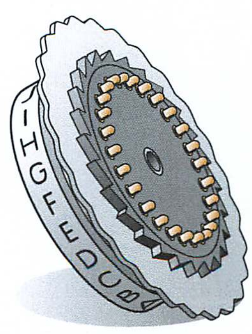
EC Mark III

Easy Chair Extended Range
Final Research Report

10 January 1958

Project Easy Chair



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Project Easy Chair Extended Range.

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Typed in twofold.
Ex. No. 2 Page 1

Contents.

1. Introduction.

2. Theoretical and experimental investigations.

   2.1 Subcarrier P.E.characteristics.
   2.2 Receiver and transmitter noise-level.
   2.3 Squelch circuit.
   2.4 Automatic duplexer.
   2.5 Antennas.

3. Considerations in connection with the system engineering.

   3.1 Consequences of the use of frequency modulation.
   3.2 Receiver bandwidth.
   3.3 The choice of tubes and/or transistors in the receiver.
   3.4 The transmitter.
   3.5 System balance, duplexer choice and range evaluation.

4. Preliminary specifications for an equipment with extended range.

1. Introduction.

A research and development program under contract, dated 6th March 1957, has been carried out in the laboratory by the contracting party during the period of March 1957 till 31st December 1957. As agreed upon in the above mentioned contract the preliminary part of the work, being of a theoretical and experimental nature, should be completed in ten months time, while the following eight months (January 1958 till August 1958 inclusive) were scheduled for the production of a prototype equipment, based upon the results obtained during the first period.
According to the "Agreements and Conclusions" of 8th August 1957 the major effort in this ten months period was concentrated on the design of a new P.E. device, incorporating a local oscillator with the audio frequency modulated on this subcarrier.
In this report the results of theoretical and experimental investigations into the system are laid down. The methods used are described and the results are discussed. Finally conclusions are drawn as to the choice of system parameters of an improved type of Easy Chair system having extended range, improved operational performance, decreased detection possibility by third parties, offering at the same time simpler handling by non-technical users.

2. Theretical and experimental investigations.

2.1 Subcarrier passive element characteristics.

2.1.1 Circuit diagram. The circuit diagram is given in fig. 1.
Transistors V_3 and V_4 are connected together as an oscillator, the load and coupling impedance of output transistor V_4 being the internal impedance of the crystal detector at the d.c. terminals.
A more or less rectangular waveform with a frequency of the order of 100 kc/s is generated across said d.c. terminals. Hereafter this frequency will be referred to as the subcarrier frequency. The subcarrier voltage, developed across the d.c. terminals of the crystal detector, is in fact a load modulation for the detector and will cause a corresponding modulation of the reflected r.f. power at the r.f. terminals of the detector. The frequency of oscillation is governed by a number of factors, amongst others the impedance offered by the output circuit of modulating transistor V_2.
The output impedance of V_2 depends on its d.c. working conditions and it will be clear that, when these working conditions are altered, e.g. by an amplified microphone voltage, the corresponding alteration of output impedance of V_2 will cause the generated subcarrier frequency to be modulated.
Transistor V_1 is a microphone voltage amplifier.

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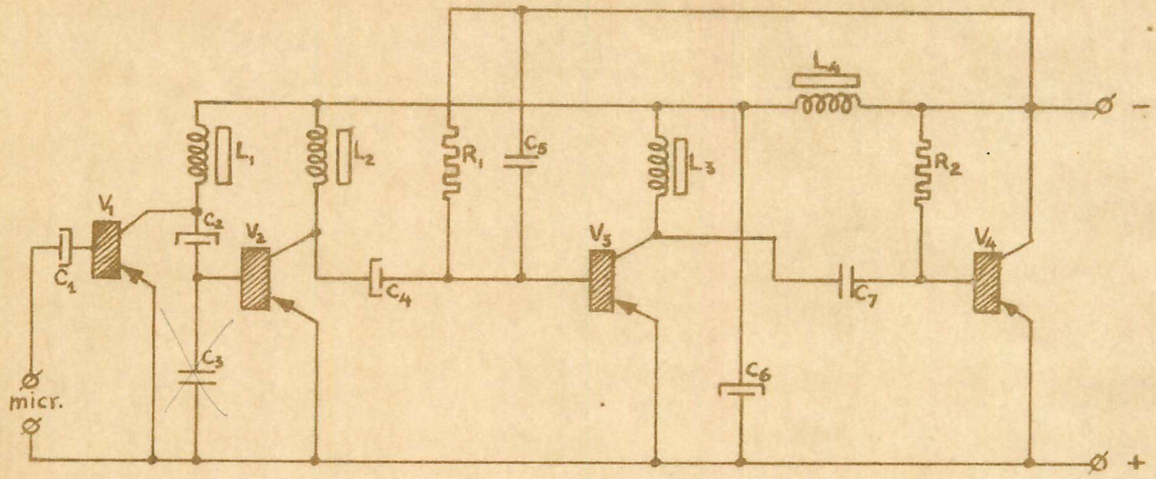

FIG. 1. PASSIVE ELEMENT

V_1, V_2	junction transistor, Philips,	OC 71
V_3, V_4	" " "	OC 44
L_1, L_2, L_4	inductance, Fortiphone	EX 192, 1.5 H - 150 Ω
L_3	" on ferroxcube core,	6 mH
R_1	resistor, Vitrohm	22.000 Ω, ½ W
R_2	" "	33.000 Ω, ½ W
C_1, C_2, C_4	condensor, tantalum, Philips,	2 μF, 6 V.
C_3	" ceramic, T.C.C.,	0.01 μF
C_5, C_7	" " Philips,	0.0012 μF
C_6	" tantalum, "	10 μF, 3 V.

This explanation of working principles is greatly simplified and in fact several other factors are contributing to the result, some of which are:

a. The oscillating condition is governed to a large extent by the voltage- and current-limiting properties of transistors at very low operating voltages. On this subject only a very limited amount of information is available.

b. The waveform generated departs considerably from either a true rectangular wave or a true sine wave. The calculated behaviour, if undertaken, is at its best only a very rough approximation.

c. Subcarrier frequency modulation is not only effected by the varying output impedance of V_2, but also, and to a not negligible amount, in V_3 by the amplified microphone voltage present in the output of V_2.

d. The modulating effect of the output impedance of V_2 is largely dependent on the α-cut-off-frequency of V_2.

These complications in fact prevent any exact and universal analysis. Therefore an experimental and empirical approach was considered most suitable and for this reason previous progress reports stressed the desirability to have at hand a measuring equipment, offering at the same time and at once several related results.
With this equipment and the simple explanation of operation, given before, in mind, an experimental and empirical development was performed, resulting in the unit described.

2.1.2 <u>Video-frequency test set-up.</u> A block diagram of the video-frequency test set-up is given in fig. 2.
This test set-up has proved to be extremely useful and almost indispensable for the purpose, because of its fast and simultaneous indication of results, which fact made it possible to devote almost all attention to the passive element under test.
The passive element under test is fed from a variable and metered d.c. voltage source having an internal resistance of 680 ohms, simulating the d.c. output circuit of a crystal detector.
The microphone input terminals of the passive element were fed from an audio oscillator with calibrated attenuator.
The frequency-modulated subcarrier output voltage waveform, amplitude and spurious amplitude modulation of the passive element is shown by oscilloscope 1.
The subcarrier output from the passive element is also fed to an amplifier-limiter-frequency-detector. The average d.c. output voltage from this detector is an indication for the mean subcarrier frequency, whilst the audio frequency component, measured with a V.T.V.M., is a measure for the frequency modulation width. Oscilloscope 2 shows modulation non-linearity.

