G J Phillips

BBC

ENGINEERING DIVISION

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A Band V Signal-frequency Unit and a Correlation Detector for a VHF/UHF Field-strength Recording Receiver

Part I

A Band V Signal-frequency Unit

by

G. J. PHILLIPS, M.A., Ph.D., B.Sc., A.M.I.E.E.

and

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(Research Department, BBC Engineering Division)

Part II

A Correlation Detector

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BRITISH BROADCASTING CORPORATION

PRICE FIVE SHILLINGS

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FOREWORD

THIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

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A BAND V SIGNAL-FREQUENCY UNIT AND A CORRELATION DETECTOR FOR A VHF/UHF FIELD-STRENGTH RECORDING RECEIVER

SUMMARY

The monograph is divided into two parts, each describing additional equipment for use with a VHF/UHF field-strength recording receiver which was designed for propagation measurements and has been described in Monograph No. 6.

Part I deals with the design of a pre-tuned signal-frequency unit covering Band V. This is basically similar to the previously described Band IV unit in that the local oscillator is crystal-controlled, using several stages of frequency multiplication, and the final oscillator and signal-frequency circuits are coaxial-line resonators. It differs from the Band IV design by using a coupled pair of resonators at the signal frequency; the design for the coupling elements is discussed in the Appendix.

Part II describes a correlation detector which can be added after the post-detector circuits of the receiver in order to increase the effective sensitivity. This device has the principal property that, in the absence of the signal to be received, the output due to noise contains no d.c. component, so that the signal threshold is more clearly defined and is less subject to drift. The device derives its name from an analogy with the mathematical operation involved in forming the auto-correlation function from a time-function with a known spectrum; this is also explained in the Appendix.

Part I

A BAND V SIGNAL-FREQUENCY UNIT

1. Introduction

A receiver for recording field-strength in three of the frequency bands allocated for television broadcasting in the United Kingdom, namely Band III (174–216 Mc/s), Band IV (470–582 Mc/s), and Band V (614–854 Mc/s),* has been described in an earlier monograph.[1] This receiver, which is of the superheterodyne type, consists of three interchangeable signal-frequency (s.f.) units and a main unit containing the intermediate-frequency (i.f.) amplifier, detector, and post-detector circuits.[2] The Band V s.f. unit was not fully described in the earlier account, and now becomes the subject of the first part of the present monograph.

2. Specification

The special purpose for which this receiver was designed and the specification evolved to make it suitable for this task were stated and discussed in the earlier monograph. The specification is repeated in Section 2.1 below.

2.1 *Specification for a VHF/UHF Field-strength Recording Receiver*

The receiver must be able to record the field strength of special transmissions on any frequency in the ranges 174–216 Mc/s and 470–960 Mc/s. These transmissions will be square-wave modulated to a depth of 100 per cent at 1,000 c/s ± 1 c/s and the carrier frequency controlled to within $0 \cdot 01$ 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.

2.2 *Specification for the Band V Signal-frequency Unit*

Since the Band V s.f. unit is an integral part of the receiver its general form and certain detailed requirements were already determined. For example, it was necessary that it should be interchangeable with the other two s.f. units and suitable for use with the existing main unit for which an i.f. of $10 \cdot 7$ Mc/s had been chosen. The following specification for the Band V unit was derived from these requirements and the specification for the whole receiver.

(i) The unit shall be suitable for use with an aerial and feeder system providing a nominal source impedance of 71 ohms.

(ii) The i.f. circuit must be suitable for direct connexion to the standard main unit.

(iii) When used with the standard main unit, having an i.f. noise factor of $3 \cdot 2$ dB, the overall noise factor shall be less than 15 dB.

(iv) The local oscillator shall be maintained within ± 100 kc/s of the nominal frequency. This requires a stability of better than $0 \cdot 01$ per cent at the highest frequency.

* The frequency ranges given for Bands IV and V are those adopted at the Stockholm Conference of 1961.

(v) The rejection of the image-frequency shall be not worse than that provided by the Band IV unit, that is not worse than 20 dB.

(vi) The long-term gain stability of the Band V unit must be such that, when it is used with the standard main unit, the overall stability of the receiver complies with the specification.

3. Description of the Band V Signal-Frequency Unit

3.1 *General*

A photograph of the unit is shown in Fig. 1. Fig. 2 gives the circuit diagram.

The basic design of the Band V unit is similar to that of the Band IV unit. A diode frequency changer is preceded by a coaxial line resonator signal-frequency tuner without a signal-frequency amplifier. The local oscillator consists of a crystal-controlled oscillator followed by several stages of frequency multiplication, the final tuned circuit of which is also a coaxial line resonator. However, the higher frequency and greater frequency range of Band V required considerable development, particularly in ensuring adequate image-frequency rejection and local-oscillator power.

The final design of the unit shows two major differences from that of the Band IV unit:

(i) Signal-frequency tuning is provided by a coupled pair of coaxial line resonators, whereas the Band IV unit has only a single signal-frequency resonator.

(ii) An extra stage has been provided in the local-oscillator frequency-multiplier chain.

The main chassis is of strip and plate construction and has a 19-in. by 7-in. (48 cm by 18 cm) front panel for rack mounting like the units for Bands III and IV. The r.f. input socket and metering jack are on the front panel while the i.f. output and power input sockets are at the rear.

The two main sub-units—the tuning sub-unit and the

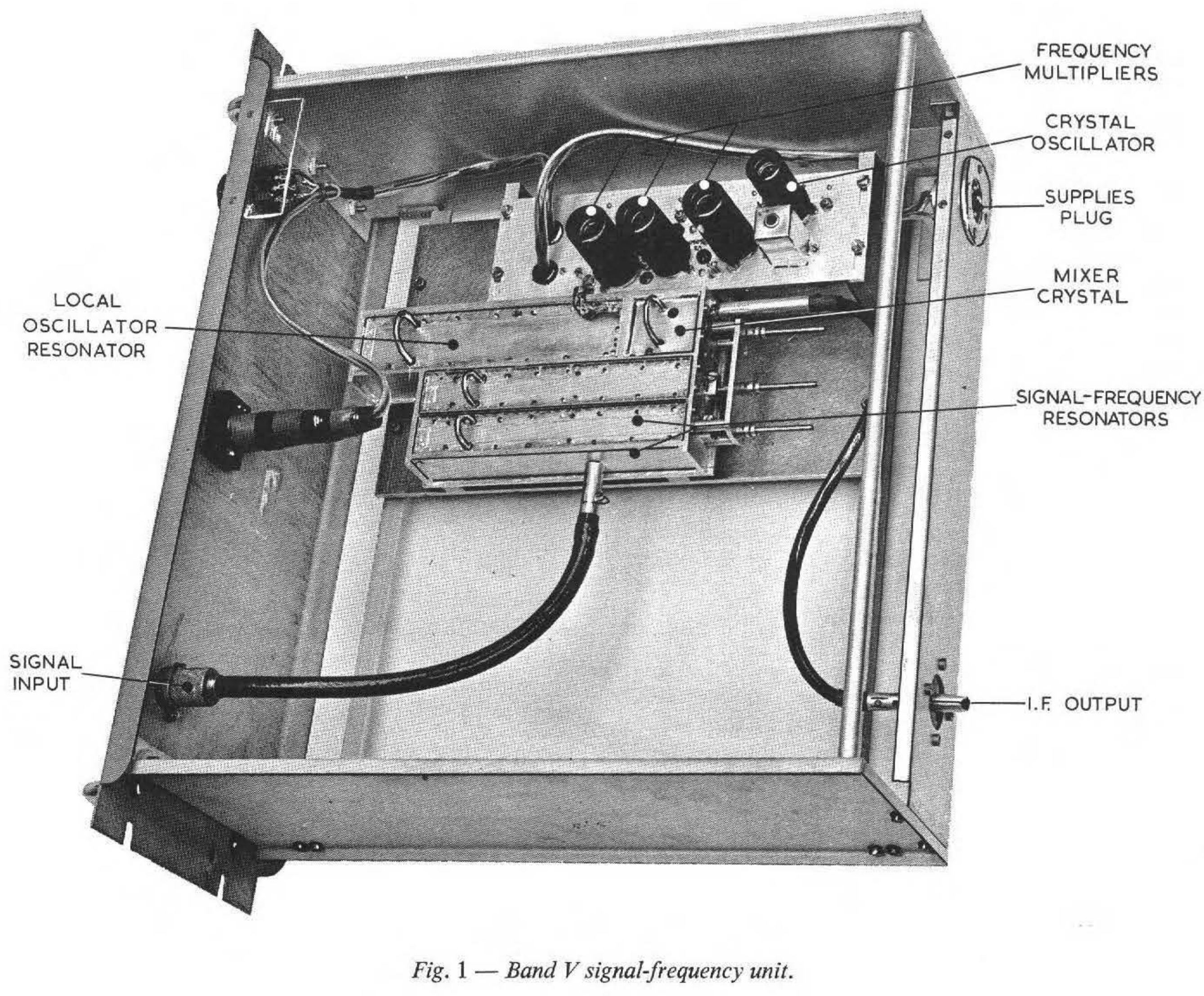


Fig. 1 — Band V signal-frequency unit.

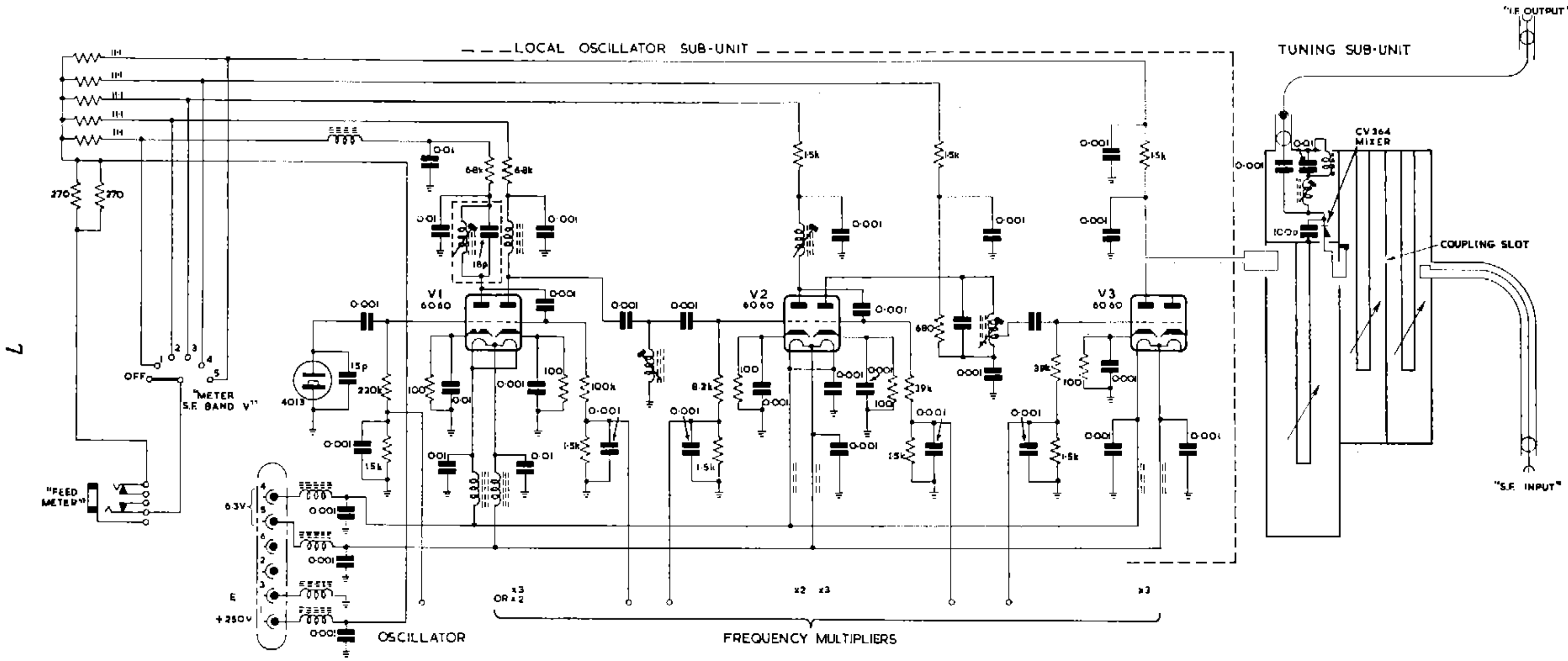


Fig. 2 — Circuit diagram of Band V unit.

local-oscillator sub-unit—are mounted together on a $\frac{3}{16}$-in. (0·48 cm) thick duralumin plate, which is itself fixed to the main chassis strips by shock-absorbing mountings.

