PART 4

CIRCUIT DESCRIPTION

receiver output noise. The 136B–2 has its own aerial input and is tuned to 40 Mc/s, the principle of operation being that spurious noise signals in the 40 Mc/s region of the h.f. spectrum occur simultaneously in the 3 to 30 Mc/s region.

4. A plan view of the KWM-2A removed from its case is shown in fig. 2, and under chassis views in fig. 3, 4 and 5. A component layout diagram of the chassis mounted components (underside view) is shown in fig. 6.

5. Details of the operating controls on the KWM–2A are listed in Table 1. All except three of these controls are located on the front panel (fig. 1), the remaining three (associated with the VOX circuit) are located inside the unit underneath the top cover (fig. 2).

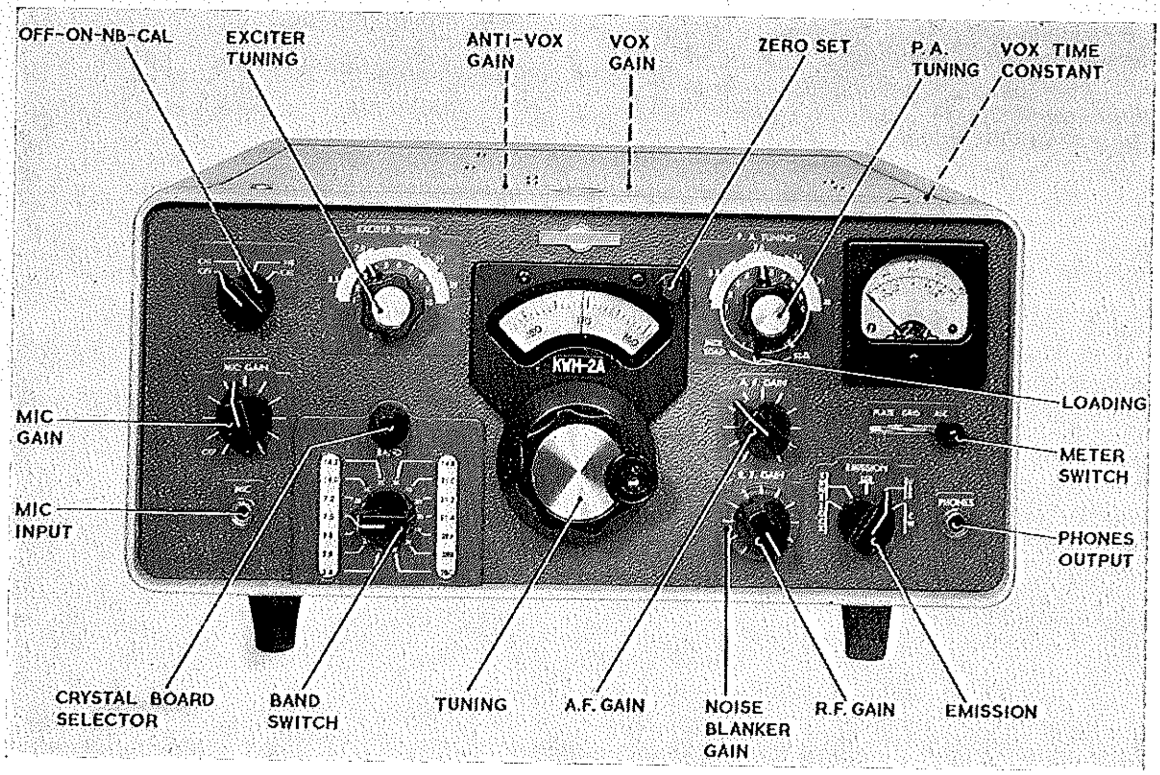

Fig. 1. Transceiver KWM-2A: operating controls

TABLE 1

Transceiver KWM-2A: operating controls

Control	Reference	Function
OFF-ON-NB-CAL	S11	Switches on the a.c. mains supply to the PM-2. In the NB position it brings the 136B–2 into operation, and in the CAL position it switches on the 100 kc/s calibration oscillator.
MIC GAIN	R8 and S10	Controls the gain of the transmitter channel a.f. input stage.
Crystal board selector switch	S15	Selects one of two 14-position crystal boards.
BAND	S2–S8, S13 and S14	Selects one of fourteen 200-kc/s frequency bands.
Tuning	L303	Tunes the variable frequency oscillator 70K–2 over the 200–kc/s frequency band.
Zero set	—	Adjust the position of the hairline on the tuning dial.
EXCITER TUNING	T,3 T4, L10, L13, L14	Tunes the transmitter second mixer, transmitter-receiver r.f. amplifier, drive amplifier input, and power amplifier input.

TABLE 1 (contd.)

Control	Reference	Function
P.A. TUNING	C150	Tunes power amplifier output.
INCR LOAD	C151	Controls the amount of shunt capacitance across the power amplifier output.
R.F. GAIN	R84	Controls the overall gain of the receiver channel.
A.F. GAIN	R92	Controls the gain of the receiver channel a.f. output stage.
Noise blanker gain	—	Controls the gain of the 136B–2 sub-assembly unit.
PLATE-GRID-ALC	S12	Selects the meter function.
EMISSION	S9	Selects the mode of operation, lower or upper sideband or c.w. In the LOCK and TUNE position it maintains the KWM–2A in the transmit function with the 1500 c/s side tone oscillator on.
VOX GAIN	R39	Controls the level at which the audio input switches the KWM–2A to the transmit function.
ANTI-VOX GAIN	R45	Prevents loudpseaker output picked up by the microphone from switching the KWM–2A to the transmit function.
VOX TIME CONSTANT	R43	Controls the transmit to receive release time of the VOX circuit.

BRIEF DESCRIPTION

6. A block diagram of the KWM–2A is shown in fig. 7 from which it will be seen that the unit consists basically of a double-conversion receiver and a double-conversion exciter-transmitter. Certain stages, notably the oscillators, are common to both the receiver and the transmitter channels.

7. A facility referred to as VOX (voice-operated transmission) is incorporated in the KWM–2A whereby when the unit is operational its normal quiescent state is in the receive function. It switches to the transmit function only when the a.f. input from the external microphone or phone patch circuit exceeds a predetermined level. The VOX circuit can, however, be bypassed if necessary by the use of a microphone fitted with a 'press-to-transmit' (p.t.t.) switch. The unit can then be maintained in the transmit function by holding the switch closed.

Transmitter channel

8. The a.f. signal input from the microphones SM–2 or the phone patch circuit on the station control 312B–4 (Part 4, Chap. 4) is first amplified and then applied as one input to a balanced modulator, the second input to which is the output from a crystal-controlled beat frequency oscillator (b.f.o.). This is either 453·65 kc/s for lower sideband operation, or 456·35 kc/s for upper sideband. The balanced modulator output, consisting of upper and lower sidebands but negligible carrier signal, is amplified and fed through a mechanical filter. This filter has a band-width of 2·1 kc/s centred about 455 kc/s, i.e. a pass band of 453·95 to 456·05 kc/s, and therefore passes only one of the two sidebands. The other sideband is rejected and any residual carrier signal is further attenuated. It will be seen, therefore, that the mechanical filter also determines the audio frequency response of the KWM–2A, this being 300 to 2400 c/s.

9. It will be noted that when the function switch on the KWM–2A is set to select upper sideband operation the carrier frequency is 456·35 kc/s and the mechanical filter selects the lower sideband output from the balanced modulator, i.e. an increase in the a.f. signal frequency causes a decrease in the sideband frequency. Similarly, when lower sideband operation is selected the carrier frequency is 453·65 kc/s and the mechanical filter selects the

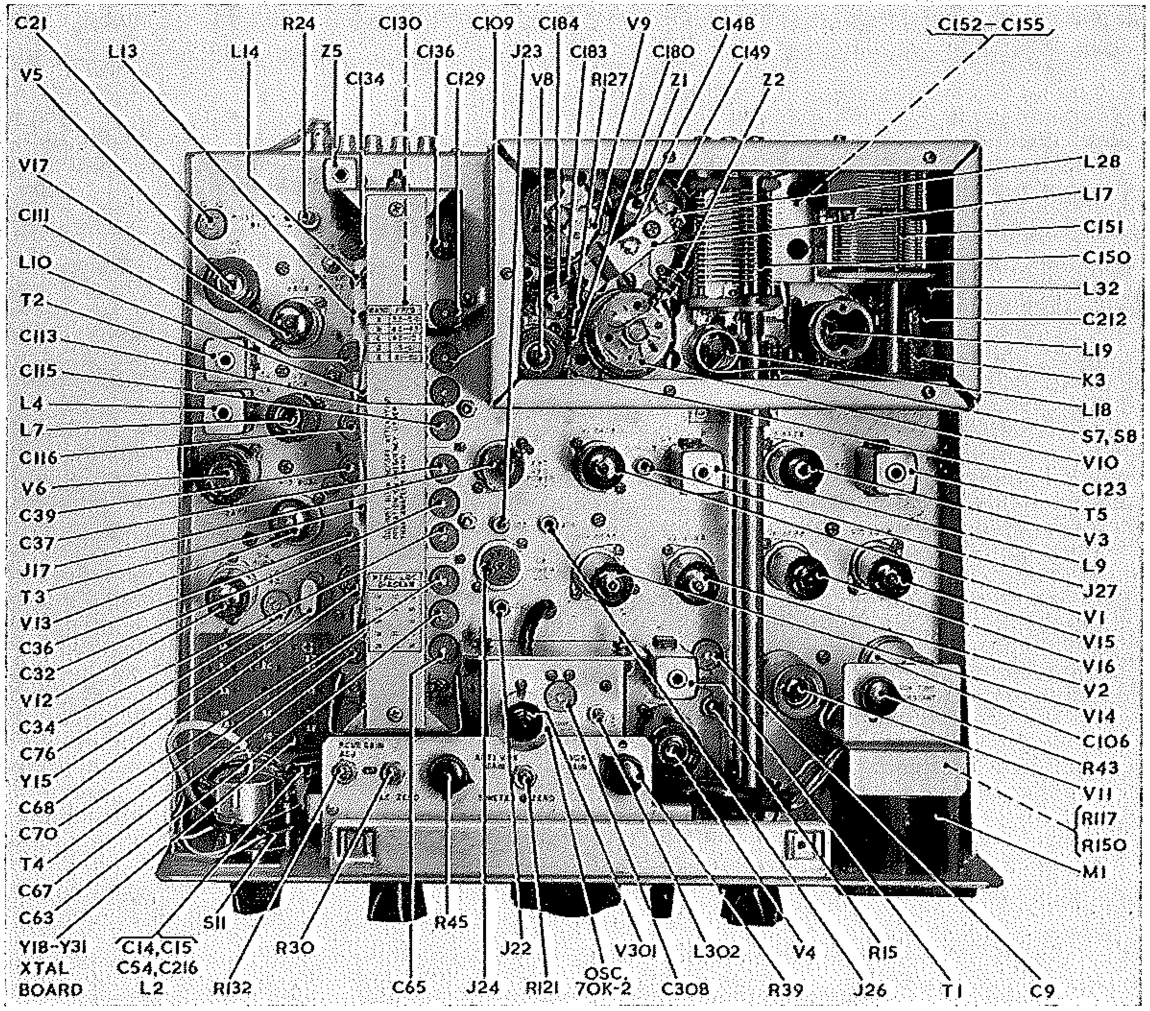

Fig. 2. Transceiver KWM-2A: plan view

upper sideband, i.e. an increase in the a.f. signal frequency causes an increase in the sideband frequency. This apparent anomaly is corrected in the transmitter second mixer when the difference frequency is selected and what was previously a frequency increase appears as a decrease, and vice versa.

10. The mechanical filter output is fed to the transmitter first mixer, the other input to which is the 2·7 to 2·5 Mc/s output from the variable frequency oscillator 70K–2. This is a sub-assembly unit mounted behind the KWM–2A front panel and is tuned over its 200kc/s range by the front panel tuning control. This control is calibrated 0 to 200 in increments of 1 kc/s and the v.f.o. incorporates a switching circuit to ensure that when changing from one sideband to the other its frequency is automatically changed by 2·7 kc/s (the difference between the two operating frequencies of the b.f.o.) to avoid upsetting the dial calibration. The mixer selects the sum frequency and the 3·155 to 2·955 Mc/s output, referred to as the variable i.f. is coupled to the transmitter second mixer via broadband tuned circuits.

11. The other input to the second mixer is the 6·555 to 32·955 Mc/s output from a crystal-controlled master oscillator. This oscillator employs one of a possible total of 28 crystals as selected by the crystal board selector and BAND switches on the front panel (Table 1). The method of selecting a crystal for a particular frequency range is discussed in para. 19. In this case the mixer selects the difference frequency which can therefore be anywhere in the range 3·4 to 30·0 Mc/s.

12. The second mixer output is amplified and coupled via a driver stage to the power amplifier, the output from which is fed via the receive-transmit transfer relay to the antenna or, in a medium power station configuration, to the r.f. linear amplifier 30L–1. The transfer relay is energized by the VOX circuit (para. 7). The a.f. signal input is amplified and rectified and applied to a relay switching circuit, so that when the audio input exceeds a predetermined level the relay is energized and the KWM–2A is in the transmit function. There is a possibility, however, that when the unit is in the receive function the output from the loudspeaker picked up by the microphone may be sufficiently strong to operate the VOX circuit. To obviate this an ANTI-VOX circuit is incorporated whereby the a.f. output from the receiver channel is rectified and used to neutralize the VOX rectifier output.

13. Automatic load control (a.l.c.) is employed in the transmitter channel in order to maintain a high average power output without exceeding the peak power limitation of the power amplifier. An a.l.c. negative feedback signal is produced when the power amplifier valves are driven into excessive grid current. This signal, together with the a.l.c. signal produced by the r.f. linear amplifier 30L–1, is applied as negative bias to the 455 kc/s i.f. amplifier and the output r.f. amplifier.

14. Also included in the transmitter channel is a tone oscillator. This produces a 1500 c/s sinewave signal when the EMISSION switch is set to the LOCK, TUNE or CW position. In the LOCK and TUNE positions of the switch, used during initial setting-up, the KWM–2A is maintained in the transmit function and produces a continuous 1500 c/s upper sideband signal. In the CW position, however, the unit is switched to the transmit function only for the period that the external morse key is depressed and a ground line is applied to remove negative bias on the second microphone amplifier and the transmitter second mixer. Due to the unit transmitting the upper sideband in the c.w. mode, the actual transmitted c.w. signal is 1500 c/s above the tuning dial setting.