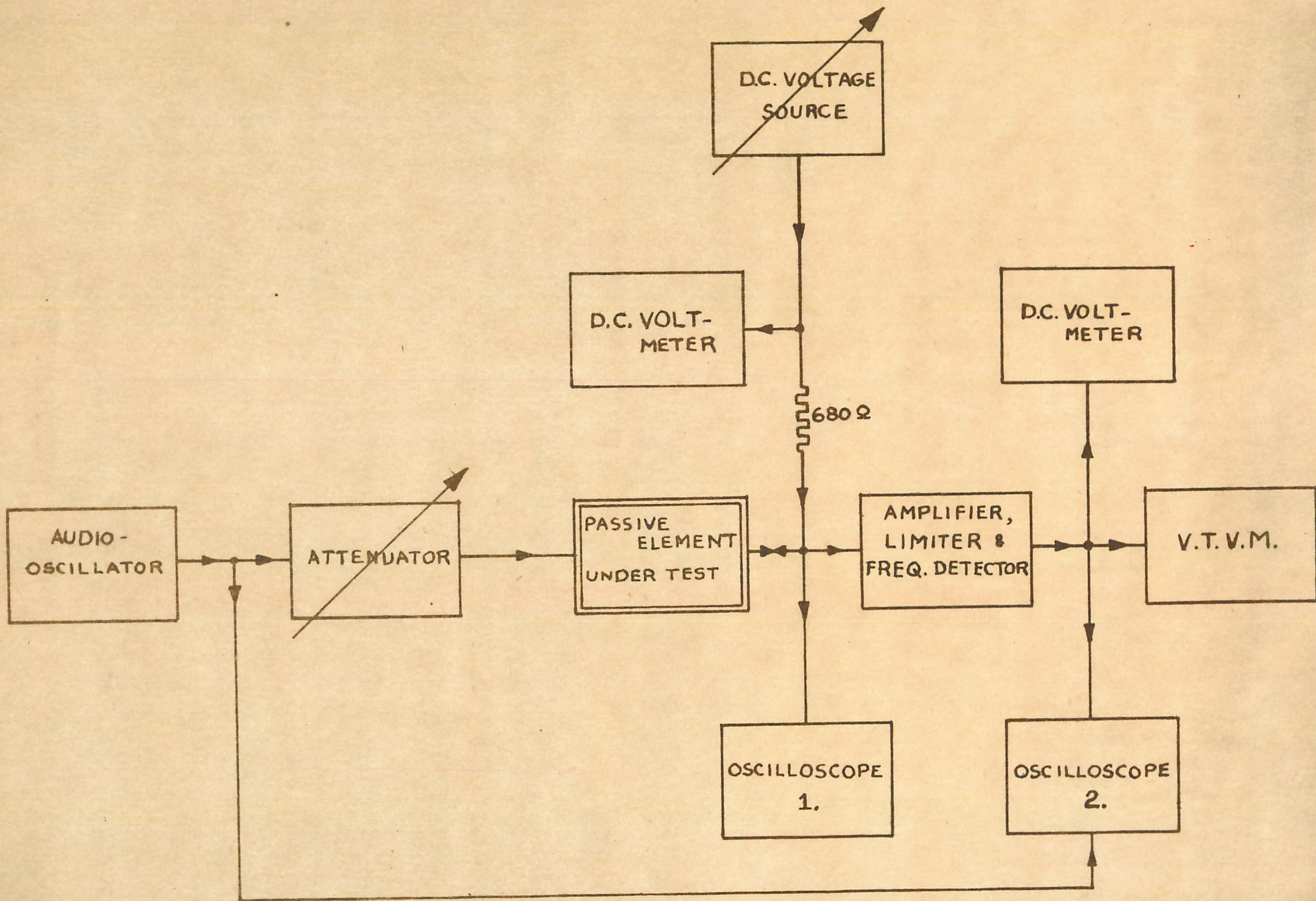

FIG. 2 VIDEO-FREQUENCY TEST SET-UP

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2.1.3. Radio-frequency test set-up. A block diagram of the radio-frequency test set-up is given in fig. 3.
This test set-up quite accurately simulated actual circumstances and results obtained with it were considered decisive and final. All results from measurements given in paragraph 2.1.4 of this chapter have been obtained with this set-up.
The video-frequency test set-up was used mainly to set up a working passive element or to work up to some specific result in detail, whilst the radio-frequency test set-up showed the practical and actual outcome.
In our opinion both set-ups should be used together and an integrated set-up might be even better.
A brief explanation of the radio-frequency test set-up follows now.
A 378 Mc/s transmitter can be modulated to an adjustable and accurately known modulation percentage by an oscillator at subcarrier frequency. The resulting sideband power is therefore also known and is used in a substitution method.
Some of the transmitter power is fed to the passive element and crystal detector via a directional coupler, two attenuators and a line stretcher. The amount of power fed to the passive element and crystal detector can be determined with a power monitor, which in this instance is connected to one of the directional coupler outlets.
Some of the transmitter power is also fed via the directional coupler and two attenuators to the receiver detector to act as a local oscillator for homodyne operation.
The reflected and modulated power from the passive element and crystal detector goes via the line stretcher, the directional coupler and two attenuators also to the receiver detector.
The output at subcarrier frequency from the receiver detector passes via an amplifier on to V.T.V.M. 1. This V.T.V.M. indicates either the reflected subcarrier modulation power from the passive element and crystal detector or the substitution sideband power from the transmitter.
Either condition is selected by opening or closing the path to the passive element and crystal detector or the path from the substitution oscillator at subcarrier frequency, as schematically indicated in the block diagram.
The correct phasing of passive element subcarrier sideband- and local oscillator-vectors can be found by adjustment of the line stretcher for maximum deflection of V.T.V.M. 1.
The attenuation in every path is known and fixed, which results in a very simple conversion from attenuator 1 reading and power monitor reading to the reflection loss offered by the passive element and crystal detector. In this instance we define the reflection loss as:

$$\frac{\text{Subcarrier frequency sideband power from crystal detector of P.E.}}{\text{Available r.f. power at crystal detector of P.E.}}$$

Audio frequency modulation of the passive element is effected by a 1000 c/s 50 µV R.M.S. oscillator. The resulting frequency modulation width is indicated by V.T.V.M. 2.

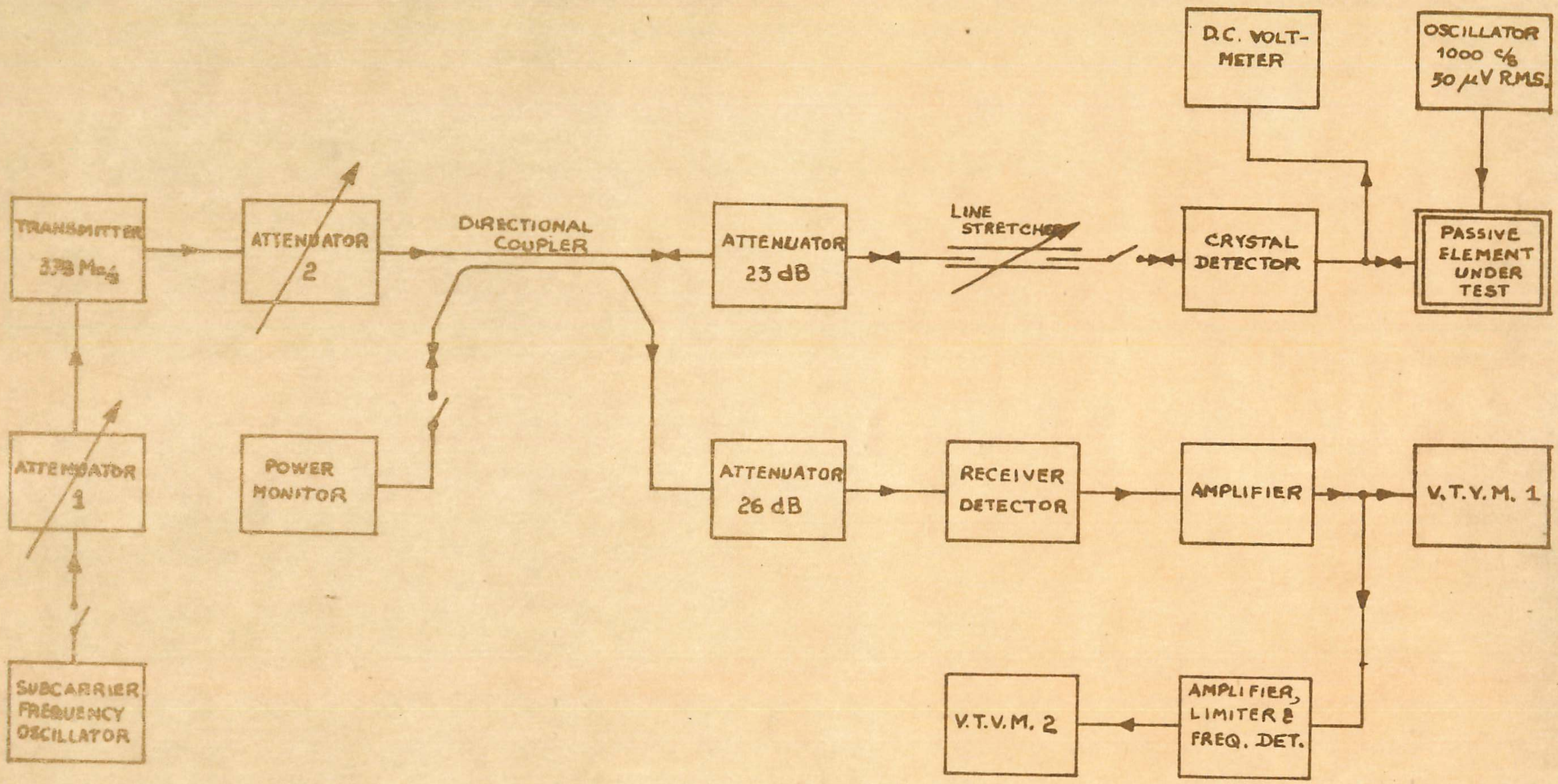

FIG. 3 RADIO-FREQUENCY TEST SET-UP

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2.1.4. Results from measurements. The results from the measurements by the methods described are illustrated in fig. 4.
The relative modulation depth is defined with respect to a frequency modulation index of 1 for an audio modulation frequency of 4 kc/s maximum and with a microphone voltage of 100 μV R.M.S.
As can be seen the usable range of activating radio-frequency power is between approximately -43 dBW and -37 dBW.
An average and recommended working power level would be -40 dBW (= 100 μW).
At this level the experimental passive element had a mean subcarrier frequency of 92 kc/s, a reflection loss of 16.3 dB and a relative modulation depth of + 3 dB.
The audio input noise level of the passive element was about 0.45 μV R.M.S., which limits the obtainable audio signal-to-noise ratio for 100 μV R.M.S. microphone voltage to about 47 dB.

