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No. 6

A V.H.F./U.H.F. FIELD-STRENGTH RECORDING RECEIVER USING POST-DETECTOR SELECTIVITY

by R. V. HARVEY, B.Sc., G. F. NEWELL, A.M.I.E.E.
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FOREWORD

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A V.H.F./U.H.F. FIELD-STRENGTH RECORDING RECEIVER USING POST-DETECTOR SELECTIVITY

Summary: This Monograph describes a pre-tuned receiver for radio-frequency field-strength recording, designed specifically for propagation tests in Band III, Band IV and, ultimately, Band V.

Part I includes a discussion of the requirements for such a receiver and gives the general design and performance details. A separate signal-frequency unit is used for each frequency band, followed by an intermediate-frequency unit working at 10·7 Mc/s. The main features of the receiver are its high sensitivity and stability of calibration. Stability is particularly important as the recording is expected to remain accurate to within ±1 dB for periods of one month without attention, in spite of normal changes in valves, supply voltage and temperature.

The output signal/noise ratio has been increased by restricting the bandwidth of the circuits following the detector, enabling a higher sensitivity to be achieved. This feature is discussed in Part II and the results of tests compared with those of a theoretical analysis based on an idealized detector. It is shown that, over a large part of the range of measurable signals, the signal/noise ratio at the input to the detector is less than unity and its value determines the effective gain of the receiver. Consequently, the change of gain resulting from the insertion of fixed attenuators before the detector will vary with the output signal level and a new calibration will be required for each value of attenuation.

PART I
THE RECEIVER

1. Introduction

The receiver described was designed for recording field strength in two of the frequency bands allocated for television broadcasting in the United Kingdom, namely Band III (174–216 Mc/s) and Band IV (470–585 Mc/s). Several of these receivers are to be used for obtaining information on propagation conditions over long distances. Since only relatively low-power transmitters were available for these tests it was necessary to make the receiver sufficiently sensitive. To achieve a high output signal/noise ratio it was decided, at the outset, to modulate the transmissions at an audio frequency and take advantage of post-detector filtering, while at the same time making provision for recording the field strength of other signals which might not be modulated in this manner.

The investigation of propagation conditions requires accurate and continuous records of field strength for long periods at sites where it may not be convenient for skilled engineers to inspect apparatus more frequently than once a month. The reliability and gain stability of a receiver for this purpose must therefore be high, so that valuable recording time is not wasted either through obvious failure or unsuspected errors.

The following specification was eventually evolved.

2. Specification

The receiver must be able to record the field strength of special transmissions on any frequency in the ranges 174–216 Mc/s and 470–585 Mc/s, with provision for later adaptation to operate up to 960 Mc/s. These transmissions will be square-wave modulated to a depth of 100 per cent at 1000 c/s ±1 c/s and the carrier frequency controlled to within 0·01 per cent of the nominal frequency. Continuous tuning is not necessary and the signal-frequency circuits of the receiver may be fixed before installation.

For normal operation, the output of the receiver shall be proportional to the logarithm of the input signal, but facilities shall be provided for a linear relationship if required. The recording error, referred to the input, shall be less than ± 1 dB during a period of up to one month after inspection and calibration, provided that the supply voltage remains within the range 190–250 V r.m.s. The useful gain shall be limited only by fluctuation noise and the noise factor be as low as is reasonably possible, with a maximum of 15 dB. Provision shall also be made for recording the field strength of unmodulated transmissions.

3. Discussion of Specification

In general, a receiver can be no more reliable than its least reliable component, and the designer, while choosing the best component for a particular purpose, must be aware of its limitations and ensure that its reliability is not impaired by the conditions in which it operates. Certain changes in the values of components must however be tolerated; these arise both from ageing and from changes in supply voltage and temperature. The effect of these changes can often be minimized by care and ingenuity of design. Negative feedback, for example, can provide a high degree of gain stability in some parts of the receiver, so permitting a greater variation in those parts which cannot conveniently be stabilized in this way.

The effect of frequency drift can be overcome by ensuring that the gain of the receiver is constant over a range of frequencies equal to the sum of the maximum drifts expected at the transmitter and the receiver.

To obtain a logarithmic amplitude characteristic, the receiver gain must vary as a function of the input signal. This may be achieved by using one stage of variable gain, but negative feedback cannot then be used to protect the stage from the changes mentioned above. There are also difficulties in applying negative feedback to the signal frequency and mixer stages. Hence, though the effect of supply voltage changes can be minimized by a voltage stabilizer, some gain changes due to ageing and temperature may remain unless suitable precautions are taken.

4. Brief Outline of Receiver Design

In order to obtain the specified performance, a superheterodyne receiver was designed having two interchangeable signal-frequency (s.f.) units and a main unit containing the intermediate-frequency (i.f.) amplifier, detector and post-detector circuits, negative feedback being used wherever convenient. The main unit determines the bandwidth of the receiver, the gain being sensibly constant over a range of ± 200 kc/s to accommodate the maximum frequency drift of $0 \cdot 01$ per cent at both transmitter and receiver at the maximum frequency of 960 Mc/s.

An intermediate frequency of $10 \cdot 7$ Mc/s was selected after consideration of the following points. A higher frequency would have simplified image rejection but would also have increased difficulties in development, particularly in obtaining the desired degree of negative feedback. While no special techniques are necessary to obtain the required coil Q-factor of 100 at about 10 Mc/s, a higher frequency would demand higher values of Q for the same bandwidth, and consequently greater stability with temperature. The actual frequency of $10 \cdot 7$ Mc/s, rather than any other in this region, was chosen in the hope that it might eventually become cleared of transmissions, as it is a popular intermediate frequency for v.h.f. broadcast receivers in both Europe and America.

The i.f. amplification is provided by three cascaded three-valve amplifiers, the gain of the second being controlled by a direct voltage derived either from a manual control, or from the output of the receiver; in the latter case a close approximation to the required logarithmic law is achieved. These amplifiers are separated by band-pass filters giving a frequency response constant to within $\pm 0 \cdot 2$ dB over the range ± 200 kc/s and falling rapidly outside this range.

The output of either of two separate amplifiers can be selected by a switch to provide post-detector amplification, the choice depending on the type of transmission to be recorded.

For the special test transmissions mentioned earlier, the recorded output is taken from an audio frequency (a.f.) amplifier and detector. This amplifier is preceded by an a.f. filter having an energy bandwidth of about 30 c/s at the

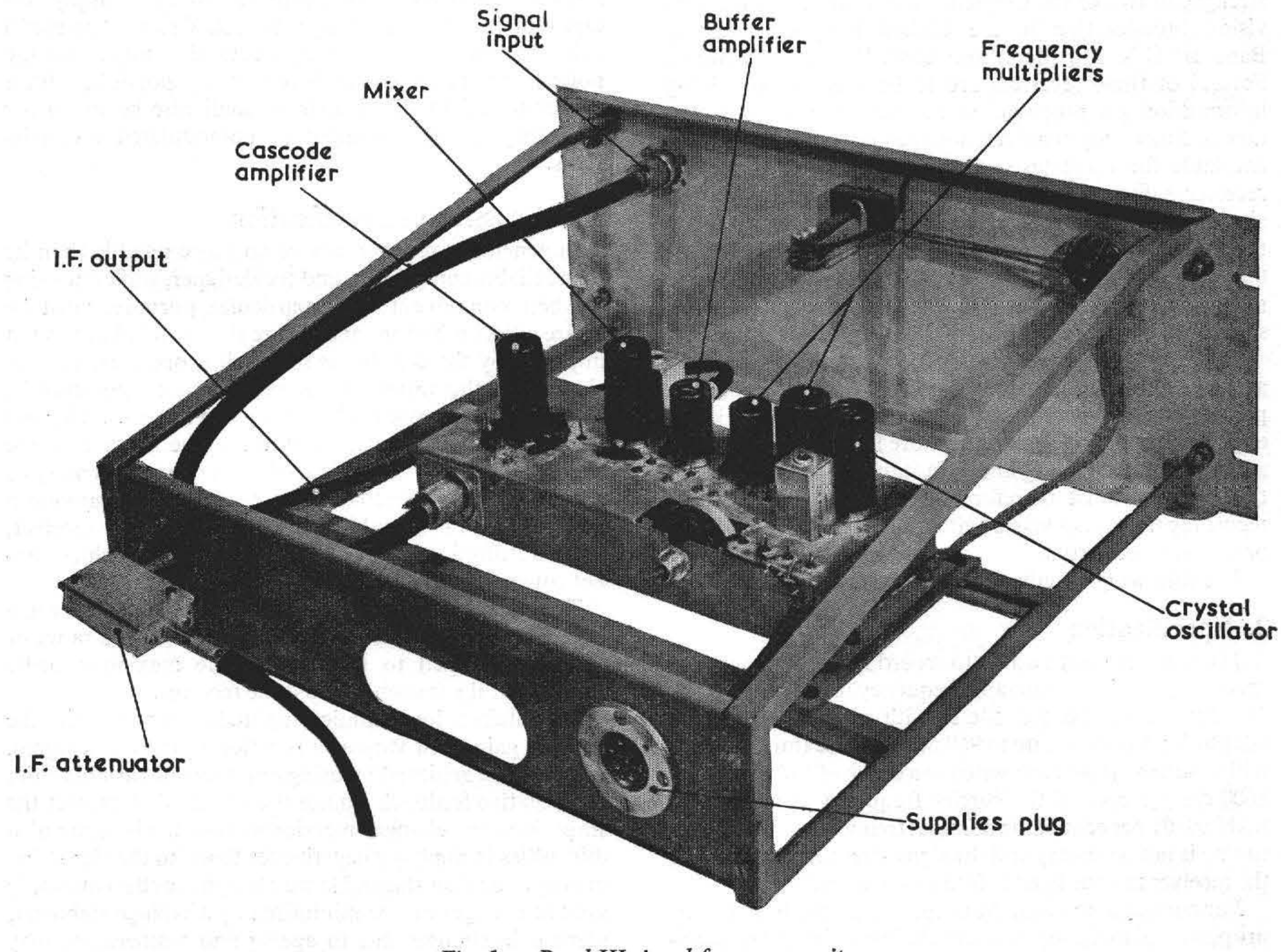

Fig. 1 — Band III signal-frequency unit

modulation frequency of 1000 c/s, resulting in an improved output signal/noise ratio when receiving the special transmissions. For other signals, such as C.W. or programme, the output is taken from a d.c. amplifier, the mean signal level being recorded with negligible error. The receiver is not suitable for receiving transmissions with pulse modulation or steady sinusoidal modulation to a depth greater than 50 per cent, though it can be adapted to record signals which are 100 per cent square wave modulated at frequencies other than 1000 c/s by changing the a.f. filter. Either of the two amplifiers mentioned above can supply an output of 0–1 mA d.c. to a recorder having an impedance of 3000 ohms.

A conventional d.c. power unit provides stabilized anode, screen and bias voltages; the r.m.s. mains voltage supplied to the whole receiver is stabilized to within ±1 per cent by a special unit designed for this purpose.

5. Signal-Frequency Units

5.1. Band III s.f. unit (174–216 Mc/s)

A photograph of the unit is shown in Fig. 1. The circuit diagram, Fig. 2, shows a double triode type 12AT7 as a cascode s.f. amplifier connected, by a transitionally-coupled pair of tuned circuits, to one section of a second 12AT7 operating as a triode mixer. Coils of silver wire are used for the s.f. circuits, which are tuned by slugs of iron dust or silver-plated copper. The nominal input impedance is 75 ohms and the voltage standing-wave ratio is better than 0·9; the output impedance at 10·7 Mc/s is 470 ohms. Provision is made for inserting attenuators between the s.f. and i.f. units if required.

The local oscillator section of the unit comprises a 12AT7 crystal-controlled oscillator and quadrupler, a pentode type EF95 used as a trebler and a second EF95 buffer amplifier, its output being coupled to the grid of the mixer by a variable capacitance. The tuning range of the oscillator and multipliers is sufficient to allow the final frequency to be set either above or below that of the signal; this facility may be required to avoid interference resulting from spurious responses.

5.2. Band IV s.f. unit (470–585 Mc/s)

The unit comprises a tuned input circuit followed by a crystal mixer, and a local oscillator. The mechanical arrangement may be seen from Fig. 3; there are two concentric-line resonators and a small chassis containing the crystal oscillator and multiplier stages. The circuit diagram is shown in Fig. 4.