Receiver channel

15. The signal input from the antenna, having a frequency in the range 3·4 to 30·0 Mc/s, is connected through the de-energized contacts of the transfer relay to the common receiver--transmitter r.f. amplifier and thence to the receiver first mixer. The other input to the mixer is the 6·555 to 32·955 Mc/s output from the crystal oscillator (para. 11). The mixer selects the difference frequency of 3·155 to 2·955 Mc/s, referred to as the variable i.f., and this is coupled via broadband tuning circuits and the noise blanker 136B–2 (para. 3) to the receiver

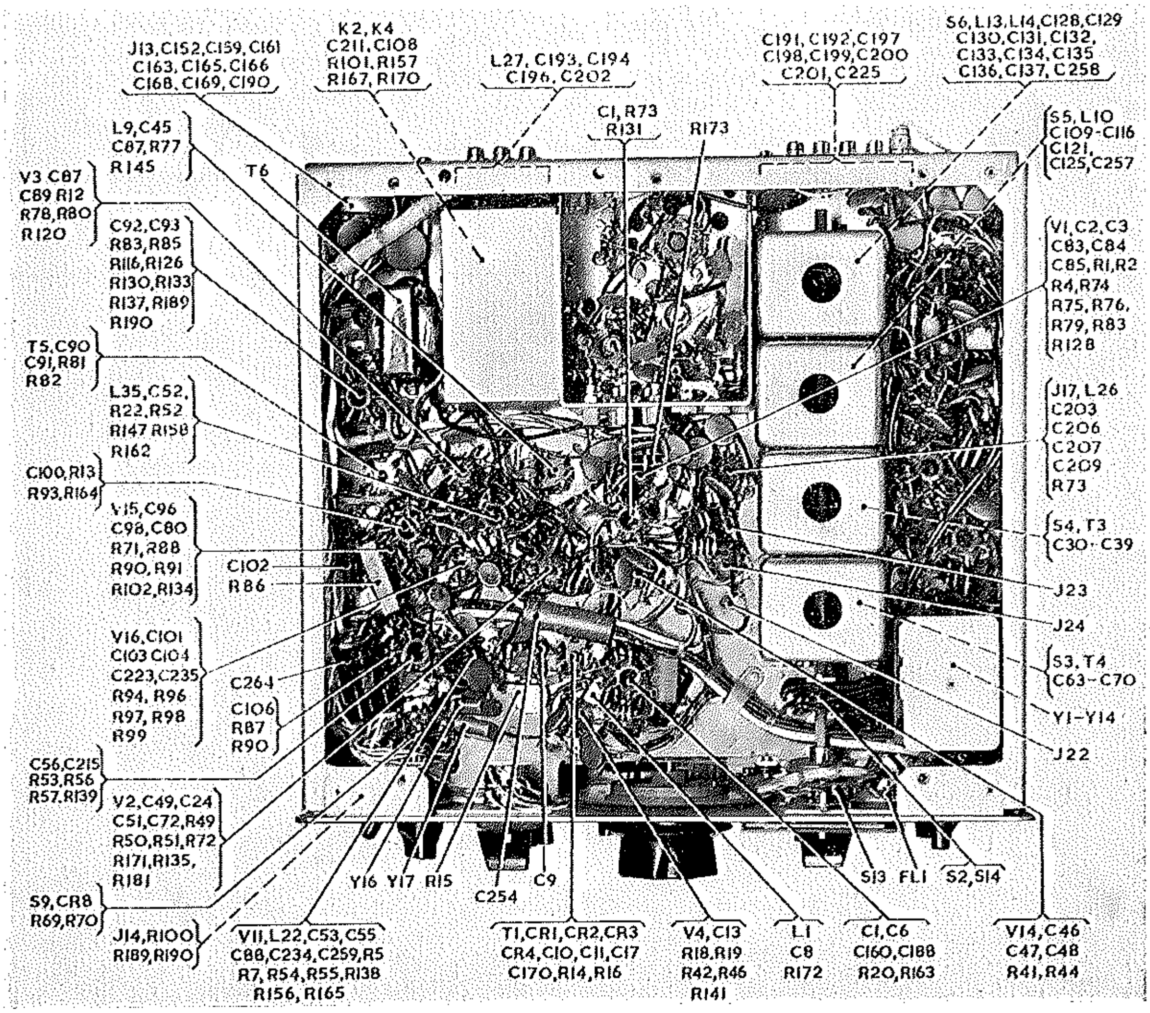

Fig. 3. Transceiver KWM-2A: under chassis view (1)

second mixer. It will be noted that what was previously an upper sideband signal now appears, as a result of the mixing, as a lower sideband signal, and vice versa.

16. The other input to the second mixer is the 2·7 to 2·5 Mc/s output from the v.f.o. (para. 10) and the mixer output selected is the difference frequency of 456·35 kc/s (upper sideband operation) or 453·65 kc/s (lower sideband). It will be noted that a decrease in the frequency of the variable i.f. signal still appears as a decrease in frequency, i.e. what was originally at the antenna an upper side-band signal appears now as an i.f. lower sideband signal, and vice versa.

17. The second mixer output is filtered by the mechanical filter (453·95 to 456·05 kc/s pass band) to obtain the required receiver selectivity of 2·1 kc/s bandwidth at —6dB and after amplification is applied as one input to a product detector. The other input to the product protector is the 456·35 kc/s (upper sideband and c.w. operation) or 453·65 kc/s (lower sideband) output from the b.f.o. (para. 8). The resulting 300 to 2400 c/s output from the product detector is amplified and fed out to the external loudpseaker or headphones and also to the ANTI-VOX circuit (para. 12).

18. The receiver channel employs a conventional automatic volume control (a.v.c.) circuit. The signal input to the product detector is also rectified and applied as a negative feedback signal to both receiver 455 kc/s i.f. amplifiers and the common receiver-transmitter i.f. amplifier. The receiver channel also incorporates a 100 kc/s calibration oscillator. This is switched on when the function switch is set to the CAL position and is used in the initial setting-up of the v.f.o. dial calibration.

Operating frequencies

19. The KWM–2A provides a possible total of twenty-eight 200 kc/s wide frequency bands divided into the five band groups detailed in Table 2. The lower edge of each 200 kc/s band is determined by the frequency of the crystal-controlled master oscillator (para. 11) and the required crystal for this oscillator is selected in the following manner:

(1) If the lower edge of the desired 200 kc/s band is 11·8 Mc/s or less, the crystal frequency is equal to the lower edge frequency plus 3·155 Mc/s.

(2) If the lower edge of the desired 200 kc/s band is 12·0 Mc/s or greater, the crystal frequency is half the sum of the lower edge frequency plus 3·155 Mc/s.

TABLE 2

Transceiver KWM–2A: frequency band groups

Group	Total possible frequency coverage
A	3·4 – 5·0 Mc/s
B	6·5 – 9·5 Mc/s
C	9·5 –15·0 Mc/s
D	15·0 –22·0 Mc/s
E	22·0 –30·0 Mc/s

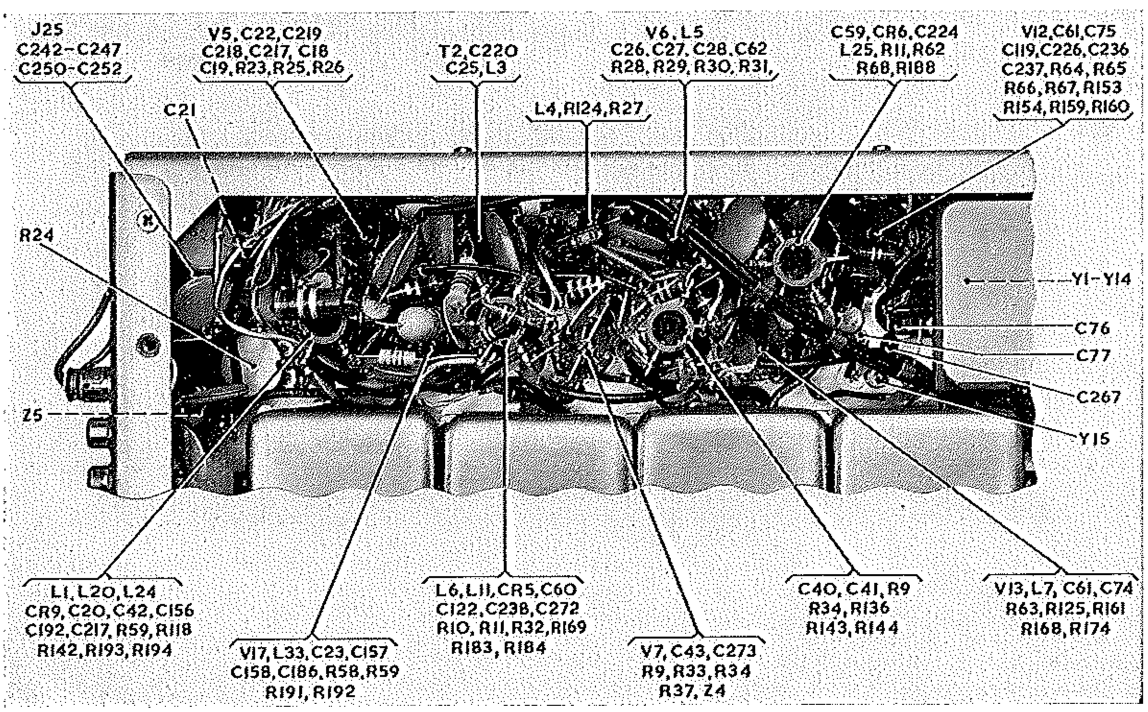

Fig. 4. Transceiver KWM-2A: under chassis view (2)

Note . . .

It is inadvisable to operate the KWM–2A in the frequency range 5·0 to 6·5 Mc/s since in this range the second harmonic of the variable i.f. is almost the same as the desired frequency and this could cause spurious emission in the transmit function.

DETAILED CIRCUIT DESCRIPTION

General

20. A complete circuit diagram of the KWM–2A is shown in fig. 8–10. The KWM–2A is designed to operate in certain equipment configurations that are not discussed in this publication and consequently a number of plugs and sockets are provided that are either not used or have short-circuiting links inserted. Briefly, these are as follows:

(1) J18 is left open-circuit and a dummy plug is inserted in J17 short-circuiting pins 2 and 3 and 6 and 7. These are provided to enable the use of an additional external, variable frequency oscillator 70K–2 when it is required to transmit and receive on different frequencies.

(2) J20 is connected to J21. These are provided for when it is required to employ an external receiver mute switch.

(3) J5 is connected to J6, and J12 is left open-circuit. These are provided for use with an external v.h.f. converter.

(4) J25 is left open-circuit.

21. The switching of the unit between the receive and transmit functions is controlled by the three relays K2, K3 and K4. In the de-energized condition the equipment is receiving, and in the energized condition it is transmitting.

Power supplies

22. All power supplies are obtained from the a.c. power supply PM–2 mounted at the rear of the KWM–2A (para. 2). Plug J13 connects to socket P2 on the PM–2 when the two units are pressed firmly together, and the supplies are switched on by means of S11 (OFF-ON-NB-CAL) on the KWM–2A. The a.c. mains live line is fed in from the PM–2 at pin 7 of J13 and out again at pin 5 via S11 in the ON, NB or CAL position. Pin 6 on J13 is not used with the PM–2, it being provided for external switching purposes when the KWM–2A is used with certain other types of power units (not discussed in this publication).

+800V h.t.

23. The +800V at pin 2 of J13 is used exclusively as the anode supply for the two power amplifier valves, V9 and V10. R.F. decoupling is provided by L17, L28, C148 and C168.

+275V h.t.

24. The incoming +275V at pin 1 of J13 is referred to as the TR+275V supply and is used by those circuits common to both the receiver and transmitter channels. A +200V supply is obtained from TR+275V by means of the voltage dropping resistor R86 and the smoothing capacitor C106b. It is desirable, in order to reduce power consumption and to increase component life, that when the unit is in the receive function the h.t. supply should be removed from the transmitter channel, and vice versa. This is achieved by the use of two h.t. lines referred to R+275V for the receiver circuits and T+275V for the transmitter circuits. TheTR+275V line is therefore fed via R157 to relay contact K4/5. In the receive function, when this relay is de-energized, the R+275V line is at+275V and the T+275V line is open-circuit. In the transmit function, however, T+275V is at +275V and R+275V is open-circuit.

—70V bias

25. The negative bias supply (approximately —70V) at pin 4 of J13 is adjustable over the range —50 to —90V by means of a potentiometer on the PM–2. It is set-up when the equipment is first switched on in order to obtain the correct 'no signal' anode current in the two power amplifier valves.

6·3V a.c.

26. 6·3V a.c. is supplied at pins 8 and 9 of J13 with the ground return lines connected to pins 10 and 11. All valve heaters, the meter illumination lamp DS2 and the tuning dial lamp DS1 are parallel connected between 6·3V and ground. The reason for connecting the valve heaters in the manner shown in fig. 10, and for certain apparently unnecessary decoupling capacitors connected to the ground lines, is that in certain equipment configurations (not discussed in this publication) using a power supply other than the PM–2, the valve heaters are connected in groups across a tapped 24V supply. R.F. isolation for the driver and power amplifier valves, V8–V10, is provided by L29, L30 and L34. The 6·3V outlet at J19 is provided for use in the station control 312B–4 (Part 4, Chap. 4).

Transmitter channel

27. We will consider first the operation of the KWM–2A transmitter channel in the single side-band mode, i.e. the EMISSION switch, S9, in either the LSB or USB position. For the purposes of this description, assume that the transfer relays K2, K3 and K4 are energized.

