point C in Fig. 2, Q_3 is biased into saturation and Q_1 and Q_2 are cut off completely. When the voltage V is between points C and D, the circuit breaker then acts as a high-value resistance. The current flow in this region consists of the collector current of Q_3 flowing through R_3, plus the bias current through R_5. The maximum voltage limit at point D is determined by the breakdown voltage of Q_1 and Q_2; for the transistors shown it is 30 volts.

The circuit breaker remains in its high-impedance state as long as the voltage across it is higher than the value shown at point C. The voltage across the breaker in this region (C to D) is determined by the power-supply voltage and the total loop resistance (breaker resistance plus external series resistance). For

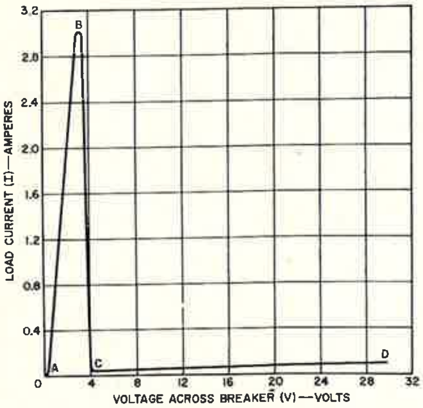


FIG. 2.—Voltage-current characteristic for breaker shown in Fig. 1.

[redacted]

a given breaker resistance and power-supply voltage the external resistance must be low enough to maintain the voltage across the breaker above the value shown at point C so that the breaker will not reset until the power supply is turned off. For the circuit shown in Fig. 1 and a 30-volt supply, a series resistance of 1200 ohms or less is recommended.

To prevent wide variation of the tripping-current level with variations in transistor beta (common-emitter forward current transfer ratio), it is desirable that the drop across R_1 be large compared with that across Q_1. Transistors Q_1 and Q_2 should be held in saturation to minimize their voltage drop and power dissipation in the normal operating region (A to B in Fig. 2). The use of the high-beta Darlington configuration for Q_1 and Q_2 makes it possible to use a high value of resistance for R_3 and still hold Q_1 and Q_2 in saturation. As a result, both the power requirement for R_3 and the current and power requirements for Q_3 are reduced. The tripping-current level for the circuit of Fig. 1 at a temperature of 25° C. varied for 3 to 3.2 amperes for minimum and maximum values of beta for the transistor types shown.

Temperature variations also affect the tripping-current level, as shown in Fig. 3, primarily because the bias voltage required to turn on Q_3 varies with temperature. The silicon diode D_1 is included in the circuit to reduce the effect of this temperature dependence. High temperatures also increase the collector leakage current flowing from Q_3 through R_3, and produce an added fixed voltage drop across the circuit breaker. This added drop reduces the tripping-current level and also increases the power dissipation in Q_1 and Q_2. For the circuit shown in Fig. 1, this added voltage drop is less than one volt at 55° C. Some degree of temperature dependence can be used to advantage in protection circuits because it is generally desirable for current limiting to start at lower power (lower currents) when class B stages are operated at high ambient temperatures.

[redacted]

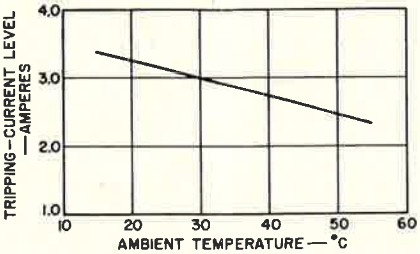


FIG. 3.—Tripping-current level as a function of ambient temperature.

The switching speed of the circuit breaker is determined primarily by the speed of the semi-conductor devices. If the current I in Fig. 1 increases very rapidly, the full voltage across R_4 is applied to the emitter-to-base junction of Q_3 because of the capacitance in diode D_1. As a result, the breaker starts to limit at a lower current because Q_3 can be biased on before the forward drop of D_1 is overcome. The addition of capacitor C_1 across R_4 slows down the voltage increase across R_4 when the current I has a high rate of change, and helps to control the triggering point (within limits) for fast-rise-time waveshapes of current.

Fig. 4 shows the current-limiting and turn-off characteristics for the circuit of Fig. 1 when a transient current pulse which would reach 3.5 amperes in about 3 microseconds is applied. (This waveshape is typical of the square-wave rise time in a high-quality, high-power amplifier.) Current limiting (point B) occurs at about 2.8 amperes in the circuit of Fig. 1, and the current begins to decrease within a few microseconds after limiting occurs. The current then drops from 90 to 10 per cent. of its peak value in about 80 microseconds. The most significant features of this performance are the current-limiting level and the time required for the current to begin to drop. The relatively long turn-off time is usually not a problem because the current is safely below its normal maximum. In the event that faster turn-off time is required, a high-frequency transistor such as the AWV-2N2148 can be used for Q_1.

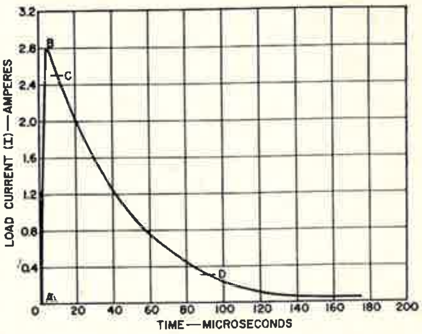


FIG. 4.—Current-limiting and turn-off characteristics of breaker circuit.