Spurious amplitude modulation and duty cycle modulation of the subcarrier waveform as a result of audio modulation of the passive element was checked. For normal levels of modulation these unwanted effects were negligible. At very high levels of audio modulation depth noticable spurious modulation occurred, but in practical circumstances no modulation levels of that order are to be expected. These spurious modulations might impair the secrecy of the passive element when a third party tries to locate it.

2.2 Receiver and transmitter noise-level.
A test set-up was made to determine the noise figure of a receiver under typical Easy-Chair conditions.
These conditions might be referred to as homodyne operation, in which the transmitter itself supplies the local oscillator power to the receiver detector.
The test set-up, intended to simulate subcarrier type Easy-Chair conditions, is shown in block diagram fig. 5.
The transmitter power is fed via a directional coupler and an attenuator to the receiver detector. The output at subcarrier frequency of the detector is amplified and indicated on a V.T.V.M.
The amount of local oscillator power fed to the receiver detector is deduced from the power monitor reading and from the attenuator 2 setting.
Normally the V.T.V.M. indicates the noise output of this chain, but means are provided to amplitude-modulate the transmitter with a subcarrier frequency to an accurately known percentage.
The modulation depth is adjusted during the measurement with attenuator 1 until a 3 dB increase in V.T.V.M. reading is obtained.
In this way the actual noise power at the receiver detector input is equalled and because the amount of added sideband power can be deduced from the attenuator readings, the actual noise power is known.
Determination of the receiver bandwidth, calculation of the theoretical minimum noise power (= kTB) and comparison directly yields the receiver noise figure. It is important to notice that in this case the bandwidth must be considered to be twice the receiver

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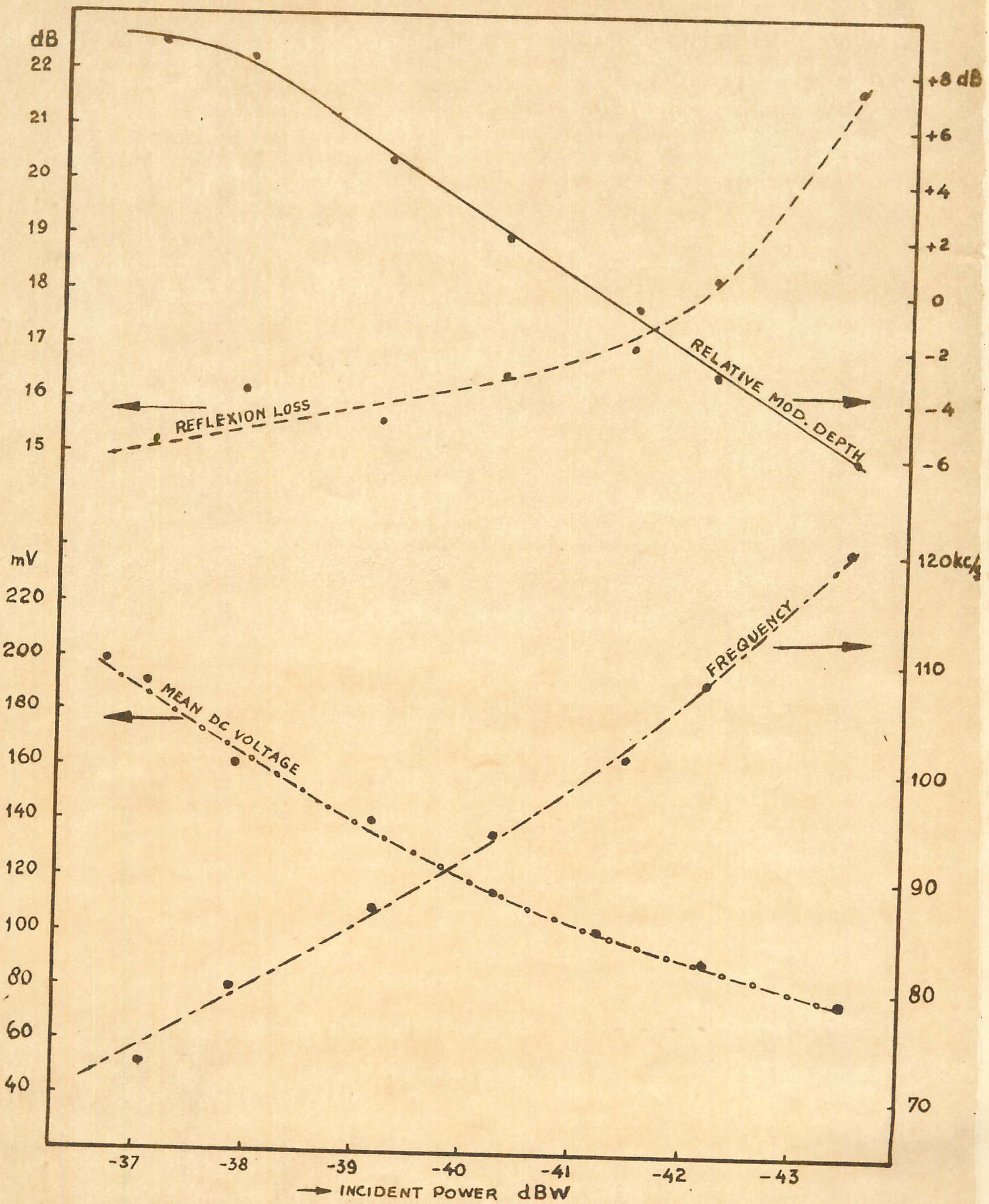

FIG. 4 PASSIVE ELEMENT CHARACTERISTICS

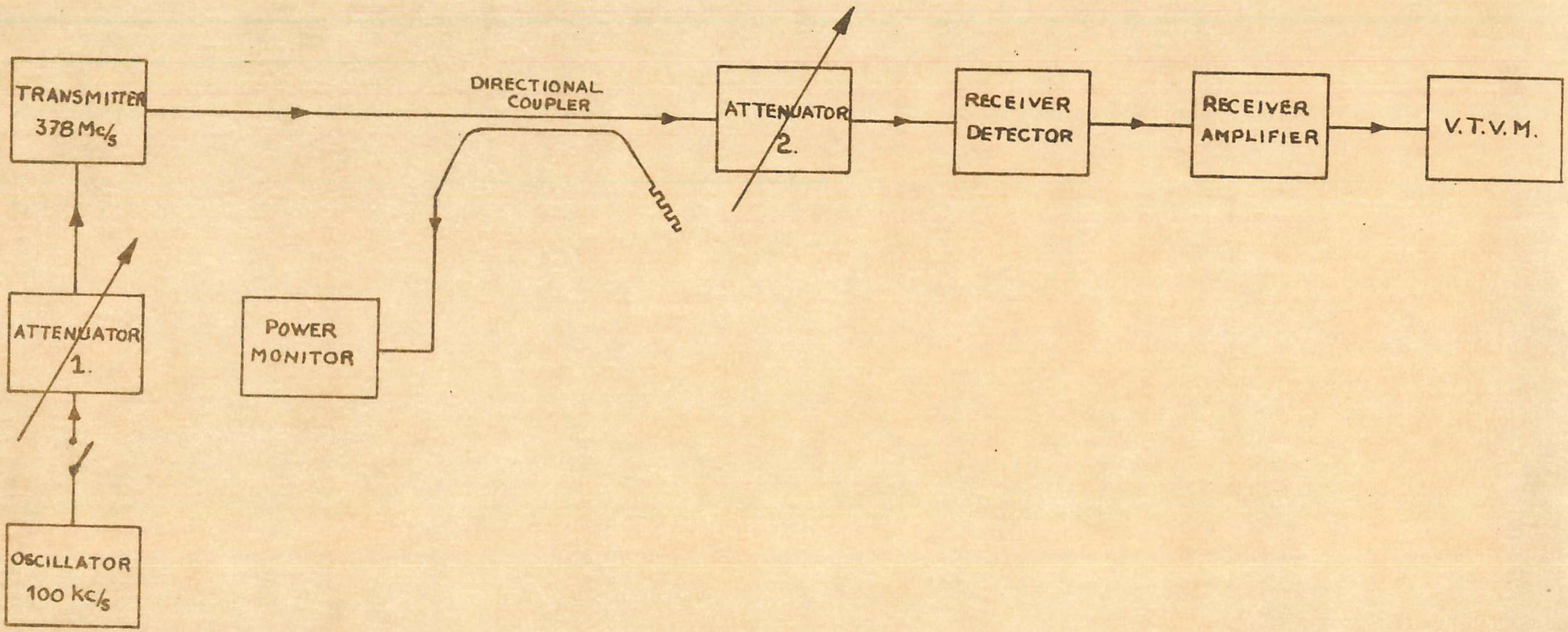

FIG. 5 NOISE FIGURE TEST SET-UP

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intermediate frequency amplifier bandwidth, due to the absence of image-frequency rejecting elements.
Further it must be remembered that this method is only absolute if the V.T.V.M. has a true square-law characteristic. For comparative measurements a discrepancy from the square-law characteristic has no consequences. In the measurements described here, not much effort was done to obtain a true square-law characteristic.