3.2 *Tuning Sub-unit*

3.2.1 *Resonator Construction*

The tuning sub-unit consists of three coaxial line resonators and a compartment for the mixer crystal and its i.f. circuit. Two identical resonators of 1-in. by 1-in. (2·54 cm by 2·54 cm) cross-section form the signal-frequency tuned circuit while the third, of $1\frac{1}{2}$-in. by 1-in. (3·81 cm by 2·54 cm) cross-section, is the local oscillator resonator. The main structure of this sub-unit is fabricated in brass sheet which is then silver-plated to a minimum thickness of 0·0005 in. (0·013 mm) to ensure high conductivity. Milled slots in the base plate are used to position the vertical walls and ensure adequate dimensional accuracy, while removable lids with spring contacts allow ready access to components inside the cavity.

Compared with the tube construction employed for the Band IV unit, the present construction is simpler (bearing in mind that three tubes would be needed) especially when inter-cavity coupling arrangements are considered.

3.2.2 *Tuning*

The system for tuning the three resonators is the same as that used in the Band IV unit, which has proved very satisfactory. Each resonator is a coaxial quarter-wave line short-circuited at one end and open-circuited at the other. The inner conductor consists of a tube which telescopes inside a larger diameter fixed tube. Adjustment is made by means of a rod of 'Nilo 36' low-thermal-expansion alloy running inside the tubes and pinned to the smaller tube at the open-circuited end of the line. After adjustment, the rod is clamped to the cavity wall at the short-circuited end, thus ensuring stability of the inner conductor length. The tubes are of silver-plated brass and the contact between them is a ring of silver wire located in a groove at the end of a fixed tube. For ease of tuning the rod passes, with a friction grip, through a threaded bush; it may thus be pushed or pulled directly to give a coarse adjustment, or rotated by a knob so that the thread operates in a fixed plate to give fine adjustment.

3.2.3 *Coupling Circuits*

The coupling to the first signal-frequency cavity is by a small adjustable loop connected to the receiver input socket by a flexible coaxial cable. The two signal-frequency cavities are loosely coupled by a slot in the wall between them. This acts as a common inductance in the outer conductor of the two resonators. The signal is finally coupled to the mixer diode by a loop which passes through the opposite wall of the second cavity into the oscillator cavity, where it makes contact with one terminal of the diode. By this means, oscillator power is also coupled to the mixer. The mixer is loaded by an i.f. circuit in a separate compartment at the rear of the oscillator cavity. The diode holder incorporates a plate and sheets of mica which form a capacitance to the rear wall of the oscillator cavity, thereby completing the r.f. circuit.

The coupling loops into and out of the signal cavities, as well as the coupling slot between them, are placed at some distance from the short-circuited end of the lines, rather than at the point of short-circuit. This enables the circuits to be tuned over the whole band with a minimum of coupling adjustments and to have an absolute-frequency bandwidth which is nearly constant. (See Appendix, Section 13.1.)

The tuning condition recommended is one close to the 'transitionally coupled' condition for which the r.f. response has the flattest maximum. However, as shown in the Appendix, Section 13.2, the match at the r.f. input can be perfect for this condition only if the cavities have an infinite (unloaded) Q factor, i.e. if they are completely loss-free. The receiver is therefore aligned for a good response shape when fed from a source impedance equal to the nominal input impedance (71 ohms). The input v.s.w.r. is then checked and, if poorer than 0·67, slight readjustment of the input coupling loop and the first cavity tuning is made to improve the match. The disturbance to the response curve is slight and is in the direction of an 'under-coupled' response, i.e. one with a single maximum not quite as flat as for transitional coupling. Since the r.f. response shape will be affected by changes in the loading of the first cavity, it is recommended that the aerial v.s.w.r. as seen by the receiver should be not lower than 0·75.

During development there was some difficulty in designing a fixed mixer coupling loop which gave satisfactory loading of the signal cavity as well as adequate oscillator coupling over the whole band, but a design that appeared to achieve this was finally evolved. Subsequently some difficulty arose in a fairly narrow range of frequencies around 775 Mc/s, apparently due to insufficient loading of the signal cavity. It was found that the difficulty could be overcome by adding, for operation in this frequency range, a small capacitance to earth at the point of connexion to the mixer diode. This suggests that some resonance effect in the loop is coming into play, but the matter has not been investigated further.

3.3 *Local-oscillator Sub-unit*

3.3.1 *Construction*

The local-oscillator sub-unit employs a small flat chassis bolted to two duralumin blocks attached to the main base plate. The output is coupled into the oscillator cavity of the tuning sub-unit by a wire loop which projects into the cavity resonator through a small hole. Adjustment of the coupling is achieved by movement of the local-oscillator sub-unit, the fixing holes being slotted to allow for this. By varying the penetration of the loop into the resonator in this way, the optimum mixer crystal current may be obtained.

3.3.2 *Type of Oscillator and Choice of Multiplication Factor*

The local oscillator consists of a crystal-controlled oscillator followed by four stages of frequency multiplication. The local-oscillator resonator forms the anode tuned circuit of the final multiplier. With the type of circuit employed the maximum frequency for which crystals were readily obtainable was 20 Mc/s and so an overall multiplication factor of 54 was chosen for the higher local-oscillator frequencies. However, at the lowest local-oscillator frequencies an overall multiplication by 54 requires a crystal-oscillator frequency of about 11 Mc/s. Signals at frequencies separated by this frequency might be generated in the multiplier chain and produce spurious signals in the i.f. amplifier close to 10·7 Mc/s with sufficient power to affect the receiver performance. For this reason an overall multiplication by 36 was chosen for local-oscillator frequencies less than 675 Mc/s.

3.3.3 *Multiplication Scheme and Valve Type*

The Band IV multiplier used two double triodes type 12AT7, the crystal oscillator being followed by three stages multiplying by factors of 4, 3, and 3. With the larger overall multiplication factor, 54, required for the higher-frequency portion of Band V, and the known difficulty of obtaining sufficient grid drive for the final multiplier when using large factors in the earlier stages, it was found necessary to use four multiplier stages. The highest multiplication factor in any one stage could then be limited to 3.

Three double triode valves type 6060 (a special quality version of type 12AT7) are used. V1 is used for the crystal oscillator and first multiplier circuits, V2 for the second and third multipliers, and a half of V3 for the fourth multiplier stage. The multiplication scheme is given in Table 1.

3.3.4 *Circuit Design*

The quartz crystal is a BT-cut plate in an evacuated glass envelope with a B7G-type base. The total frequency tolerance is 0·01 per cent of the nominal frequency. The crystal is connected as the grid element of a tuned-anode-tuned-grid oscillator.

The multipliers are class-C amplifiers with single-tuned anode circuits to select the required harmonic of the grid circuit frequency. The anode tuned circuit of the final multiplier is the local-oscillator coaxial line resonator; the other circuits consist of coils tuned entirely by stray circuit capacitances. The range of inductance values required for tuning over the complete frequency range is obtained in the anode circuit of the crystal oscillator and of the first and second multipliers by using iron-dust cores at the lower frequencies and copper cores at the higher frequencies.

The design of the third and fourth multiplier stages is discussed separately below.

The grid current of each triode may be measured by external meters connected to test points on the upper side of the chassis plate. Facilities for monitoring anode currents are provided by a selector switch and a meter jack on the front panel.

Decoupling of the heater leads of V2 and V3 is provided by ceramic condensers and by ferrite beads threaded over the supply wires. Wire-wound chokes are used for the heater supply of V1.

3.3.5 *Design of the Third and Final Multiplier Stages*

The crystal oscillator and the first two multipliers follow

TABLE 1

Stage	Local-oscillator frequency less than 675 Mc/s		Local-oscillator frequency greater than 675 Mc/s	
	Stage multiplication factor	Output frequency of stage	Stage multiplication factor	Output frequency of stage
Crystal oscillator	1	f_c	1	f_c
First multiplier	2	$2f_c$	3	$3f_c$
Second multiplier	2	$4f_c$	2	$6f_c$
Third multiplier	3	$12f_c$	3	$18f_c$
Fourth multiplier	3	$36f_c$	3	$54f_c$

the design of the Band IV unit fairly closely, but the design of the third and fourth multipliers necessitated considerable development. It is convenient to discuss these two stages together.

The final local-oscillator frequency required lies in the band 589–971 Mc/s, where special UHF valve types are normally used. Nevertheless a more conventional valve such as the type 6060 (12AT7) is useful as an oscillator at frequencies up to about 500 Mc/s. This suggested that it could function satisfactorily as a multiplier for input frequencies up to at least 300 Mc/s. The advantages of the use of a conventional valve rather than a special UHF type are as follows:

(*a*) They are readily available, inexpensive, and possess good reliability.

(*b*) Being constructed on a normal pin-type base they allow a more compact and convenient mechanical layout than, for instance, disc-seal valves.

(*c*) The number of valve types used in the receiver need not be increased.

Again, for reasons of compactness and ease of construction, a lumped tuned circuit, rather than a line resonator, was used at the grid of the final multiplier valve. This means, in practice, an upper frequency limit of about 300 Mc/s for this circuit; a multiplication factor of 3 is therefore required for the final stage and the grid circuit must be tunable from 200 to 325 Mc/s to cover Band V. It was found that the same coil could be used over the complete frequency range in both the third and fourth multiplier stages, using copper or iron-dust cores for adjustment provided that, in the case of the final stage, different tapping points are used in different parts of the frequency range, and suitable changes are made to the grid coupling capacitor. Also, one part of the range requires the addition of a capacitor in parallel with the coil. Several changes are thus required in this scheme, but it was preferred to an earlier design in which, although there were fewer combinations over the band, the coil had to be changed.

One of the difficulties met was that, at certain frequencies, the provision of a reasonable drive voltage at the final multiplier grid did not always ensure adequate third harmonic output. This is believed to be due to resonance effects associated with the grid-anode capacity giving rise to a low impedance across the output at the third harmonic frequency; since the grid-to-earth impedance would form part of this low-impedance network, changes in the grid circuit arrangement could affect the output considerably. The circuit adjustments described above enabled this difficulty to be overcome.

4. Results of Performance Tests

The Band V s.f. unit has been found to operate satisfactorily at all frequencies within the specified range including the maximum frequency of 960 Mc/s. Results of performance tests at three representative frequencies are given below.