Microphone amplifier

28. The microphone input at J15 or the phone patch input at J11 (both high impedance) is applied to a 2-stage a.f. amplifier consisting of V1a, V11b and the associated circuit. The cathode circuit of V1a is connected to ground via S9c, and V11b is returned to ground via R5 and S9d. The tone

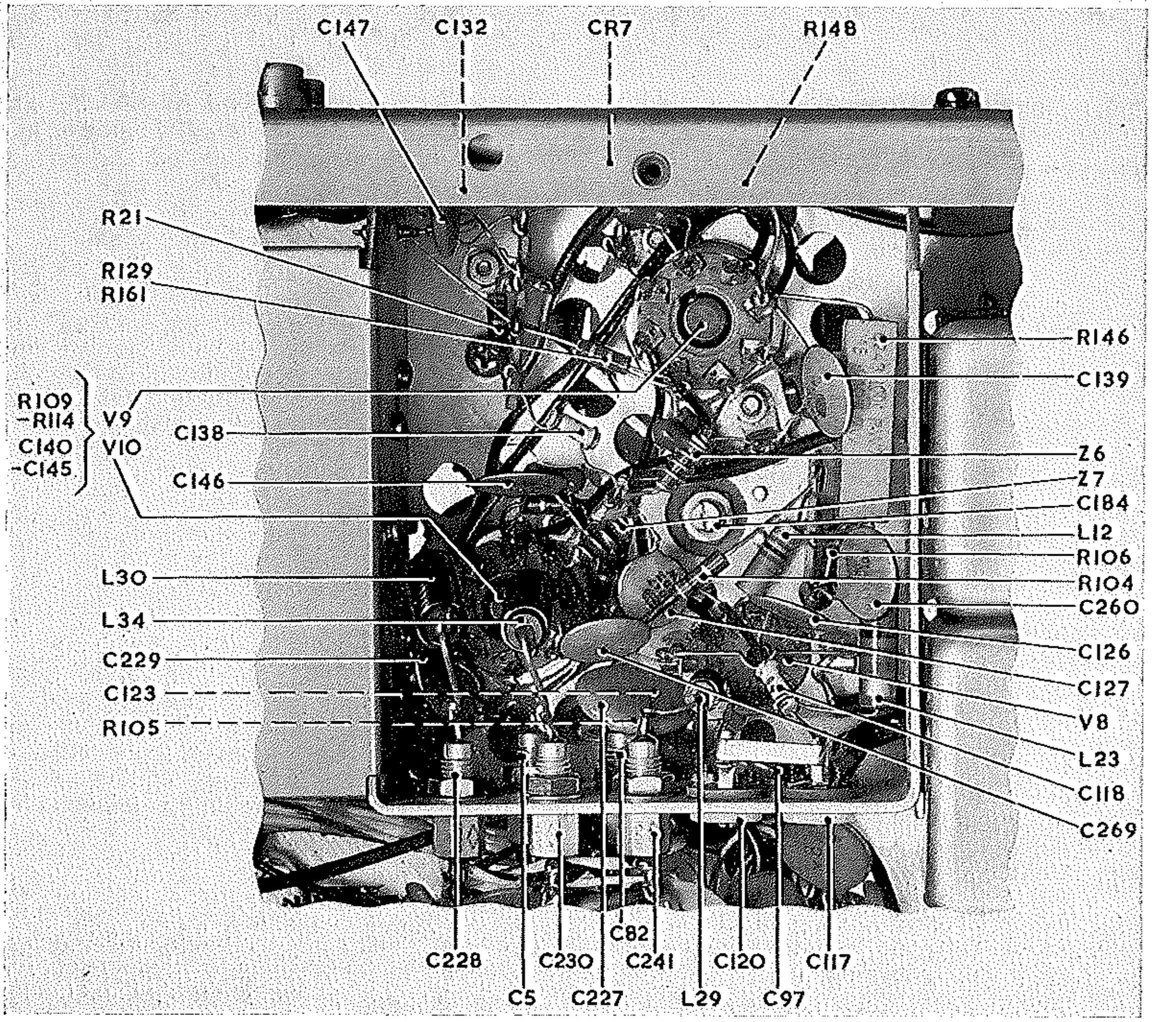

Fig. 5. Transceiver KWM-2A: P.A. grid compartment

oscillator (para. 62) associated with this amplifier is inoperative as V2b cathode is open-circuit in the LSB and USB positions of S9. The microphone amplifier output from V11b anode is coupled via C4 and the MIC GAIN control, R8, to the cathode follower V3a. In the extreme counter-clockwise position of R8, S10 is closed and the input to the VOX amplifier is grounded. A second a.f. output from V11b, via R156, is grounded by S9g when this switch is in the LSB or USB position. When the switch is in the CW position an a.f. output is fed via R155 to the receiver first a.f. amplifier (para. 65).

29. The microphone amplifier uses the +200V h.t. and is therefore still operational when the unit is in the receive function. This is necessary as it is the amplified a.f. input signal applied to the VOX circuit that switches the unit to the transmit function. The h.t. supply for V3a, however, is obtained from the T+275V line via CR10. The diode is included to isolate the T+275V line from C264 in the receive function as this capacitor discharges slowly through R177. The cathode of V3a is connected via R122 to the R+275V line so that in the receive function approximately +5·5V bias is applied to this point. The purpose of this is to prevent clipping of the a.f. input to the VOX amplifier by grid-cathode diode effect in V3a.

Beat frequency oscillator

30. The b.f.o., consisting of V11a and the associated circuit, is a crystal-controlled oscillator operating at one of two possible frequencies dependent on the setting of S9. In the LSB position, Y17 is short-circuited by S9h and the frequency, 453·65 kc/s, is determined by Y16. In all other positions of S9, however, Y16 is short-circuited and the frequency, 456·35 kc/s, is determined by Y17.

31. The circuit is a modified Pierce oscillator. The crystal is effectively coupled between V11a screen and grid operates at a frequency sufficiently below parallel resonance to give an inductive reactance that resonates with C53 and C55. This type of oscillator has the advantage that no circuit adjustments are necessary when switching crystals, and also, by coupling the oscillatory circuit to the screen instead of to the anode, the frequency is unaffected by the anode load impedance.

32. The b.f.o. uses the +200V h.t. so that it is operational in both the transmit and receive functions. The carrier output is developed across L22, tuned by C88, and R56, and the output to the balanced modulator and the receiver channel product detector is taken from the junction of L22 and R56.

Balanced modulator

33. The bridge rectifier CR1–CR4 and its associated components form a balanced modulator circuit to which the unbalanced a.f. signal output from V3a is applied via C6 and L38. Capacitor C7 is part of an r.f. filter and the two r.f. chokes L38 and L39 present negligible impedance at audio frequencies. The b.f.o. carrier output, applied to both sides of the bridge rectifier via C181 and C182, is developed across L38 and L39.

34. On positive half-cycles of the a.f. signal, the bridge conducts only during positive half-cycles of the b.f.o. carrier and the current path is CR3, T1 primary, CR1, L39. On negative half-cycles of the a.f. signal, the bridge conducts only during negative half-cycles of the carrier and the current path is CR4, T1 primary, CR2 and L39. When the a.f. input signal is zero the bridge conducts on both half-cycles of the carrier, but the resultant current through T1 primary during each half-cycle is zero and the carrier is effectively cancelled. The modulator output signal applied to T1 therefore consists of a series of pulses whose repetition frequency is determined by the b.f.o. and whose amplitude and polarity are determined by the a.f. signal. A spectrum analysis of this signal shows the presence of upper and lower sideband signals displaced about the carrier frequency, but negligible carrier signal. Also present are a similar pair of sidebands displaced about the carrier second harmonic and various other high frequency components. The unwanted frequencies are rejected by tuning T1 secondary to the required centre frequency of 455 kc/s.

35. A grounded centre-tap for T1 primary is simulated by the resistor-capacitor network R14–R16, C8–C10. Resistor R15 and capacitor C9 are set-up initially to obtain an accurately balanced output with negligible carrier content. The carrier suppression with this circuit is approximately 30dB. The value of C10 is selected during production test and in some equipments, due to large stray capacitances at this point, this capacitor may not be fitted or may be connected instead in parallel with C9.

Low i.f. circuit

36. Transformer T1 secondary is tuned with C171 to a centre frequency of 455 kc/s. A damping resistor, R140, the value of which is selected during production test, is placed across the tuned circuit to obtain a flat-top response pattern over the entire double sideband frequency range for the two possible b.f.o. frequencies. Capacitor C11 provides an effective r.f. ground return, and the unbalanced output from T1 is applied to the amplifier V4a. The gain of this amplifier is controlled by the negative a.l.c. bias (para. 60) fed to the grid circuit via R17. The resistor network R141, R150 and R30 in the screen and cathode circuit of this valve is part of the meter a.l.c. circuit to be discussed later (para. 95).

37. The amplified double sideband signals deloped across L2 are fed to the mechanical filter FL1 which, because of its 453·95 to 456·05 kc/s sharp cut-off response pattern, rejects the unwanted sideband and further attenuates any residual carrier signal. Since the input and output transducers of a mechanical filter are inductive, FL1 is tuned to its operating frequency by C14 and C54 at the input and C217 and C218 at the output.

Variable frequency oscillator 70K–2

38. The v.f.o. is a tuned grid-tuned cathode oscillator and uses fixed capacitance and variable inductance to tune the range 2·7 to 2·5 Mc/s. The front panel tuning control varies L303, and L302 is preset initially for accurate dial calibration, i.e. a full scale sweep of 200 kc/s. A further control of the tuning dial calibration is provided in the form of a mechanical set zero adjustment for the dial hairline. The parallel tuning capacitor C301 is selected during production test for correct oscillator tuning.

39. The series combination of C308 and CR301 is connected in parallel with C303. Diode CR301 then switches the preset capacitor C308 in or out of circuit depending on the polarity of the bias voltage applied to the anode. In all positions of S9 except LSB, CR301 anode is connected via R303, S9b and R35 to the +200V line, and C308 forms part of the tuned circuit. In the LSB position of S9, however, a negative voltage derived from the divider network R130 and R137 across the —70V supply is applied to the diode and C308 is effectively open-circuit. This causes an increase in the v.f.o. frequency of 2·7 kc/s to compensate for a 2·7 kc/s decrease in the b.f.o. frequency in this position of S9. This ensures that switching from one sideband to the other does not upset the tuning dial calibration.

40. The oscillator output is developed across V301 anode load, T301, and the advantage of this type of oscillator is that the output circuit is isolated from the oscillatory circuit and therefore the output load impedance cannot affect the frequency. The h.t. supply for V301 is obtained from the junction R73 and R131 connected across the TR+275V supply via the dummy plug inserted in J17 (para. 20).

41. The 2·7 to 2·5 Mc/s output is taken from a low impedance tapping on the auto transformer T301 via the attenuator R22 and R162 to the cathode follower V2a. The reason for transforming the v.f.o. output to low impedance in this manner is that in certain equipment configurations (para. 20) a second, external, variable frequency oscillator

70K–2 may be connected to J18 via a low impedance coaxial cable. The grid, and hence the cathode, of V2a is at a potential of approximately +120V as determined by the divider network R134 and R71 connected between the anode and ground. This is to ensure that the valve is operating on the most linear portion of its characteristic. The cathode-follower output is then fed to the transmitter channel first mixer and the receiver channel second mixer. A filter consisting of C248, C249 and L35 is provided to reject any spurious harmonics generated by the v.f.o. and to pass only the fundamental frequency.

First mixer

42. The single sideband output from FL1 is applied as a push–pull signal to the grids of the double-triode, V5. The unbalanced v.f.o. output from V2a, developed across L20 and C16, is coupled via C217 and C218 to both grids of V5, and as both anodes are connected in push-pull to the tuned output transformer T2, the v.f.o. signal energy is cancelled. Adjustment of the mixer balance is achieved by R24 in V5 cathode circuit. Transformer T2 is tuned to accept the sum frequency and the resulting 3·155 to 2·955 Mc/s output from T2 secondary is fed to the transmitter second mixer. The tuning of T2 in conjunction with L4 in the second mixer input circuit is sufficiently broadband to accommodate the 200 kc/s range of adjustment in the variable i.f.

43. Since T2 is also used in the receiver channel variable i.f. circuit, the h.t. for V5, applied to T2 primary centre-tap is derived from the TR+275V line to avoid any change in the associated stray capacitance. Therefore, in order to ensure that V5 is cut-off in the receive function, the grid circuit of this valve is connected to —70V via R26, L24 and K2/1. In the transmit function, however, K2 is energized and the grid d.c. circuit is connected to ground via K2/2 so that V5 conducts. Resistor R70 is placed across the relay contacts for spark suppression purposes.

TABLE 3

Crystal oscillator band groups

Band group	Crystal references
A	Y1, Y2, Y3, Y18, Y19, Y20
B	Y4, Y5, Y21, Y22
C	Y6, Y7, Y8, Y23, Y24, Y25
D	Y9, Y10, Y11, Y26, Y27, Y28
E	Y12, Y13, Y14, Y29, Y30, Y31

Crystal oscillator

44. The crystal oscillator formed by V13a and the associated circuit operates at any one of a possible total of 28 frequencies in the range 6·555 to 32·955 Mc/s as determined by the crystal selected by the 14-position BAND switch, S2 or S14, and the 2-position crystal board selector switch, S15. The

TABLE 4

Crystals supplied as standard components

BAND switch position	KWM–2A frequency band	Crystal frequency	Crystal reference
1A–3·4	3·4–3·6 Mc/s	6·555 Mc/s	Y1
2A–3·6	3·6–3·8 Mc/s	6·755 Mc/s	Y2
3A–3·8	3·8–4·0 Mc/s	6·955 Mc/s	Y3
1B–7·0	7·0–7·2 Mc/s	10·155 Mc/s	Y4
2B–7·2	7·2–7·4 Mc/s	10·355 Mc/s	Y5
1C–14·0	14·0–14·2 Mc/s	8·5775 Mc/s	Y6
2C–14·2	14·2–14·4 Mc/s	8·6775 Mc/s	Y7
3C–14·8	14·8–15·0 Mc/s	8·9775 Mc/s	Y8
1D–21·0	21·0–21·2 Mc/s	12·0775 Mc/s	Y9
2D–21·2	21·2–21·4 Mc/s	12·1775 Mc/s	Y10
3D–21·4	21·4–21·6 Mc/s	12·2775 Mc/s	Y11
1E–28A	28·5–28·7 Mc/s	15·8275 Mc/s	Y12

crystals appropriate to each of the five band groups of the KWM–2A are detailed in Table 3. Of the 28 possible crystals, only 12 are supplied as standard components and these are mounted on the under chassis crystal board (fig. 3). Details of these crystals and the particular KWM–2A frequency band that they cover are given in Table 4. The selection of the appropriate crystal for a particular band not listed in Table 4 is discussed in para. 19.

45. The crystal oscillator is a Pierce oscillator of the type discussed in para. 13. In this case, however, it will be noted that there is no capacitor from V13a grid to ground. This is because there is sufficient stray capacitance appearing at this point for an additional capacitor to be unnecessary. The tuned anode load of V13a is formed by T4 and the capacitor and damping resistor combination selected by S3. In the A band group, C69, C70 andR152 are selected; in the B band group, C63, C64 and R151; in the C band group, C65 and C66; in the D band group, C67; and in the E band group, C68. At KWM–2A operating frequencies of up to 12 Mc/s, V13a anode circuit is tuned to the crystal fundamental frequency. At all higher signal frequencies, V13a anode load selects the second harmonic of the crystal frequency. Fine tuning of V13a anode load is effected inductively by the adjustment of T4 with the front panel EXCITE TUNING control.

46. The crystal oscillator is common to both the transmitter and the receiver channels and therefore derives its h.t. from the TR+275V line via an r.f. filter network consisting of L7, C224, and L25. Two outputs are taken from the oscillator. One is from T4 secondary to the transmitter second mixer, and the other is from V13a anode to the receiver first mixer via the cathode follower V12b. The grid, and hence the cathode, of V12b is at a potential of approximately +120V in the receive function (when the R+275V supply is on) as determined by the divider network R160 and R154 connected between the anode and ground. This ensures that the valve is operating on the most linear portion of its characteristic.

Second mixer

47. The transmitter second mixer consists of the double-triode V6 and the associated circuit. The grid circuit of V6 is returned to ground via S9d in all positions of S9 except CW, and the h.t. supply used is the T+275V line, decoupled by C221 and R143, so that the mixer is inoperative in the receive function.

48. The variable i.f. output from T2 secondary is developed across the tuned circuit L4, C175, C176, and applied to V6b grid. A damping resistor, R124, is included to obtain the required 200 kc/s bandwidth. The high frequency output from the crystal oscillator at T4 secondary is coupled both to V6a grid and V6b cathode, and as the anodes are connected together to a common load, the crystal oscillator signal energy is cancelled in the output circuit. Any residual signal appearing as a result of slight unbalance in the two halves of V6 is cancelled by negative feedback via C26 from the common anodes to V6b grid.