The aerial input is loop-coupled to the first concentric-line resonator which is tuned to the signal frequency. A silicon mixer crystal, mounted at the base of the second resonator, is coupled to the local oscillator and the signal by a loop common to both resonators. The crystal mounting embodies a 100pF r.f. bypass condenser which is

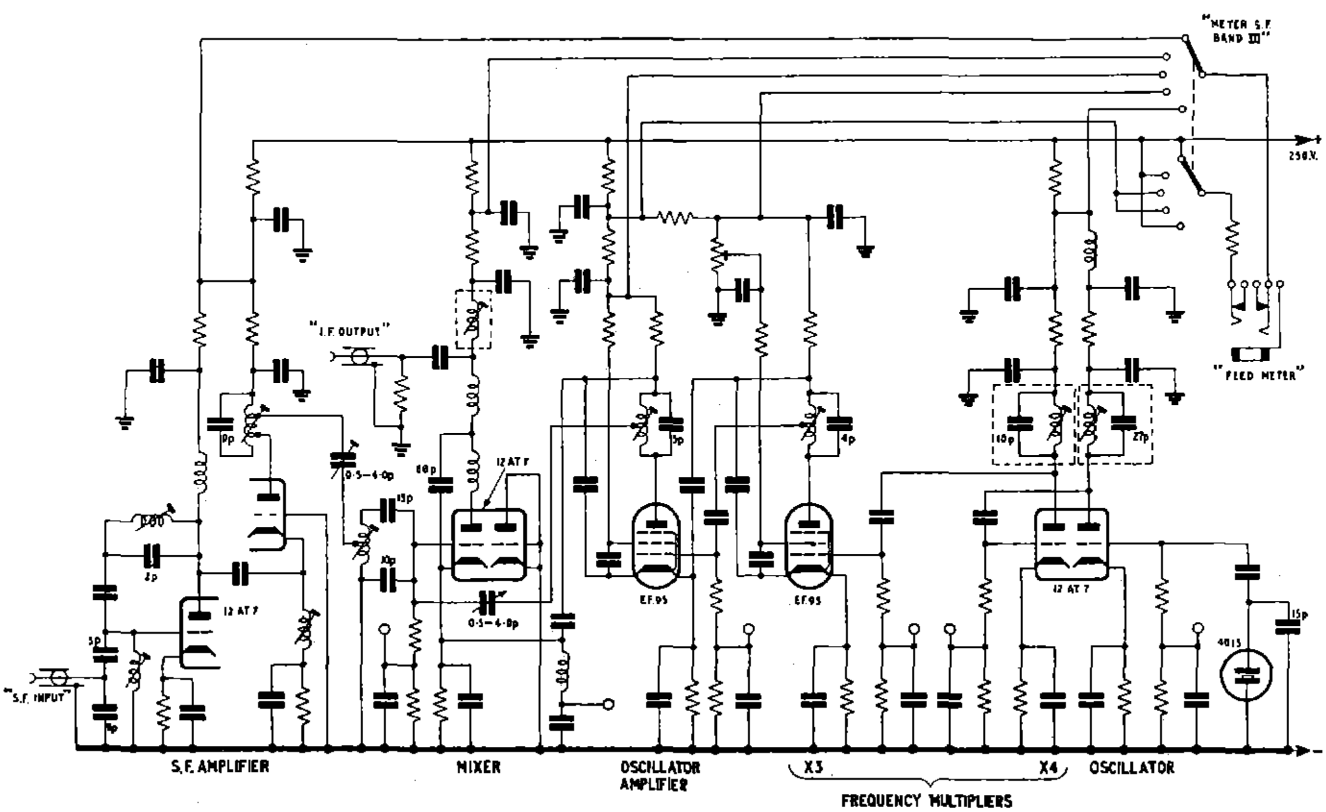

Fig. 2 — Band III signal-frequency unit: circuit diagram

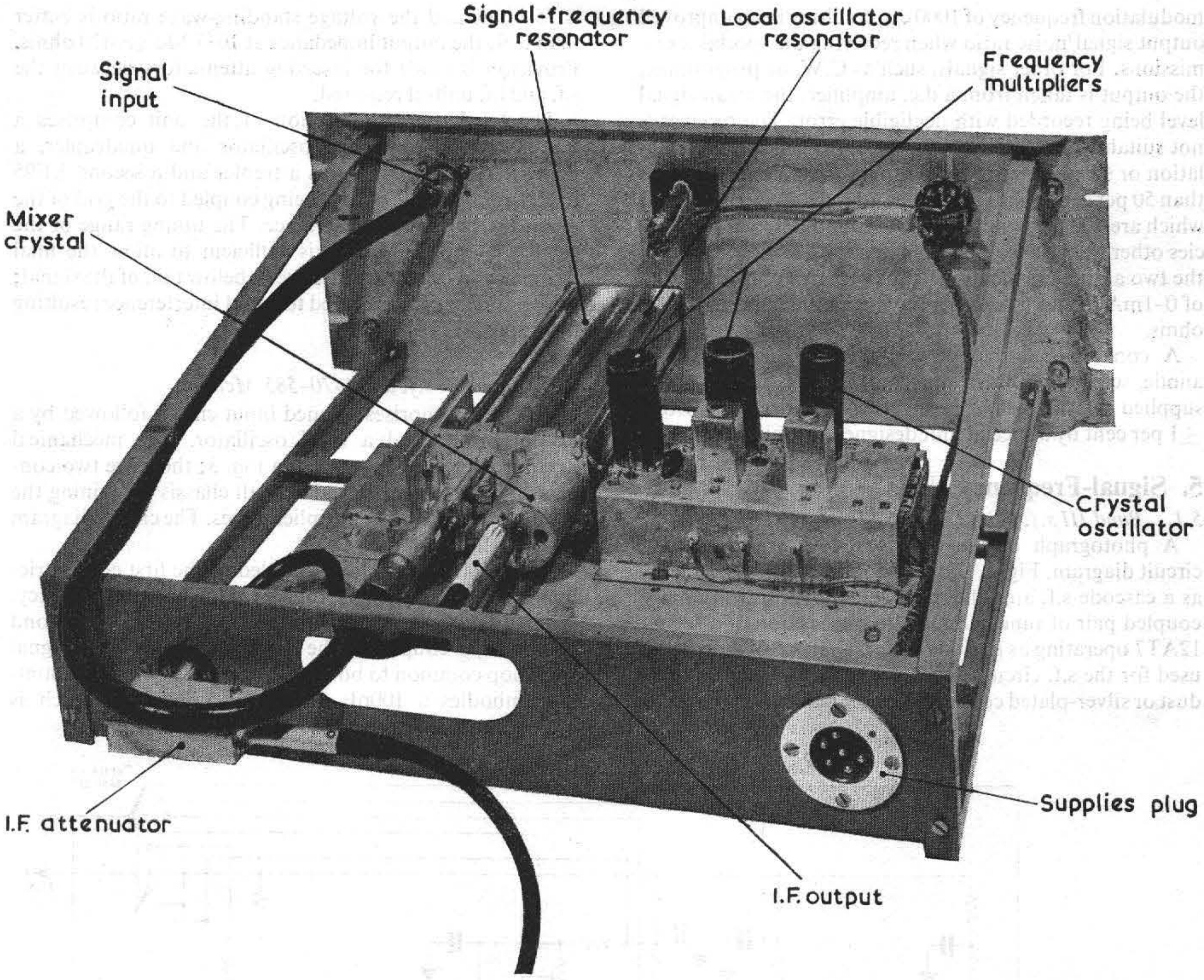

Fig. 3 — Band IV signal-frequency unit

tuned by a shunt inductance to resonate at the intermediate frequency; the i.f. output developed across this circuit is fed to the i.f. unit by a short length of 100 ohm cable.

The s.f. resonator has an unloaded Q-factor of about 3000 which is reduced to about 300 by the loading of the source and the mixer, giving a s.f. response which does not materially affect the overall response of the receiver. The image rejection is about 28 dB, which is considered adequate in view of the facility of choosing a final oscillator frequency either above or below that of the signal.

The local oscillator section employs two double triodes type 12AT7, the first as a crystal-controlled oscillator and quadrupler, and the second as two cascaded treblers, the final output being coupled by a loop into the second concentric line, resonant at a frequency 36 times that of the crystal oscillator. The high Q-factor of this resonator affords adequate rejection of other harmonics of the crystal oscillator, and so helps to reduce interference resulting from spurious responses.

Each resonator is tuned by a pre-set adjustment of the length of the inner conductor, this being determined by the position of a rod of Nilo 36, an alloy having a low thermal coefficient of expansion.

As in the Band III unit, the input impedance is arranged to be nominally 75 ohms, with an input voltage standing-wave ratio at mid-band of better than 0·9, while the output impedance at 10·7 Mc/s is approximately 470 ohms. Provision is made for the insertion of attenuators between the s.f. and i.f. units if required.

6. Intermediate-frequency Unit

Each of the i.f. amplifiers and filters is built on a separate chassis; these are arranged in cascade on an insulating frame with the post-detector circuits in the centre, as shown in Fig. 5. The circuit diagram is shown in Fig. 6.

6.1. Input circuit and i.f. amplifier No. 1

The input transformer has a Q-ratio of 16 and a transitional coupling-coefficient of 0·1, giving a bandwidth of about 1 M/cs at -3 dB, and a level response in the passband of the receiver. The impedance presented by this circuit to the first amplifier is about 10,000 ohms.

The amplifier incorporates three cascaded pentodes type Z77 with overall negative feedback, the feedback voltage is derived from the cathode circuit of the third valve and applied, through a phase-correcting circuit, to the cathode of the first valve. As this valve is fed from a high source impedance, the input impedance of the valve itself must be made high to avoid loss of feedback; this necessitates tuning out the grid-cathode capacitance by means of an inductance. The reduction in the gain of the amplifier due to the application of feedback is 32 dB at the intermediate frequency, and the resultant gain can be expressed as a transfer conductance of 40 mA/V. Only this transfer conductance is stabilized by the negative feedback loop, which does not include the anode load circuit of the last stage. The components of this circuit, therefore, in addition to those of the feedback loop, must be particularly stable with time and temperature.

The output is delivered to the next band-pass filter, which has an input impedance of 10,000 ohms, either directly or through a built-in attenuator which may be inserted by changing two soldered connections. This attenuator introduces a loss of about 15 dB and is normally inserted only when operating in Band III, in order to compensate for the higher gain of the Band III s.f. unit without impairing the overall noise factor.

6.2. I.F. amplifier No. 2

This amplifier contains the stage of variable gain required to achieve a logarithmic amplitude characteristic, the valve being a variable-mu pentode type 6BA6. As the use of overall negative feedback would defeat the effect of gain control, this valve is not included in the feedback circuit; instead, a feedback voltage is derived from the output of another 6BA6, in parallel with the controlled valve, but operating at fixed gain. In this way, although any change in the gain of either valve separately will have a direct effect

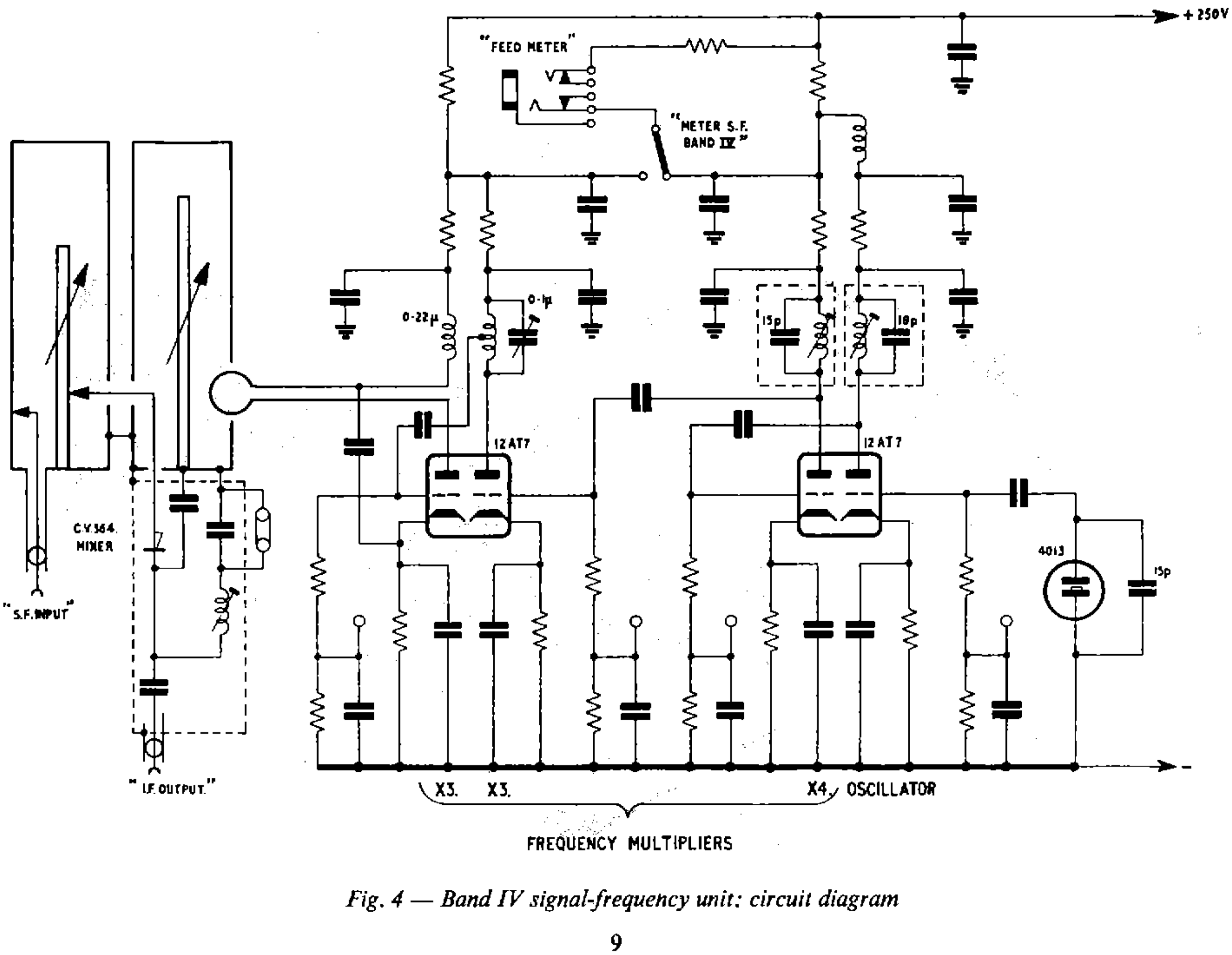

Fig. 4 — Band IV signal-frequency unit: circuit diagram

on the overall gain, the system may be made to compensate for any changes that affect both valves simultaneously.