A typical noise figure curve for a selected crystal detector type CS 2 A is shown in fig. 6, curve A.
An optimum noise figure is obtained for local oscillator power levels of about -41 dBW.
For lower values of local oscillator power the conversion loss of a crystal detector increases and thereby the noise figure as well. Moreover the deteriorating effect of intermediate frequency amplifier noise becomes important.
For higher levels of local oscillator power than the optimum one, two factors can contribute.
One factor might be the transmitter noise, about whose magnitude no figure was known, although some suspicion existed.
The other better-known factor is the increasing noise temperature ratio of the crystal detector at higher level of local oscillator power.
The power level at which the increase in noise temperature ratio becomes significant in superheterodyne receivers with intermediate frequencies of 30 Mc/s or higher is known to be some 10 dB higher than the level experienced here. This might be explained by the known tendency of crystal detectors to exhibit higher virtual noise temperature ratios at lower values of intermediate frequency. The value of intermediate frequency in the experiments described was about 100 kc/s, which may be considered relatively low.
No sharp separation could be made yet to determine the contribution of transmitter noise.
One indication, however, was obtained when transmitters of radically differing lay-out and operation did not show any clear difference in noise-figure measurements.
This receiver used a GL 446-A lighthouse tube as a grounded-grid amplifier with flat lines as tuning elements, followed by a GL 446-A lighthouse tube as diode detector. The results obtained with this combination are illustrated in fig. 6, curve B.
A considerable improvement was obtained at high local-oscillator power levels.
A similar unit, built around regular miniature tubes with NOVAL bases, using a Philips EC 80 as a grounded-grid radio-frequency amplifier and a Philips EC 81 as an anode detector, gave results as shown in fig. 6, curve C.
Both curves show the same tendency, although the curve for the lighthouse tubes is superior, due to the better high-frequency behaviour of these tubes.

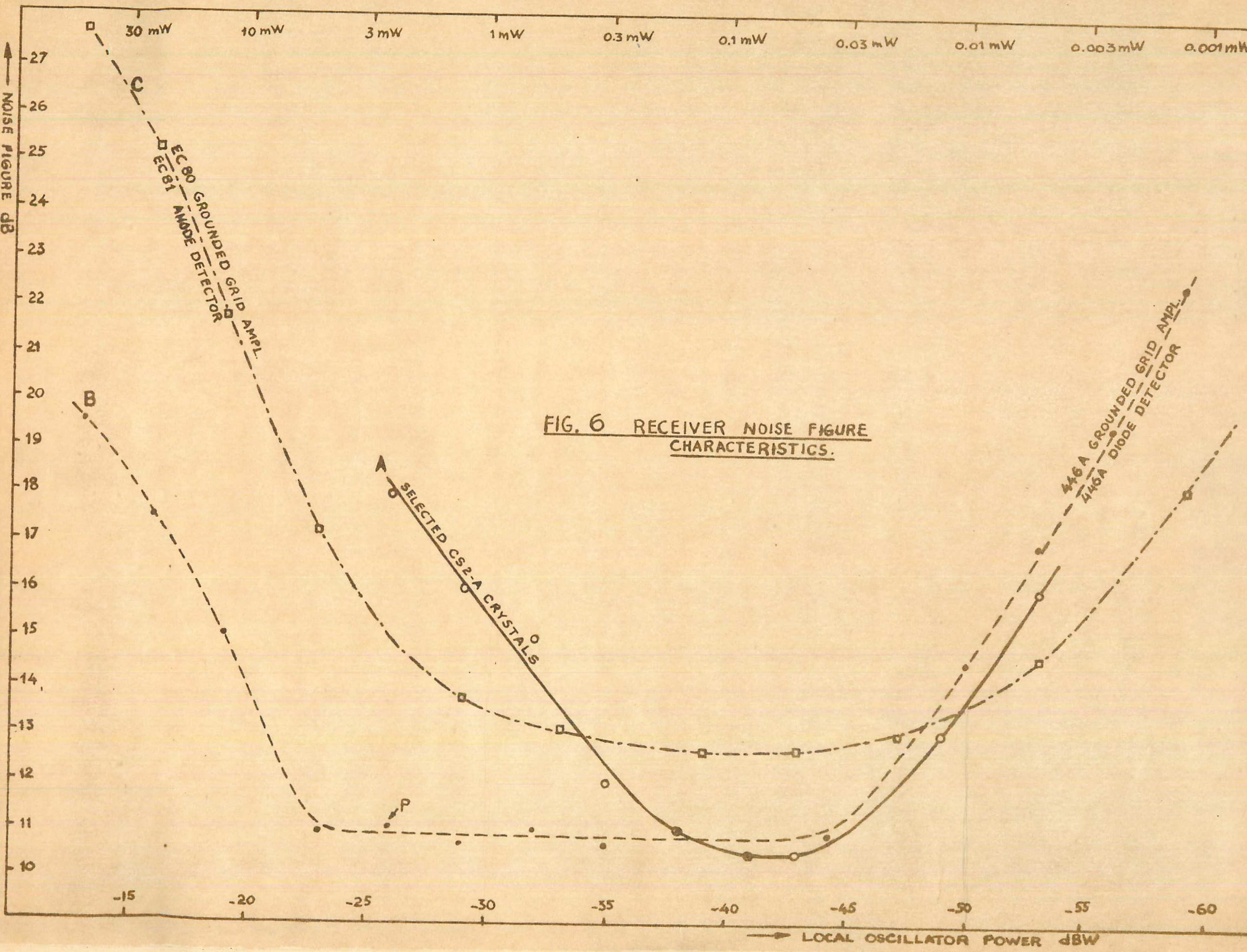

FIG. 6 RECEIVER NOISE FIGURE CHARACTERISTICS.
NOISE FIGURE dB
LOCAL OSCILLATOR POWER dBW
30 mW
10 mW
3 mW
1 mW
0.3 mW
0.1 mW
0.03 mW
0.01 mW
0.003 mW
0.001 mW
27
26
25
24
23
22
21
20
19
18
17
16
15
14
13
12
11
10
-15
-20
-25
-30
-35
-40
-45
-50
-55
-60
A
B
C
P
EC 80 GROUNDED GRID AMPL.
EC 81 ANODE DETECTOR
SELECTED CS2-A CRYSTALS
446 A GROUNDED GRID AMPL.
446 A DIODE DETECTOR

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The increase in noise figure at high local oscillator power levels is caused by overloading of the radio-frequency amplifier stage, also noticable by an increase of average d.c. anode current.
A vacuum tube diode detector was tried, which showed a good consistency of noise figure up to high levels of local oscillator power, but the noise figure was in all respects worse than that obtained with a radio frequency amplifier and detector combination. In addition input matching over a relatively wide range of local oscillator power levels was more critical, due to the higher intrinsic input impedance.
With the amplifier detector type of unit again no separation could be made between receiver noise figure and transmitter noise contribution. A further check was made on this point by using both amplifiers together. The input power to both amplifiers was made equal and derived from the same transmitter. The output voltages of both detectors were combined in antiphase with suitable attenuation in the circuit to equalize gain differences.
Any modulation on the transmitter carrier would cancel in the output, whilst all noise contributions in the radio-frequency amplifiers would add, because of their incoherence. Two exactly similar amplifiers could give a 3 dB increase in actual noise output compared to a single amplifier. Results indicated an increase of some 2 dB in actual noise output for almost all values of local oscillator power considered. Due to the difference in amplifier construction this was considered sufficient evidence to conclude that the noise figures measured were not greatly influenced by transmitter noise. Therefore it can be deduced that the transmitter signal to noise ratio over a bandwidth of about 40 kc/s (= 2 x 20 kc/s receiver amplifier bandwidth), centered at about 100 kc/s from the carrier frequency, is better than some 130 dB.
This figure is deduced for transmitters of the power oscillator as well as the master oscillator-power amplifier variety.
It might be mentioned here that previous Easy Chair transmitters from the laboratory, incorporating an anti-microphony circuit, showed a signal to noise ratio considerably worse than the figure mentioned above, due to the introduction of noise by the anti-microphony circuit itself.
A better optimum noise figure can be expected if a second stage of radio frequency amplification would be added to the radio frequency amplifier-detector configuration. Overloading will, however, occur at lower levels of local oscillator power, requiring more critical matching adjustments to be made.
As a preliminary conclusion from the measurements described here, it might be said that the vacuum tube receiver input circuit, as compared to the crystal detector input circuit, has a number of very desirable features. One of these features is the improved protection against crystal burn-out. Another good thing is the preservation of lower noise figures up to rather high values of local oscillator power. This allows a far less critical balancing adjustment for the duplexer and results in easier handling or the possibility of using higher transmitter powers.