4.1 *Selectivity*

The frequency responses of the s.f. circuit of the Band V unit are shown in Fig. 3. Over a frequency range of ± 200 kc/s about the nominal frequency, which corresponds to the 400-kc/s pass-band of the main unit, the responses are within $0 \cdot 2$ dB of the value at band centre.

Table 2 gives the type and amplitude of spurious responses which occur when using the Band V unit. 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.

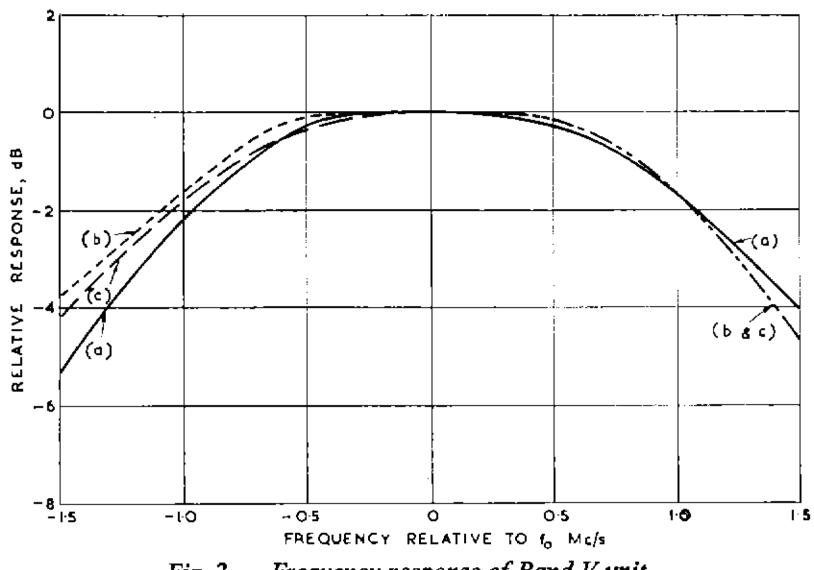


Fig. 3 — Frequency response of Band V unit.
(*a*) f_o=659·75 *Mc/s.*
(*b*) f_o=774·0 *Mc/s.*
(*c*) f_o=874·65 *Mc/s.*

These tests were made with the receiver operating on c.w. at a fixed gain. The open-circuit voltage of the reference signal applied to the input of the Band V unit was 10 μV. No i.f. attenuators were inserted.

4.2 *Sensitivity and Input Impedance*

Table 3 shows the measured values of noise factor, conversion loss, and input v.s.w.r. at the three test frequencies. The i.f. noise factor was $3 \cdot 2$ dB.

The earlier monograph[1] on the whole receiver includes typical calibration curves obtained for various conditions of operation. Such curves are largely determined by the characteristics of the main unit, except at low values of input signal where the noise factor has an important influence. The calibration curves for a receiver consisting of the Band V unit and the main unit are very similar to those obtained when the Band IV unit is in use. Curves for the latter case, when the special test transmissions described in the specification are being received, have been given in Part II of this monograph, which discusses the improvement in sensitivity which may be expected to be obtained by the use of increased a.f. gain and a correlation detector working at 1 kc/s.

4.3 *Local-oscillator Radiation*

Local-oscillator radiation has been measured in terms of the open-circuit voltage available at the aerial input socket and was found to be 1 mV. This test was carried out with the receiver tuned to receive a signal at 774 Mc/s, with the local oscillator working at a frequency of $763 \cdot 3$ Mc/s.

4.4 *Stability*

A Band V unit was tested in an oven over the temperature range 20°C to 50°C. The overall change of gain of the receiver, including the main unit which was operated at room temperature, was not more than $\pm 0 \cdot 3$ dB. The

TABLE 2

f_o = signal frequency
f_b = final local-oscillator frequency
f_c = crystal frequency
f_i = intermediate frequency
n = overall local-oscillator multiplication factor

Type of response	Input Voltage at spurious frequency relative to a 10 μv reference signal at f_o for 1-dB error, dB		
	$f_o = 659 \cdot 75$ Mc/s, $f_b > f_o$, $n = 36$	$f_o = 774 \cdot 0$ Mc/s, $f_b < f_o$, $n = 54$	$f_o = 870 \cdot 65$ Mc/s, $f_b < f_o$, $n = 54$
$f_b + f_i$(image)	43	—	—
$f_b - f_i$(image)	—	49	38
$(n-1)f_c - f_i$	>78	>90	>84
$(n-1)f_c + f_i$	48	>90	83
$(n+1)f_c - f_i$	>79	62	54
$(n+1)f_c + f_i$	>79	76	78
f_i (i.f. breakthrough)	71	62	>90
Other responses	>60	>60	>60

TABLE 3

Test	Measured value		
	$f_o = 659 \cdot 75$ Mc/s	$f_o = 774 \cdot 0$ Mc/s	$f_o = 870 \cdot 65$ Mc/s
Noise factor, dB	$10 \cdot 9$	$12 \cdot 2$	$11 \cdot 6$
s.f. unit loss, dB	$8 \cdot 8$	$7 \cdot 2$	$7 \cdot 0$
Input v.s.w.r.	$0 \cdot 72$	$0 \cdot 68$	$0 \cdot 78$

time required, after switching on, before the receiver had reached its final calibration, was approximately five minutes.

5. Conclusions

The results of the tests show that the performance of the Band V s.f. unit meets the main requirements derived from the specification of the whole VHF/UHF field-strength recording receiver. In particular the noise factor of less than 13 dB is satisfactory.

The unit is, in effect, a further development of the design of the Band IV unit and could, if necessary, be modified easily to operate in Band IV.

The adoption of a two-resonator signal-frequency tuning system has provided a very useful improvement in the image-frequency rejection performance as compared with that of the Band IV unit. The success in extending the range of the frequency multiplier sub-unit to cover Band V without resorting to the use of special UHF valves is very satisfactory from an operational point of view.

Part II

A CORRELATION DETECTOR

6. Introduction

The field-strength recording receiver,[1] referred to in Part I of this monograph, is employed for propagation measurements where the field strength may be very low, requiring directional aerials and post-detector selectivity to improve the signal-to-noise ratio. The lowest measurable signal level is, in practice, $0 \cdot 4\ \mu V$ in Band III, $1 \cdot 0\ \mu V$ in Band IV, and $1 \cdot 1\ \mu V$ in Band V, these being open-circuit values from a 75-ohm source.

For some applications a higher sensitivity would be desirable (particularly in the higher frequency bands) provided that it could be obtained by a simple optional addition resulting in no reduction of the range of measurable signal levels.

In order to explain the problems involved, the original receiver is first described before dealing with its limitations and methods of overcoming them. For convenience, the figures given are those for the Band IV receiver, but the same limitations and methods apply to the Band III and Band V receivers.

7. Description of the Original Band IV Receiver

The receiver is a superheterodyne using automatic gain control to enable a large range of signal levels to be recorded. Its primary function is to record transmissions which are modulated by a 1,000-c/s square wave; the 1,000-c/s fundamental is extracted after detection and rectified to produce the recorded output. A schematic diagram is shown in Fig. 4.

The recording meter gives a full-scale deflection of 1 mA when the r.m.s. level of the modulated input signal is $700\ \mu V$ (open-circuit voltage from a 75-ohm source); an input of $0 \cdot 7\ \mu V$ gives an output of $0 \cdot 03$ mA. For a 60-dB change of input signal, therefore, the receiver d.c. output changes by 30 dB. The recording law is roughly logarithmic above $0 \cdot 05$ mA, as shown in Fig. 5, curve (a).

The noise bandwidth of the receiver is about 600 kc/s, and its noise factor is 11 dB, or $12 \cdot 5$ in power ratio. The r.m.s. noise voltage in the i.f. amplifier is therefore $\sqrt{12 \cdot 5}$ times that due to a 75-ohm source resistor at 290°K in a bandwidth of 600 kc/s. This is given by the expression $\sqrt{4KTR\Delta f}$ (where K is Boltzmann's Constant), and is $0 \cdot 85\ \mu V$ r.m.s. The noise level in the i.f. amplifier is therefore equivalent to a noise input of $3\ \mu V$ r.m.s. In this condition, it can be shown that an unmodulated signal input of $3\ \mu V$ r.m.s. will produce an increase of about 3 dB in the d.c. output of the i.f. detector.[2] However, by using a modulated transmission and post-detector selectivity, the r.m.s. value of the modulated signal input required to produce a 3-dB rise in the d.c. output of the a.f. detector is reduced to $0 \cdot 35\ \mu V$, representing a signal-to-noise ratio improvement of 19 dB. The mechanism of this is as follows.

The i.f. detector is followed by a narrow-band filter of 30 c/s noise bandwidth centred on the modulation frequency of 1,000 c/s; its output is amplified and rectified, the rectified current forming the output of the receiver. If a signal produces an output of 3 dB above the no-signal value, the signal-to-noise ratio must be 0 dB at the output of the a.f. filter. Owing to the bandwidth reduction of the filter, the signal/noise ratio at the i.f detector output must then be lower; its value can be shown to be −38 dB. Unfortunately, this does not mean that the input signal is 38 dB below noise level, but only 19 dB, since, in the presence of such a relatively large noise voltage, the i.f. detector—a so-called 'linear diode'—behaves as a square-law detector.[1]

8. Limitations of the Original Design

The range of measurable signal levels is limited at both extremes partly because, in the interests of a stable calibration, a.g.c. is applied to one stage only. As the signal approaches the upper limit, the a.g.c.-stage gain falls to a minimum and both the i.f. amplifier preceding the a.g.c. stage and that preceding the i.f. detector approach their overload conditions. When the signal falls to the lower limit, the gain reaches maximum and the gain of the last i.f. amplifier cannot be further increased without causing

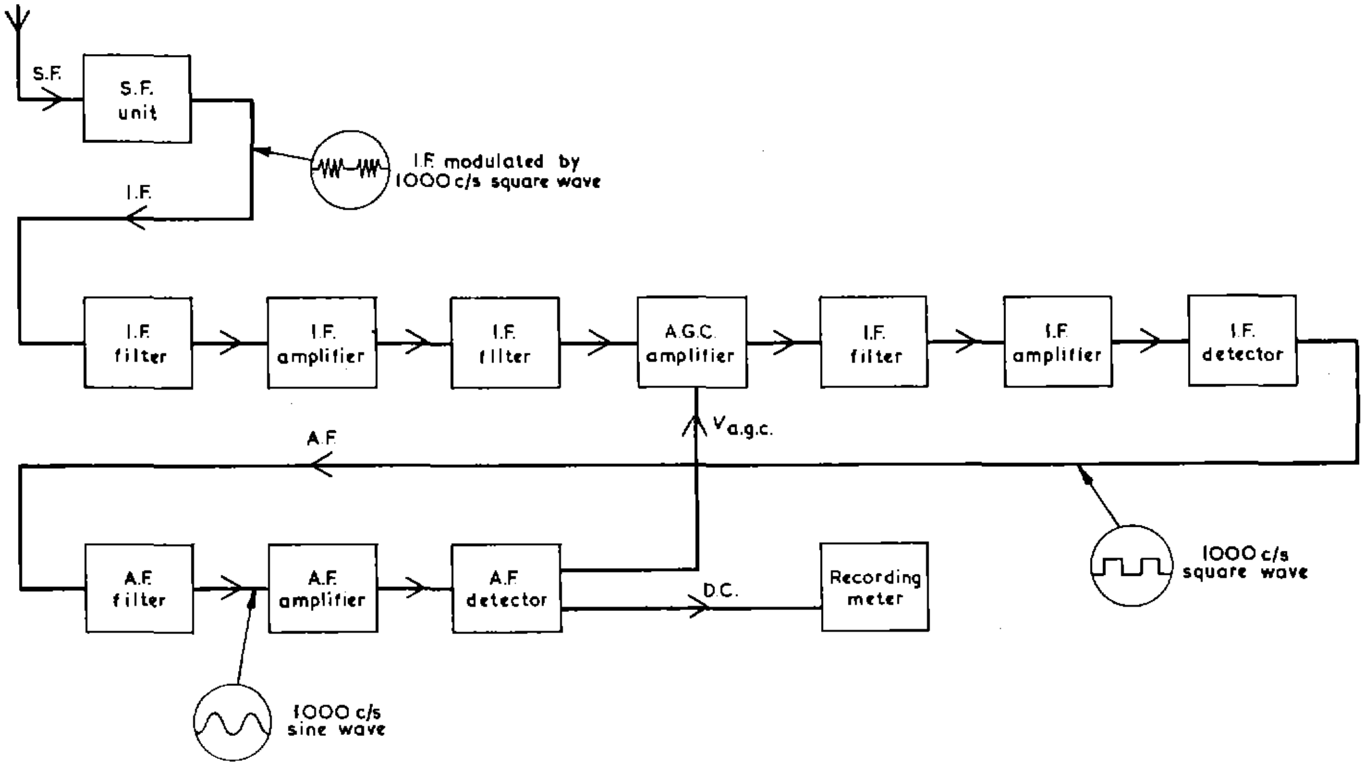


Fig. 4 — Schematic diagram of VHF/UHF receiver.