As shown in Fig. 6, the amplifier comprises a pentode type Z77 driving two valves type 6BA6 with their grids in parallel at the intermediate frequency. The first 6BA6 grid is maintained at a pre-set direct voltage, while its cathode circuit provides an i.f. feedback voltage which is applied to the Z77 grid through the input circuit, the reduction in gain due to feedback being about 28 dB. The grid voltage of the second 6BA6 is derived either from the rectified output of the receiver, as an a.g.c. voltage, or from a manually operated potential divider; this valve forms part of the main amplifier chain, and its output is developed across an anode load of 1000 ohms, this being the input impedance of the next i.f. filter. The overall voltage gain of the amplifier can be varied between the limits of 1·5 and 0·02.

6.3. I.F. amplifier No. 3 and i.f. detector

As can be seen from Fig. 6, the circuit arrangement of this amplifier is identical with that of amplifier No. 1, but in this case the amount of feedback applied is 26 dB and the voltage gain is 2500. The final i.f. voltage appears across a 2500 ohm resistor in the anode circuit of the third valve and is then rectified by a diode, the signal passing to the post-detector stages.

6.4. I.F. filters

The selectivity of the receiver is principally determined by the two filters connected between the i.f. amplifiers, shown in Fig. 6 as Filters Nos. 2 and 3. The filter design is based on a confluent band-pass π-section, using coil Q-factors of about 100; the input and output circuits are coupled to the centre circuit by mutual inductance both to provide d.c. isolation and to allow convenient component values to be used. Since the amplifiers have a relatively high input impedance, it has been possible to improve the filter response by underloading, i.e. terminating the filter by a resistance higher than the mid-band iterative impedance. The resultant response of each filter is uniform to within $\pm 0 \cdot 1$ dB over the pass-band of ± 200 kc/s, centred on $10 \cdot 7$ Mc/s, falls to -10 dB at ± 400 kc/s and -21 dB at ± 600 kc/s.

Filters Nos. 2 and 3 have mid-band iterative impedances of 10,000 ohms and 1000 ohms respectively; the lower impedance of filter No. 3 is necessary both to minimize break-through across the grid-anode capacitance of the