These improved characteristics may outweigh the drawbacks of increased weight, size and power supply demands of vacuum tube receiver input circuits as compared to crystal detector inputs. A more recent type of lighthouse tube, Philips EC 56, will be available to the laboratory in a few weeks time and even better figures might then be expected.

2.3 <u>The squelch circuit.</u>
The output noise of the frequency modulation base station receiver is governed by the input subcarrier frequency signal-to-noise ratio.
Below a certain threshold subcarrier signal-to-noise ratio, the noise output of the receiver increases very rapidly and fully out of proportion with the decrease in subcarrier signal-to-noise ratio.
Subcarrier signal-to-noise ratios below this threshold value may therefore be considered to be of little or no use and a squelch circuit operating according to this criterium will be justified fully by the prevention of annoyance, due to excessive audio noise levels during r.f. balancing or aiming adjustments.
A simple solution, such as a subcarrier amplitude-dependent switch breaking the audio path of the receiver would not operate very well under these circumstances. This is caused by the fact that the noise level at the input of the receiver, being just as important as the subcarrier signal strength in meeting this criterium, is not constant, but dependent on the amount of transmitter power, coupled or reflected into the receiver input.
The requirements were met very satisfactorily, however with a circuit which was activated directly by the noise output of the frequency detector.
From this noise a narrow frequency band is selected, amplified and rectified. The narrow frequency band is chosen somewhere intermediate between the highest audio frequency and the lowest subcarrier frequency to be expected. A suitable value for this application would be about 20 or 25 kc/s.
The rectified noise voltage is used as a d.c. switching signal for an audio amplifier stage in the audio chain following the frequency detector.
This circuit needs no adjustment or presetting at all under greatly varying conditions of noise level and/or signal strength.
Reliable switching-over is effected by subcarrier signal-to-noise ratio changes of about 2 dB.
The additional circuitry required, including the switched audio amplifier stage, consists mainly of 2 transistors and 1 tuned circuit.

2.4 <u>Automatic Duplexer.</u>

2.4.1 <u>The duplexing principle.</u> A duplexer permits the simultaneous passage of signals in both directions along a transmission line, thereby discriminating for the direction of travel.
Instead of using two antennas, one for transmitting and one for receiving, it is possible to use only one antenna if a duplexing circuit is used. The duplexer may incorporate a directional

coupler or a magic T (which is in fact a directional coupler with a coupling ratio of 3 dB).
The range which can be achieved with a duplexing system incorporating a magic T is the same as that of the two antennas system, provided that the one antenna used is twice as large as one of the antennas used in the two antennas system.
The minimum possible total loss of continuous wave duplexers is 6 dB (viz. 3 dB in each direction). The main advantage of a duplexer incorporating a directional coupler is that by proper choice of the coupling ratio the system can be brought into balance, thereby increasing the range. If e.g. the receiver sensivity is very good, the coupling ratio can be increased above 3 dB, thereby decreasing the transmission loss. It will be clear that this applies only to passive modulators, for which the sideband power versus incident power characteristic is non-linear.
A continuous wave system, which is optimum in all respects, will incorporate a directional coupler with a coupling ratio of 3 dB.
Other advantages of the duplexer system over the two antennas system are:

a. Due to the fact that the power is transmitted by an antenna which is twice as large, the power is confined in a sharper beam, resulting in less power in unwanted directions. For the same reason the discrimination against interference coming from directions other than the direction of the main beam is increased.

b. The receiving system is based upon the principle of synchronous detection, in which part of the transmitter power has to reach the receiver directly. In the two antennas system local oscillator power is derived from reflection against surrounding objects and by leakage from one antenna into the other. The duplexing system permits the local oscillator power to be diverted internally by the insertion of mismatching units in the main transmission line or in the coupling arm. In this way, the local oscillator power level will be less dependent on antenna position and will be less frequency-sensitive. Furthermore the adjustment for the right phase and amplitude of the mixer excitation can be performed by tuners instead of by shifting of the antenna positions.

c. In addition to the adjustment of the right local oscillator power level the tuning elements in the transmission line will cancel antenna mismatch, thereby leading to a small increase in range.

d. An operational advantage is that one antenna requires less space and can be orientated much easier than two antennas having half the dimensions, but being separated for a certain minimum distance.

2.4.2 Automatic duplexing. By using two directional couplers, spaced $1/8\lambda$ or an integral multiple of $1/4\lambda$ more, along the main transmission line, it is possible to achieve a duplexing circuit which is phase-insensitive. This means that tuning for the right phase is no longer a requirement, the only concern being to keep the reflected power within limits.

In combin ation with the use of vacuum tube mixer diodes this means that tuning of the antenna mismatch will not be necessary under many circumstances and will be greatly facilitated under conditions where reflections in the vicinity of the antenna are large or when full power is used.

The main advantage of the automatic system is gained when aiming the base stat ion anetnna.

Apart from the second directional coupler the automatic system requires a second mixer and phase shifting networks on the subcarrier frequency. A block diagram is given in fig. 7. Local oscillator power is obtained by inserting mismatching elements, e.g. shunt capacitances, before the loads in the coupling arms. The reflected power travels via the secondary arms to the mixer. The tuners in the transmission line to the antenna are no longer used for the deviation of local oscillator power, but for the matching of residual antenna mismatch, in order to keep the power reflected into the receiver within limits. The displacement of one directional coupler $1/8\lambda$ along the transmission line towards the antenna increases the local oscillator path with $1/8\lambda$, at the same time decreasing the signal path by the same amount. Therefore, if the phase difference between local oscillator power and signal is φ at mixer 1, it will be $\varphi - \frac{\pi}{2}$ at mixer 2.

If the input to mixer 1 is written as:

$$A_o \cos \omega t + A_r (1 + m \cos pt) \cos (\omega t + \varphi)$$

the input to mixer 2 will be:

$$A_o \cos (\omega t + \frac{\pi}{4}) + A_r (1 + m \cos pt) \cos (\omega t + \varphi - \frac{\pi}{4})$$

where A_o = amplitude of local oscillator power,

A_r = amplitude of reflected component,

m = modulation depth of reflected component,

p = modulation frequency of reflected component,

ω = carrier frequency.

The output components of the mixers of frequency p are:

Mixer 1: $M_1 = c\ m\ A_o A_r \cos\varphi \cos pt$

Mixer 2: $M_2 = c\ m\ A_o A_r \sin\varphi \cos pt$

In these formulae c is a mixer constant.