overloading on peaks of noise. This noise has the full bandwidth of the receiver, whereas the receiver output only responds to those noise components in the narrow bandwidth of the a.f. filter. In fact, the d.c. output due to noise alone is only about 1 per cent of full output and is barely perceptible on the output record. The same limitation applies to the Band III receiver, in which the i.f. gain has to be reduced on account of the higher gain of the Band III r.f. unit.

In practice, the lowest measurable signal is about 1 μV r.m.s., giving 5 per cent of full output; below this level the curvature of the characteristic renders measurement inaccurate. It is evident, then, that the effective sensitivity is limited by lack of gain. This may be remedied as will be described later, but it is sometimes in the interests of accuracy to minimize the output due to noise. If there is a large output indication due to noise alone, some of the available output scale is wasted; if the noise deflection is biased to zero the stability of the bias must be very high.

When adequate gain without overloading is ensured, the sensitivity for practical purposes is given by the input signal at which the a.f. signal-to-noise ratio is approximately unity. This is because the d.c. output of the final detector becomes insensitive to signal changes for signal/noise ratios less than unity. In the present application this limit of sensitivity is actually the governing one; fluctuations of output due to noise are considered later.

The a.f. signal-to-noise ratio is, of course, dependent on many factors, and the fact that the signal-to-noise ratio at the i.f. detector is already less than unity over an appreciable part of the recording range leads to some unexpected results. Thus the input signal voltage for a given a.f. signal-to-noise ratio varies only as the fourth root of either the i.f. bandwidth or the a.f. bandwidth, so that halving either bandwidth affords an improvement in sensitivity of only 1·5 dB. Substantial bandwidth reductions are made difficult by problems of frequency stability. The noise factor of the receiver, which controls the sensitivity directly, is already within 2 dB of the lowest practicable. It is true that a large improvement would be possible if a coherent i.f. detector were used, but (in the absence of suitable communication between sender and receiver) the improvement would be stultified by the time required to synchronize the frequency and phase of an oscillator.

One further factor affecting the measurement of small signals must be mentioned. The response time of the receiver—determined primarily by the bandwidth of the a.f. filter—is about 0·02 sec, whereas the recording meter has a response time of about 0·3 sec. This means that the r.m.s. value of the fluctuating part of the noise output of the receiver is attenuated by some 10 dB in the meter itself, so that the meter behaves as a second stage of post-detector selectivity. The detailed theory has been given elsewhere;[3] but the result is that if the d.c. output is in-

(*a*) -·-·- *Original Band IV receiver.*

(*b*) - - - - *Original Band IV receiver with 20 dB extra a.f. gain.*

(*c*) — — *Original Band IV receiver with 20 dB extra a.f. gain and extended a.g.c. range.*

(*d*) —— *Original Band IV receiver with correlation detector unit.*

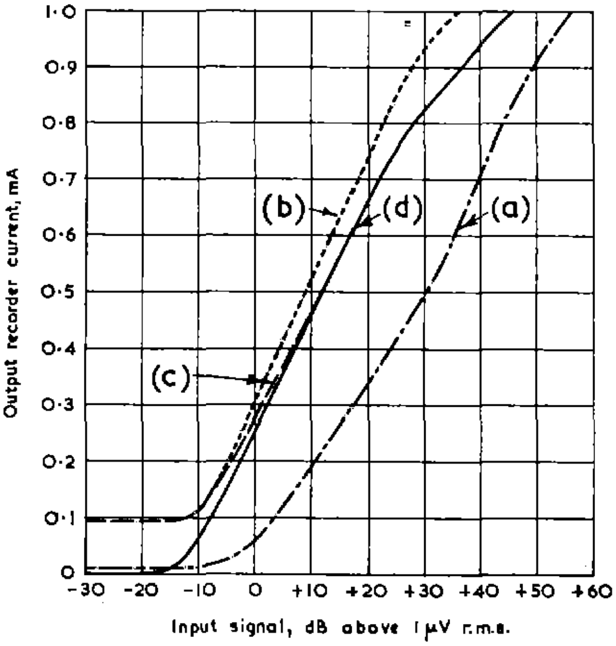


Fig. 5 — *Output calibration of Band IV receiver.*

creased from 1·0 to 1·4 by applying an input signal, the r.m.s. value of the fluctuation on the recording meter is about 0·2.

To summarize, the sensitivity of the original receiver in Bands IV and V is not limited by the presence of noise, but by the lack of gain. In practice, the lowest measurable signal is about 1 μV; but if sufficient gain were available the signal could be reduced to about 0·4 μV before its measurement became impaired by noise—any further reduction in noise level is impracticable.

9. Problems of Increasing the Sensitivity

9.1 *Increasing Gain*

As the final i.f. stages are nearly overloaded by noise at maximum gain, the requisite increase in gain must be obtained in the a.f. stages.

An increase in gain of about 20 dB is required to raise the d.c. output due to noise to a level at which a 3 dB change can be observed. From Fig. 5 curve (b), it can be seen that the increased sensitivity at low levels is at the expense of a reduction of 20 dB in the maximum signal, with a consequent reduction in the total range of measurable signals.

9.2 *Modifying a.g.c. Range*

Having increased the a.f. gain, the receiver could be made to handle the same maximum signal as before if the minimum gain of the a.g.c. stage were reduced by 20 dB. This could be achieved by using a maximum negative a.g.c. voltage of about 40 V instead of 30 V. Unfortunately, the gain characteristic of the a.g.c. valve (6BA6) has an undesirable curvature at grid voltages of between −35 V and −40 V. By using a maximum of −35 V, a 10-dB reduction in minimum gain can be made without introducing excessive curvature. The result is shown in Fig. 5, curve (c).

Comparing curves (a) and (c), the signal required to produce a 0·05-mA change in d.c. output has been reduced by about 7 dB, with a 10-dB reduction at 1·0 mA. This represents a considerable improvement, though 10 per cent of the output meter deflection is now occupied by the d.c. output due to noise.

9.3 *Suppression of the d.c. Output due to Noise*

The a.f. noise spectrum, narrow though it is, extends over a considerably greater band of frequencies than that occupied by the a.f. signal. It is possible, therefore, to arrange that noise components at the edges of the a.f. passband produce a d.c. output opposing that produced by the components near the 1,000-c/s signal frequency. The immediate advantage of such a system is that the d.c. output may be made zero in the absence of signal in spite of changes in the noise input due, for example, to wide-band interference. This allows the signal calibration to occupy the whole output meter deflection and further improves the sensitivity, as shown in Fig. 5, curve (d).

This result can be achieved by using a suitable adaptation of a circuit known as an auto-correlation detector; the modified form described in this monograph will be referred to as a correlation detector. An experimental detector of this type is described in the next section; the theory of its operation is given in the Appendix.

10. An Experimental Correlation Detector

10.1 *General Description*

A unit was designed to combine the features discussed in Section 9, namely increased a.f. gain, increased a.g.c. voltage range, and noise suppression by using a correlation detector. The unit, with its own power supplies, can be mounted in the same cabinet as the existing receivers and, when desired, can be put into operation by changing a few leads.

The circuit consists essentially of a phase-shift network and a phase-sensitive detector. A schematic diagram is shown in Fig. 6, its operation is illustrated in Fig. 7, and a circuit diagram is shown in Fig. 8.

The input, which is taken from the existing a.f. amplifier, consists of a narrow spectrum of noise centred on a signal at 1,000 c/s. The network introduces a phase shift, ϕ, varing from 0 to 2π radians throughout the a.f. pass-band, with a value of π radians at 1,000 c/s. The phase-sensitive detector is arranged to give a positive d.c. output in the range $\phi=\pi/2$ to $3\pi/2$ and a negative output elsewhere. The a.f. response of the system is therefore positive in the neighbourhood of 1,000 c/s and negative at the edges of the pass-band. By choosing the correct phase slope, the d.c. output due to the noise components in these two domains can be made to cancel. There is, in addition, a reduction in the r.m.s. noise current output, due to the effectively reduced bandwidth of the above system.

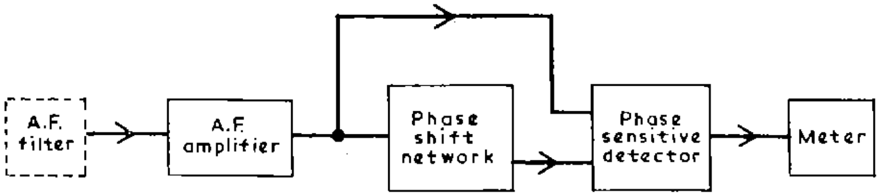


Fig. 6 — Schematic diagram of correlation detector.

(*a*) *Characteristic of phase-shift network.*

(*b*) *Response of phase-sensitive detector without a.f. filter.*

(*c*) *Response of a.f. filter.*

(*d*) *Overall response of correlation detector.*

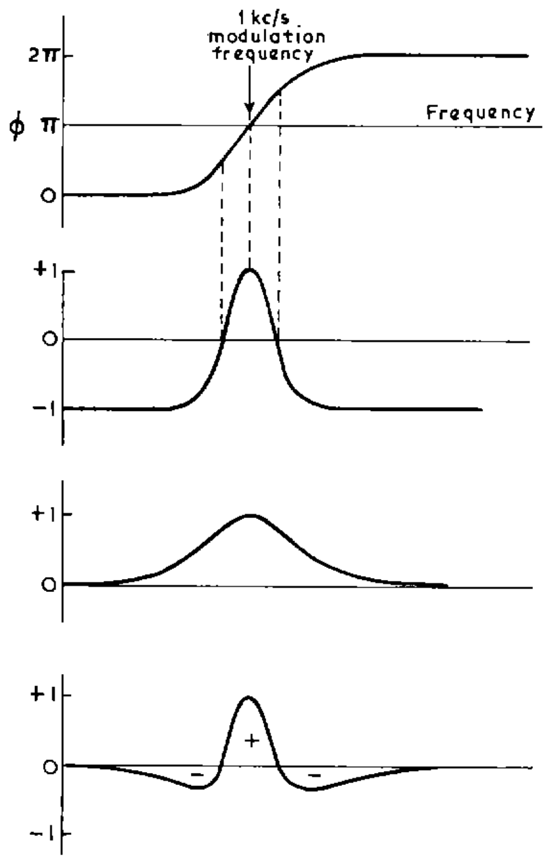


Fig. 7 — Operation of correlation detector.