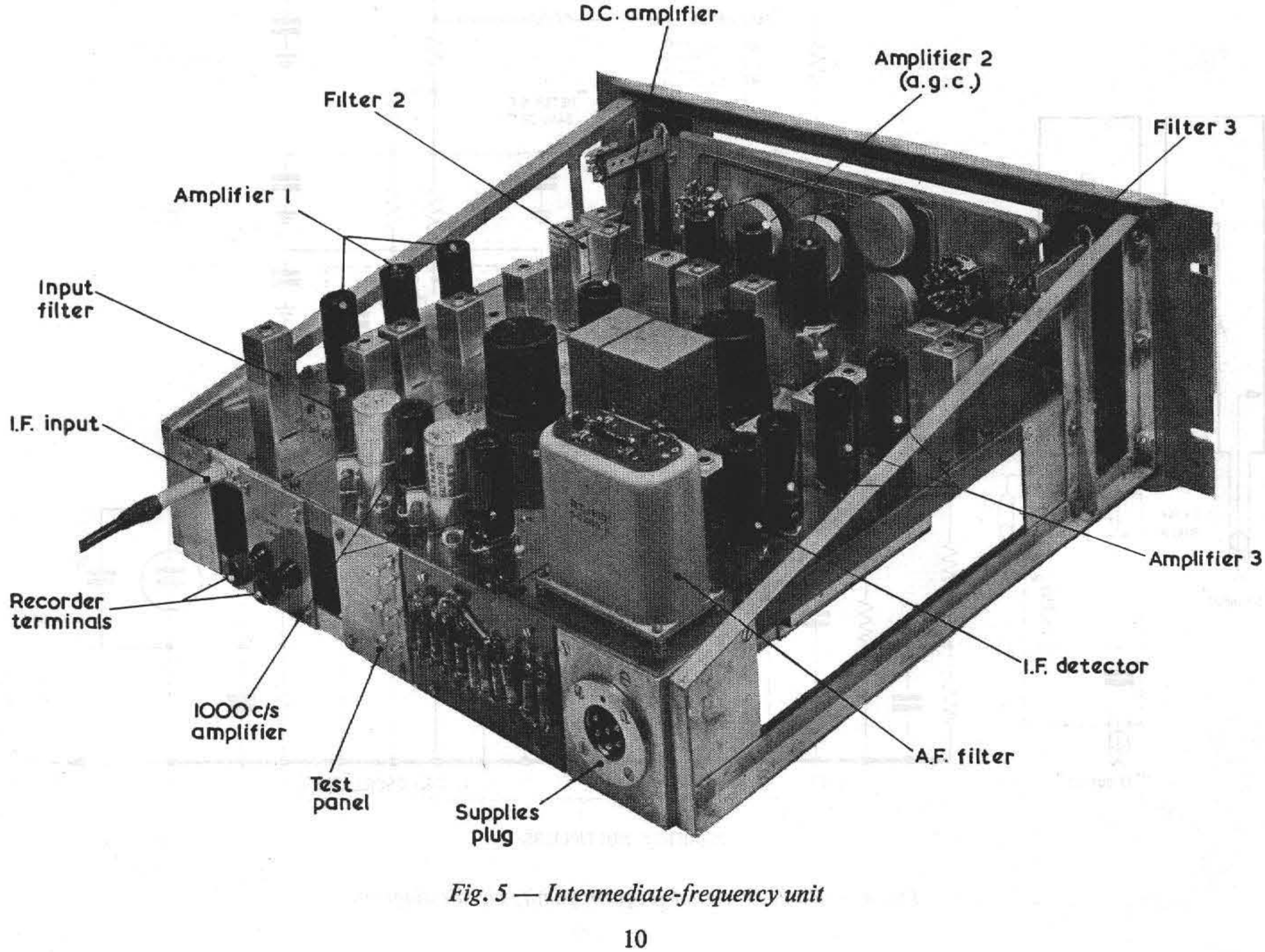

Fig. 5 — Intermediate-frequency unit

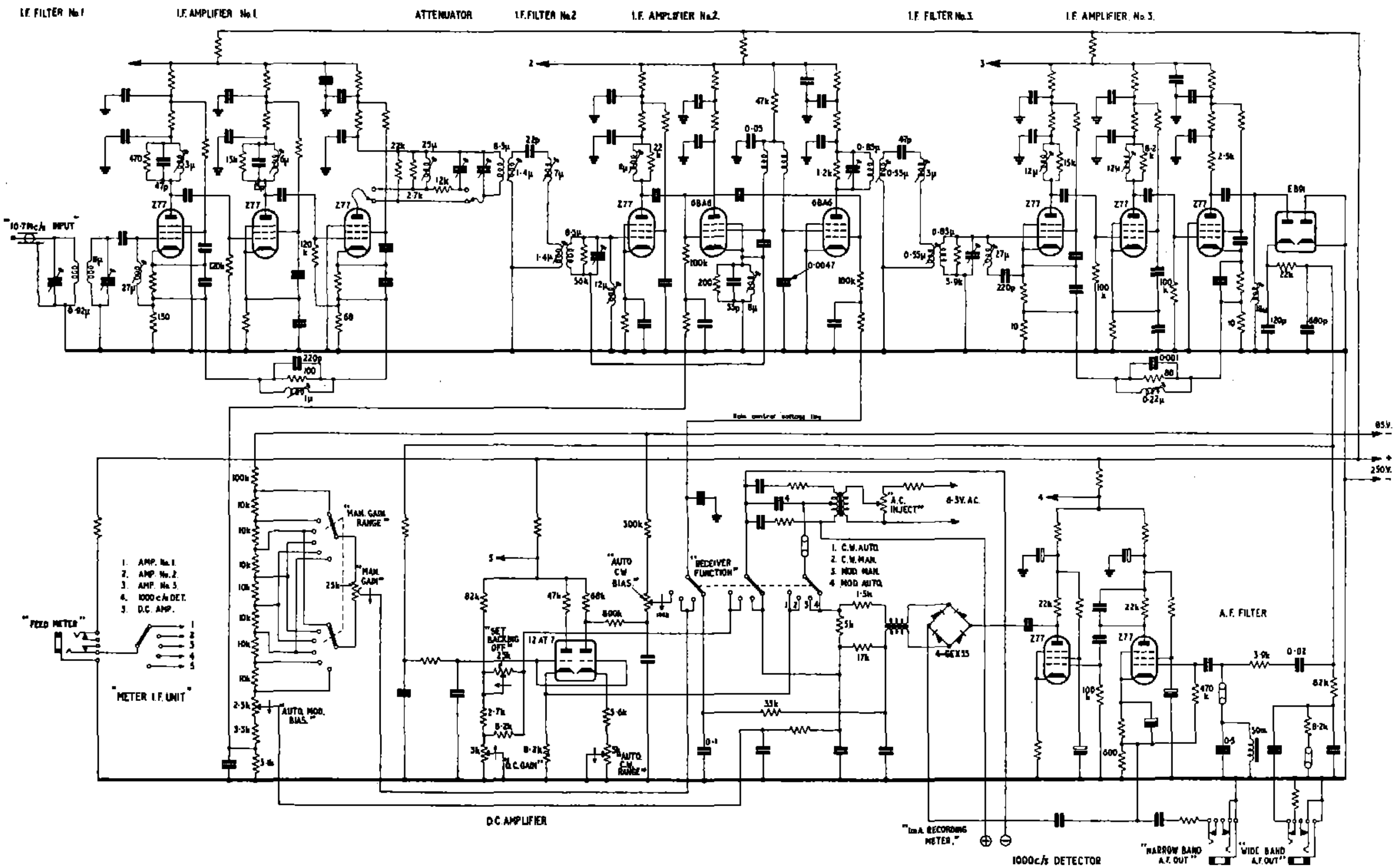

Fig. 6 — Intermediate-frequency unit: circuit diagram

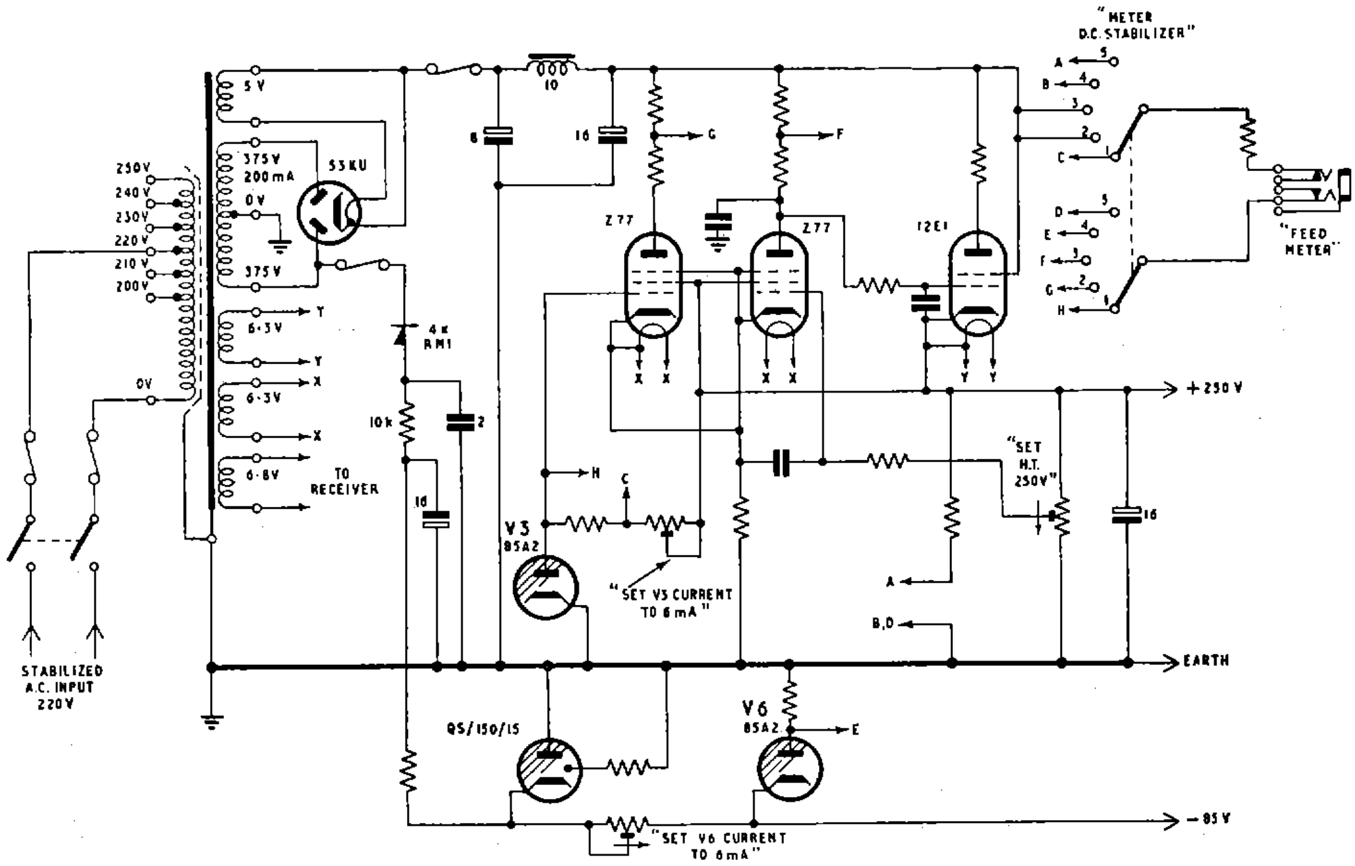

Fig. 7 — Stabilized d.c. power unit: circuit diagram

preceding 6BA6, when operating at low gain, and to simplify the application of negative feedback to amplifier No. 3.

6.5. *Post-detector stages*

The output from the i.f. detector is separated into its a.f. and d.c. components which are fed to separate amplifiers shown in the lower half of Fig. 6. The switches selecting the operation and output of the receiver are described in Section 8.

6.5.1. *1000 c/s filter, amplifier and detector*

The a.f. component of the rectified i.f. output is first passed through a narrow-band filter, consisting of a resistor in series with a parallel tuned circuit resonant at 1000 c/s, having a bandwidth of ± 12 c/s at -3 dB. This filter passes the fundamental component of the modulation on the special test transmissions described in Section 2 and, in so doing, gives an a.f. signal/noise ratio of 0 dB for an i.f. signal/noise ratio of -19 dB. The output is amplified by two Z77 pentodes in cascade and rectified by a bridge circuit using four GEX55 germanium diodes, while the a.f. current through the bridge is returned to the cathode of the first Z77, giving about 28 dB of negative feedback round the whole circuit.

The d.c. load, consisting of three resistors in series, is connected across the bridge output terminals through a bifilar choke, which allows the a.f. current to flow freely between the bridge terminals while presenting a high impedance to current flowing from the bridge to earth. In this way, the load may be connected to earth at any point without disturbing the feedback at 1000 c/s. A negative a.g.c. voltage of 0–30 V is provided by this d.c. load and part of the load current provides the 0–1 mA output for the recording meter.

6.5.2. *D.C. amplifier*

The d.c. component of the rectified i.f. output is passed through a low-pass filter to remove any modulation frequencies and applied, as a positive voltage, to both grids of a 12AT7 double triode. One section of this valve supplies 0–1 mA for the recording meter, connected between its cathode and a point at a pre-set positive voltage, while the other section supplies the a.g.c. voltage. The anode voltage of the second section falls as the signal increases, the negative a.g.c. voltage of 0–30 V being obtained from a point on a potential divider connected between its anode and a stabilized negative supply.

6.5.3. *A.C. injector*

With the conventional type of 3000 ohm d.c. pen recorder, some of the fine structure of the record is distorted due to static friction, and it is sometimes desirable to improve the response by adding a small alternating current to the recorded direct current. This facility is provided by a bridge circuit using a centre-tapped transformer across the output to the recording meter, as shown in Fig. 6. The injected current can be varied up to a maximum of 0·1 mA r.m.s. for this type of recorder, at the supply frequency of 50 c/s. The bridge is balanced by a dummy load so that the a.c. is not fed back into the d.c. source.

7. Power Supplies

7.1. Stabilized d.c. power unit

The circuit (shown in Fig. 7) is that of a conventional stabilized d.c. power supply using a full-wave rectifier followed by a condenser input filter and a series regulator valve. The impedance of the series valve is controlled by a two-stage d.c. amplifier to compensate for changes in the output voltage, the voltage reference being provided by a neon stabilizer type 85A2.

The unit will give a constant 250 V output at up to 150 mA for all input voltages from 190–250 V r.m.s. In addition, a metal rectifier and neon stabilizers provide a constant 85 V negative supply for the gain-control circuits of the receiver.

7.2. A.C. stabilizer

This unit will accept a mains input at any voltage between 190 V and 250 V r.m.s. at any frequency between 40 c/s and 60 c/s. It will deliver an output power of up to 1·7 kW at 220 V r.m.s. ± 1 per cent, the voltage difference of up to ± 30 V being added or subtracted automatically using a 'buck-and-boost' transformer.

As shown in Fig. 8, the secondary of this transformer is connected between the live terminals of the input plug and output socket, the neutral terminals being directly connected. The primary winding is supplied from the slider of a variable auto-transformer fed from the input, the slider being adjusted by a small motor until the output returns to 220 V. The rate at which the unit corrects for a change of input is approximately 3 volts per second.

An error voltage is derived from the output of the unit by means of a lamp bridge which is adjusted to balance at 50 c/s when the output voltage is 220 V r.m.s. When the balance is disturbed, a signal proportional to the error voltage is passed through an amplifier, a limiter, a 50 c/s filter and a 'flip-flop' circuit, finally arriving as a train of pulses