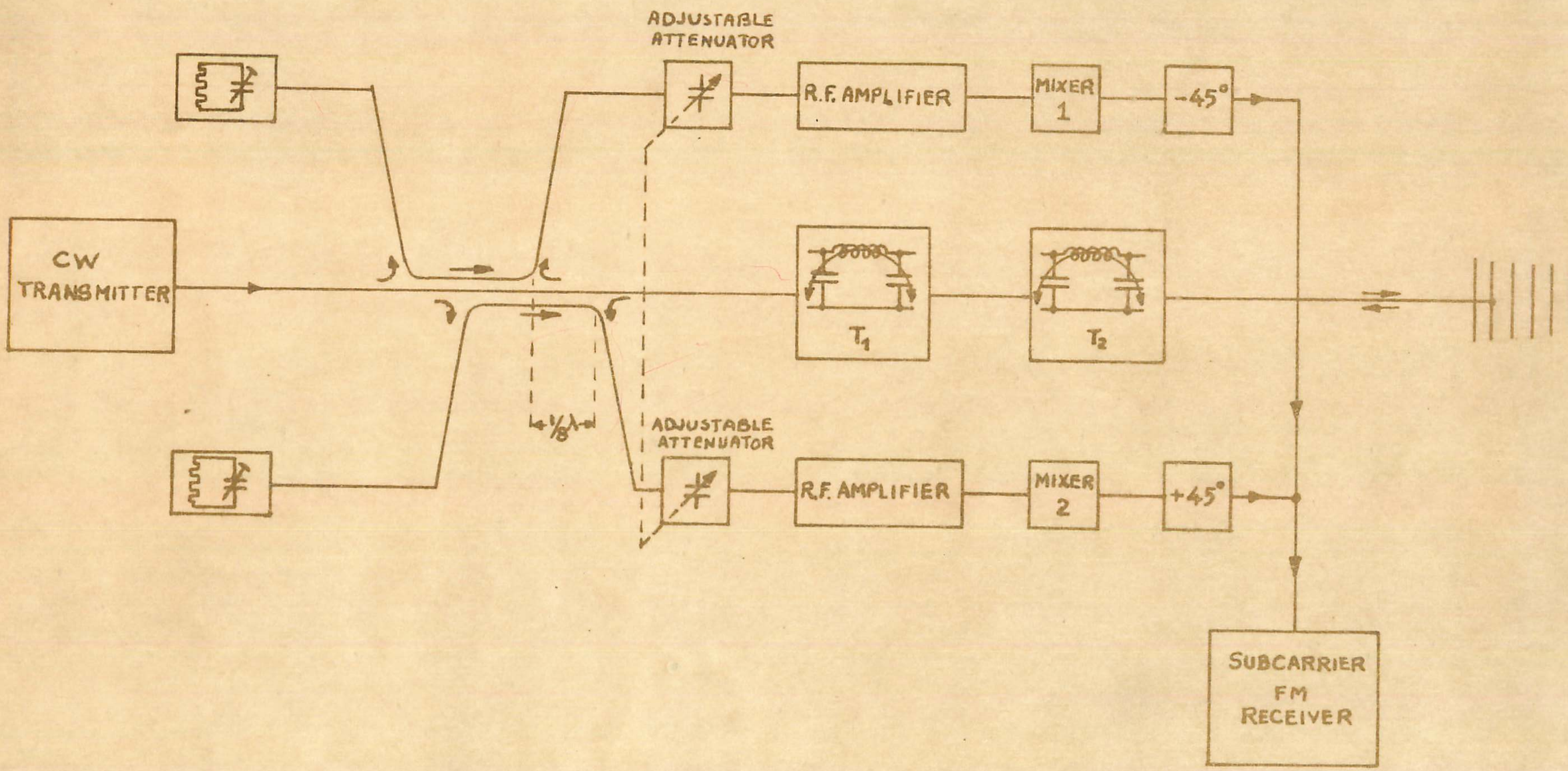

FIG. 7 AUTOMATIC DUPLEXER SYSTEM

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If the output component of mixer 2 is retarded 90 degrees relative to M_1 the two output components become:

$$c\ m\ A_o A_r \cos\varphi \cos pt$$

and

$$c\ m\ A_o A_r \sin\varphi \sin pt$$

Addition of these two outputs results in:

$$c\ m\ A_o A_r \cos\ (pt + \varphi\ ),$$

the amplitude of which is independant of φ .
Whereas the absolute value of the phase of a signal is of no importance and rapid movements of passive element or antenna, causing a shift in frequency of p, are not likely to be met, it will be clear that the final output is independant of the number of wavelengths between base station antenna and passive element. This means that a change of transmitter frequency or a change of antenna location does not cause a decrease of signal output due to out-of-phase effects.
In the block diagram fig. 7 is also shown an adjustable attenuator before the mixers. This attenuator can be used when, while aiming the antenna, the power reflected from objects in the vicinity is so large as to overload the mixers. The overloading can now be eliminated at the cost of a reduction in the signal to noise ratio, but without loosing the signal completely, which would be the case if transmitter power was decreased.
When a suitable antenna location is found the reflected power can be reduced with the tuners T_1 and T_2. The mismatching units before the loads in the coupling arms, which determine the local oscillator power level, are presets, adjusted for the right mixer excitation at maximum transmitter power. From a vector diagram it can be shown that an amount of reflected power equal to the local oscillator power level can be tolerated before the phase shifting action of the displacement of the two directional couplers is seriously deteriorated. The mixer excitation will therefore be adjusted on a level 3 dB below the point of overloading. This working point is indicated as P on curve B in fig. 6. The automatic duplexer will have in total three knobs, which generally will be used only to optimize the signal to noise ratio.

2.5 Antennas.

Two different antennas have been developed.
The first one has a gain of approximately 17 dB and is of medium size. It can be used as a replacement type for the E.C. 01 and "carrier Pigeon" antennas and should be regarded as an intermediate step towards the development of a larger antenna. This 17 dB antenna is a two-bay Yagi array of the same construction as the "carrier Pigeon" type of antenna, but with longer elements and a center support with base plate. When packed in a suitcase only 6 cm in height is required. The total weight is 2.2 kg ($\approx$ 5 lbs). The aiming facilities have been improved, especially when working in a downward direction. Irrespective of the polarization used, the antenna can be aimed in all directions.

The second one is a large antenna, being an array of four Yagi's in broadside configuration. The gain of this antenna is estimated to be 19 dB, which is the same gain as can be expected from a flat antenna (e.g. a "pine tree" antenna with reflector) with an area of 2 x 2 meters. This Yagi array has however a number of advantages over a flat antenna, taking into account the special requirements to be met under operational conditions. The dimensions of the array are about 1 x 1 x 1 meters. The array is mounted on a center support with base plate and can be aimed in all directions, irrespective of the polarization used.
When disassembled the antenna requires about 8 cm in height when packed in a suitcase.
The total weight of the antenna is approximately 3.4 kg ($\approx$ 7.5 lbs).
All Yagi elements as well as the T-supports are matched to 50 Ohms.
In case the ultimate range is not required it is therefore possible to use two Yagi elements instead of four, thereby decreasing the size of the assembled array, the estimated gain still being 17 dB.
A further possibility is to connect only one Yagi to the center support, realizing in this way a very handy antenna with a gain of approximately 14 dB.
By this conception a large, but at the same time very flexible antenna is obtained at the cost of only a small increase in size and weight.
If still more gain is required it is suggested -provided enough space is available- to connect an even number of these antennas in parallel by using coaxial cables of the correct lengths and matched T-joints.

3. Considerations in connection with the system engineering.

3.1 Consequences of the use of frequency modulation.

The use of a frequency-modulated subcarrier system for Easy Chair purposes brings along a number of attractive features, which are listed here:

a. The secrecy of the system is increased manyfold. A third party trying to locate a passive element of the new type will find that some of the commonly used countermeasures, such as regular search receivers, will be of little or no use.
Some of the possible jamming methods for Easy Chair will be quite ineffective.

b. Susceptibility to hum being reradiated by electrical wiring, conduit pipes and fluorescent lighting is greatly reduced.

c. Susceptibility to hum and microphony in the transmitter and radio-frequency parts of the base station is greatly reduced.

d. The signal-to-noise ratio for signals above the improvement threshold is improved.

e. A major part of the passive element consists of an oscillator, which in general is able to operate satisfactorily at lower supply voltages than amplifiers. The new passive element is therefore able to operate at lower levels of radio-frequency excitation than previous models, resulting in extended range.

The experimental passive element of the subcarrier type has a modulation index of about 1.4 for normalized operating conditions. The corresponding improvement in signal-to-noise ratio is in that case 7.6 dB as compared with an amplitude-modulated system.

The modulation index of 1.4 and the highest audio frequency to be passed being 4 kc/s, require a minimum receiver bandwidth of about 15 to 20 kc/s to accomodate the major modulation sidebands.

The audio bandwidth of the system is assumed to be 4 kc/s. The passive element has a bandwidth much wider than 4 kc/s, but the microphone used, a miniature magnetic type, will not respond appreciably to frequencies higher than about 4 kc/s.
Moreover it may be doubted whether a larger bandwidth will always ensure a better intellegibility in view of the unproportionaly increased noise power. In frequency modulation systems the noise power in the output of the receiver is proportional to the square of the audio bandwidth if other constants remain unchanged.

The improvement in signal-to-noise ratio mentioned will only be effective for signals having a definite minimum level compared to the noise present. Below this minimum level the intelligibility falls off very rapidly. For signal levels near the minimum level the output noise takes on a intermittent character, for which reason this minimum level has been termed the "sputter point".

Full use of the noise-reducing properties of frequency-modulation can only be expected when the frequency modulation detector is preceded by an effective amplitude limiter.
This limiter also maintains the output signal level at a constant value, irrespective of the incoming signal strength. A stronger incoming signal will therefore only result in a reduction of noise background at the output of the receiver.

3.2 Receiver bandwidth.

In the following discussion the actual parameters of the experimental subcarrier type of passive element will be used as a basis.
The subcarrier mean frequency is amongst others dependent on the radio-frequency excitation level and is about 93 kc/s at the recommended operating level of -40 dBW. For extreme values of activation power the frequency may be as high as 115 kc/s or as low as 70 kc/s.
The modulation index of the frequency modulation is about 1.4 for normalized conditions and for a highest modulation frequency of 4 kc/s.
The modulation index is also dependent on excitation level and may in extreme cases reach a value either two times higher or lower.

If the receiver bandwidth is chosen as narrow as possible, e.g. 20 kc/s, the optimum signal-to-noise ratio will be obtained in the output of the receiver. In view of the range of center frequencies likely to be encountered a provision must be made to tune the bandpass center frequency through the range of e.g. 55 to 135 kc/s. This range is appreciably wider than the range required by the experimental passive element, but allows for production tolerances of different units.
The narrow bandwidth makes the setting-up of a system rather difficult, due to the extra tuning knob to be adjusted along with other setting-up adjustments.
Moreover, in an already set-up system, the narrow bandwidth will cause the system to be susceptible to breaks in operation under varying propagation conditions, such as people walking through the antenna beam.