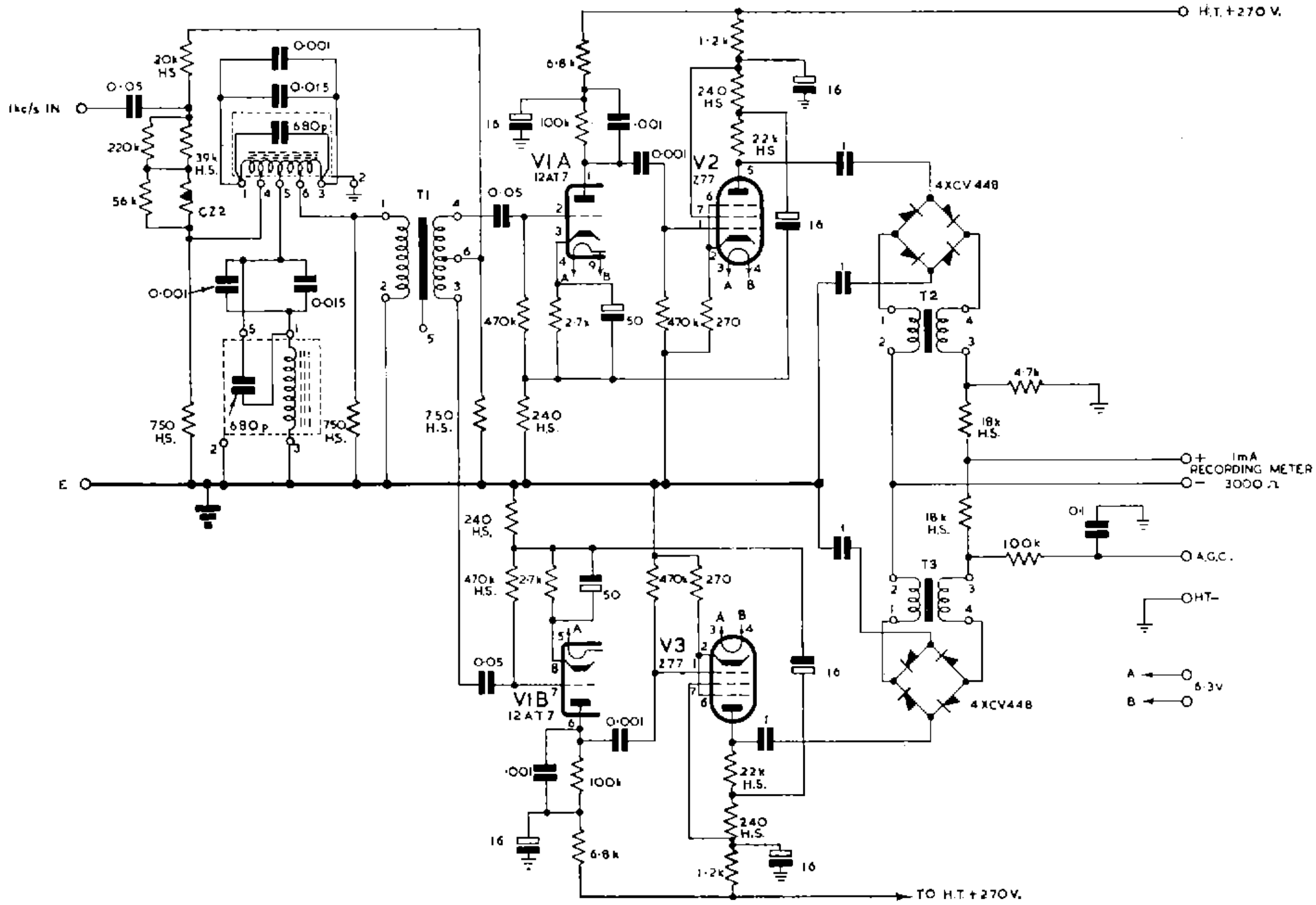


Fig. 8 — Circuit diagram of correlation detector.

10.2 *The Phase-shift Network*

The input signal is derived from the existing a.f. amplifier, working at increased gain, and provides two outputs. One output passes directly to the phase-sensitive detector, as a reference voltage, and the other traverses the phase-shift network.

The phase shift is obtained from an all-pass network known as a second-degree phase equalizer. As shown in Section 13.4.1, in order to obtain zero d.c. output in the absence of signal the reciprocal of the parameter s, which determines the phase slope, must be equal to the effective Q of the a.f. filter, namely 40 in the case considered. Owing to loss in the elements, the network has appreciable attenuation near 1,000 c/s, where the phase slope is greatest. This reduces the positive contribution to the d.c. output in this region and it is necessary to increase s to about $\frac{1}{30}$ in order to broaden the positive domain and so redress the balance. Adequate Q values (of the order of 100) were obtained for the inductors of the network by using Ferroxcube cores.

The voltage across the inductor in the centre branch of the network is $1/s$ times that at the input. The input voltage

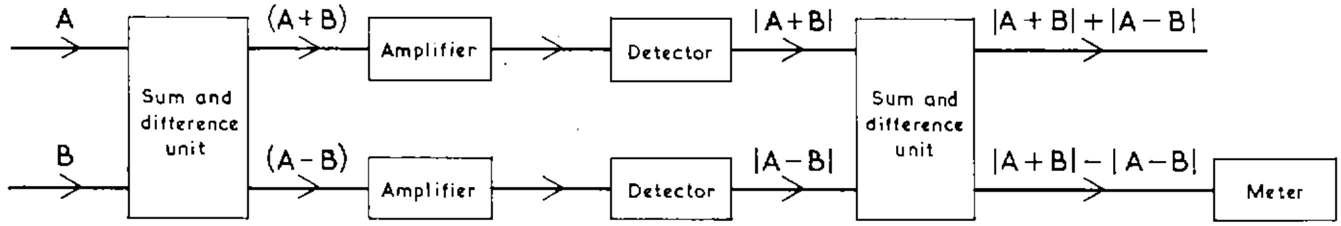


Fig. 9 — Schematic diagram of phase-sensitive detector.

is kept below 0·5 V in order to avoid the effect of core saturation in this inductor, and the subsequent amplification required is embodied in the phase-sensitive detector.

10.3 *The Phase-sensitive Detector*

A schematic diagram, Fig. 9, shows the operation of the detector. The output of the a.f. amplifier, voltage A, is passed through the phase-shift network to produce voltage B. B is then added to and subtracted from A in a centre-tapped transformer ($T1$ in Fig. 8). Voltages $(A+B)$ and $(A-B)$ are then amplified and rectified separately, the difference between the rectified currents forming the output for the recording meter. For sinusoidal voltages, the d.c. output is a maximum when A and B are in phase or in antiphase, and is zero when they are in quadrature. As shown in Section 13.4.3, by making the magnitude of B about half that of A, this process gives an output nearly proportional to the cosine of the phase difference of A and B, as in a multiplier, but a response linearly related to the amplitude of the input. Since the gain is all obtained before rectification, the necessary stability is achieved by negative feedback.

10.4 *The Detector Load and a.g.c. Output*

The two rectifier circuits, as shown in Fig. 8, are connected to their loads through the output meter as a common 3,000-ohm resistance. The d.c. coupling between the circuits is negligible; the a.c. coupling between the rectifiers and between each rectifier and earth is minimized by the bifilar chokes $T2$ and $T3$.

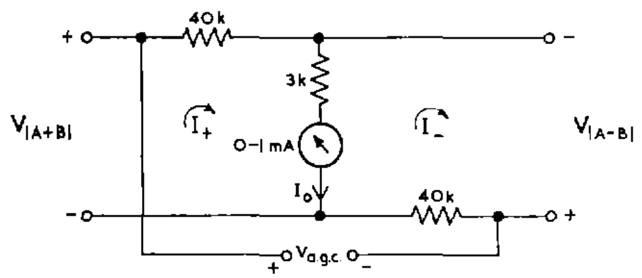
Now the outputs of the rectifier circuits are proportional to the moduli of $(A+B)$ and $(A-B)$. Fig. 10(a) shows how the a.g.c. voltage and the recorder current may be derived, both being proportional to the difference of these moduli. Fig. 10(b) shows an alternative arrangement as used in the circuit of Fig. 8; here the recorder output remains as before but the a.g.c. voltage is proportional to the sum of the moduli of $(A+B)$ and $(A-B)$. These two circuit arrangements produce different effects, which are best explained by referring to the modulation-frequency responses of the circuits concerned.

The difference between the rectifier voltages varies with frequency in the same way as the recorded output, as illustrated in Fig. 7(d); the response characteristic is narrower than that of the a.f. filter alone, due to its dependence on phase, and is reversed in sign at the edges of the pass-band. The sum of the voltages, however, is almost independent of phase, as explained in Section 13.4.3; its response characteristic is simply that of the a.f. filter, as in the original receiver, and does not reverse its sign.

The curves of Fig. 11 represent the error incurred in measuring a signal if there is a relative drift between the tuning of the a.f. circuits and the modulating frequency. They are equally valid for this purpose whether the receiver is working at fixed gain or using a.g.c., provided that the a.g.c. voltage is proportional to the output indication. Thus, curve (a) represents the error for the original receiver and curve (b) for the receiver with the a.g.c. derived as shown in Fig. 10(a); they also represent the modulation-frequency response determining the most rapid signal fluctuations which can be faithfully recorded, assuming no limitation in the recorder itself.

If the circuit of Fig. 10(b) is used, the a.g.c. voltage is no longer uniquely related to the output. In this case the output responds to a frequency drift to a greater extent than the a.g.c. voltage so that the overall error is exaggerated, as shown in Fig. 11, curve (c). The stability of the a.f. circuits, particularly that of the phase-shift network, must therefore be high to avoid appreciable error. However, the response to rapid signal fluctuation is still determined by curve (b) in Fig. 11.