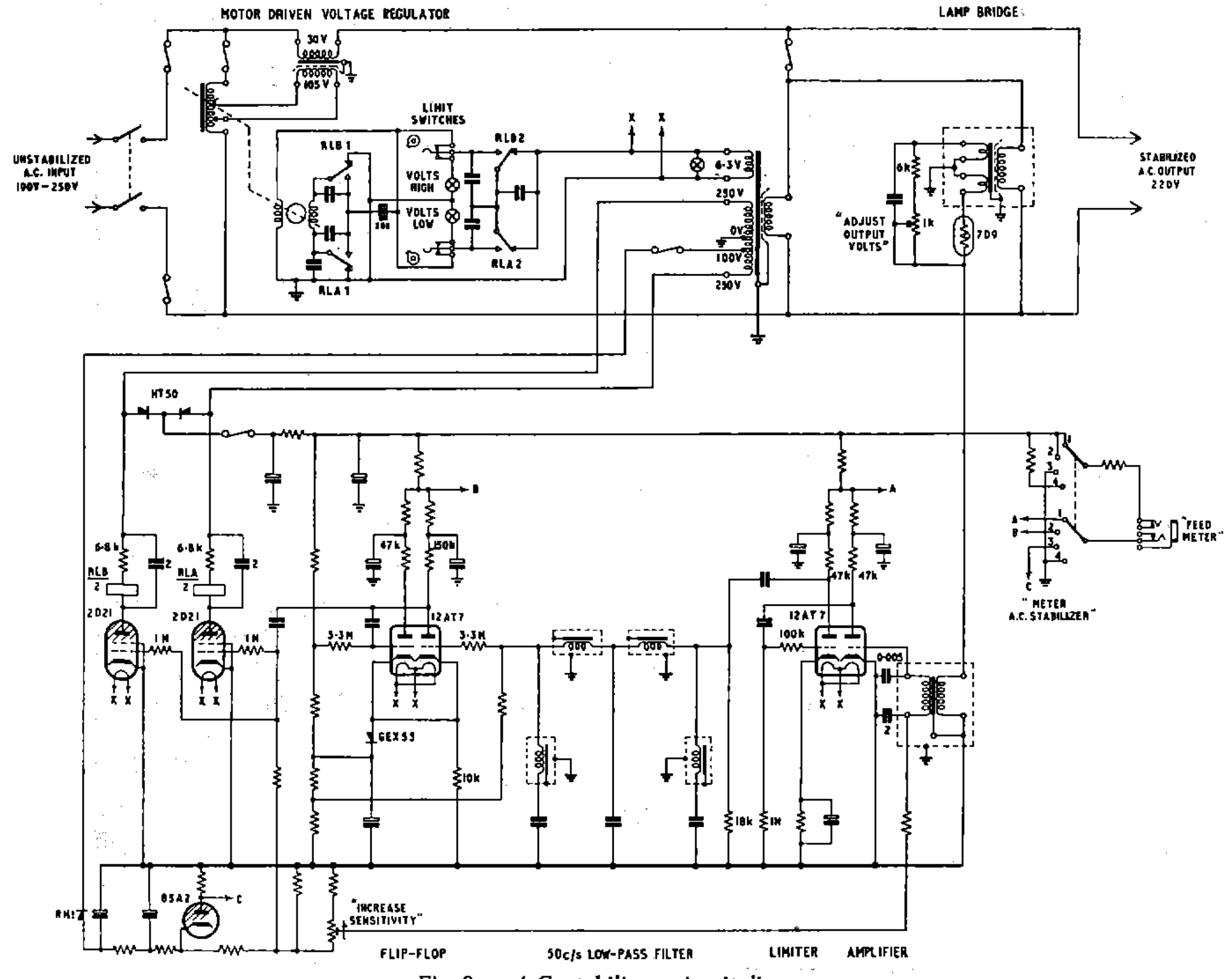

Fig. 8 — A.C. stabilizer: circuit diagram

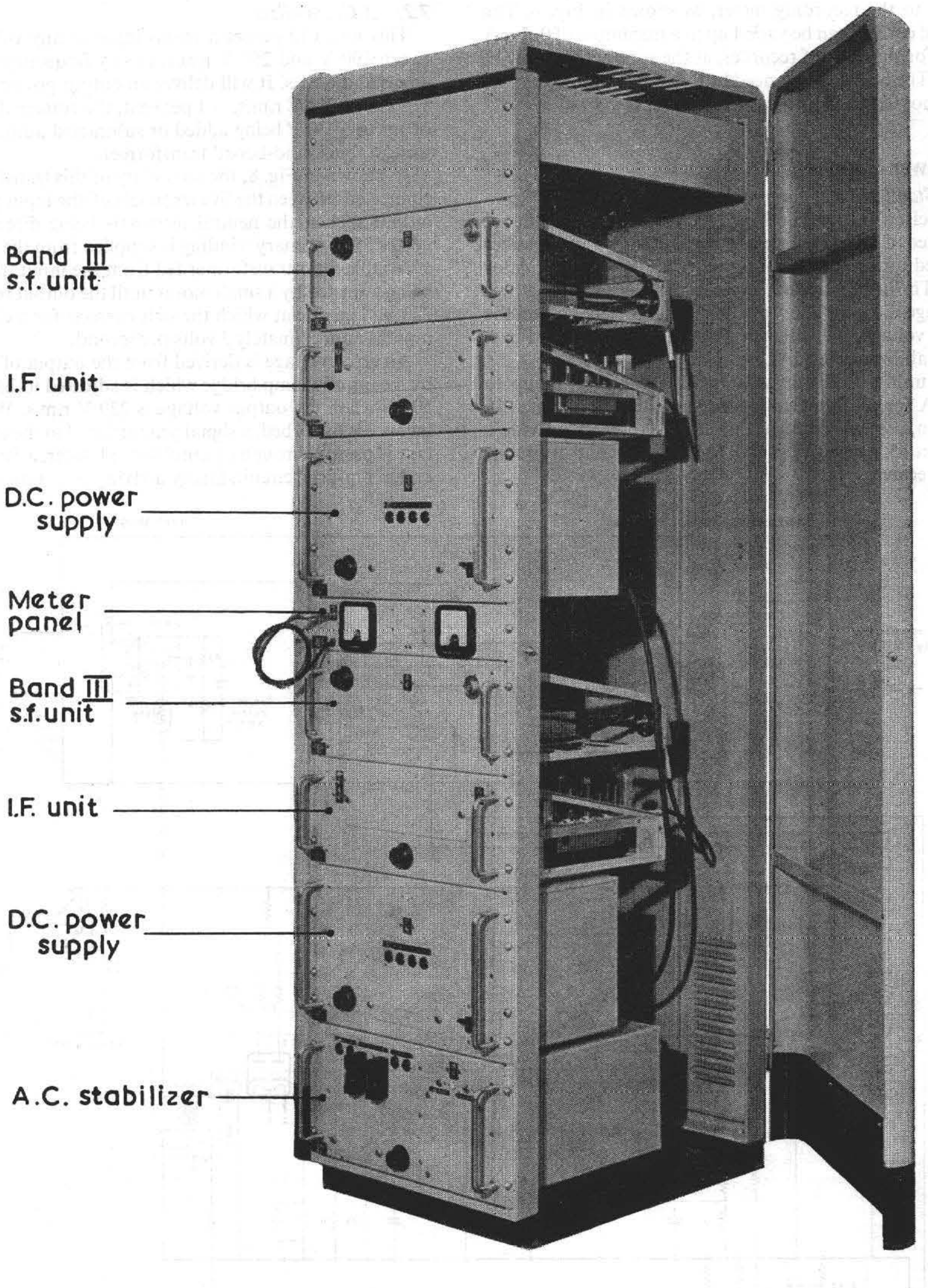

Fig. 9 — Cabinet containing two complete Band III receivers

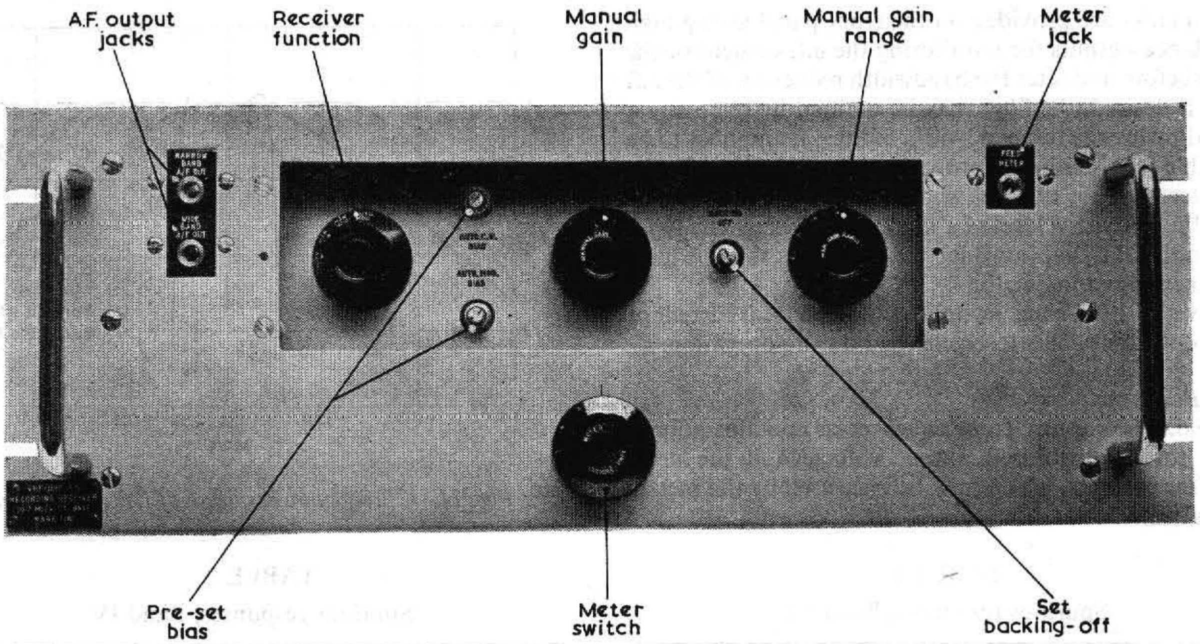

Fig. 10 — Intermediate-frequency unit: front panel controls

at the grids of two thyratrons, their anodes being fed in antiphase at 50 c/s through two relays controlling the motor. In this way, the sense of the error voltage determines which thyratron conducts and the direction in which the motor adjusts the correcting voltage.

Since it is the r.m.s. value of the output voltage which is stabilized by this unit, the cathode temperature of the valves in the receiver which it supplies will remain sensibly independent of the mains input for a wide range of voltage, frequency and waveform.

8. Operation, Controls and Metering

It is intended that the receivers be mounted in pairs in steel cabinets approximately 6 ft (2 m) high and 2 ft (0·6 m) square, each receiver consisting of one s.f. unit, one i.f. unit and one d.c. power unit; the two receivers are fed from a common a.c. stabilizer. There is also a panel containing two meters, mounted between the receivers. One of these meters gives a permanent indication of the stabilized supply voltage and the other may be connected by a flying plug lead to any of the units for metering purposes.

The photograph in Fig. 9 shows the arrangement of the units in the cabinet. The principal controls and some of the pre-set adjustments are recessed in the front panels and are normally covered by detachable metal plates. The functions of these controls and other features are described in the following sub-sections.

A major adjustment of the amplitude and frequency characteristics of the receiver is regarded as a laboratory operation and is not intended to be carried out in the field.

8.1. *Signal-frequency units*

The principal controls are all pre-set internally and locked on test. Other internal features include test points for checking the s.f. response and the grid drive of the multipliers. The anode currents of all valves may be checked during operation using the flexible plug lead and a rotary panel-switch.

8.2. *Intermediate-frequency unit*

The function of the receiver is controlled by the four-position switch on the left of the front panel recess, shown in Fig. 10. This determines whether the quantity recorded is the mean carrier level (C.W.) or the modulation of the special test transmissions (Mod.) and whether the amplitude characteristic is linear (Man.) or logarithmic (Auto). On the C.W. and Mod. positions, the recorder current is taken, respectively, from the d.c. amplifier and the 1000 c/s amplifier, while the gain-control voltage is taken from the appropriate amplifier output on the Auto. positions and from a manually controlled potential divider on the Man. positions. The remaining two knobs in the recess are for this manual gain control, one being a six-position switch giving approximately 6 dB steps in gain, and the other giving a continuous coverage between steps.

Other pre-set controls are provided for adjustment of the amplitude characteristic, a.c. injection and backing-off the recorder current. The latter feature is only operative when recording C.W. since only then is the output current due to noise undesirably large at maximum gain. It is arranged to suppress the recorded noise level to zero on Auto C.W. without disturbing the input level required for full scale deflection. The 0–1 mA output for a 3000 ohm recording meter is taken from a pair of terminals at the rear of the unit.

Two jacks are provided on the front panel giving high impedance outputs for monitoring the a.f. content of the signal before and after the bandwidth reduction of the a.f. filter. If desired, the filter may be omitted by removing a link, allowing the full bandwidth of the a.f. amplifier to be available for listening to programme; as this disturbs the i.f. detector load, this link should not be removed while recording the mean level of a modulated signal. The types of modulated signal for which the receiver is suitable have been mentioned in Section 4.

A five-position meter switch enables the total current of each of the amplifiers to be checked during operation.

8.3. Power supplies

The d.c. power unit front panel recess contains adjustments for the stabilized output voltage and the neon stabilizer currents, which may be measured by the metering system.

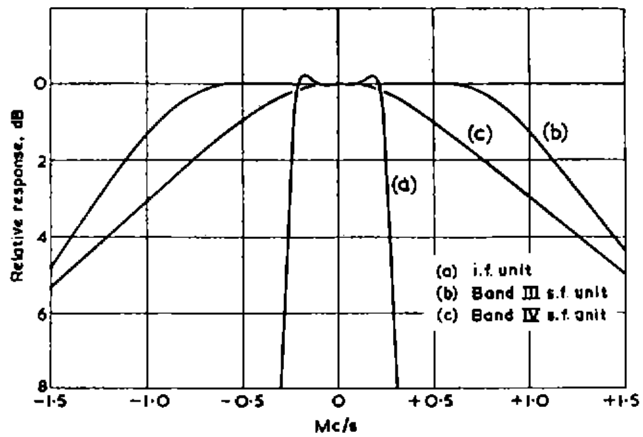

Fig. 11 — Frequency response of i.f. and s.f. units

TABLE 1

Spurious responses, Band III

Type of response		*Input voltage at spurious frequency relative to a 10µV reference signal at f_o for 1 dB error, dB*
$f_b<f_o$*	$f_b>f_o$	
f_b-f_i	f_b+f_i (image)	52
$13f_c-f_i$	$11f_c+f_i$	58
$2f_b-f_i$		70
$2f_b+f_i$		60
$3f_b-f_i$		54
$3f_b+f_i$		55
Other responses		>60

where f_o = signal frequency
f_b = final oscillator frequency
f_c = crystal frequency
f_i = intermediate frequency

* For the above measurements, the final oscillator frequency was below that of the signal. If their relative position were reversed, the first two types of response would be changed, as shown in the Table, with little change in the magnitudes of the responses.

The a.c. stabilizer has a similar metering system, and pre-set adjustments for the stabilized output voltage and the sensitivity of the correcting circuits.

TABLE 2

Spurious responses, Band IV

Type of response		*Input voltage at spurious frequency relative to a 30µV reference signal at f_o for 1 dB error, dB*
$f_b<f_o$	$f_b>f_o$**	
f_b-f_i	f_b+f_i (image)	20
$37f_c-f_i$	$35f_c+f_i$	58
$37f_c+f_i$	$35f_c-f_i$	65
$56f_c+f_i$		59
$56f_c-f_i$		61
$60f_c+f_i$		57
$60f_c-f_i$		56
f_i (i.f. break-through)		60
Other responses		>60

where f_o = signal frequency
f_b = final oscillator frequency
f_c = crystal frequency
f_i = intermediate frequency

** For the above measurements, the final oscillator frequency was above that of the signal. If their relative position were reversed, the first three types of response would be changed, as shown in the Table, with little change in the magnitudes of the responses.

9. Results of Performance Tests

9.1. Selectivity

The frequency response of a typical i.f. unit is shown in Fig. 11(a), while the response of the individual Band III and Band IV units is shown by curves b and c respectively. The response does not vary more than ±0·2 dB over the 400 kc/s pass-band, though a slight change, within these limits, occurs when attenuators are inserted between the s.f. and i.f. units.

Tables 1 and 2 give the type and amplitude of spurious responses which occur with the Band III and Band IV receivers. They show the relative amplitude of a signal at the spurious frequency which causes an error of 1 dB in the indicated amplitude of a fixed reference signal in the pass-band. These tests were made with the receiver operating on C.W. at a fixed gain to produce an i.f. detector current of

TABLE 3

Band	*Noise factor dB*	*S.F. unit gain dB*	*Internal attenuator loss dB*	*R.M.S. signal input µV for*			
				Unity output signal/noise ratio		*0·01 mA meter deflection*	
				C.W.	*Mod.*	*Auto C.W.*	*Auto Mod.*
III	7	+15	15	1·1	0·13	0·75	0·18
IV	11	−10	0	3·5	0·4	3·5	0·8

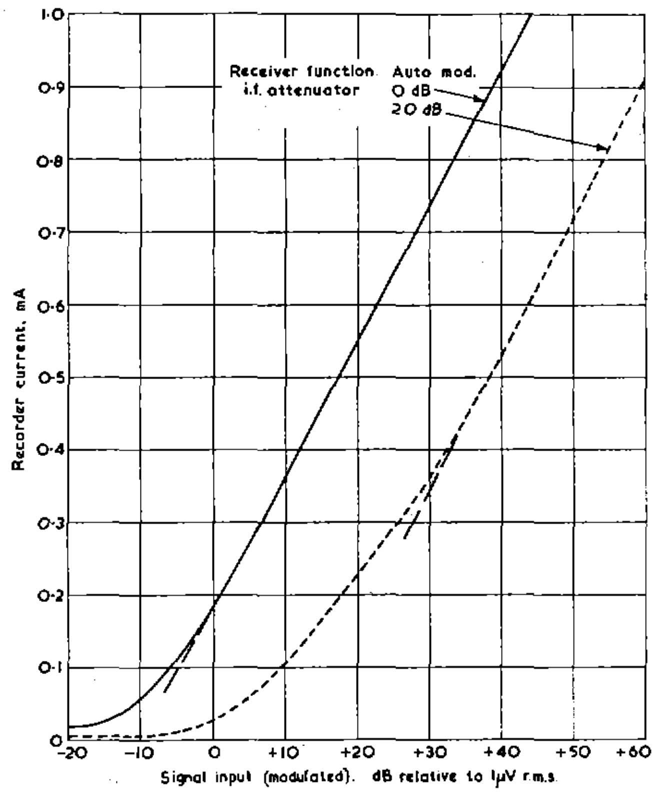

Fig. 12 — Calibration curve for a typical Band III receiver: Auto Mod.