49. The mixer output, consisting of the sum and difference frequencies plus various higher order components, is developed across L5 and the difference frequency, in the range 3·4 to 30·0 Mc/s, is selected by the tuned circuit T3, C30, C31 and the capacitor(s) selected by the BAND switch S4. These are C37 and C38 in the A band group, C32 and C33 in the B band group, C34 and C35 in the C band group, C36 in the D band group, and C39 in the E band group. Fine tuning of the mixer is effected inductively by the adjustment of T3 with the front panel EXCITER TUNING control in common with the tuning of the crystal oscillator output (para. 45). A primary winding on T3 associated with the receiver channel input circuit is effectively open-circuit in the transmit function.

R.F. amplifier

50. The 3·4 to 30·0 Mc/s output from the second mixer is coupled via C40 to the grid of the common receiver-transmitter r.f. amplifier V7. The gain of this amplifier is controlled in the transmit function by the negative a.l.c. bias (para. 60) fed to the grid circuit via R34 and K4/4 (energized). Resistor R37, in the screen circuit of V7, is associated with the meter a.l.c. circuit to be discussed later (para. 95). As this amplifier is common to both the receiver and the transmitter channels, the h.t. used is the TR+275V supply via the r.f. decoupling circuit C222, R144 and L25.

51. A parasitic oscillation suppressor, Z4, consisting of the inductor L8 wound on the resistor R185, is connected in the anode circuit of V7, the effective anode load of this valve being formed by a tuned circuit consisting of L10 and the capacitor(s) selected by the BAND switch S5. These are C109, C110 and C125 in the A band group, C112 and C113 in the B band group, C114 and C115 in the C band group, C116 and C257 in the D band group, and C111 in the E band group. Fine tuning of the r.f. amplifier is effected inductively by the adjustment of L10 with the front panel EXCITER TUNING control in common with the tuning of the crystal oscillator and the second mixer. Also, connected across V7 anode circuit, and therefore effectively in parallel with the tuned circuit, is a series capacitive network consisting of C272 and the varactor diode CR5. This is a reverse-biased diode whose self-capacitance is a function of the reverse voltage applied. In the transmit function this voltage is approximately +110V derived from a resistor divider network across the T+275V supply consisting of R183, R184 and R58 (cathode load of the receiver second mixer, V17b). In the receive function, however, when the T+275V line is open-circuit, the voltage applied to the diode via

R184 is the voltage appearing at V17b cathode, approximately +4V. The reduction in voltage causes an increase in the capacitance of CR5 which in turn compensates for the reduction in stray capacitance across the tuned circuit caused by the drive amplifier, V8, being cut-off in the receive function of the unit.

Driver and power amplifier

52. The r.f. amplifier output is further amplified in the driver stage, consisting of V8 and the associated circuit, before being fed to the grids of the power amplifier valves V9 and V10. This is a conventional circuit employing negative feedback from the power amplifier anodes to the driver cathode to reduce distortion in the output circuit to a minimum, and both stages are neutralized to ensure stability.

53. In the transmit function and grid circuit of V8 is returned to ground via R103, the r.f. filter C122 and L11 and relay contact K2/2 (energized). In the receive function, however, when K2 is de-energized, —70V is applied to this circuit via K2/1 and V8 is cut-off.

54. The signal developed across L10 is applied direct to the grid of V8, the effective anode load of which is formed by a tuned circuit consisting of L13 or L13 and L14 in parallel and the capacitor(s) selected by the BAND switch S6. In the first five positions of the switch (covering the lower frequency band groups A and B) L14, connected to S6b, is open-circuit, but in the remaining nine positions it is shunted across L13. This reduces the tuned circuit inductance to about 40% of its original value thereby minimizing the effect of stray capacitance in the higher frequency bands. The tuning capacitors selected by S6 are C130 an C131 in the A band group, C128 and C129 in the B band group, C133 in the C band group, C135 and C138 in the D band group, and C134 in the E band group. As in the preceding stages the tuning of the driver is effected inductively by adjustment of L13 and L14 with the EXCITER TUNING control.

55. The anode-grid and grid-cathode inter-electrode capacitances in V8 are neutralized by the series capacitor combination C117 and C118 in conjunction with C121. Negative feedback is applied from the power amplifier valve anodes to V8 cathode, and developed across L23, via the capacitor divider network C180 and C123 (an additional capacitor, C270, is added in parallel with C180 on some units during production test). When employing negative feedback in this manner it is essential that the grid-cathode capacitance of the driver valve should be neutralized to prevent undesirable feedback coupling to the input grid. This is achieved by the inclusion of the parallel feedback neutralizing capacitors C97 and C120.

56. The two power amplifier valves V9 and V10 are connected in parallel. Parasitic oscillation suppressors Z1, Z2, Z6 and Z7, each consisting of a small inductor wound on a 47-ohm resistor, are connected in the anode and grid circuits. The anode h.t. supply for these valves is the +800V via the filter network L17, L28, C148 and C168. As this supply is not switched off in the receive function, V9 and V10 are cut-off when the unit is operating in this function by the removal of the screen h.t. supply. This is obtained from the divider network R148 and R146 to which the T+275V line is connected via S9a and the gating diode CR7. The diode serves to isolate the open-circuit T+275V line in the receive function from any positive potential developed at the screens due to the +800V supply still being applied to the anodes. The d.c. grid circuit for V9 and V10 is completed via L13, R115, the meter shunt R117, and R116 to the —70V line. In the initial setting-up of the equipment, the correct operating point on the power amplifier valve characteristic is obtained by adiustment of the —70V output from the a.c. power supply PM-2.

57. The output of the driver, V8, is applied to the grids of the two power amplifier valves. The anode-grid and grid-cathode inter-electrode capacitances in V 9 and V10 are neutralized by C184 in conjunction with C137 in a circuit similar to that employed in the preceding driver.

58. The amplified output at the power amplifier valve anodes is developed across the tuned circuit formed by C150 (P.A. TUNING), L18, the tapped inductor L19, C155, the capacitor selected by the BAND switch S8, and C151 (INCR LOAD). The amount of inductance in the circuit is selected by the BAND switch S7. In the lowest frequency band group, A, all of L19 is in circuit and this is progressively reduced as the switch is rotated through the bands until in the highest frequency group, E, L19 is completely short-circuited. Similarly, the amount of shunt capacitance is also reduced with increasing operating frequency. The INCR LOAD front panel control is a twin-gang capacitor one-half of which C151b, is connected in the A and B band groups, and in the highest frequency bands, groups D and E, no additional trimming capacitor is selected by S8. The circuit is tuned to the frequency in use by means of C150 and matched to the impedance of the output line and antenna by means of C151.

59. The power amplifier output is fed via K3/1 (energized) to the R.F. OUT socket, J1. A d.c. path to ground for the output line is provided by the high impedance choke L21. A choke is used here in preference to a resistor since it is important that all components in the matching circuit should, ideally, be pure reactances so that the r.f. power output is dissipated entirely in the antenna, this being the only pure resistance present.

Automatic load control

60. A negative a.l.c. bias signal is obtained from the double-diode rectifier V17a. When the positive signal peaks applied to V9 and V10 grids drive these valves into grid current, current is drawn through the resistor chain R116, R117 and R115, and the waveform developed at the junction of R115 and R117 is coupled via C157 to the anode of the first diode. On large amplitude signals when this waveform is of sufficient amplitude to overcome the threshold bias of approximately +1·5V derived from R191 and R192 at the diode cathode, the diode detects the waveform and produces a negative going envelope signal that is applied to the second diode cathode. The anode of this diode is at approximately —1·5V as determined by R189 and R190 across the —70V supply and is maintained at this potential by the catching diode CR8. The negative-going signal is rectified by the second diode and smoothed by the capacitor network C158–C160 and added to the —1·5V already existing on the a.l.c. line. Diode CR8 is then cut-off so that the a.l.c. signal has a fast rise time as the capacitors are charged through the diode and a slow decay time as they discharge through R118 and R119.

61. Also added to the —1·5V steady bias on the a.l.c. line is the a.l.c. signal developed by the external r.f. linear amplifier 30L–1 (Part 4, Chap. 3) and fed into this unit at J4. The full a.l.c. signal is then applied to the grid of the r.f. amplifier V7 (para. 50) via K4/4, and an attenuated a.l.c. signal, from the junction of R118 and R119, to the grid of the first i.f. amplifier V4a (para. 36).

Tone oscillator

62. We have considered so far the operation of the transmitter channel with the EMISSION switch, S9, in the LSB or USB position only. In the other three positions of this switch a 1500 c/s tone oscillator consisting of V2b and the associated circuit is switched on. This is a phase shift oscillator and the cathode circuit of V2b is connected to ground by S9f in the LOCK, TUNE and CW positions. The feedback path from anode to grid is via the d.c. blocking capacitor C232 and a complex advance and delay R–C phase change network. The tone oscillator output is sinusoidal with negligible harmonic content and is taken via R128 to the anode of the first microphone amplifier V1a.

LOCK and TUNE

63. These two positions of the EMISSION switch, S9, are used in the initial setting-up and tuning of the equipment. Setting the switch to the LOCK positions has the following effects:

(1) Switch S9c removes the ground connection from V1a cathode circuit thereby rendering the microphone or phone patch inputs inoperative. A ground line is routed instead to the grid of the receiver channel second a.f. amplifier V16b to render this stage inoperative also.

(2) Switch S9f switches on the tone oscillator.

(3) Switch S9g connects the p.t.t. line to ground, thereby maintaining the VOX relay amplifier V4b (para. 84) conducting and hence the unit in the transmit function.

(4) Switch S9h selects the upper sideband b.f.o. frequency.

(5) Switch S9e short-circuits CR8 in the a.l.c. circuit (para. 60) so that C158–C160 are able to discharge through R190 and the a.l.c. signal has a short decay time.

64. With S9 in the LOCK position therefore, the unit transmits continuously at full power and at a frequency equal to the tuning dial setting plus 1500 c/s. Setting the switch to the TUNE position has a similar effect except that S9a applies at a reduced screen voltage, obtained from the junction of R69 and R70, to the power amplifier valves thereby reducing the output power.

C.W. operation

65. For c.w. operation of the unit, a morse key is connected to J7 and S9 is switched to CW. Setting the switch to this position has the following effects.

(1) Switch S9c removes the ground connection from V1a cathode circuit thereby rendering the microphone or phone patch inputs inoperative. A line is routed instead to the VOX circuit to connect C188 directly to ground and so reduce the VOX release time (para. 86).

(2) Switch S9f switches on the tone oscillator.

(3) Switch S9d removes the ground connection to the junction of R165 and R136. The second microphone amplifier, V11b, and the transmitter second mixer, V6, are thus cut-off by —35V bias applied to the grid circuits. A ground line is routed instead to the receiver a.f. output transformer, T6, so that this circuit is operative in the transmit as well as the receiver function.

(4) Switch S9g removes the ground connection from the junction of R155 and R156 so that an attenuated output from V11b is fed to the receiver first a.f. amplifier V16a. This enables the transmitted c.w. signal to be monitored on the local loudspeaker or headphones.

(5) Switch S9h selects the upper sideband b.f.o. frequency.

66. With S9 in the CW position, therefore, the unit remains in the receive function until the morse key is depressed. This connects a ground input to J7 which removes the —35V bias applied to V11b and V6. These valves now conduct and the amplified 1500 c/s signal from V11b actuates the VOX circuit and switches the unit to the transmit function. The transmitted signal is at a frequency equal to the tuning control dial setting plus 1500 c/s. Due to the shortening of the VOX circuit release time (para. 65 (1)) the unit reverts to the receive function immediately the morse key is released.

Receiver channel

67. The following assumptions are made for the purposes of this description.

(1) The EMISSION switch, S9, is in the LSB, USB or CW position.

(2) S11 is in the NB position.

(3) The transfer relays K2, K3 and K4 are de-energized.

R.F. amplifier

68. The 3·4 – 30·0 Mc/s received signal input from the antenna at J1 is fed via K3/1 and K2/3 to the primary of the tuned transformer T3 in the input circuit of the common receiver-transmitter r.f. amplifier V7 (para. 50 and 51). In the receive function the gain of this amplifier is controlled by the negative a.v.c. bias signal fed to the grid via R34 and K4/4. An amplified r.f. output is taken from V7 anode to the first mixer V13b.

69. An alternative receiver input is provided at J2. This is not used in this particular equipment configuration, its purpose being to provide a separate receiver antenna input when the transmitter and receiver channels are operating at different frequencies.

First mixer

70. The amplified 3·4 to 30·0 Mc/s signal from V7 is coupled via C60 to V13b grid, and the 6·555 to 32·955 Mc/s output from the crystal oscillator (para. 44–46) is coupled via the cathode-follower V12b and C61 to V13b cathode. The anode load of V13b is formed by the broadband tuned transformer T2 which selects the 3·155 to 2·955 Mc/s mixer difference frequency signal. This transformer is common to both the transmitter and receiver first mixer stages, but since both V13b and the input cathode follower V12b use the R+275V h.t. line, the receiver mixer is rendered completely inoperative in the transmit function and cannot affect the operation of the transmitter mixer or feed signals back into the transmitter channel. As an additional precaution, V13b grid circuit, which is returned to ground via K2/2 in the receive function, is connected to —70V via K2/1 in the transmit function.

Variable i.f. circuit

71. The first mixer output, referred to as the variable i.f., is fed via J22 to the input of the noise blanker 136B–2 sub-assembly unit (para. 3). The processed output from the 136B–2 is then fed back again at J23. In the absence of the 136B–2, J22 and J23 are connected together by a short-circuiting link. Power supply and gain control connections to the 136B–2 are made via the 9-pin socket J24. The external noise blanker antenna is connected to J28, terminated by L27, at the rear of the KWM–2A chassis and thence via J26 to the 136B–2 r.f. amplifier input. The 136B–2 functions as a noise blanking circuit only when S11 is in the NB position. In the ON and CAL positions of this switch the 40 Mc/s noise amplifier is inoperative and the variable i.f. signals are passed without attenuation to the second mixer V17b.

Second mixer

72. The variable i.f. signal at J23 is coupled via the high frequency blocking circuit Z5 to V17b grid, the purpose of Z5 being to eliminate spurious high frequency signals that may be generated in the 136B–2. The 2·7 to 2·5 Mc/s v.f.o. signal is coupled via the cathode follower V2a, K4/6 and C186 to V17b cathode and the resulting mixer output, consisting of the sum and difference frequencies and various other higher order components, is developed across L33. This is coupled via CR9 and C23 to the mechanical filter FL1 which selects the 455 kc/s difference frequency and determines the receiver selectivity bandwidth of 2·1 kc/s at —6dB.