If the circuit of Fig. 10(a) is used, the required stability of the a.f. circuits is not so great as for the circuit of Fig. 10(b), being little more than that required for the original a.f. filter; but another effect arises in this case. A signal at the edge of the pass-band would give an output of reversed polarity, and, if the a.g.c. voltage follows the output, it also will be reversed in sign. It is possible for the receiver to become completely paralysed by a sharply rising signal of the correct modulation frequency, if the a.f. circuits overload before the a.g.c. has reacted. Harmonics of the modulation generated between the a.f. filter and the phase-shift network will produce a reversed output, reduce the a.g.c. voltage, and enhance the paralysis. This effect would

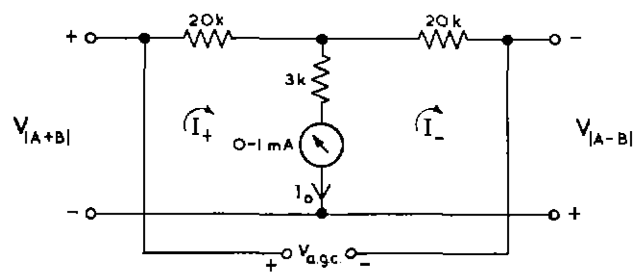


Max. d.c. values:

I_o	1·0mA
I_+	1·5mA
I_-	0·5mA
$V_{a.g.c.}$	40V

Max. a.c. power input to (A+B) detector: 110mW

(*a*) *a.g.c. voltage proportional to output current.*



Max. d.c. values as for (a)
Max. a.c. power input to (A+B) detector: 55mW

(*b*) *a.g.c. voltage proportional to signal plus noise.*

Fig. 10 — *Alternative methods of deriving a.g.c. voltage.*

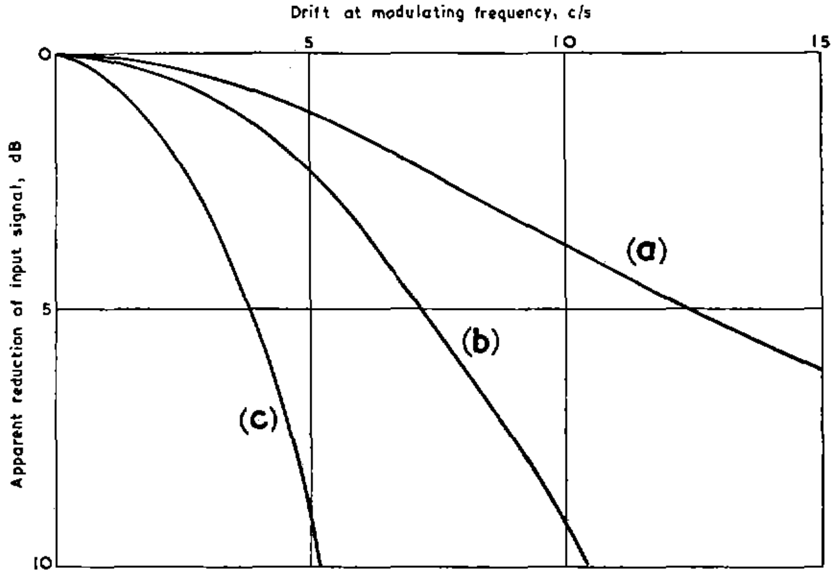


Fig. 11 — Effect on the calibration of drift at modulating frequency.

not occur if the amplitude of the signal were limited before passing through the a.f. filter.

10.5 *Choice of a.g.c. Circuit*

As can be seen from Fig. 10, the circuit (b) has the advantage of requiring only half the detector power input. Tests on the experimental unit using this circuit indicated that the required stability could be achieved without undue complication; in addition, the steep slope of the output calibration, shown in Fig. 5(d), was preserved at a lower signal level than with the alternative circuit. These considerations led to the adoption of the a.g.c. circuit of Fig. 10(b) for the final detector.

11. Conclusions

The effective sensitivity of the VHF/UHF receiver in its original form can be increased by about 10 dB by the use of increased a.f. gain and a correlation detector working at 1 kc/s; about 7 dB improvement is possible by the increased gain alone. The new detector also gives a stable zero of d.c. output in the absence of signal, even when the noise input is varying due, for instance, to wide-band interference.

A modified form of auto-correlation detector has been constructed in which an all-pass phase-shifting network has been used instead of a constant-delay network, thus simplifying the circuit requirements.

12. References

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13. Appendix

13.1 *Optimum Positions for the Coupling Elements in a Pair of Coaxial-line Resonators*

The arrangement considered is that used for the Band V signal-frequency tuner described in Section 3 and shown diagrammatically in Fig. 12. It is assumed that the impedances of the source and load are purely resistive and independent of the tuned frequency, f_o, of the resonators. It is also assumed that the losses of the resonators are negligible. It may be shown that this arrangement is equivalent to the circuit of Fig. 13, the components of which are found by evaluating the equivalent elements of Fig. 12 at the short-circuited ends of the coaxial lines.

Neglecting disturbances of the field in the resonators due to the coupling elements and line imperfections, the current in the line follows a cosine law. Therefore a mutual inductance M placed at a distance x from the short-circuited end of a resonator is represented in Fig. 13 by a value $M \cos\theta$ where

$$\theta = \frac{2\pi f_o x}{v}$$

and v is the velocity of propagation in the line.

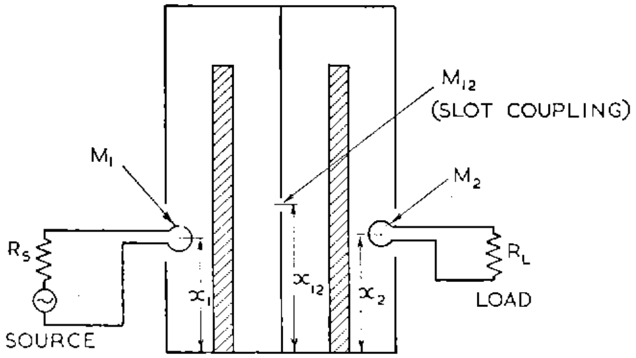


Fig. 12 — *Diagrammatic representation of signal-frequency cavities.*

In the tuner it was desirable that, for any chosen value of f_o,

(*a*) the shape of the response curve should be close to that of a transitionally coupled (maximally flat) pair of circuits, and

(*b*) the bandwidth, B, at the half-power points of the response curve should be substantially constant.

If possible, these requirements should be achieved without making any mechanical adjustment to the coupling elements when f_o was changed.

It is shown in Section 13.2 that, when the losses of the tuned circuits are negligible, the optimum power match is obtained when the damping from the source and load gives the two circuits the same magnification factor, Q, and we shall for convenience consider this case only. The coupling coefficient between the two circuits, referring to Fig. 13, is

$$k=\frac{M_{12}\cos\theta_{12}}{L}$$

and, if

$$kQ=1 \tag{1}$$

the response is maximally flat and has a half-power bandwidth of

$$B=\sqrt{2}\frac{f_o}{Q}. \tag{2}$$

To satisfy the conditions (*a*) and (*b*) above, the bandwidth and response shape are only independent of f_o if

$$Q\propto f_o \tag{3}$$

and

$$k\propto\frac{1}{f_o}. \tag{4}$$

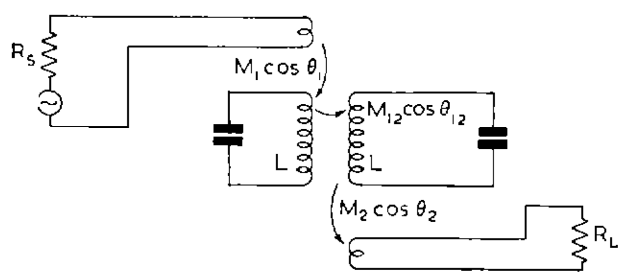


Fig. 13 — *Equivalent circuit of signal-frequency cavities.*

The inductance L in Fig. 13 can be shown[4] to have a reactance

$$2\pi f_o L=\frac{\pi Z_o}{4} \tag{5}$$

where Z_o is the characteristic impedance of the line.

The magnification factor of this circuit is

$$Q=\frac{2\pi f_o L}{R}=\frac{\pi Z_o}{4R} \tag{6}$$

where R is the effective series resistance coupled into either circuit from the source or load. Since the circuits have the same values of Q and Z_o, they have the same values of L and R. R can be shown to be

$$R=(2\pi f_o M_1 \cos\theta_1)^2/R_S$$
$$=(2\pi f_o M_2 \cos\theta_2)^2/R_L$$

and if, as is convenient, $x_1=x_2=x$ in Fig. 12, then

$$\theta_1=\theta_2=\theta=\alpha f_o$$

where

$$\alpha=\frac{2\pi x}{v}$$

and so, from equation (6),

$$Q\propto\frac{1}{f_o^2\cos^2\alpha f_o}. \tag{7}$$

From Fig. 13 and equation (5)

$$k=\frac{M_{12}\cos\theta_{12}}{L}=\frac{8f_o M_{12}\cos\theta_{12}}{Z_o}.$$

Therefore

$$k\propto f_o\cos\alpha_{12}f_o \tag{8}$$

where

$$\alpha_{12}=\frac{2\pi x_{12}}{v}.$$

From (7) and (8) we have

$$kQ\propto\frac{f_o\cos\alpha_{12}f_o}{(f_o\cos\alpha f_o)^2} \tag{9}$$

and from (2) and (7)

$$B\propto f_o^3\cos^2\alpha f_o. \tag{10}$$

Equation (10) is strictly true only if the transitionally coupled condition is maintained.

Although kQ and B are not independent of frequency as desired, the variation may be reduced by operating in the region where the rate of change with respect to f_o is zero. Assuming that, by this means, the variation of kQ can be kept small, equation (10) may be used to find the condition for zero rate of change of bandwidth. By differentiating (10) with respect to f_o the required condition is found to be

$$\theta=\alpha f_o=57°. \tag{11}$$

Then, assuming that the constancy of expression (10) is fulfilled, equation (9) may be written

$$kQ\propto f_o^3\cos\alpha_{12}f_o \tag{9a}$$

and the condition for zero rate of change of kQ with f_o is

$$\theta_{12}=\alpha_{12}f_o=62°. \tag{12}$$

If, therefore, x and x_{12} are so chosen that α and α_{12} fulfill equations (11) and (12) when f_o is equal to the mean frequency of Band V, a near-optimum performance will be obtained. The bandwidth, B, at the extreme frequencies of Band V is reduced by only 30 per cent of its value at the mean frequency and kQ lies between the values 1 and 0·935 over the whole band. These results may be compared with the situation when the coupling elements are placed as near as practicable to the short-circuited end, when there would be an increase of bandwidth by a factor of more than three and a reduction of kQ from 1 to 0·64 when f_o is changed from the lowest to the highest frequency of Band V.

The value of x used in the Band V tuner is that given by the above analysis but the value of x_{12} is rather less than the optimum. The effect is to allow a slight increase in bandwidth at the higher frequencies, and to require the ratio of external damping to the small internal circuit loss to vary less over the frequency range than in the theoretically optimum design.

13.2 *Effect on Impedance Matching of the Frequency Response Characteristic of a Pair of Coupled Circuits*

For a discussion of the response characteristic we consider the simplified representation of the receiver input circuit given in Fig. 14. This shows a pair of coupled circuits in which resistive components due to both circuit losses and external connexions are represented by conductances across the tuned circuit. The damping of the first circuit has been represented by two components, namely G_s arising from the aerial circuit and G_d which represents loss in the tuned circuit itself. The terminal pair AA' therefore represents, in an idealized form, the input connexion to the receiver. In the second circuit, all losses (including the effect of the load) are represented by a conductance G_2. If we take $G_1=G_s+G_d$, the input and output voltages, V_1 and V_2, may be expressed as follows in terms of the input current I:

$$V_1=\frac{I}{G_1}\cdot\frac{1+jx}{(1+jx)(1+jqx)+n^2}$$

$$V_2=\frac{I}{G_1}\cdot\frac{-jn\sqrt{p/q}}{(1+jx)(1+jqx)+n^2}.$$

It is assumed that the combinations C_1, L_1, and C_2, L_2 are both tuned to the frequency $\omega_o/2\pi$ and that $L_m^2 \ll L_1L_2$. The other symbols have the following significance:

$$k=L_m/\sqrt{L_1L_2}$$

$$n=k\sqrt{Q_1Q_2} \text{ where } Q_1=\frac{\omega_o C_1}{G_1} \text{ and } Q_2=\frac{\omega_o C_2}{G_2}$$

$$p=L_1/L_2=C_1/C_2$$

$$q=Q_1/Q_2$$

$$x=2Q_2\cdot\frac{d\omega}{\omega_o},$$

where $\omega_o+d\omega$ is the angular frequency with $d\omega \ll \omega_o$.