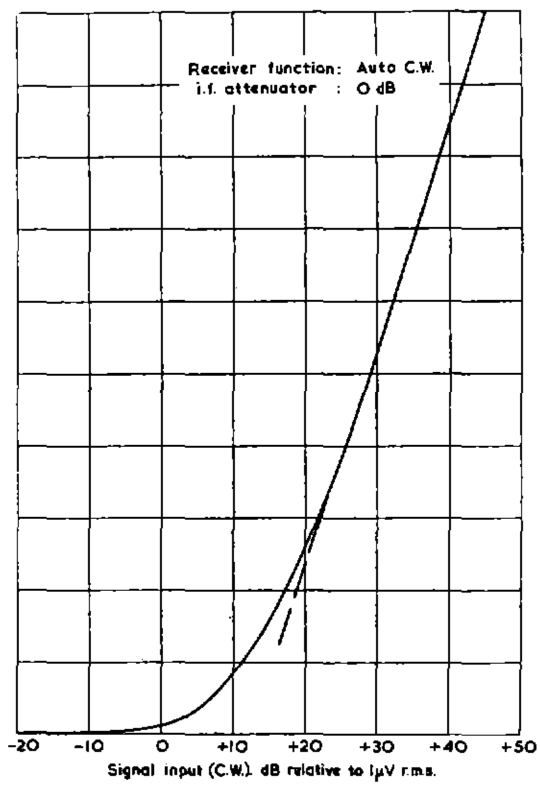

Fig. 13 — Calibration curve for a typical Band III receiver: Auto C.W.

50µA with the reference signal alone. No i.f. attenuators were inserted.

9.2. Sensitivity

The noise factors of the receivers are given in Table 3, with figures relating to the limiting sensitivity. The i.f. unit noise factor was 3·2 dB.

Figs. 12 and 13 show typical calibration curves for a Band III receiver operating on Auto Mod. and Auto C.W. respectively. The relationship on Auto Mod. is a good approximation to a logarithmic law over most of the recording range, which extends for 60 dB above the input level required for a unity signal/noise ratio output; the law is similar for the Band IV receiver. As shown by the dotted curve, however, there is a change in the characteristic when an attenuator is placed between the s.f. and i.f. units. This effect is analysed in Section 16.4.

When recording a modulated signal with the receiver set to Auto C.W., the amplitude characteristic remains unchanged for constant modulation depths less than 50 per cent. With programme modulation, and the receiver gain set to give an output of 1mA, modulation peaks were severely distorted in the a.f. output, but the effect on the recorded level was negligible.

9.3. Stability

Six complete Band III receivers, three recording the output of each of two signal generators, were run continuously for a period of 4 weeks; two of these receivers were run continuously for an additional ten weeks and intermittently for two weeks. No significant relative change in the output of the receivers in each group was observed, provided the signal generator frequency remained in the pass-band. Any drift in their absolute output could be attributed to the effect on the signal generator of fluctuating temperature and mains voltage.

A similar test on the Band IV receiver has not yet been performed, but all the units have been tested in an oven over the temperature range 20°C to 50°C; the changes observed in the resonant frequency of tuned circuits and in the gain of each unit were such as to produce a maximum change well within the specified limit of ± 1 dB. The time required, after switching on, before the receiver has reached its final calibration, is approximately 5 minutes.

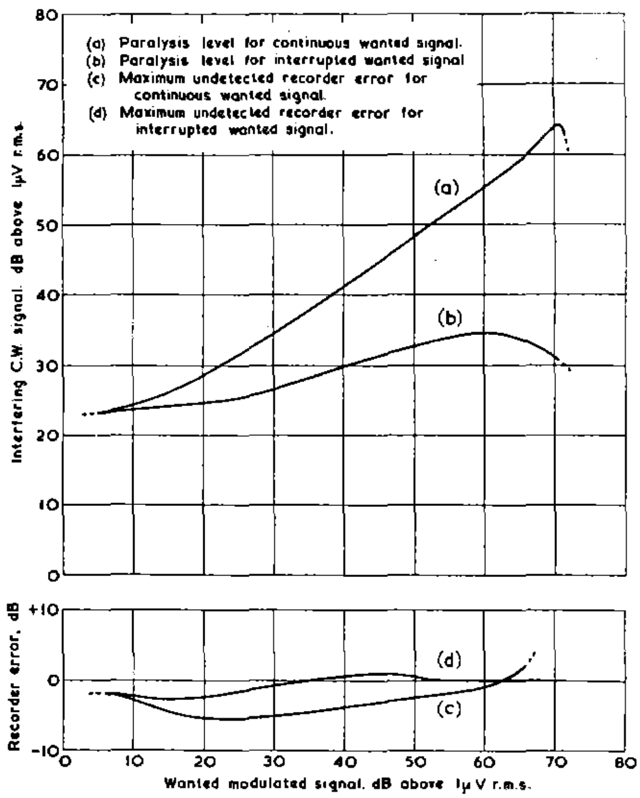

Fig. 14 — Co-channel interference characteristic

9.4. Co-channel interference rejection

When receiving the modulated test transmission with the receiver switched to Auto Mod., the gain of the receiver is controlled by the amplitude of the modulation, so that if a signal not modulated at 1000 c/s is present in the pass-band, there will be no appreciable change in gain or output until the amplitude of this unwanted signal approaches the overload level in the i.f. amplifier. When overload occurs, the wanted signal output will suddenly decrease and the gain will increase, thus enhancing the effect and causing paralysis. This overload level for a typical receiver is shown in Fig. 14 curve (a). If the unwanted signal is then reduced in amplitude, paralysis will continue to occur until curve (b) is reached, when the modulation reaching the a.f. detector will be sufficient to take control of the receiver gain. The recording error at these levels is shown on the same figure in curves (c) and (d).

Thus, if an interfering C.W. signal is present in the pass-band while recording a steady, modulated test transmission, the maximum undetected error will be given by curve (c), since it will be clear from the record if overload has occurred by the sharp and sustained fall in output. If, on the other hand, the test transmission is broken at regular intervals for identification marks, the error becomes obvious at a lower value, given by curve (d) since, if the interference is greater than (b) paralysis will occur at the next break in the test transmission and continue until the interfering signal falls below curve (b).

The curves in Fig. 14 refer to a C.W. signal differing in frequency from the wanted signal by 50 kc/s or greater, and will be modified if the frequency difference falls below this value, or if the interference contains modulation at 1000 c/s.

10. The Design of Future Receivers

In view of the satisfactory performance of this receiver to date, the i.f. unit may become the basis for the design of other receivers, with signal-frequency units designed for other bands. This would ease the problems of maintenance and training in operational service. If a special application requires a different i.f. bandwidth, it would be a simple matter to replace the existing i.f. filters by others having bandwidths ranging from ± 50 kc/s to ± 1 Mc/s.

11. Conclusions

The results of tests show that the performance of the receiver meets the specification for stability, amplitude characteristic and noise factor in Bands III and IV. It is expected that, when the development of the Band V signal-frequency unit is complete, its performance will be similar. Apart from meeting the requirement for noise factor, an improvement in the effective signal/noise ratio has been obtained by the use of a modulated test transmission and post-detector selectivity.

It is suggested that the receiver might become the basis for the design of receivers for other bands, by the use of other signal-frequency units.

12. Acknowledgments

Advice on the theoretical design was given by C. G. Mayo. The preliminary design of the A.C. stabilizer and the preliminary work on the i.f. amplifiers was carried out by D. J. Whythe. The lining-up and testing of the equipment was carried out with the assistance of R. J. H. Slaughter and P. Gill. R. G. Dilley assisted in the construction of the prototype receiver.

PART II

THE IMPROVEMENT IN SIGNAL/NOISE RATIO DUE TO POST-DETECTOR SELECTIVITY

13. Symbols Used in Text

v_i	r.m.s. signal voltage at input to receiver
v_s	r.m.s. signal voltage at input to i.f. detector
v_n	r.m.s. noise voltage at input to i.f. detector
i_{dc}	total d.c. output of i.f. detector
i_{dc}^o	d.c. output of i.f. detector in absence of signal
i_s	increase in d.c. output of i.f. detector due to presence of signal
i_m	r.m.s. current output of i.f. detector at modulating frequency f_o'
i_{dc}'	total d.c. output of a.f. detector
f_o	intermediate frequency
Δf	noise bandwidth of i.f. amplifier, centred on f_o
f_o'	modulating frequency
$\Delta f'$	noise bandwidth of a.f. filter, centred on f_o'
s	r.m.s. signal/noise ratio at input to i.f. detector
$C(s)$	function relating i_{dc} to s for a linear detector
$D(s)$	function relating P_s to s for a particular receiver
$F(s)$	function relating i_{dc}' to s for a particular receiver
S_o	r.m.s. ratio of signal output from a.f. filter to noise output, the noise being measured in the absence of signal
S_s	r.m.s. ratio of signal output from a.f. filter to noise output, the noise being measured in the presence of signal
S_{mo} S_{ms}	$\left.\begin{matrix} S_o \\ S_s \end{matrix}\right\}$ for a signal square-wave modulated at f_o' and a band-pass a.f. filter of width $\Delta f' \ll f_o'$
p_o	power density of noise spectrum at input to i.f. detector
P_o	power density of noise spectrum at output near $f = 0$ in absence of signal
P_s	power density of noise spectrum at output near $f = 0$ in presence of signal
P_m	mean value of P_s for modulated signal
N	overall noise factor of receiver
N_o	overall noise factor of receiver with no i.f. attenuator
A	insertion loss of i.f. attenuator
R_n	reduction in r.m.s. noise voltage at i.f. detector on inserting attenuator

14. Introduction

The accurate measurement of the amplitude of a small variable signal requires a receiver having sufficient gain, no internal sources of thermal noise and a bandwidth only just sufficient to pass the required signal information. In practice, the maximum useful gain is limited by the presence of thermal noise sources; these add fluctuations to the signal which reduce the accuracy of each observation, the accuracy depending on the r.m.s. signal/noise ratio at the output of the receiver. If this ratio is to be as large as possible, all those noise components having frequencies not occupied by the signal must be excluded from the detector. This can be done in some systems, such as communications links, by restricting the bandwidth of the circuits preceding the detector to that occupied by the signal information. But in a system designed for the measurement of relatively slow variations in the amplitude of a signal this may be impossible, due to the frequency drift of the transmitter relative to the receiver. The pass-band of the receiver must then be sufficiently wide to accommodate the maximum frequency drift and, for a given minimum signal input, the extra noise components included will affect the accuracy of measurement.

These unwanted noise components can be reduced by restricting the bandwidth of the circuits following the detector. The improvement with the conventional types of incoherent detector is less than would be achieved by a similar restriction of the i.f. bandwidth; this is due to intermodulation of the unwanted components in the detector. A practical detector of this type is difficult to represent theoretically, but its main characteristics are thought to be represented by the so-called ideal linear detector. Bennett[1] has derived theoretical expressions for the behaviour of such a device, taking account only of first-order intermodulation products of signal and noise. He has shown that a closer approximation may lead to an increase of about 0·5 dB in the calculated noise output, giving closer agreement with experimental results. Using the first-order approximation, Smith[2] has derived expressions for the theoretical improvement in signal/noise ratio using a low-pass a.f. filter, and his results are given in Section 15. The theory is extended in Section 16 to apply to a particular system using a band-pass a.f. filter and a square-wave modulated signal.

An important feature of the linear detector is that, for a small input signal, the signal output is inversely proportional to the noise input; this effect is discussed in Section 16.4 with special reference to the use of attenuators in the i.f. amplifier of the receiver.

15. Theoretical Results for an Idealized System

15.1. *Behaviour of the detector*

Consider an idealized system, which might be approached in a practical receiver, consisting of an ideal linear detector preceded by a band-pass filter with a rectangular response of bandwidth Δf, centred on f_0. We shall initially consider this to be followed by a low-pass filter which removes input frequencies and their harmonics but passes all frequencies up to $2\,\Delta f$.

The detector may be represented by the expression

$$\left.\begin{aligned} i &= bv \quad v>0 \\ &= 0 \quad v\leqslant 0 \end{aligned}\right\} \tag{1}$$

where i and v are the instantaneous values of the output current and input voltage respectively.

When a sinusoidal signal of r.m.s. voltage v_s is applied to such a device, the d.c. output is

$$i_{dc} = \frac{\sqrt{2}\,b}{\pi}\,v_s \tag{2}$$

For convenience the detector load is assumed to be unity; i_{dc}^2 therefore represents the total power output of the detector since all other currents are returned by a low impedance path.

If, instead of a signal, a normally distributed noise voltage of r.m.s. value v_n is applied, the d.c. output will be

$$i_{dc} = \frac{b}{\sqrt{2\pi}}\,v_n \tag{3}$$

Here, the d.c. power output i_{dc}^2 represents only a part of the total power output, the remainder being delivered in a low-frequency noise spectrum generated by intermodulation of components of the input spectrum and extending from $f = 0$ to $f = \Delta f$, as shown in Fig. 15(a) (ii).