73. The purpose of the gating diode CR9 connected to the potential divider, R193 and R194, across the TR+275V supply is to ensure that in the transmit function when the R+275V line is open-circuit and V17b is inoperative the transmitter channel first mixer, V5, is completely isolated from the receiver channel second mixer circuit. The grid of V17b, which is returned to ground via R59 and K2/2 in the receive function, is connected to—70V via K2/1 in the transmit function.

Low i.f. circuit

74. The 455 kc/s output from FL1 is coupled via C83 to the grid of the first i.f. amplifier V1b, the output of which is developed across the tuned anode load formed by L9 and C117, and fed to the grid of the second i.f. amplifier V3b. The damping resistor R145 is included in order to obtain a flat-top response pattern over the required 2·1 kc/s bandwidth. The gain of the 2-stage amplifier is set-up initially by adjustment of R132 in V1b cathode circuit and in operation is controlled by the a.v.c. negative bias signal (para. 80) applied to both valves via R74 and R78 respectively. The resistor network in the screen circuit of the two valves is part of the meter circuit to be discussed later (para. 91).

75. The anode load of V3b is formed by the tuned transformer T5, the required 2·1 kc/s bandwidth being obtained by critical coupling between the two windings of the transformer. One side of T5 secondary is effectively grounded through C91 which, with V15a, forms part of the a.v.c. rectifier circuit (para. 81), and the output signal is taken from the opposite end via C94 to the grid of the product detector V15b.

Product detector

76. The product detector, consisting of V15b and the associated circuit, is basically a similar mixer to the two already described (para. 70 and 72) in that the signal is applied to the valve grid and the b.f.o. output is applied to the cathode. Components C96 and L31 form a high-pass filter to prevent any stray a.f. signals picked-up on the b.f.o. signal line in the transmitter balanced modulator being applied to the product detector. The low-pass filter R91, C98, C99 selects the 300 to 2400 c/s difference frequency component of the output appearing at V15b anode and this is then coupled to the a.f. amplifier via C100.

77. The product detector uses the +200V h.t. supply since in the c.w. mode of operation V15b anode circuit forms part of the second microphone amplifier circuit and a transmitter monitor signal is fed to the receiver a.f. amplifier (para. 65). As in the case of the preceding two mixers, V13b (para. 70) and V17b (para. 73), V15b grid is returned to —70V via K2/1 in the transmit function.

A.F. amplifier

78. The 300 to 2400 c/s output from the product detector is applied via the A.F. GAIN control R92 to a 2-stage R-C coupled a.f. amplifier consisting of the triode-pentode V16 and the associated circuit. The +200V line is used as an h.t. supply for this amplifier since it is used during the transmit function in the c.w. mode of operation. Negative feedback is supplied to the second stage by means of the un-decoupled cathode bias resistor R97.

79. The output transformer T6 provides a 500-ohm output at J9 that is used as an input to the external phone patch on the station control 312B-4 (Part 4, Chap. 4) and a 4-ohm loudspeaker output at J10. If high impedance headphones are connected to J14, the output to J10 is open-circuited and the load impedance is suitably adjusted by R100 being connected in parallel with R182 across the 4-ohm section of the secondary. In the receive function, one side of T6 secondary is grounded via K2/4. In the transmit function, however, K2 is energized and this ground connection is removed. All audio outputs are then inoperative and negative feedback is applied from T6 to the input of V16a via R164 to cancel any signal applied to the amplifier. When S9 is in the CW position, however, the ground connection is not removed but is obtained instead from S9d (para. 65) to that the a.f. amplifier continues to function normally in order to provide a means of monitoring the transmitted c.w. signal.

Automatic volume control

80. The gain of the r.f. amplifier in the receive function and of the two receiver i.f. amplifiers is controlled by the negative bias a.v.c. signal. The steady d.c. potential on the a.v.c. line is derived from the R.F. GAIN control R84 in a potential divider network between —70V and ground. In the receive function, the junction of R126 and R85 is connected to ground via K3/2 and L32. In the transmit function, however, this ground connection is removed with the result that an additional resistor, R85, is inserted in the ground end of the divider network. This increases the negative bias in order to ensure that the i.f. amplifier valves are cut-off immediately the unit switches to transmit.

81. Superimposed on the steady d.c. level is a negative-going signal produced by the a.v.c. rectifier V15a. In the quiescent state this valve section, a parallel-connected double-diode, is reverse biased by the negative a.v.c. voltage derived from R84 applied to its anodes, and the positive bias voltage of approximately +2·8V obtained from V16a cathode applied to its cathodes. When the 455 kc output from T5 secondary is of sufficient amplitude to overcome the reverse bias, V15a conducts on positive peaks and charges C91 negatively. The circuit has a fast rise time, determined by the forward resistance of V15a, and a slow decay time, determined by the discharge path R180 and R83. Additional smoothing of the a.v.c. signals is provided by C265 and C93.

Calibrator oscillator

82. We have so far considered the operation of the receiver channel with S11 in the ON or NB position only. When this switch is set to the CAL position a 100 kc/s crystal calibrator and harmonic generator, consisting of V12a and its associated circuit, is switched on by the completion of V12a cathode circuit to ground. The calibrator is a 100 kc/s Pierce oscillator similar to that already described (para. 31) and uses the R+275V h.t. supply so that it is operative only in the receive function. Fine adiustment of frequency is by means of the preset capacitor C76.

83. The 100 kc/s sinusoidal output at V12a anode is coupled via C119 to the harmonic generator CR6. This clips the negative half cycles of the 100 kc/s sine wave so that the input fed via C79 to T3 primary is a complex waveform consisting of harmonics at 100 kc/s intervals throughout the operating range of the KWM–2A. The front panel tuning dial calibration can therefore be checked by tuning for zero beat with the calibration signal at any of the 100 kc/s points (0, 100 and 200) on any of the 28 frequency bands.

VOX and ANTI-VOX circuits

VOX rectifier

84. The transmitter microphone amplifier employs the +200V h.t. and is therefore operational in both the transmit and receive functions. The a.f. signal from the second microphone amplifier V11b is applied via the VOX GAIN control, R39, to the VOX amplifier V14b, the output of which is recti-by one half of the double-diode V14a. The smoothing capacitors associated with the rectifier are C188, connected to ground via R13 and K2/2, and C47, connected to ground via the T+275V line and the various resistor paths across this line. The positive charge developed on these capacitors is applied via R42 to the grid of V4b, the anode load of which is formed by K2 operating coil. In the quiescent state, V4b is maintained cut-off by a potential of approximately +18V at the cathode derived from the divider network R46, R47 and R20 across the TR+275V supply. When the rectified signal is of sufficient amplitude to cause V4b to conduct, K2 is energized and a ground line is connected via K2/4 and R167 to K3 and K4 operating coils, thereby energizing these relays also. The unit is then in the transmit function.

85. With K2 energized and the unit in the transmit function, C188 is connected to —70V via R13 and K2/1 instead of to ground, and +275V is applied to C47. The resulting potential at the junction of C188 and C47 is then in excess of +275V due to the positive charge already existing on C47. A complex charge path then exists for the two capacitors: R43 (VOX TIME CONSTANT), the second half of the double-diode and R45 (ANTI-VOX GAIN) to ground for both capacitors, and additionally R13 to the —70V line for C188. When the potential at the junction of C47 and C188 has fallen to a sufficiently low potential to cause V4b to cut-off, assuming that there is no longer any audio input, K2 is de-energized and the unit reverts to the receive function. The potential at the junction of the two capacitors is then negative and a complex discharge path exists: the VOX rectifier diode and R44 to ground for both capacitors, and additionally R13 to ground for C188. Resistor R43 is adjusted to ensure a sufficiently long time constant so that when the operator is speaking into the microphone the unit stays in the transmit function and does not switch back to the receive function between words.

86. When the EMISSION switch, S9, is set to the CW position, C188 is connected permanently to ground direct via S9c. Consequently, when the unit switches to the transmit function the potential at the cathode of the VOX rectifier diode is determined by the capacitive divider C47 and C188 between +275V and ground. This is considerably less than that obtained in the single-sideband (speech input) mode of operation so that there is only a very short delay in switching back to the receive function after releasing the morse key.

ANTI-VOX rectifier

87. When the unit is in the receive function it is possible that the loudspeaker audio output picked up by the microphone could cause the VOX circuit to operate. To avoid this, an ANTI-VOX circuit is employed to provide a threshold signal in opposition to the VOX signal. The output from the receiver channel second a.f. amplifier, V16b, is coupled via C235 and the ANTI-VOX GAIN control R45 to the second of the diodes of V14a. This provides a negative signal across C48 that is applied via R48 to V4b grid.

Press-to-transmit facility

88. The VOX circuit can be bypassed if necessary and the unit maintained in the transmit function by the use of an external p.t.t. switch. This is connected either to J16 or to the top of the microphone plug inserted in J15. When the switch is operated a ground line is applied to the cathode of the relay amplifier V4b, thereby removing the bias from this valve and maintaining it in a conducting state. The circuit operates in a similar manner when S9 is set to the LOCK or TUNE position (para. 63).

External relay circuits

89. When K3 is energized in the transmit function, the ground line that was previously routed from K3/2 to the junction of R126 and R85 in the a.v.c. voltage divider network (para. 80) is routed instead to J3 (ANT. RELAY). The purpose of this output is to operate the changeover relay on the r.f. linear amplifier 30L–1 (Part 4, Chap. 3).

Meter circuit

90. The front panel meter, M1, is a 400mA f.s.d. d.c. instrument the connections to which are made via the relay contacts K4/1 and /2. In the receive function, M1 is connected to a circuit measuring receiver signal strength, while in the transmit function, when K4 is energized, it is connected instead to the PLATE-GRID-ALC switch, S12. This selects the meter function and M1 can be used to measure power amplifier anode current, power amplifier grid drive signal level, or the a.l.c. bias level respectively.

Receiver signal level

91. In normal operation, the R.F. GAIN control R84 (para. 80) is tuned fully clockwise so that the bias appearing on the a.v.c. line due to this control is only approximately —1·4V. Therefore, the greater part of the negative bias appearing on the a.v.c. line is derived from the a.v.c. rectifier V15a, i.e. the a.v.c. potential is proportional to receiver signal strength. The a.v.c. signal is applied to the control grids of the two pentode amplifiers V1b and V3b. Screen grid current in these valves is a function of the control grid d.c. potential, so that therefore the potential appearing at the screens can be used as a relative measure of receiver signal strength, i.e. an increase in signal strength causes an increase in screen potential. When used in this manner, M1 is referred to as an 'S-meter' and is calibrated in 'S' units.

92. M1 is connected across the meter shunt R158 in a resistor network to which is applied, via R79 and R120, the screen potentials of V1b and V3b respectively. One side of the network is returned to ground via R89 and the cathode load of V1b, while the other side is grounded through R121 (S METER ZERO) and R123. Resistor R121 is adjusted in conjunction with the preset gain control R132 so that with no signal input the meter reads zero.

Power amplifier anode current

93. With S12 in the PLATE position, M1 is connected in series with the parallel resistors R129 and R161 to measure the power amplifier cathode potential, this being proportional to anode current.

Power amplifier grid drive

94. With S12 in the GRID position, M1 is connected across R117 in the power amplifier grid circuit. The voltage developed across this resistor due to grid current is a measure of the grid drive applied to the power amplifier valves.

ALC signal level

95. As in the case of the a.v.c. meter circuit (para. 91), the screen potentials of the transmitter channel i.f. amplifier, V4a, and the r.f. amplifier, V7, is the function of the a.l.c. bias potential applied to these valves. With S12 in the ALC position therefore, M1 is connected across the meter shunt R150 in a resistor network to which is applied, via R41 and R37, the screen potentials of V4a and V7 respectively. One side of the network is returned to ground via V4a cathode load, R19, and the other side via the ALC ZERO control R30. In this position of S12, M1 is set to zero by adjustment of R30 with zero input signals (MIC GAIN, R8, turned fully counter-clockwise) and the transfer relays energized (p.t.t. line grounded).

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Chapter 2

NOISE BLANKER 136B-2

LIST OF CONTENTS

LIST OF ILLUSTRATIONS

INTRODUCTION

1. The noise blanker 136B-2 is a sub-assembly unit in the transceiver KWM-2A (Part 4, Chap. 1) and is used to convert random noise signals to bias pulses for gating the KWM-2A receiver channel 2·955 to 3·155 Mc/s variable i.f. signals, thereby reducing the receiver output noise to a minimum. The 136B-2 is tuned to 40 Mc/s and ideally is fed from its own 40 Mc/s quarter-wave whip aerial, the principle of operating being that spurious noise signals (particularly those caused by the motor transport electrical ignition systems) in the 40 Mc/s region of the h.f. spectrum occur simultaneously in the 3 to 30 Mc/s region.

2. The 136B-2 is mounted inside the hinged top cover of the transceiver KWM-2A (fig. 1). All connections, including power supplies, between the 136B-2 and the main unit are made via four plugs (three coaxial and one 9-pin) and their mating sockets on the top of the KWM-2A chassis. The operation of the 136B-2 is controlled by the OFF-ON-NB-CAL switch on the KWM-2A front panel and the noise blanker gain control mounted immediately behind the R.F. GAIN control knob, also on the KWM-2A front panel. A twin potentiometer assembly with concentric spindles is used for the r.f. and noise blanker gain controls. Inside views of the 136B-2 showing component locations are provided in fig. 2.

Limitations

3. The 136B-2 has the following three limitations which decrease the noise blanking efficiency of the system.

(1) Noise pulses having no energy distribution at 40 Mc/s can occur in the frequency range of the KWM-2A. The 136B-2 will, therefore, not generate a blanking pulse and so will permit the passage of these noise pulses.

(2) A strong signal in the pass band between the first and second mixers in the KWM-2A can be modulated by blanking pulses. This modulation process will cause sideband signals in the pass band which result in decreased blanking efficiency and can be reduced only by either reducing the noise blanker gain or switching off the 136B-2 completely.

(3) Some corona noise and static disturbances have a repetition rate in excess of 100 000 p.p.s. The blanking efficiency decreases as the pulse repetition rate exceeds 5000 p.p.s.

BRIEF DESCRIPTION

4. A block diagram of the 136B-2 is shown in fig. 3. The input from the noise blanker aerial is applied to a stagger-tuned 40 Mc/s amplifier. Ideally the amplifier should have a broadband (1 to 2 Mc/s) flat top response pattern centred about 40 Mc/s and should not be tuned for a sharp response at any particular frequency. The amplifier output is detected in a circuit incorporating a pulse stretching circuit so that the duration of any given detected pulse is partly a function of its peak amplitude. The negative-going 'white noise' signals thus produced are applied to a 3-stage pulse amplifying and shaping circuit producing positive-going gating pulses.