At the mid-band frequency ω_o, $x=0$ and we have

$$V_1=\frac{I}{G_1(1+n^2)}.$$

Thus the admittance presented to the generator is $G_1+n^2G_1$. In the case of a receiver fed from an aerial which has been matched to the feeder, the generator conductance G_s represents the characteristic admittance of the feeder.* The admittance presented by the circuit to the right of AA' is $G_1+n^2G_1-G_s$ or $(1+n^2)G_d+n^2G_s$ and the degree of

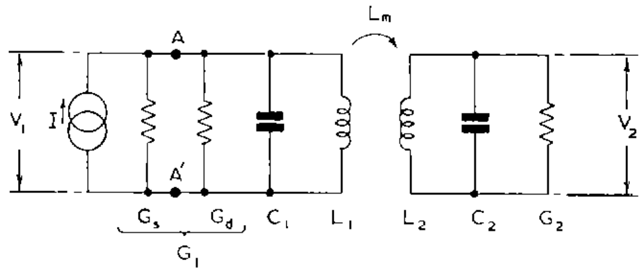


Fig. 14 — *Representation of coupled circuits, including losses.*

matching corresponds to a standing-wave ratio equal to the ratio of this conductance to the feeder impedance, i.e.

$$\text{s.w.r.}=(1+n^2)\frac{G_d}{G_s}+n^2.$$

A s.w.r. of unity (perfect match) requires

$$n^2=\frac{G_s-G_d}{G_s+G_d}.$$

On the other hand it may be shown that the condition for a frequency response which is maximally flat is

$$n^2=\tfrac{1}{2}\left(\frac{q+1}{q}\right).$$

It will be seen that the first condition, assuming a finite circuit loss ($G_d>0$), requires $n<1$ while the second condition requires $n \geqslant 1$. It is thus impossible to obtain a perfect match and a maximally flat response simultaneously, although both conditions can be nearly fulfilled if G_d is kept small compared with G_s and the circuit Q factors are equal ($q=1$). In general, adjustment of n for a good match results in a single-peak response corresponding to an 'undercoupled' pair of tuned circuits, while the adjustment for a double-peak or 'overcoupled' response will tend to degrade the input s.w.r.

It has hitherto been assumed in the discussion that the aerial has been matched to the feeder so that the effective generator admittance is a conductance of constant value G_s. It is, of course, apparent that a mismatch of the aerial will change the admittance presented to the first circuit and may affect the response by detuning of the first circuit, or by a change of conductance which will affect the normalized coupling parameter n. It is for this reason that it is recommended that the match of the aerial to the feeder should be as good as possible (s.w.r. at least 0·75), even though the receiver may be allowed to present a somewhat poorer match (s.w.r. not less than 0·67) in order to obtain a frequency response close to the maximally flat condition.

13.3 *The Principle of the Correlation Detector*

If either the time-function $I(t)$ or its power spectrum $W(f)$ is known, the auto-correlation function $\psi(\tau)$ can be found; this is defined[5] by

$$\psi(\tau)=\underset{T\to\infty}{Lim}\frac{1}{T}\int_o^T I(t).I(t+\tau).dt \qquad (13)$$

and is related to the power spectrum $W(f)$ by

$$\psi(\tau)=\int_o^\infty W(f)\cos(2\pi f\tau).df. \qquad (14)$$

* In practice, the input connexion will be more nearly represented by a tapping on the first circuit inductance so that the generator conductance will be transformed by a fixed ratio. To simplify the discussion, this transformation has been omitted.

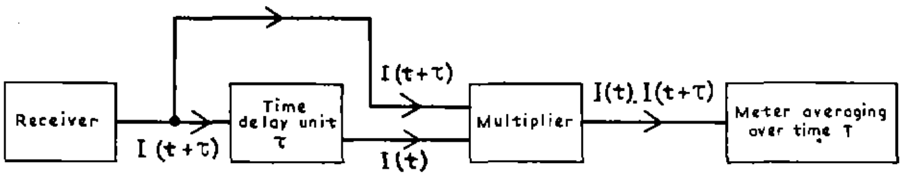


Fig. 15 — Mechanism of basic auto-correlation detector.

The operation of equation (13) can be performed by the circuit of Fig. 15 where the meter indicates the value of the auto-correlation function of a current derived from the output of a receiver. If $I(t)$ is a sinusoidal signal of frequency f_0, the response of the circuit may be made a maximum by a suitable choice of the delay time, τ. In the absence of signal, the noise output from the receiver has a power spectrum governed by the characteristics of the filters in the receiver; these would be chosen, where possible, to make the response $\psi(\tau)$ zero. Two types of filter are discussed below.

13.3.1 *Result for a Rectangular Band-pass Filter*

If the power spectrum, $W_1(f)$, of the noise output from a filter is uniform and equal to P_o in the interval $(f_o-\frac{1}{2}\Delta f_o)<$ $<(f_o+\frac{1}{2}\Delta f_o)$ and zero elsewhere, then the auto-correlation function is

$$\psi_{N1}(\tau)=\frac{P_o\sin(\pi\Delta f_o\tau)}{\pi\tau}\cos(2\pi f_o\tau). \qquad (15)$$

As can be seen from Fig. 16(a), this function has a decaying oscillatory envelope with repeated zeros

$$\psi_{N1}(\tau)=0 \text{ at } \tau=\frac{n}{\Delta f_o} \quad n=1, 2, 3. \qquad (16)$$

A sinusoidal signal of frequency f_o and amplitude A will have an auto-correlation function $\psi_s(\tau)$ with repeated maxima given by

$$|\psi_s(\tau)|=\tfrac{1}{2}A^2 \text{ at } \tau=\frac{m}{2f_o} \quad m=1, 2, 3. \qquad (17)$$

as shown in Fig. 16(b).

Thus, if f_o is chosen so that

$$\frac{\Delta f_o}{f_o}=\frac{2n}{m}$$

both equations (16) and (17) may be satisfied with the same

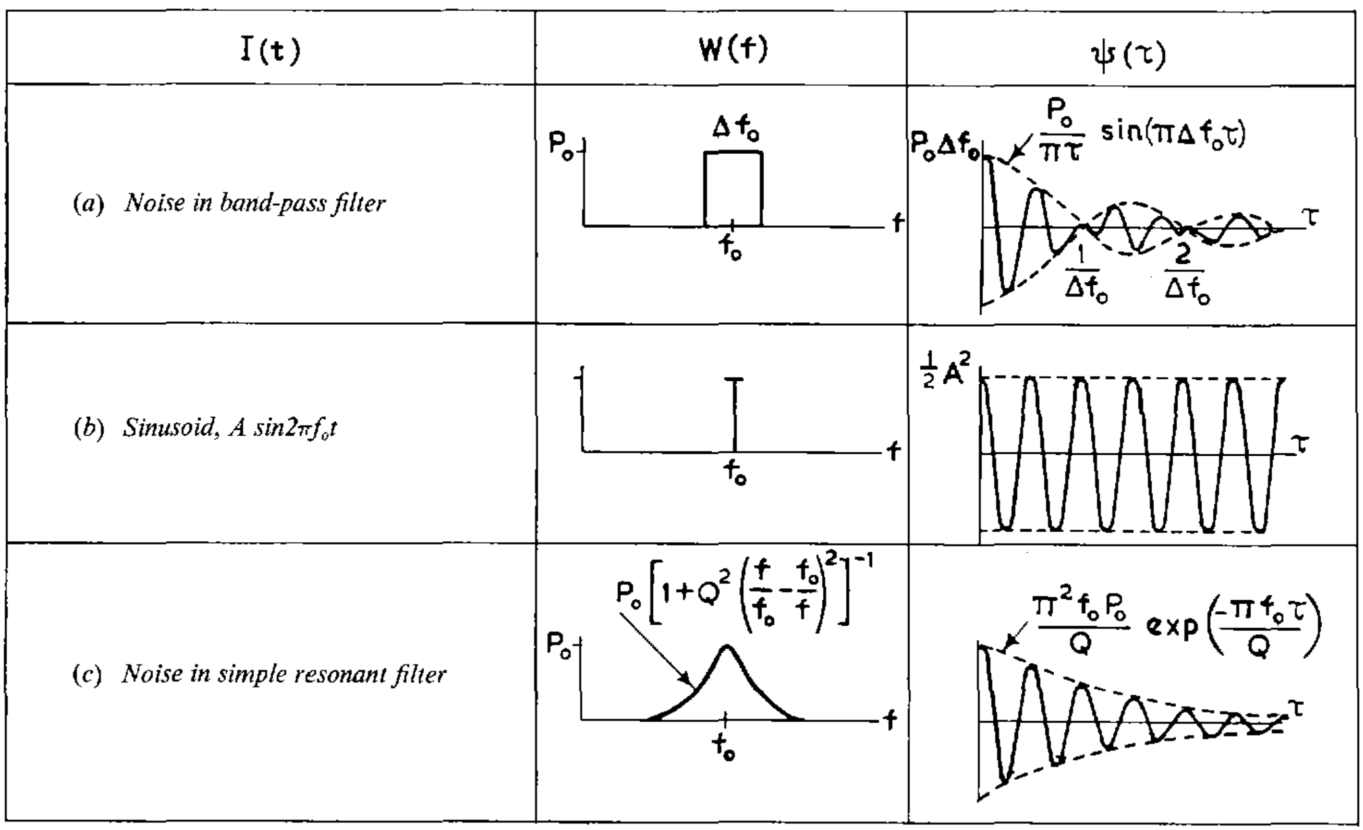


Fig. 16 — Auto-correlation functions for various types of time function.

value of τ. The output of the detector in Fig. 15 will then be proportional to the mean square value of the signal but identically zero in the presence of noise alone. In practice, $\Delta f_o \ll f_o$, and n is usually unity, in which case

$$\tau = \frac{1}{\Delta f_o} = \frac{m}{2 f_o}.$$

13.3.2 *Result for a Resonant Filter*

The field-strength recording receiver embodies an output filter consisting of a series resistor and a shunt LC circuit, resonant at f_o. Assuming no other restriction, if the input noise to this filter has a uniform power spectrum, P_o, the output power spectrum in the presence of noise only is

$$W_2(f) = \frac{P_o}{1 + Q^2\left(\frac{f}{f_o} - \frac{f_o}{f}\right)^2} \qquad (18)$$

where Q is the effective magnification factor of the circuit, assumed to be large. Using this type of spectrum, it can be shown from equation (14) that the auto-correlation function is

$$\psi_{N2}(\tau) = \frac{\pi^2 f_o P_o}{Q} \exp\left(\frac{-\pi f_o \tau}{Q}\right) \cos(2\pi f_o \tau). \qquad (19)$$

The envelope of this function is shown in Fig. 16(c); it decays exponentially, the function itself having zeros for the same values of τ as the curve in Fig. 16(b). Thus for a finite value of τ, the output of the detector of Fig. 15 cannot be made identically zero in the presence of noise only, yet finite in the presence of a signal of frequency f_o.

The conventional auto-correlation detector circuit cannot therefore be used to achieve the desired results in the case of the receiver under consideration, since this receiver includes a post-detector resonant filter. This difficulty can be overcome without altering the filter by using a modified detector circuit described in Section 13.4.

13.4 *A Practical Form of Detector*

13.4.1 *The Use of an All-pass Network instead of a Constant-delay Network*

It was shown in Section 13.3.2 that the noise output from a resonant filter has an auto-correlation function with no convenient zeros in its envelope.

The network of time delay τ in Fig. 15 can be considered as a network of phase shift ϕ, where ϕ is a linear function of frequency. This restriction is essential if the output of the detector is to be identified with the auto-correlation function but is of no significance in achieving the result desired here. If we remove the restriction (i.e. use a network whose phase shift is not proportional to frequency), a type of detector which can follow a resonant filter and yet have the desired properties becomes possible. We will refer to this as a 'correlation detector', the auto-correlation detector being a special case of this general class. A convenient network for the particular application in view is an all-pass network having the phase/frequency characteristic shown in Fig. 17.