On the other hand the receiver can have a large bandwidth, e.g. 55 to 135 kc/s to accomodate all operating conditions without necessity of retuning. Tuning facilities can be omitted and the setting-up procedure be facilitated, at the cost however of a worse signal-to-noise ratio. For the figures mentioned, e.g. a bandwidth of 80 kc/s, the signal-to-noise ratio is 6 dB worse than the optimum obtainable.
Omitting the tuning provisions saves some size, weight and circuit complexity.

The choice seems to be between higher operating complexity and decreased performance.

A compromise might be considered.
Suppose a receiver bandwidth of 40 kc/s were chosen, whose center frequency could be switched to three different values, e.g. 75, 95 and 115 kc/s.
The loss in signal-to-noise ratio between this compromise and the optimum obtainable would be 3 dB.
The bandwidth of 40 kc/s would however be wide enough to accomodate most of the subcarrier frequency range to be encountered with a particular passive element during the setting-up procedure, all with the switch in a fixed position. Working with passive elements at the low or high side of production limits, or operation at extreme activation levels would require another position of the switch to be chosen.
It is also expected that the size and weight of this compromise will be smaller than a tunable narrowband circuit.
The 3 dB lower-than-optimum signal-to-noise ratio may be greatly compensated by a suitable choice of duplexer parameters, the reduction in range probably not exceeding some 5%.

Utmost-performance systems might reject any compromise resulting in range reduction and choose a narrow bandwidth.
Here again there might be a choice, in this case between higher adjustment complexity by the addition of a tuning control, or a higher circuit complexity by the addition of an automatic frequency control system.
Equipments designed for relatively short ranges and requiring the utmost saving in size and weight might be of the simple wide-band variety.

3.3 The choice of tubes and/or transistors in the receiver.

In par. 2.2 two types of radio frequency receiver front ends were considered, either a silicon crystal diode detector or a radio frequency amplifier and detector using vacuum tubes. In choosing between these two possibilities the following facts must be born in mind:

a. The vacuum tube front end has a bigger size, weight and circuit complexity.

b. The vacuum tube front end requires filament and anode voltage supplies and produces heat.

c. The vacuum tube front end is superior with respect to noise figure, protection against damage by overloading and it allows the operation at considerably higher local oscillator power levels, facilitating operation.

When extended range in combination with improved operational facilities is aimed at, the vacuum tube front end is strongly advisable.
The use of a radio frequency amplifier of the lighthouse variety, although slightly bigger in size than the regular miniature type with Noval base, is considered justified by the better characteristics, illustrated in fig. 6 of par. 2.2.
The remainder of the receiver, consisting of a subcarrier frequency amplifier, amplitude limiter, frequency detector and audio amplifier can be built either with regular vacuum tubes or with transistors.
One advantage of a fully transistorized receiver, viz. the possibility to provide maximum operational flexibility by using an independent power supply of dry batteries, must be discarded in view of the high power demands of the vacuum tube front end. The use of a hybrid power supply system has no attraction in this case.
Another advantage of the use of transistors remains valid: the smaller size and weight may weigh out the increased size and weight of the vacuum tube front end.
The use of transistors however in an integrated receiver design requires thermally well-stabilized transistor circuits and a very careful mechanical design with respect to heat isolation and heat exchange.
A good solution for these problems might eventually lead to an integrated design, in which transmitter, receiver and duplexing means are combined into one unit, the entire power supply constituing another unit, in order to distribute heat production and weight.
This integration and subdivision would provide a maximum in operating facility and reduce the number of interconnecting cables to a minimum. Moreover all controls and metering facilities would be centralized as far as possible.
The realization of this possibility depends mainly on the degree of success in solving the temperature control problems mentioned.

3.4 The transmitter.

In view of the desire of the contracting party to direct the effort not in first instance to a system using higher transmitted power, it was decided to stick for the time being to the existing power level of about 40 W.

The decreased susceptibility of the subcarrier frequency modulation system to hum and microphony originating in the transmitter, as compared to the EC-02 and CP-01 equipments, allows the omission of anti-microphony circuits and a reduction of filtering in the power supply system.

Comparable transmitters are the EC-02 transmitter of the master oscillator-power amplifier type and the CP-01 transmitter of the crystal oscillator-frequency multiplier-power amplifier type.
The EC-02 transmitter, although relatively simple and compact, was not the last word in operating stability, easy maintenance and logical servicability.
The CP-01 transmitter on the other hand might surpass electronically the requirements set for EC-03 and is more complicated as to circuitry and number of stages.

A compromise transmitter of the master oscillator-buffer amplifier-power amplifier type was considered. The master oscillator in this case would operate at a relatively low power level and therefore use tubes better suited for oscillator applications and with lower microphony than the tubes used in the EC-02 master oscillator.
Frequency stability and independence of tube tolerances would be improved upon.
This set-up would miss a great deal of the circuit complexity of the CP-01 transmitter, yet would very probably meet the requirements as to frequency and power stability for EC-03 purposes.
Overall size, weight and efficiency are expected to be intermediate between those of EC-02 and CP-01.
Experiments along this line, although not yet fully completed, tend to confirm the views expressed above.

Still another possibility was considered, being the use of a cavity power oscillator operating directly at the ultimately required power level and using e.g. a coaxial tube type such as the 4 X 150 A.
No significant increase in overall efficiency is expected if minimum stability requirements have to be met. Continuous output power control over a wide range will be more difficult to achieve. It might however be a practical proposition in cases where a variable frequency transmitter is required.

With these considerations in mind it was decided to give preference to the master oscillator-buffer amplifier-power amplifier type of transmitter. If the experiments on this type might not lead to a satisfactory solution, the CP-01 type of transmitter will be regarded as an alternative and will be re-engineered slightly.

3.5 System Balance, Duplexer choice and Range Evaluation.

Using the results of the measurements given in section 2, the performance of the system can be calculated for several possible arrangements. The following parameters are involved. All powers are expressed in dB relative to 1 Watt and the ratio's in dB.

Transmitter power	P	dBW
P.E. excitation level = r.f. power level at P.E.	P_e	dBW
Gain of base station antenna	G	dB
Gain of P.E. antenna	2.1	dB
Transmission loss = $\frac{\text{r.f. power at input base station ant.}}{\text{r.f. power at output P.E. ant.}}$	L	dB
Total subcarrier sideband power generated by P.E.	P_r	dBW
P.E. modulation depth = $\frac{P_r}{P_e}$	M_s	dB
P.E. f.m. modulation index	M_a	dB
Total sideband power at input receiver	P_s	dBW
Receiver noise figure	F	dB
Receiver bandwidth	B	c/s
Insertion loss directional coupler	I	dB
Coupling loss directional coupler	C	dB
Signal to noise ratio at output receiver	S	dB

Considering the transmission path from transmitter to passive element it will be clear that:

$$P - P_e = L + I \quad \text{.......................} \quad (1)$$

Consideration of the transmission path from passive element to receiver leads to:

$$P_r - P_s = L + C \quad \text{.......................} \quad (2)$$

The relation between P_e and P_r is given by:

$$P_r = P_e + M_s \quad \text{.........................} \quad (3)$$

The signal to noise ratio at the receiver input is P_s - 2 kTB, where kTB also is expressed in dBW. The factor 2 had to be introduced because there is no image rejection. Taking into account the receiver noise figure, the f.m. improvement of 4.6 dB for a modulation index of 1 (see par. 3.1) and M_a, the following expression is found for the signal to noise ratio at the output of the receiver:

$$S = P_s - kTB - F + 4.6 + M_a \quad \text{..........} \quad (4)$$

With the equations (1) to (4) the performance of the system can be evaluated.

The transmitter power and the gain of the base station antenna are in the first instance governed by operational considerations such as the maximum allowable weight and storage space and for the antenna also the maximum allowable dimensions when erected. For relatively high transmitter powers the fact that tuning becomes more critical, due to the necessity of better cancellation of reflected power and the degree to which this can be performed under practical circumstances, becomes an important factor. With the advent of the automatic duplexer and the vacuum tube r.f. amplifier and detector, this effect is considered not to set a limit for the transmitter power within the range of powers now under consideration.
By the use of the subcarrier system the problem of hum interference is eliminated to a great extent.
When choosing transmitter power and antenna gain it seems best to start with the antenna and to choose the antenna gain as high as is consistent with the operational requirements. The space and weight requirements for the Yagi antennas are small in comparison with those of other types. Furthermore it should be borne in mind that an increase in antenna gain is effective in the transmitting path as well as in the receiving path.
For transmitter power and antenna gain thus given, the duplexer parameters can be calculated for the system to be in balance. The system balance problem comes in here because the passive element modulation characteristic is non linear; this means that the modulation depth, and in this system also the modulation index, are dependent on the excitation level.
For linear passive elements, such as those of the cavity type, the system is always in balance and an optimum 3 dB directional coupler should be used. In that case one half of the transmitter power is being lost in the load (instead of in "free space" as in the two antennas system) and the total duplexer loss is 6 dB, as is the inherent minimum for all cw-systems, whether they use two antennas or one.