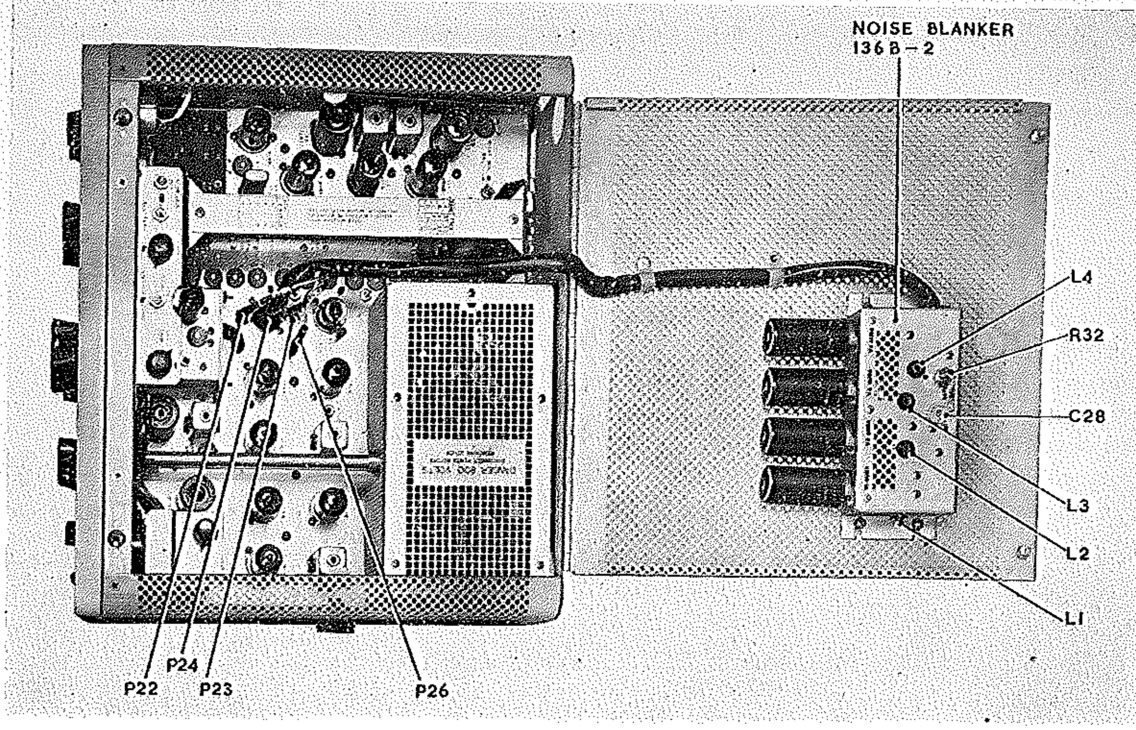

Fig. 1. Noise blanker 136B–2 fitted in transceiver KWM–2A

5. The 2·955 to 3·155 Mc/s variable i.f. signals from the receiver channel of the KWM–2A are fed into the gating circuit via a cathode follower. The gating circuit is arranged in a balanced modulator configuration so that any noise generation resulting from the gating action is cancelled and prevented from entering the receiver circuits. The positive-going gating pulses from the pulse amplifier are applied to this circuit via a threshold gate to eliminate all small amplitude pulses. This is necessary to ensure that the circuit operates only for noise pulses that are likely to cause appreciable interference in the receiver, and also because many small amplitude noise pulses need not necessarily be present at the aerial but may be generated internally in the first stage of the high gain 40 Mc/s amplifier.

6. The output from the gating circuit is fed to a low gain i.f. amplifier to compensate for the signal attenuation in the cathode-follower and the gating circuit so that the overall i.f. gain of the 136B–2 is approximately unity. The output from this amplifier is fed back to the variable i.f. circuit in the KWM–2A receiver channel.

DETAILED CIRCUIT DESCRIPTION

7. A circuit of the 136B–2 is shown in fig. 4. The plugs P22, P23, P24 and P26 connect to sockets J22, J23, J24 and J26 respectively on the transceiver KWM–2A. Socket J26 then connects to J28 NB ANT at the rear of the KWM–2A to which the 40 Mc/s noise blanker aerial is connected. In the absence of a 136B–2 sub-assembly unit, J22 and J23 are connected together via a short-circuiting link.

Power supplies

8. The 275V h.t. and —70V bias supplies for the 136B–2 are obtained from the KWM–2A via P24. The 6·3V a.c. valve heater supply is also obtained from the KWM–2A via pins 6 and 8 of P24, the ground return line being connected to pins 5 and 7. The reason for series—connecting the valve heaters in the manner shown in fig. 4 and for the apparently unnecessary decoupling capacitors C34 and C36 is that in certain equipment configurations (not discussed in this publication) the 136B–2 valve heaters are connected to a 24V supply tapped at 6V intervals.

R.F. amplifier

9. The noise blanker aerial input at P26 is coupled via C17 to the input of a conventional stagger-tuned r.f. amplifier consisting of V1a, V2a, V3a and their associated circuit. The necessary flat-top, broadband (approximately 1·5 Mc/s centred about 40 Mc/s) response is obtained by tuning the four tuned circuits of the amplifier in the following manner, and also, to a lesser extent, by connecting a damping resistor, R12 across the final tuned circuit.

(1) Input circuit, L1, tuned to 40·0 Mc/s

(2) V1a anode load, L2, tuned to 39·7 Mc/s

(3) V2a anode load, L3, tuned to 40·3 Mc/s

(4) V3a anode load, L4, tuned to 40·0 Mc/s

Gain control

10. The r.f. amplifier gain is controlled by an external variable resistor, R25, mounted on the KWM–2A front panel (*para.* 2) and connected

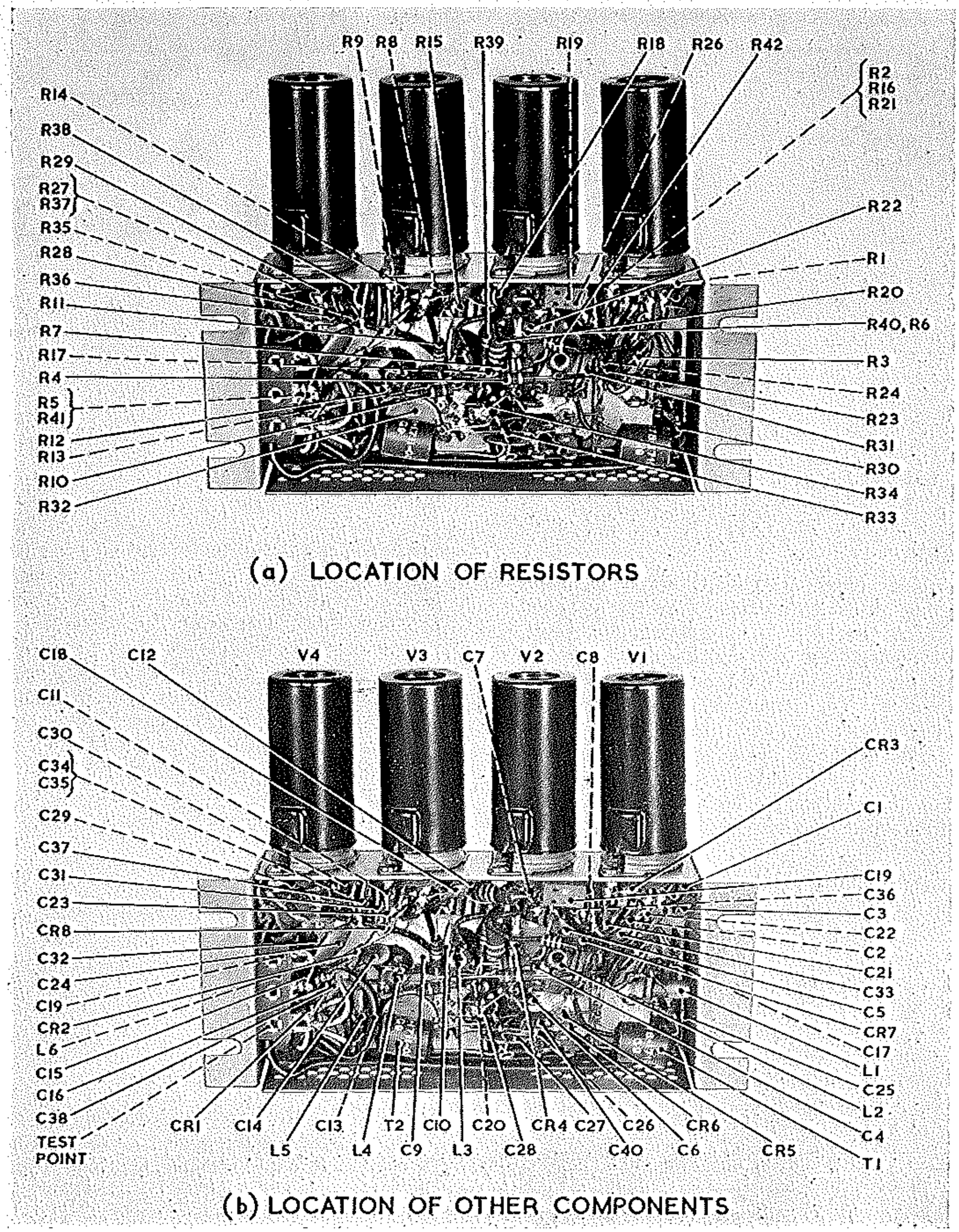

Fig. 2. Noise blanker 136B–2—inside

in the cathode circuit of V2a. The resistor is connected to ground only when S11 on the KWM–2A is in the NB position so that in any other position of S11 the potential at V2a cathode is determined by the potentiometer chain R39 and R40 across the +275V supply. The cathode potential is then approximately +80V and V2a is cut-off.

Detector

11. The amplified r.f. noise signals are developed across L5, and large amplitude signals are limited by the action of CR8 and the d.c. bias potential of approximately +2·6V at V3a cathode. Components CR1 and C15 form a detector circuit so that negative-going "white noise" signals are developed across R5. The combination of C15 and R5 form a pulse-stretching circuit so that the noise pulses have a duration that is a function of their original duration and their peak amplitude. This satisfies the requirement for large amplitude noise signals to produce longer duration blanking pulses than small amplitude signals.

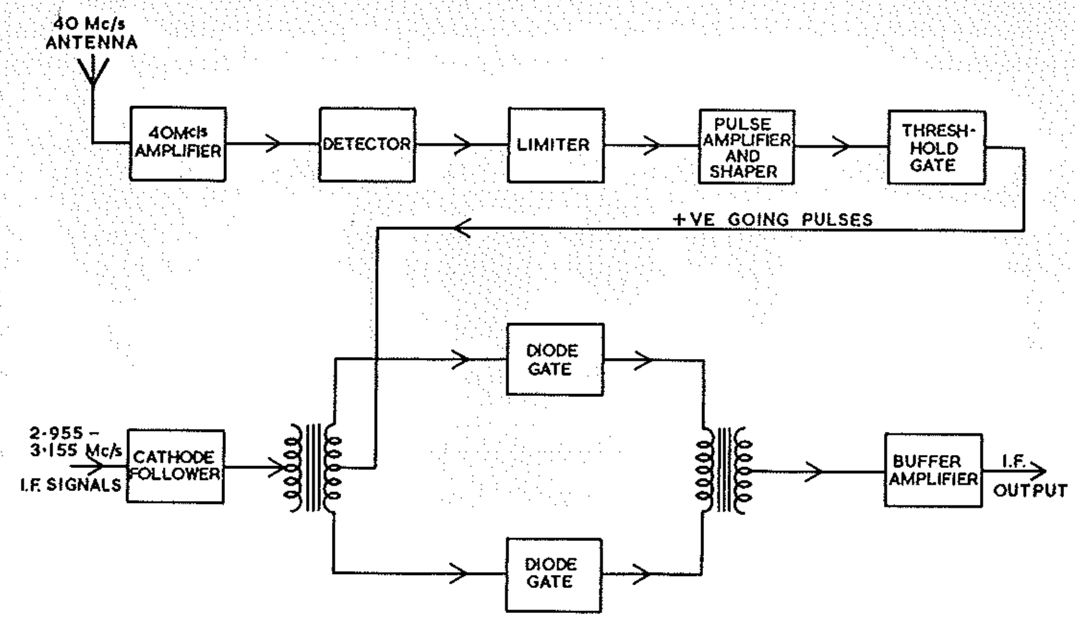

Fig. 3. Noise blanker 136B–2—block diagram

Pulse amplifier

12. The noise signals developed across R5 are negative-going with their base line at ground potential. They therefore have a mean d.c. level that is slightly negative. This d.c. level is removed by a.c. coupling through C16 so that the signals developed across R41 have their base line at a slightly positive potential. CR2 therefore conducts between signals and the waveform appearing at V3b grid consists of negative-going noise signals clipped at a low amplitude by CR2 to produce flat top pulses with short rise and fall times.

13. The triodes V3b, V2b, V1b and their associated circuit form a conventional R–C coupled pulse amplifier. Diode CR4 is included to eliminate negative overswing on the trailing edge of the positive-going pulses applied to V2b grid, and CR3 is a 'catching diode' to limit the positive-going pulse amplitude at V1b anode thereby ensuring that the gating pulse output has a flat top with fast rise and fall times. Gain stability is ensured by the provision of a d.c. negative feedback loop from V1b anode to V2b cathode via R42.

I.F. gating circuit

14. The 2·955 to 3·155 Mc/s variable i.f. signals at P22 are applied to the cathode-follower V4b, the effective cathode load of which is formed by T1. This transformer provides a balanced i.f. output at its secondary, the centre-tap being returned to earth via the low impedance of C40. The d.c. potentiometer network consisting of R32–R34, CR5, CR6, T1 secondary and R31 connected between ground and the —70V line ensures that normally the two gating diodes, CR5 and CR6, are conducting and the balanced i.f. signals are fed direct to T2 primary. The preset resistor R32 is adjusted initially to obtain equal current flow through both halves of T1 secondary and the gating diode cathodes are then at approximately —1V. Capacitor C27 provides d.c. blocking in the T2 primary circuit. It will be seen that T2 primary is not centre-tapped, but instead an effective centre-tap is provided by means of the capacitive divider network C26 and C28. The complete circuit is then accurately balanced by the adjustment of C28 in conjunction with R32.

15. The unbalanced output from T2 secondary is amplified by V4a and fed back to the KWM–2A receiver circuit via P23 and the impedance matching circuit C32, L6, C19. The gain of V4a is sufficient to compensate for the attenuation of the i.f. signals in the cathode-follower, V4b, and the following gating circuit. The two halves of V4 therefore effectively isolate the gating circuit from the KWM–2A receiver channel.

16. The positive-going gating pulses produced at V1b anode are coupled via C22 to the anode of the threshold gate diode CR7. This is at a potential of approximately —3·5V as determined by R23 and R24 across the —70V supply. The cathode of CR7, however, is at a potential of approximately —1V (*para.* 13) so that normally this diode is cut-off and is brought into conduction only by those gating pulses having an amplitude large enough to overcome the reverse bias on CR7. The large amplitude gating pulses thus applied to T1 secondary cut-off CR5 and CR6 thereby disconnecting the i.f. signals for the duration of the gating pulse.