[redacted]

Limiting at lower current levels for fast-rise-time pulses is especially desirable for protection against breakdown modes (modes in which current increases very rapidly and destruction can occur in a very short period of time). An example of a fast-rise-time, rapid destruction process is "second breakdown". An electronic circuit breaker may or may not prevent transistor failure after second breakdown starts, depending on (1) the triggering level for fast-rise-time waveshapes, (2) the speed of the breaker, and (3) the type of transistor entering breakdown and the conditions which initiated the breakdown. A breaker which would not provide protection after second breakdown starts can still protect circuits by preventing conditions which would start second breakdown.

APPLICATION IN TYPICAL CLASS B OUTPUT STAGES

The method used to incorporate the circuit breaker of Fig. 1 into a class B output stage depends on the degree of protection desired. In turn, the protection desired depends on what type of destructive conditions may occur (these conditions may be partly a function of the type of output transistor used).

The least expensive type of application is the use of one circuit breaker in the single common power supply of a stereo amplifier using capacitive-coupled loudspeakers, as shown in Fig. 5. In this arrangement, however, the breaker has no control over the energy stored in the speaker coupling capacitors C_1 and C_2, and all of this energy could be transferred to transistors Q_2 and Q_4 under abnormal conditions. Failure of Q_2 and Q_4 could result unless the transient breakdown capabilities of the transistors were sufficient to enable them to absorb this energy.

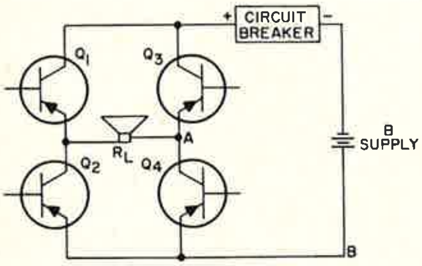


FIG. 5.—Use of circuit breaker in stereo amplifier using capacitive-coupled loudspeakers.

Another disadvantage of the arrangement shown in Fig. 5 is that the triggering level must be set above the sum of the peak currents of the two channels (because the circuit breaker shown in Fig. 1 is a peak current detector). When only one channel is drawing current, therefore, the breaker will not disconnect the power supply until the current level is equal to twice the normal peak current. As a result, the transistors used would have to be capable of reliable operation at twice normal peak currents under adverse conditions. Any type of output stage in which one circuit breaker is used to protect both channels would have this same disadvantage.

The breaker circuit can also be used in a full-bridge type of output stage, as shown in Fig. 6. If point B of this circuit is grounded, one circuit breaker can be used for both channels; however, this arrangement requires "floating" loudspeakers that create problems in the application of feedback. If the loudspeaker is grounded at point A, the power supply "floats"; two power supplies (and two circuit breakers) are then required for the separate output stages of a stereo amplifier.

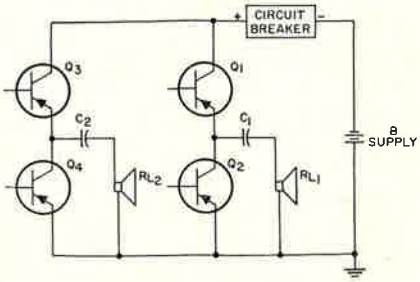


FIG. 6.—Single-channel full-bridge output stage using circuit-breaker protection.

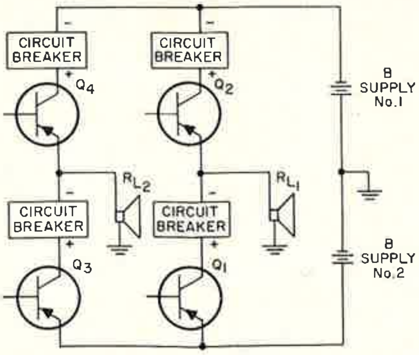


FIG. 7.—Stereo, two-supply, two-transistor-per-channel, half-bridge output stage using electronic circuit-breaker protection.

The two-supply, half-bridge type of output stage is commonly used in transistor high-fidelity equipment. This type of circuit can be built with either two or four output transistors per channel.[2] Fig. 7 shows a stereo, two-transistor-per-channel, two-supply, half-bridge-type output stage including electronic circuit-breaker protection. In this application, four circuit breakers are used, one in the collector circuit of each output transistor. If the circuit breakers are used on the emitter side of the output transistors (e.g., in B supply No. 2), stability problems might result when large amounts of loop feedback from the loudspeaker are applied. In applications where this problem does not exist, two circuit breakers could be used, one in each power supply; however, this arrangement would have the same disadvantages discussed above in connection with the use of a common circuit breaker for two channels.

When four output transistors per channel are used in a two-supply, half-bridge-type stage, the arrangement is similar to that shown in Fig. 7 except that each transistor (Q_1, Q_2, Q_3, and Q_4) is replaced by two transistors in series.

Although the two-supply output stage requires twice as many circuit breakers as the single-supply stage, the cost of adding circuit protection is not doubled because the voltage requirements for each breaker are cut in half.

OTHER APPLICATIONS

The protection circuit described can be used in many applications other than class B output stages. For example, it is a valuable tool for the protection of transistors in tests used to determine the limits of reliable operation. It is also useful for protection of many types of circuits and instruments such as current meters. In addition, it can be used as a laboratory tool to protect developmental circuits during analysis for limits of reliability.

With acknowledgment to R.C.A.

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ADDENDUM

In the July issue of this year, two transistor type members were omitted in the article on page 131 titled, "Transistor Power Supply N. 3." They are both A.W.V. 2N408's.