To evaluate the optimum directional coupler constants, the complete curve of receiver output versus transmission loss should be drawn for a number of coupling ratio's. One should start with the curve for a 3 dB coupler. This curve can be calculated from the measurements given in section 2, using the equations (1) to (4) or can be measured directly in an overall test set up incorporating a 3 dB directional coupler.
From this curve the curves for other coupling ratio's can be derived by a graphical method. As an example the curves have been drawn for a transmitter power of 40 Watts and using the passive element characteristics given in fig. 4. The receiver sensivity was assumed to be 11 dB and to be independent of the local oscillator power level (see fig. 6, curve B). The receiver bandwidth was assumed to be 40 kc/s.
The theoretical system performance is given in fig 8. When displacing curve "a" parallel to itself and keeping thereby point A on the 3 dB-curve, the values of S versus L for all possible

coupling ratio's can be found immediately. Curve "a" accounts for the effect of different duplexers at constant passive element excitation level, and can therefore be used universally, thus facilitating calculation for other passive element characteristics.
In choosing the most suitable one from the set of curves thus obtained, one should not pay attention to maximum range only. Two other factors are also of importance. The 7 dB curve fulfills the requirement of maximum range. The 10 dB curve however has only slightly less range but a 10 dB directional coupler provides about 2.5 dB better isolation between transmitter and receiver than a 7 dB coupler and will therefore make tuning easier.
Furthermore it is not advisable to use the passive element at an activation level, which is close to a cut-off point, because a slight change in the propagation conditions might interrupt the system to be operative.
If the passive element has only a limited range of powers, for which it is active the signal to noise ratio obtained over the working range of the passive element should be centered around the minimum required one. This requirement is also met by the 10 dB curve.

This graphical procedure is preferred above a direct equating of transmitting and receiving path because results obtained in that way will not always be optimum.
The automatic duplexer has not been incorporated in the considerations given above, because of the extra complications involved in the calculations. To a first approximation it can be said that, when high coupling ratio's can be used the signal to noise ratio at the output is reduced by 3 dB, because of the fact that the noise outputs of the two mixers add, whereas the signals are in phase quadrature. Moreover the transmission loss will be reduced by the insertion loss, which is about 0.5 dB for a 10 dB coupler. The curve "b" in fig. 8 indicates the expected performance of the system when an automatic duplexer is incorporated.

Finally the following statements can be made.
For a transmitter power of 40 Watts the receiver sensitivity is more than is strictly necessary, the surplus being used to obtain some increase in range by unbalancing the coupler ratio's. Because a coupling ratio of 10 dB appeared to be optimum, it will be clear that much higher powers can be used with advantage, leaving the balancing problem out for the moment.
The possibility to transmit power from transmitter to passive element, in order to reach the minimum excitation level, is still a limiting factor.

The curves of fig. 8 have been calculated as a function of the transmission loss L, because L is a more universal measure in comparing systems than the range only, as it leaves out propagation conditions and the gain of base station and passive element antennas.

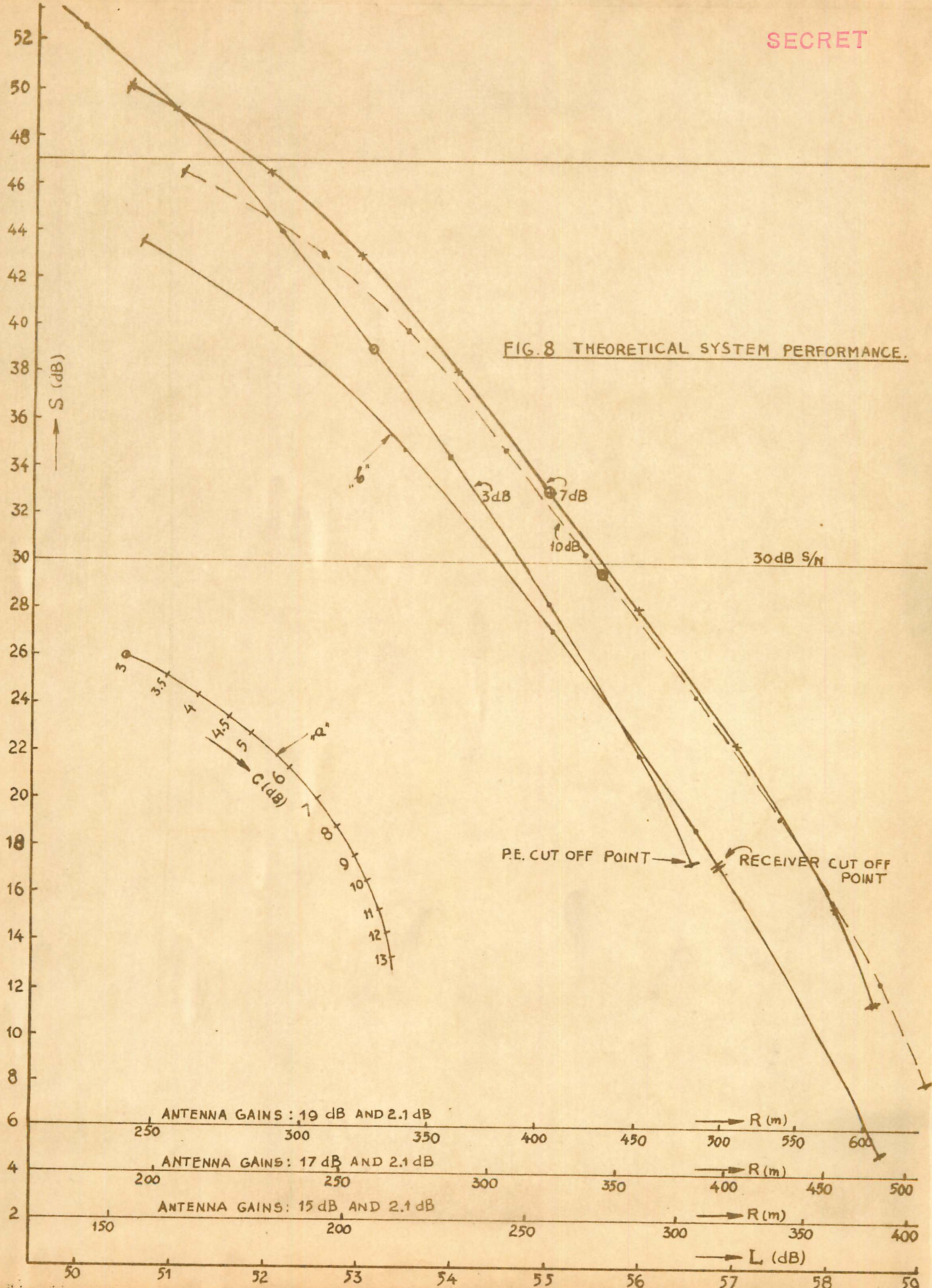

FIG. 8 THEORETICAL SYSTEM PERFORMANCE.

The relation between the free space range R and the transmission loss is (expressed in dB):

$$\left(\frac{4 \pi R}{\lambda}\right)^2 = G + 2.1 + L$$

For a wavelength of 79.5 cm and three values of G the free space ranges have been calculated as a function of L. These results can be deducted from fig. 8 by using the three additional range scales.

4. Preliminary specifications for an equipment with extended range: type EC-03.

Based upon the considerations mentioned in paragraph 3 a choice was made for the construction of a new type of Easy Chair equipment with extended range, type number EC-03. When packed, the dimensions and weight of the equipment will not exceed those of the previous types.
In this equipment full use will be made of the sub-carrier frequency modulation technique.
Automatic duplexing is incorporated leading to the use of one antenna for both transmission and reception, ensuring at the same time a greatly simplified tuning procedure.
A suitable combination of tubes and transistors will be used, the ultimate choice in each instant being governed by the specific requirements of each stage in the circuitry and the specific properties of these elements. A squelch circuit suppresses excessive noise in the receiver output under no-signal conditions improving the operational convenience during the setting-up procedure. In the receiver the sensitivity will be improved by the incorporation of an R.F. amplifier stage with a lighthouse tube and a thermionic detector, both being duplicated in connection with the automatic duplexer system.
The transmitter may be crystal controlled and will have an output power of the order of 40 Watts. Due to the sub-carrier frequency modulation technique an anti-microphony circuit is no longer a requirement.
The antenna of the EC-03 equipment is estimated to have a maximum gain of 19 dB when all the elements supplied are used, being in that case a 4-bay Yagi array. In case the ultimate range is not required, a smaller type of antenna system can be assembled by using a smaller number of elements.
The passive element of a new type will be fully miniaturized and the low frequency circuit elements will be embedded in epoxy resin. The size of the passive element will be comparable with that of the ones used with previous types of equipment.
The complete equipment will be packed in two suitcases of conventional size.
Taking reference to point 3.5 of this report the expected maximum range of the EC-03 equipment will be several times as large as the maximum range obtained with the previous system.

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