Chapter 3

R.F. LINEAR AMPLIFIER 30L-1
(Completely revised)

CONTENTS

TABLES

ILLUSTRATIONS

Modification state:

The equipment described in this chapter incorporates the approved Service modifications listed in Table 3.

Manufacturer's production changes have been introduced with the result that different 'build standard' equipments are in Service use - these changes are listed in Table 3.

Introduction

1. One of the requirements of an efficient s.s.b. transmitting system is the use of a linear power amplifier stage. It is essential that the anode output r.f. signal should be a replica of the cathode (grounded-grid circuit) input signal. Any non-linearity in the power amplifier will result in intermodulation between the frequencies of the input signal which will produce undesirable distortion within the desired channel and also an intermodulation output on the adjacent channels.

2. The r.f. linear amplifier 30L-1 is a power amplifier unit for amplifying the low power output of the transceiver KWM-2A to give a nominal r.f. output of 500 W peak envelope power (p.e.p.) for delivery to the antenna TD-1, 637T-2 or similar 50-ohm broadcast antenna.

3. Performance data of the linear amplifier is located in the Leading Particulars of this publication and the function of the front panel controls (illustrated in Pt. 1, Chap. 4, Fig. 2) is detailed in Table 1.

4. Top, rear and bottom views of the linear amplifier with the covers removed are provided in fig. 1, 2 and 3 respectively. Interlock switches are installed to disable the e.h.t. supply when the covers are removed.

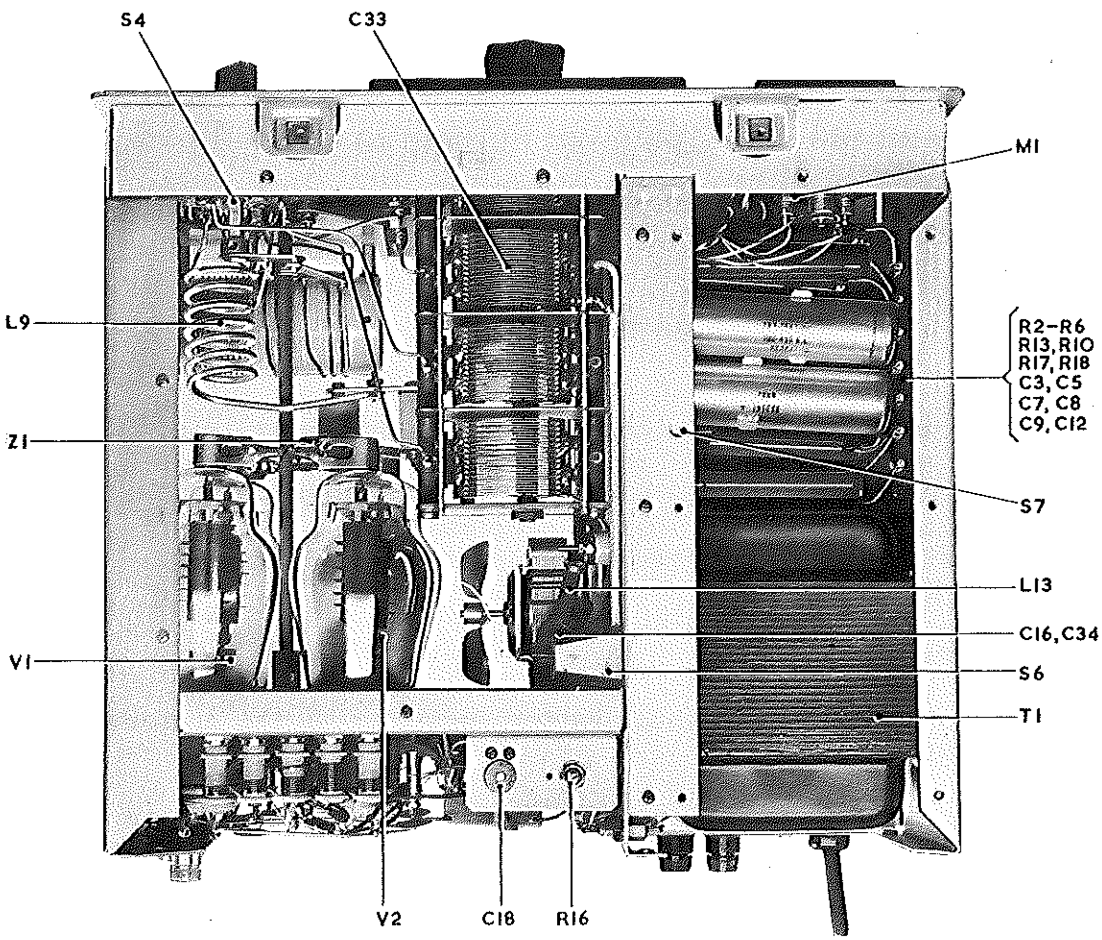

Fig. 1 R.F. linear amplifier 30L-1 - top view (cover removed)

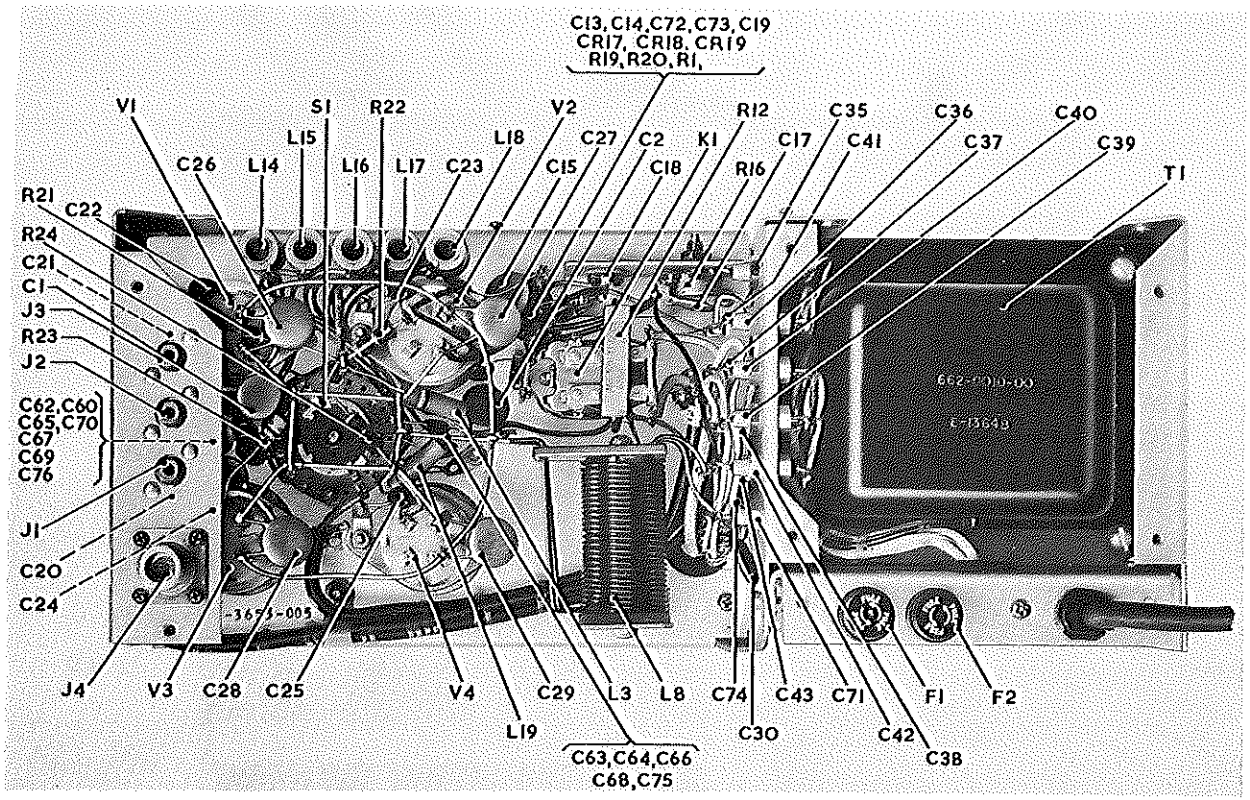

Fig. 2 R.F. linear amplifier 30L-1 - rear view (cover removed)

TABLE 1

Front panel controls

Control	Cct. ref.	Function
ON-OFF	S2	A.C. mains on-off switch
BAND	S1 and S4	Selects frequency operating band. S1 controls the input circuit and S4 the output circuit.
TUNING	C32	Tunes the power amplifier anode circuit.
LOADING	C33	Controls the shunt capacitance loading across the output.
METER	S3	Monitors power amplifier anode current, e.h.t. voltage and tuning.

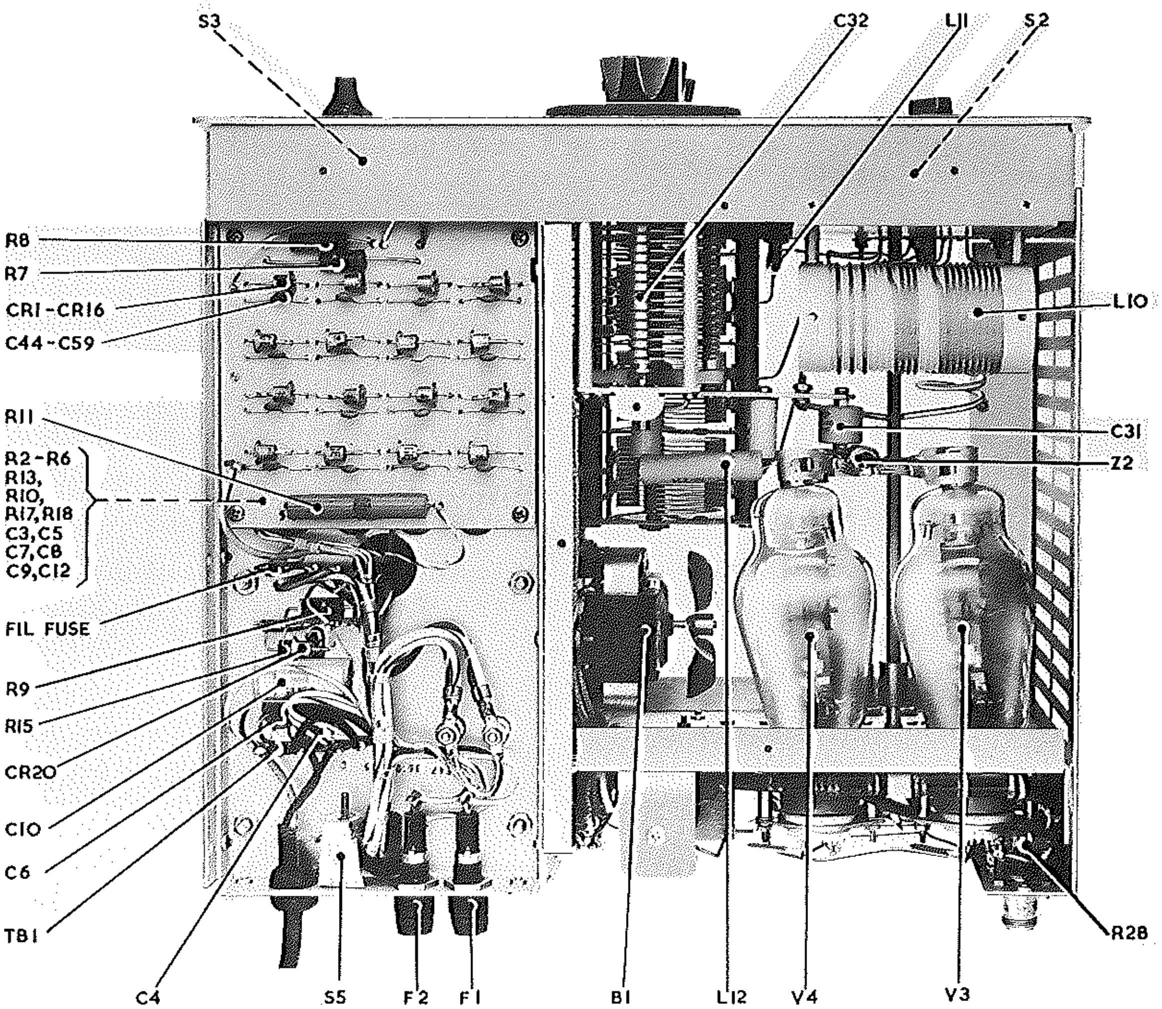

Fig. 3 R.F. linear amplifier 30L-1 - bottom view (cover removed)

DESCRIPTION

OUTLINE OF FUNCTION

5. The unit operates from a 115 V or 230 V, 50-60 Hz, single-phase power source and incorporates a power supply stage which generates the positive e.h.t., negative bias and valve heater supplies. The power supply stage uses semiconductor rectifiers and the power amplifier valves are directly heated so that the unit requires a negligible warm-up period and can be switched in or out of circuit at will.

6. The r.f. linear amplifier has the same frequency range as the KWM-2A transceiver (3.4 MHz to 30 MHz) covered in five bands corresponding to the five band groups of the KWM-2A. Table 2 details the band coverage.