If a steady signal at f_0 is added to the noise input, the d.c. output may be expressed as the sum of two terms,

$$i_{dc} = i_{dc}^0 + i_s \tag{4}$$

where i_{dc}^0 is the output due to noise alone and i_s is the increase in d.c. output due to the presence of the signal. The signal output is now no longer a linear function of the signal input; the relationship is described in Section 15.2. The output noise spectrum contains additional components generated by intermodulation of signal and noise and extending from $f = 0$ to $f = \frac{1}{2}\,\Delta f$; the result is shown in Fig. 15(b) (ii). The output noise power density in the pass-band of the post-detector filter is given in Section 15.3.

15.2. *D.C. output*

The d.c. output of the linear detector, in terms of the r.m.s. noise input, v_n, and the input r.m.s. signal/noise ratio $s = v_s/v_n$, is

$$i_{dc} = \frac{b}{\sqrt{2\pi}}\,v_n\,C(s) \tag{5}$$

where $C(s)$ is a function of s which is plotted in Fig. 16

$$C(s) \to 1 + \tfrac{1}{2}s^2 \text{ as } s \to 0$$

$$\text{and } C(s) \to \frac{2s}{\sqrt{\pi}} \qquad \text{as } s \to \infty$$

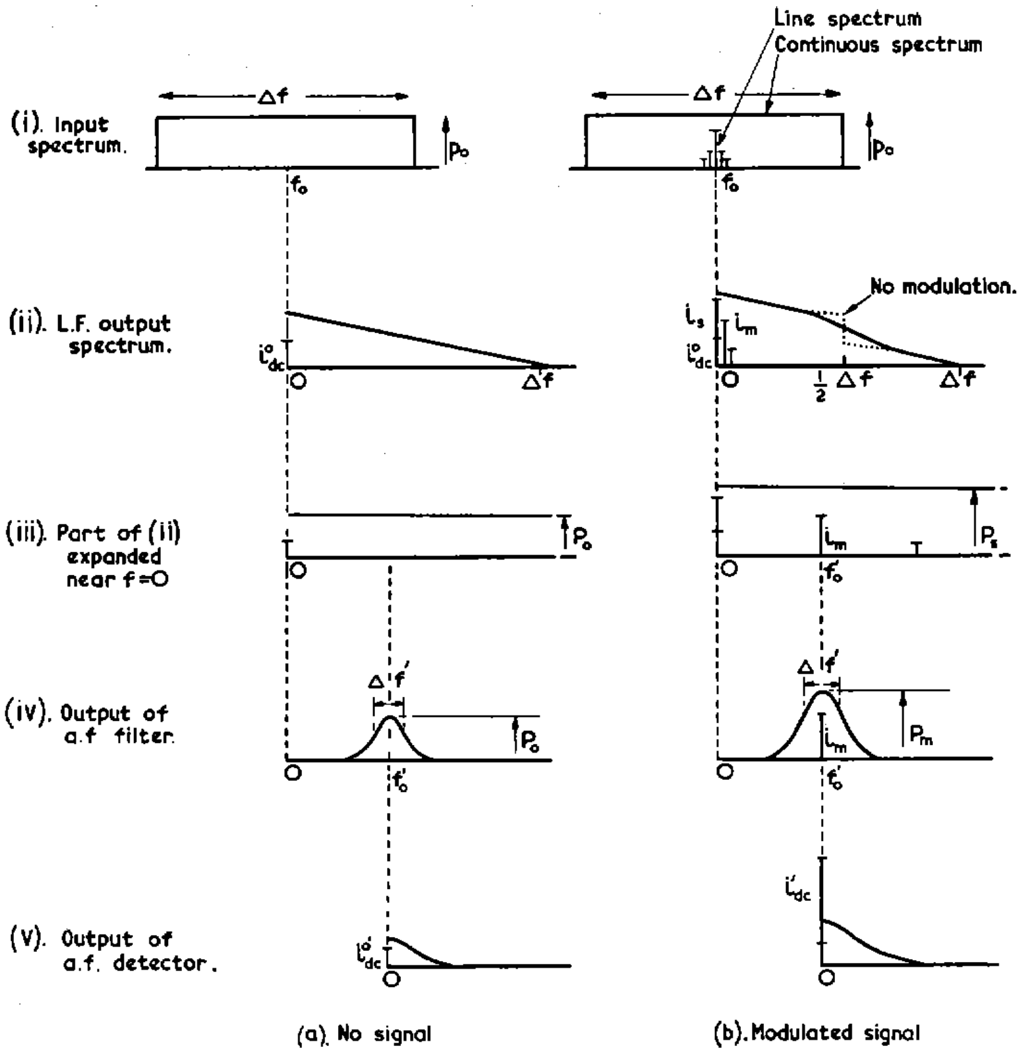

Fig. 15 — Spectra of noise and signal

Equation (5) reduces to equations (2) and (3) for large and small values of s, respectively. Substituting in equation (4) we have

$$i_s = \frac{b}{\sqrt{2\pi}} v_n \, [C(s)-1] \tag{6}$$

$$\rightarrow \frac{b}{\sqrt{8\pi}} v_n \, s^2 \text{ as } s \rightarrow 0$$

Thus, the linear detector behaves as a square-law detector for low input signal/noise ratios.

It is interesting to note, from Fig. 16, that for $s = 1$, $C(s) = 1{\cdot}44$, representing an increase in d.c. output of 3·2 dB due to the addition to the noise of a signal of the same r.m.s. value. The resemblance to r.m.s. addition is, however, fortuitous; the subsequent removal of the noise results in a reduction in the d.c. output of only 2·2 dB.

15.3. Noise output

For an input filter with a rectangular pass-band of width Δf, the r.m.s. input noise voltage is

$$v_n = \sqrt{p_o \, \Delta f} \tag{7}$$

where p_o is the power density of the input noise spectrum. The power density of the output spectrum at frequencies near $f = 0$, where the post-detector a.f. filter is to be situated, is

$$P_s = \frac{b^2 p_o}{4\pi} D(s) \tag{8}$$

where $D(s)$ is a function of s which is plotted in Fig. 16. For small values of s, $D(s) \rightarrow 1$ and for large values, $D(s) \rightarrow 8/\pi$. The value of P_s in the absence of signal is then

$$P_o = \frac{b^2 p_o}{4\pi} \tag{9}$$

and the noise power transmitted by a low-pass filter of

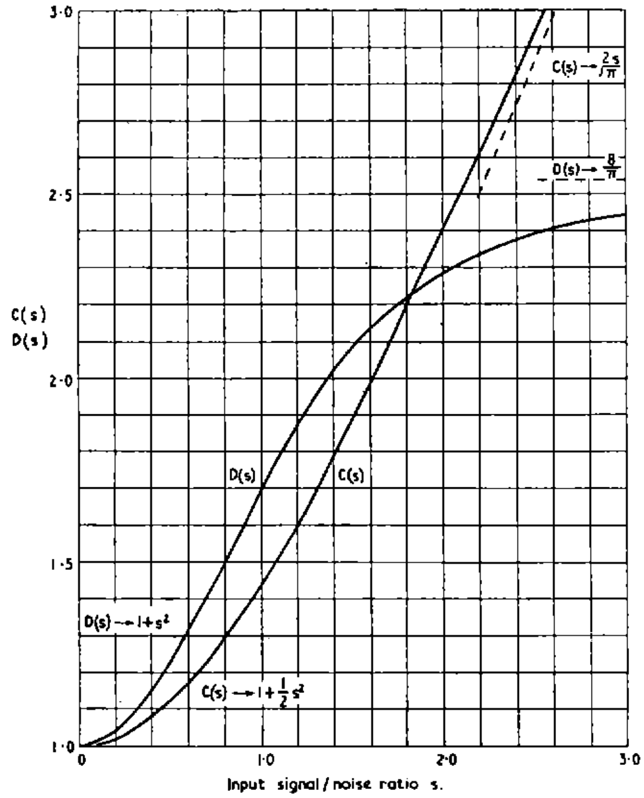

Fig. 16 — The functions C(s) and D(s)

bandwidth $\Delta f' \ll \Delta f$ in the presence or absence of signal is obtained by multiplying equations (8) and (9), respectively, by $\Delta f'$. The form of these output spectra can be seen from Fig. 15 (iii) and (iv).

15.4. Output signal/noise ratio

Expressions for the signal and noise output have been given in equations (6), (8), and (9). The noise output rises when a signal is applied; Smith[2] has therefore defined the ratio of signal output to the noise output in both the presence and absence of signal. These are, respectively,

$$S_s = \frac{i_s}{\sqrt{P_s \Delta f'}} \tag{10}$$

$$\text{and } S_o = \frac{i_s}{\sqrt{P_o \Delta f'}} \tag{11}$$

Each of these quantities forms a useful basis for comparison of various types of detector but neither represents the a.f. signal/noise ratio required here, which will depend on the modulating function superimposed on s and its effect on the output noise power. We shall, in the next section, obtain an approximate value for the output signal/noise ratio in a particular receiver when the input signal is modulated by a square wave.

16. Application to a Particular Receiver

16.1. Experimental details

The high-stability field-strength recording receiver described in Part I has a signal-frequency response within $\pm 0{\cdot}2$ dB over a bandwidth of 400 kc/s, and an energy bandwidth of about 500 kc/s. The transmitted signal used with this receiver is 100 per cent modulated with a 1000 c/s square wave. The 10·7 Mc/s i.f. amplifier is followed by a diode detector, the output of which is passed through a simple filter, resonant at 1000 c/s and having an energy bandwidth of about 30 c/s; it is then amplified and rectified in an a.f. bridge detector with negative current feedback. The d.c. component of the output of this second detector is recorded on a paper chart, and a calibration refers the recorded current to the signal input.

By means of automatic gain control a recording range of about 60 dB is achieved, that is, from 40 dB above noise level to 20 dB below noise. To extend this range to higher input levels, resistive attenuators may be inserted before the i.f. amplifier; but, as will be shown from the foregoing theory, a new calibration is required for each attenuator.

16.2. Measured signal/noise ratio improvement

A test was made on this receiver to compare the signal input required for an a.f. signal/noise ratio of unity with that required for the same ratio in the i.f. amplifier. The measurement was made using a calibrated piston attenuator and a thermocouple; the improvement ratio was found to be $-18{\cdot}5$ dB.

16.3. Theoretical signal/noise ratio improvement

If the signal input is modulated, the output spectrum will be modified as indicated by the solid line in Fig. 15(*b*) (ii). Considering only first-order intermodulation products, a low-frequency modulation will only disturb the output spectrum in the vicinity of $f = \frac{1}{2}\,\Delta f$, but if the modulation frequencies extend up to $\frac{1}{2}\,\Delta f$ the disturbance of the spectrum will extend down to the region near $f = 0$. The spectrum of the square-wave modulation used in this receiver varies as the inverse square of the frequency and the selected fundamental frequency f_o' is very small compared with Δf. The shape of the output spectrum may therefore be treated as if the signal power were all concentrated at f_o; thus we need only consider the behaviour of the detector in the presence and absence of a steady signal of amplitude equal to that of the peak input.

If the r.m.s. value of the modulated input signal/noise ratio is s, it must be alternating during the modulation cycle between zero and $\sqrt{2}s$, and the peak value of the change in d.c. output due to the signal will be, from equations (6) and (7)

$$i_s = b\left(\frac{p_o \Delta f}{2\pi}\right)^{\frac{1}{2}} [C(\sqrt{2}s)-1] \tag{12}$$

The fundamental component of the modulation at f'_o will have a peak-to-peak amplitude of $4/\pi$ times greater than this, and its r.m.s. value will be

$$i_m = \frac{\sqrt{2}}{\pi}\, i_s \tag{13}$$

The output noise spectrum will be modulated by the signal but, since the a.f. filter has a bandwidth $\Delta f' \ll f'_o$, the noise power transmitted by it will not vary during the modulation cycle; from equation (9), the mean value is

$$P_m \Delta f' = b^2 \frac{p_o \Delta f'}{4\pi} \left[\frac{1+D(\sqrt{2}s)}{2}\right] \tag{14}$$

in the presence of the modulated signal.

To check the experimental value we may use the expressions for small values of s and write

$$\left.\begin{aligned} C(\sqrt{2}s)-1 &= s^2 \\ D(\sqrt{2}s) &= 1 \end{aligned}\right\} \tag{17}$$

giving, for small values of s,

$$S_{mo} = S_{ms} = \frac{2}{\pi}\left(\frac{\Delta f}{\Delta f'}\right)^{\frac{1}{2}} s^2 \tag{18}$$

Inserting the values of $\Delta f = 5.10^5$ c/s, $\Delta f' = 30$ c/s, and putting $S_{mo} = 1$, we obtain the input signal/noise ratio $s = 0{\cdot}11$, or $-19{\cdot}2$ dB; this is in fairly close agreement with the observed improvement ratio $-18{\cdot}5$ dB.