If the network of Fig. 17 is used in place of the constant-delay network in Fig. 15, the detector output then becomes

$$\psi'_{N2}(s) = \int_o^\infty W_2(f) \cos\phi . df \qquad (20)$$

where $W_2(f)$ is the spectral power density of the noise from the resonant filter, given by equation (18), and ϕ is the phase-shift of the all-pass network of Fig. 17, given by

$$\tan\frac{\phi}{2} = \frac{s f f_o}{f_o^2 - f^2} \qquad (21)$$

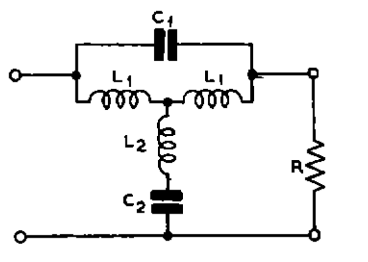
here, s is a parameter of the all-pass network and f_o is the frequency for which its phase shift is π; the response to a signal of frequency f_o is therefore a maximum, independent of s. If f_o is the same for the resonant filter and the all-pass network, the detector output can be shown from equation (20) to be

$$\psi'_{N2}(s) = \frac{\pi^2 f_o P_o}{Q} \cdot \frac{Qs - 1}{Qs + 1}. \qquad (22)$$

This expression becomes zero at $s = 1/Q$, as shown in Fig. 18; the network is therefore designed to satisfy this condition.

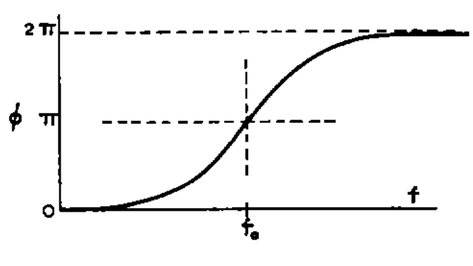
13.4.2 *The Use of a Short-time-constant Meter instead of an Averaging Meter*

In practice the averaging meter of Fig. 15 will be replaced by a meter having a time-constant sufficiently low



$$C_1 = \frac{1}{4\pi s f_o R} \qquad C_2 = \frac{s}{\pi f_o R}$$

$$L_1 = \frac{sR}{2\pi f_o} \qquad L_2 = \left(\frac{1}{s} - s\right)\frac{R}{4\pi f_o}$$



$$\tan\frac{\phi}{2} = \left(\frac{s f f_o}{f_o^2 - f^2}\right)$$

Fig. 17 — The phase-shift network.

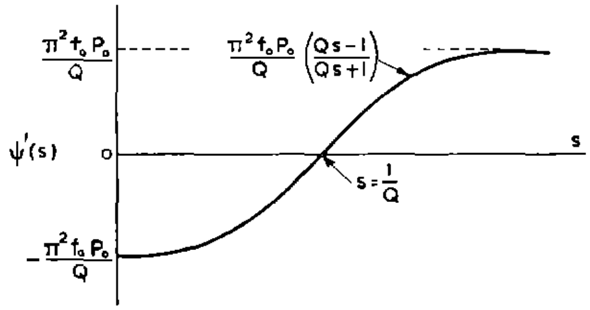


Fig. 18 — *The modified correlation function ψ' (s).*

to allow the requisite rate of response to changes in signal amplitude. The meter reading will then be subject to random fluctuations due to noise, which determine the smallest observable signal. It can be shown that the r.m.s. noise output is 3 dB lower for this type of detector than for a simple square-law detector of the same gain and fed from the same source. This change is accompanied by a corresponding reduction in effective bandwidth. The output signal/noise ratios of various types of detector are given elsewhere.[3, 6]

13.4.3 *The Use of a Linear Phase-sensitive Detector instead of a Multiplier*

The above analysis has assumed that a multiplying detector is used; Fig. 19 illustrates one possible arrangement for this purpose,[7] using two square-law amplifiers. In the presence of a sinusoidal input signal, the output is proportional to the square of the input magnitude and to the cosine of the phase difference between A and B.

For a measuring receiver it is preferable to use a detector having an output linearly proportional to the input magnitude. Fig. 20 illustrates an arrangement which fulfils this requirement, using two linear detectors; provided that k, the ratio $|B/A|$, is not greater than $0 \cdot 5$, the cosine response to the phase difference between voltages A and B is largely maintained, as shown in Figs. 21 and 22. Fig. 22 also shows the variation with phase difference of the sum of the detector currents. For $k=0 \cdot 5$, this sum is almost independent of phase so that it may be used for a.g.c. as discussed in Section 10.4.

It has been shown that in certain conditions, established in Section 13.4.1, the d.c. output from a correlation detector using a phase-shift network followed by a multiplier is zero in the presence of noise only. In this case, using the multiplier circuit of Fig. 19, the r.m.s. values of the noise voltage applied to the square-law amplifiers must

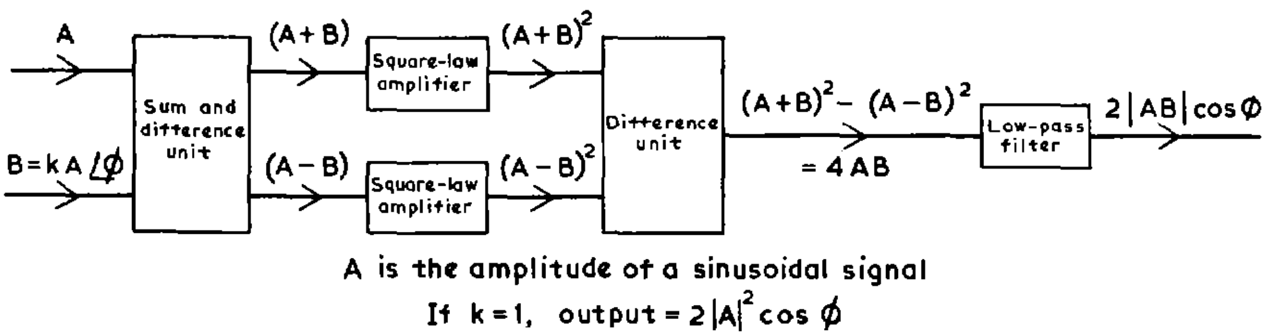


Fig. 19 — *Schematic diagram of multiplying detector.*

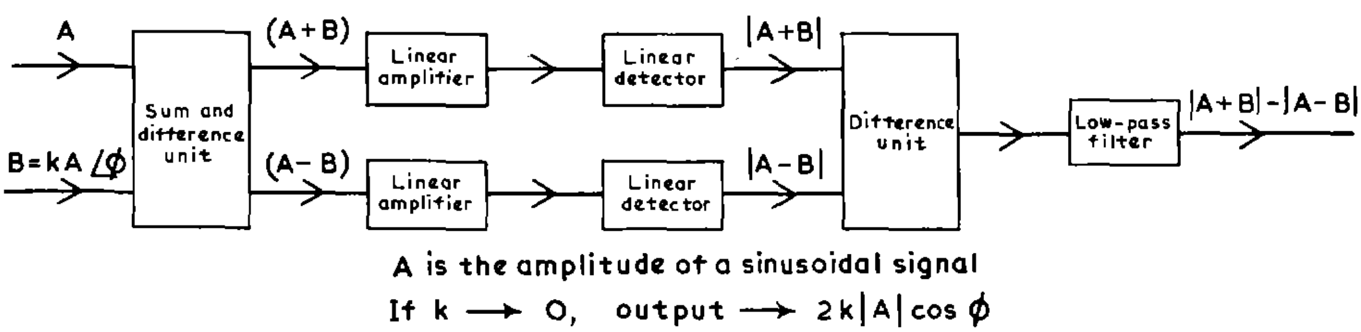


Fig. 20 — *Schematic diagram of linear phase-sensitive detector.*

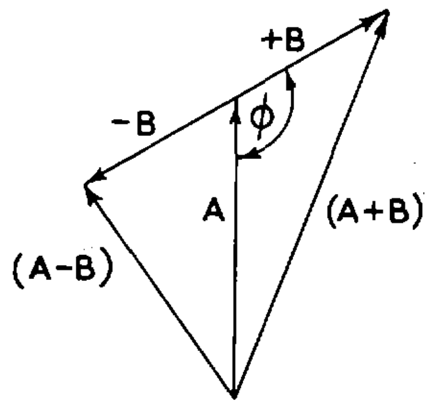


Fig. 21 — *Vector diagram for linear phase-sensitive detector.*

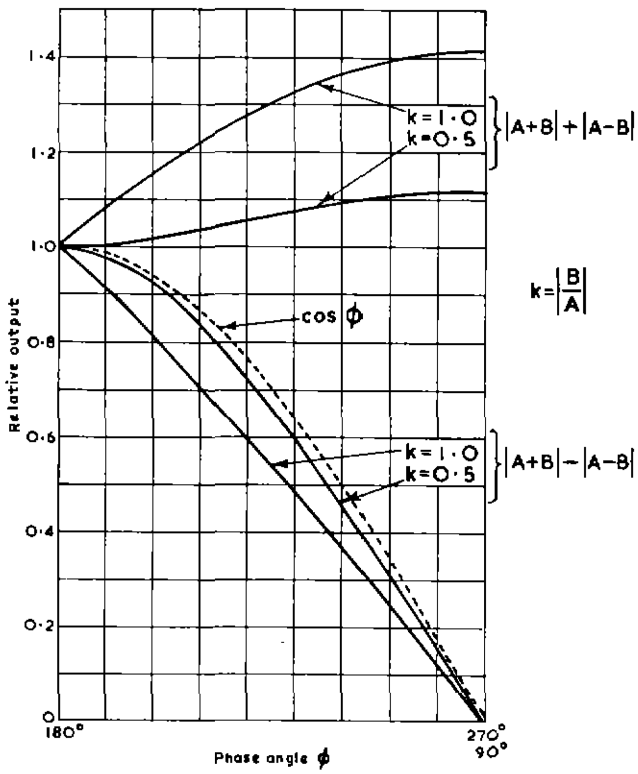


Fig. 22 — *Response of linear phase-sensitive detector.*

be equal. The same result must therefore follow if the linear system of Fig. 20 is used instead of a multiplier.

However, the two systems show a difference in their response to a small signal in the presence of noise. In the multiplying detector of Fig. 19 the d.c. output of each of the square-law amplifiers is equal to the sum of the outputs due to the signal and noise when applied separately. Since the d.c. contributions due to noise cancel in the final output, the signal output is always proportional to the square of the signal input, irrespective of the noise level. The d.c. output of a linear detector, on the other hand, is approximately proportional to the square root of the sum of the signal and noise powers.[4] This means that the output of the linear system of Fig. 20 follows a linear law only for signals which are large compared with the noise level; for signal levels below noise level it tends to follow a square law. Moreover, in this square-law region, an increase in the noise level results in a reduction in the output due to a given signal. For very low input signal/noise ratios the output is reduced in the same ratio as the increase in noise.

As described in Section 7, a linear rectifier is already used as an i.f. detector in the receiver under consideration; the signal output at this stage is therefore affected by the input noise level in a similar way to that described above, but the effect begins to occur at a signal level some 20 dB higher than at the a.f. detector. Thus the overall performance is only slightly impaired by the use of the linear rather than the multiplying form of a.f. correlation detector. More drastic changes to the design of the receiver would be necessary to eliminate the effect of changes in noise level on the calibration at all signal levels.

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