7. A block diagram of the 30L-1 linear amplifier is shown in fig. 4.

TABLE 2

Frequency bands

BAND switch setting	Total possible coverage (MHz)	Corresponding KWM-2A bands
3·5	3·4 - 6·0	A
7·0	6·0 - 9·5	B
14	9·5 - 16·0	C
21	16·0 - 22·0	D
28	22·0 - 30·0	E

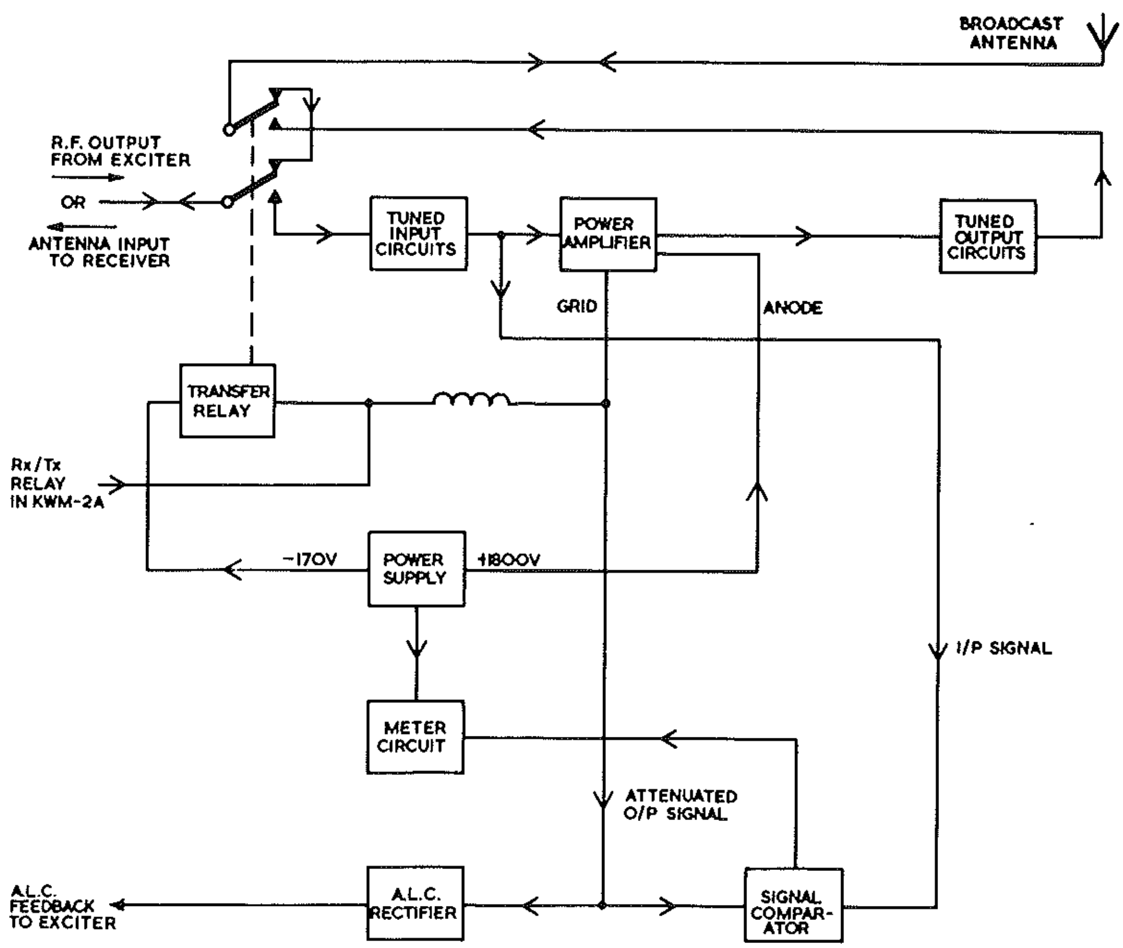

Fig. 4 R.F. linear amplifier 30L-1 - block diagram

8. The four power amplifier valves are parallel connected with the drive signal applied to their cathodes. The broad-band input circuit is tuned approximately to the drive signal frequency in order to present the correct impedance (50 ohms) to the KWM-2A output circuit, thereby improving operating efficiency and reducing distortion to a minimum. Tuning of the output circuit during the operational setting-up of the unit is simplified by a special metering circuit. This compares the input signal with an attenuated output signal and is set-up initially so that, for a given input signal level, the meter reads zero for optimum output (i.e. maximum undistorted output).

9. A transfer relay is incorporated in the 30L-1 so that, when the associated transceiver is in the receive function, the antenna is connected directly to the transceiver. Also in the receive function a negative bias voltage is applied to the grids of the power amplifier valves, thus biasing them beyond cut-off. In the transmit function, however, an earth from the antenna relay circuit in the transceiver removes the negative bias from the power amplifier valves and energizes the transfer relay. The transfer relay connects the output of the KWM-2A transceiver to the input circuit of the 30L-1 and connects the output circuit of the 30L-1 to the antenna.

10. In order to obtain a high average power output from the power amplifier while maintaining the peak power within the rated limits of the valves, an automatic load control (a.l.c.) circuit is provided in the 30L-1. This generates a negative feedback signal when the output exceeds a pre-determined level which is then applied to the transmitter r.f. amplifier and first i.f. amplifier in the transceiver KWM-2A.

FUNCTIONAL DESCRIPTION

11. The circuit diagram of the 30L-1 r.f. linear amplifier is illustrated in fig. 6 at the end of this chapter.

Power supply - input voltages

12. With the circuit connected for 230 V operation the a.c. mains is applied via F2 and one pole of S2 to the two primary windings of the mains transformer T1 which are series connected via F1 and the second pole of S2. With the circuit connected for 115 V operation, however, the two primary windings are effectively parallel connected, the a.c. mains being applied to one winding via F2 and one pole of S2 and to the other winding via F1 and the second pole of S2. The motor-driven blower, B1, for cooling the power amplifier valves is a 115 V device and is, therefore, connected in parallel with one of the T1 primary windings.

13. Earthing the 30L-1 is usually achieved by the use of a 3-wire mains supply. In the absence of a 3-wire supply, the earth terminal (GND) at the rear of the unit should be connected to the same earth line as the transceiver KWM-2A and its associated a.c. power supply PM-2.

Power supply - output voltages

14. Transformer T1 has three secondary windings the first of which provides a centre-tapped 6·3 V output for the power amplifier valve heaters and the

meter illumination lamp. The centre tap of this winding is connected to earth via a fuse consisting of a short length of No. 30 s.w.g. wire in order to protect the power amplifier valves from excessive anode current.

15. The +1800 V, e.h.t. supply is obtained from a voltage-doubler circuit consisting of CR1-CR16 and the associated capacitors C3, C5, C12 and C7, C8, C9, connected to the second of the three secondary windings. The rectifier diodes are series connected in groups of eight because of the reverse voltage limits on these components. The parallel capacitor networks C44-C51 and C52-C59 are included to equalize the reverse voltages impressed across the diodes and to protect against transient voltage surges. Similarly, the reservoir capacitors are series connected in groups of three because of the d.c. working voltage limits. Resistors R2-R7 and R13 provide a discharge path for the capacitors when the unit is switched off. R8 is a current monitoring resistor, the voltage developed across this being proportional to the total anode current taken by the power amplifier valves. The output from the voltage-doubler is approximately +1600 V on full load.

16. The output from the third secondary winding, one side of which is earthed, is applied via the surge-suppression resistor R9 to a half-wave rectifier CR20 providing the -170 V bias and relay operating supply. Smoothing is provided by C10, and R15 provides a discharge path for C10 when the unit is switched off.

Compartment screening and line filtering

17. Full electrostatic screening is provided between the power supply and power amplifier compartments of the 30L-1 and all connections, with the exception of the e.h.t. line, between these two compartments are made via feed-through capacitors. Voltage spikes appearing on the a.c. mains lines are suppressed by C4 and C6, and r.f. filtering of the 6·3 V and +1800 V supplies is provided by the decoupling circuits L8, C30, C74 and L13, C16, C34 respectively.

Safety interlocks

18. Three safety interlock switches, S5-S7, are incorporated in the +1800 V line which operate when the power supply or power amplifier compartment access covers are removed. Switches S5 and S7 are located in the power supply compartment and S6 in the power amplifier compartment. The removal of a cover, or covers, closes these switches and short-circuits the +1800 V line to earth via the current limiting resistors R17 and R18.

Receive-transmit relay

19. When the transceiver KWM-2A is in the receive function, the ANT RELAY input at J3 is open-circuit and the transfer relay, K1, is de-energized. Thus, J2 RF INPUT is connected direct to J4 RF OUTPUT via the two contacts of K1. Also , the -170 V supply is connected to the grid circuit of the power amplifier valves, V1-V4, via R12 in parallel with K1 operating coil, thereby maintaining these valves beyond the cut-off point.

20. When the transceiver KWM-2A is in the transmit function, however, an earth line is connected to J3 and K1 operating coil is connected across the -170 V supply via R28. With K1 energized, J2 is connected to the power amplifier input circuit and J4 is connected to the output circuit. Also, the -170 V bias is removed from the power amplifier grid circuit and replaced by the normal operating bias potential determined by a voltage-divider network, consisting of R28 in series with the parallel combination of R12 and K1 coil, across the -170 V supply.

21. Decoupling of the relay control line is provided by C21, C2 and L3.

Power amplifier

22. The low power (70 W) r.f. input at J2 is coupled to the cathodes of V1-V4 via one of five broadband tuned circuits selected by the switch S1. The five positions of S1 correspond to the five frequency bands detailed in Table 2. The principal purpose of the input circuits is to match the output of the transceiver KWM-2A to the input of the 30L-1. During the operational setting-up of the equipment the appropriate inductor, L14-L18, is adjusted to obtain optimum input impedance at the frequency in use.

23. The four power amplifier valves are parallel connected in a grounded-grid configuration. As these valves are directly heated the r.f. input signal is coupled via C1 to one side of the 6·3 V filament supply. An r.f. short-circuit is, therefore, placed across each valve filament by the capacitors C26 -C29. The operating grid bias potential (para. 20) is applied to each valve via a parasitic oscillation suppression resistor (e.g. R21 in the case of V1). Capacitors C22-C25 are part of the a.l.c. circuit described in para. 26.

24. The +1800 V supply is connected to the valve anodes via L12 and the two parasitic oscillation suppressors Z1 and Z2. These each consist of a small inductor (L4 and L5) wound on a 100-ohm resistor (R25 and R26). The tuned anode load for V1-V4, coupled via C31, consists of the variable capacitor C32 (TUNING) the tapped inductors L9 and L10, and the 4-gang variable capacitor C33 (LOADING). The circuit is tuned to the frequency in use by means of C32 and matched to the impedance of the output line and antenna (nominally 50 ohms) by means of C33. A d.c. path to earth for the output line is provided by the high impedance choke L11. A choke is used here in preference to a resistor since it is important that all components in the matching circuit should, ideally, be pure reactances. The r.f. power output is then dissipated entirely in the antenna this being the only pure resistance present in the output circuit.

25. The output circuit inductance is selected by one pole of the 5-position BAND switch S4. In the highest frequency band only part of L9 is connected in the circuit, the remainder of L9 and L10 being short-circuited, while in the lowest frequency band all of L9 and L10 are connected. Similarly, in the two high frequency bands only two sections of C33 are connected in the circuit, while in the centre band three sections are connected, and in the two low frequency bands all four sections are connected.

Automatic load control

26. The internal anode-to-grid capacitances of the power amplifier valves in conjunction with the grid-to-ground capacitors C22-C25 form capacitive voltage dividers. When the unit is transmitting, an r.f. voltage is developed across these dividers and, hence, across L3 via the grid-stopper resistors R21 - R24. This is then coupled via C72 to the a.l.c. rectifier circuit formed by CR19 and C73. The r.f. voltage is rectified and filtered to produce a negative d.c. control voltage at J1 (ALC) proportional to the modulation level. This is applied as negative feedback to the transceiver KWM-2A transmitter i.f. amplifier and r.f. amplifier stages to reduce the drive to the 30L-1. The time constants of the a.l.c. circuit are such that it has a fast rise time and a slow fall time. The load resistor to earth for CR19 is provided in the KWM-2A unit and the a.l.c. threshold level (i.e. the level at which CR19 starts to conduct on positive r.f. half-cycles) is determined by the positive bias on CR19 cathode. This is obtained from the voltage (approximately +16 V) developed across R7 in the +1800 V power supply circuit and is controlled by the preset potentiometer R16 and filtered by C17.

Meter circuit

27. The meter M1 on the 30L-1 front panel is a 200-0-500 μA, 190-ohm device and by means of the 3-position switch, S3, can be used to measure the total power amplifier valve anode current (D.C. AMPS) and the e.h.t. supply voltage (D.C. VOLTS) and can also be used as a tuning indicator (TUNE).

28. With S3 in the D.C. AMPS position the meter is connected in series with R10 across R8 in the +1800 V power supply circuit. The voltage developed across R8 is proportional to the total current drawn from the +1800 V supply by the four valves and in this arrangement full scale indication on the meter is 1000 mA.

29. With S3 in the D.C. VOLTS position the meter is connected in series with R11 directly across the e.h.t. supply. In this arrangement full scale indication on the meter is 2000 V.

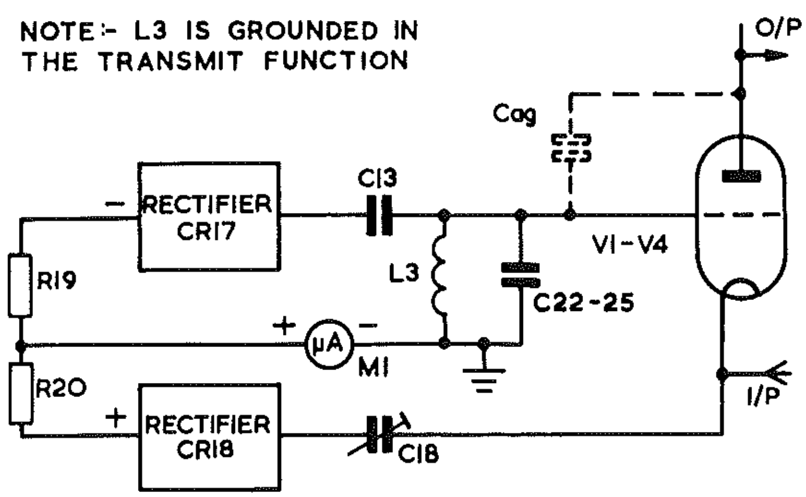

Fig. 5 Meter TUNE circuit - simplified block diagram

30. In the TUNE position of S3, the meter is connected across a comparator circuit, a simplified block diagram of which is shown in fig. 5. The r.f. input signal applied to the valve cathodes is coupled via the variable attenuating capacitor C18 to the rectifier circuit CR18 and C19. This produces across R20 and M1 a positive d.c. voltage proportional to the input signal level. Similarly, the r.f. voltage developed across L3 for a.l.c. purposes (para. 26) is coupled via C13 to the second rectifier circuit CR17 and C14. This produces across R19 and M1 a negative d.c. voltage proportional to the output signal level. Therefore, since R19 and R20 are of equal value, when the two d.c. voltages (positive and negative) are equal in amplitude the meter reads zero.

31. During the initial setting-up of the unit, J4 RF OUTPUT is terminated with a 50-ohm, non-inductive, dummy load and C32 and C33 are adjusted for maximum undistorted output. The drive input signal is then reduced to a known value (nominally 20 W) and C18 is adjusted to balance the comparator circuit, i.e. set the meter reading to zero. Therefore, during operational setting-up, the optimum tuning of the output circuit is obtained simply by inserting approximately 20W of drive signal from the KWM-2A and then alternately adjusting C32 and C33 for a meter reading of zero.

TABLE 3

Approved Service modifications and manufacturer's production changes

Cct. Ref. or location	Value/type old	Value/type new	Reason for modification or change	Effectivity
R28	39 ohm, ½ W	39 ohm, 2 W	Increase wattage rating	Mod No. A2595
TB1	-	-	Rewire of power supply input	S.B. No. 1*
F1, F2	6 A	8 A	Upgrade fuse rating	S.B. No. 1
CR19	1N252	1N458	To prevent backward deflection of meter	S.B. No. 2
R28 R12 R9	unknown 3 kohm 82 ohm	39 ohm 2 kohm 47 ohm	To lower operating temperature of p.a. valves	S.B. No. 3
C30 C74	added added	0·01 μF 0·01 μF	R.F. filter of heater 6·3 V supply	S.B. No. 3
R12/R14 R12	8·2 kohm added	deleted 3 kohm	Parallel combination of R12 and R14 deleted. Replaced by new R12	Not known

* S.B. (Service Bulletin)
Service Bulletins have been issued by the equipment manufacturer at indeterminate times and their purpose is to explain manufacturer's production changes. The effectivity of this equipment can only be determined by physical examination and by assessing the 'production build'.