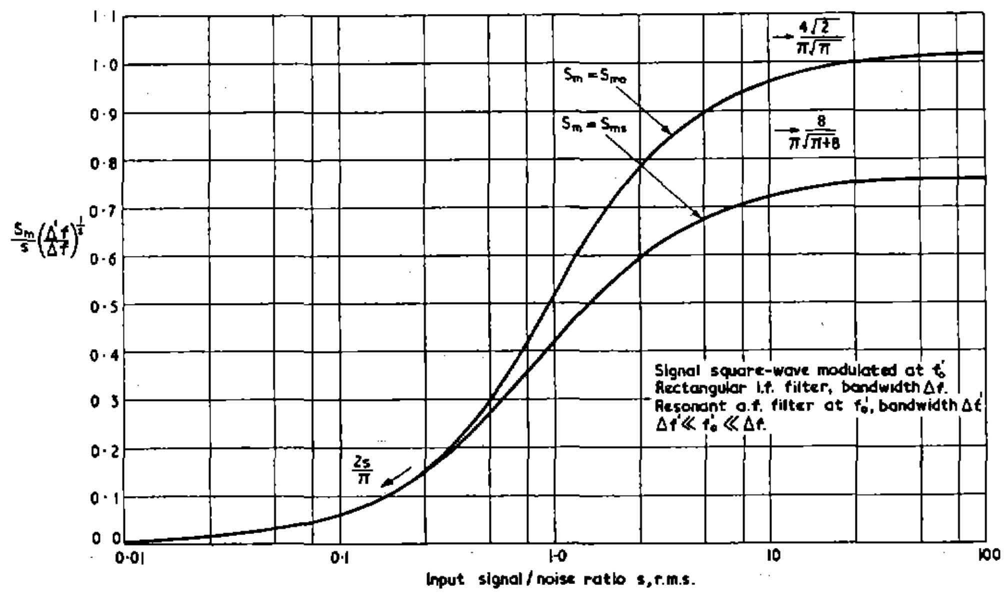

Fig. 17 — Output signal/noise ratios S_{mo}, S_{ms}

Hence, the effective r.m.s. output signal/noise ratios corresponding to those of Smith in equations (10) and (11) are

$$S_{ms} = \frac{2\sqrt{2}}{\pi}\left(\frac{\Delta f}{\Delta f'}\right)^{\frac{1}{2}} \frac{C(\sqrt{2}s)-1}{[1+D(\sqrt{2}s)]^{\frac{1}{2}}} \tag{15}$$

$$S_{mo} = \frac{2}{\pi}\left(\frac{\Delta f}{\Delta f'}\right)^{\frac{1}{2}} [C(\sqrt{2}s)-1] \tag{16}$$

Equation (15) represents the ratio of signal output to that noise output occurring in the presence of signal, and equation (16) gives the ratio of signal to that noise output occurring in the absence of signal; these are shown in Fig. 17. The experimental result obtained in Section 16.2 was for the latter quantity, the total power output being observed with and without the signal.

16.4. Distortion of calibration by i.f. attenuators

A peculiarity of the linear detector, when working at a low signal/noise ratio, may be illustrated by referring to equation (6),

$$i_s \to \frac{b}{\sqrt{8\pi}} v_n s^2 \quad \text{as } s \to 0 \tag{6}$$

This may be written

$$i_s \propto \frac{v_s^2}{v_n} \quad \text{for } v_s \ll v_n$$

showing that, for small signals, the signal output is inversely proportional to the noise voltage applied to the detector.

If, for example, a 10 dB attenuator is inserted before the detector, both v_s and v_n are reduced by the same amount and the fall in signal output is also 10 dB. But to raise the output by 10 dB requires an increase of v_s of only 5 dB. Thus, the signal input to the receiver must be changed by 5 dB to compensate for the insertion of a 10 dB attenuator.

This effect is, of course, present without post-detector filtering but its importance here lies in the fact that the signal is below noise level at the detector input over a large part of the working range.

In this receiver, an approximately logarithmic input/output law is achieved by using a.g.c., so that the gain of the i.f. amplifier varies as a function of the recorded output. To find the change in this law due to the addition of an attenuator, without defining the a.g.c. characteristic, we must find the change in the signal/noise ratio in the i.f. amplifier required to maintain a constant output from the a.f. detector at each point of the calibration.

Using the same functions as for the i.f. detector in equation (5), the d.c. output from the a.f. detector is, from equations (14) and (15),

$$i'_{dc} = b' \left(\frac{P_m \, \Delta f'}{2\pi}\right)^{\frac{1}{2}} C\,(S_{ms}) \tag{19}$$

which may be expressed in terms of the i.f. signal/noise ratio by

$$i'_{dc} = K \, v_n \, F(s) \tag{20}$$

where $K = \dfrac{bb'}{\sqrt{8\pi}} \left(\dfrac{\Delta f'}{\Delta f}\right)^{\frac{1}{2}}$

a constant,

$$v_n = (p_o \, \Delta f)^{\frac{1}{2}}$$

the r.m.s. noise input to the i.f. detector

and $F(s) = C(S_{ms}) \left[\dfrac{1+D(\sqrt{2}s)}{2}\right]^{\frac{1}{2}}$

The function $F(s)$ has been calculated for the receiver under discussion, and is plotted in Fig. 18.

If an attenuator is inserted at some point in the receiver, the r.m.s. noise input to the detector will be reduced by a factor

$$R_n = \frac{v_{no}}{v_n} \tag{21}$$

where the suffix 'o' refers to the absence of the attenuator. For any given output current i'_{dc} it follows from equations (20) and (21) that

$$F(s) = R_n \, F(s_o) \tag{22}$$

Using this result and the plotted function $F(s)$ shown in Fig. 18, it is possible to derive the change in input signal/noise ratio required to maintain a given output current when the attenuator is inserted.

Taking a practical example, let us consider the effect of inserting a 20 dB attenuator between the i.f. amplifier and the signal-frequency unit. The attenuator, as well as reducing the input signal and noise by a factor A, its insertion loss, also increases the noise factor of the receiver, N, due to the presence of noise sources in the i.f. amplifier.

The noise factor of the receiver is

$$N = N_1 + A \, \frac{N_2 - 1}{G_1} \tag{23}$$

where N_1 = Noise factor of signal-frequency unit

G_1 = Power gain of signal-frequency unit

A = Insertion loss of attenuator

N_2 = Noise factor of i.f. amplifier

With the receiver under consideration the values were:

$N_1 = 5\cdot0$ (7 dB); $N_2 = 2\cdot2$ (3·5 dB); $G_1 = 32$ (15 dB)

The reduction in the noise factor is therefore given by

$$R_n = A \, \frac{N_o}{N} \tag{24}$$

and the values for each value of A are given in Table 4.

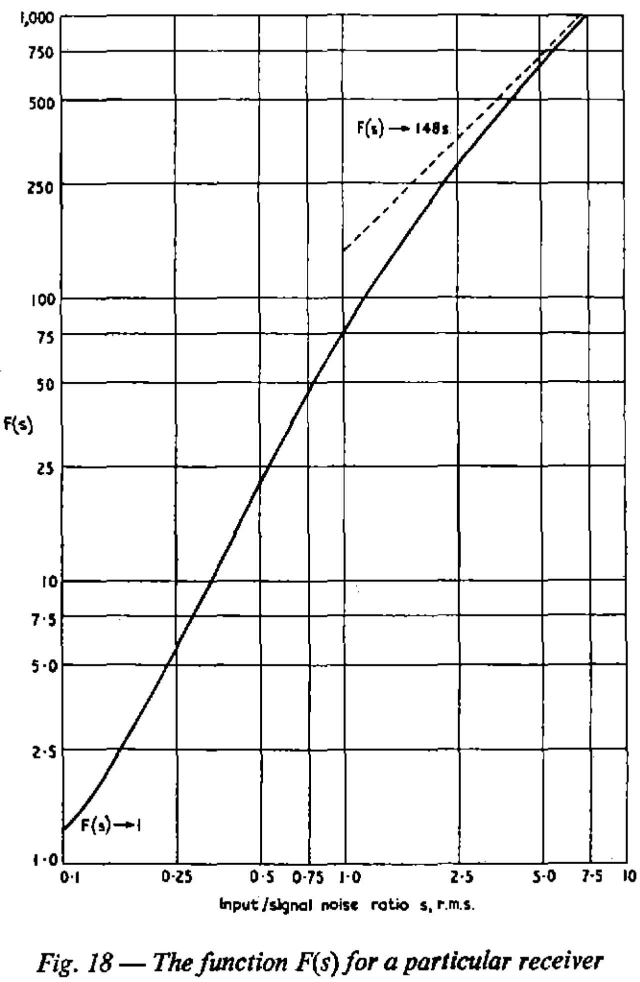

Fig. 18 — The function F(s) for a particular receiver

TABLE 4

Attenuation *dB*	*Noise factor, N* *dB*	*Increase in N* *dB*	*I.F. noise reduction factor R_n*	
			dB	*r.m.s.*
0	7·0	0·0	0·0	1·0
20	9·3	2·3	17·7	7·7

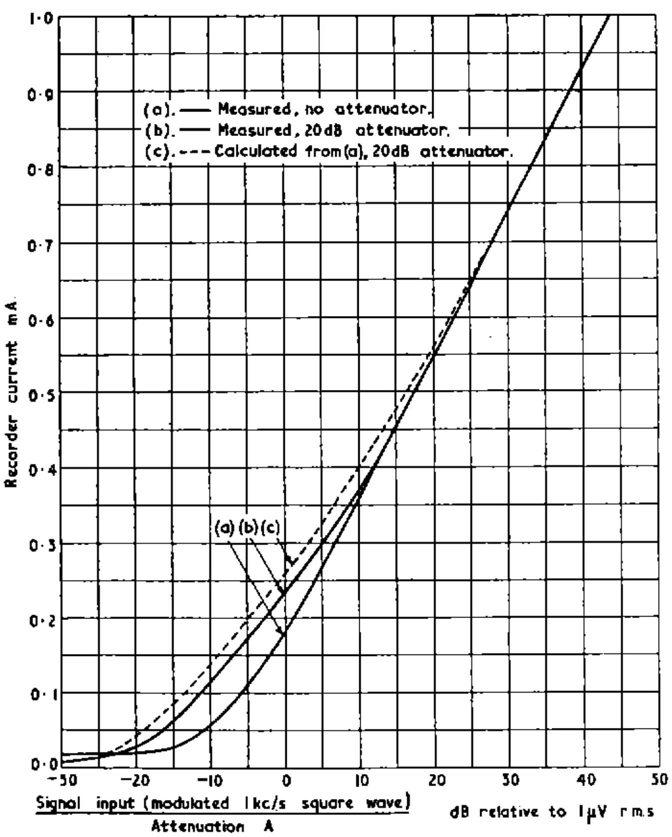

Fig. 19 — Calibration curve for a particular receiver, showing distortion by attenuator

Typical calibration curves for this receiver are shown in Fig. 19; these show the recorder output current i_{dc} as a function of the signal-frequency r.m.s. input voltage v_i. For convenience this input voltage is shown divided by the loss of the attenuator, so that the abscissa is proportional to the detector signal input, v_s, for a constant value of output current. The curves (a) and (b) were found experimentally using 0 dB and 20 dB attenuators, showing a change of up to 10 dB in the calibration over a third of the recording range.

Curve 19 (a) was used to find the input signal/noise ratio, using the expression for the r.m.s. noise voltage referred to the input of the receiver,

$$s_o = v_i(4kT\Delta f R N_o)^{-\frac{1}{2}} \tag{25}$$

where k = Boltzmann's constant
T = Temperature °K
Δf = Energy bandwidth
R = Source resistance presented to receiver
N_o = Noise factor of receiver with no attenuator

Knowing s_o and R_n, the value of s was found, using equation (22) and Fig. 18; the change of signal level for constant output was obtained from

$$\frac{v_s}{v_{so}} = \frac{s}{R_n\, s_o} \tag{26}$$

and a curve plotted for the 20 dB attenuator, Fig. 19(c). The experimental result for a 20 dB attenuator, curve (b), shows qualitative agreement with the theory, though there is a discrepancy of about 2 dB between the calculations and the measurements.

It is known from further measurements that the i.f. detector used in this receiver behaves as a linear mean detector to components of signal and noise differing in frequency by less than about 5 kc/s, but that at higher difference-frequencies the behaviour tends towards that of a peak detector, due to the effect of the bias voltage developed across the capacitance of the detector load. A satisfactory mathematical treatment for a practical detector of this type has not been produced.

17. Conclusions

The theoretical background of signal/noise ratio improvement has been given, based on an analysis of an idealized linear detector. An application of the theory to a special-purpose receiver, designed for field-strength recording, gives an agreement which is considered reasonable in view of the known differences between the practical and the ideal detector.

In a receiver using post-detector selectivity, such as the one described here, the signal/noise ratio at the detector input is less than unity over a large part of the range of measurable signals; if the noise level in the i.f. amplifier is changed by the addition of an attenuator, the amplitude characteristic of the receiver will be considerably distorted. If the range of measurable signals is to be extended by the use of calibrated attenuators, either these must be inserted at the input to the receiver or a separate calibration will be necessary for each value of attenuation.

18. References

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