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AUSTRALIAN BROADCASTING CONTROL BOARD

562-574 Bourke Street, Melbourne

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ENGINEERING SERVICES DIVISION

REPORT NO. 38

TITLE: A TWO CARRIER FREQUENCY MODULATION STEREOPHONIC RECEIVER FOR THE UHF BAND.

Issued by:

The Chairman,
Australian Broadcasting Control Board,
562-574 Bourke Street,
MELBOURNE, VIC., 3000

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AUTHOR: G.T. MARTIN - FORWARD PLANNING AND DEVELOPMENT SECTION

11 August 1975.

The Government decision to make VHF channels available for FM sound broadcasting in Australia by displacing the television services now operating in the band 88-108 MHz makes it no longer necessary to develop UHF FM services. The work carried out in the receiver development is published for the information of other workers in this field.

(E. J. Wilkinson)
Director
Engineering Services Division

SUMMARY

Design and performance details of an ultra high frequency band two carrier frequency modulation receiver are described. This receiver was developed as an experimental model to investigate the feasibility of the two carrier system where separation of the carriers depends on the selectivity of phase lock loop discriminators, and to gain insight into any possible problems of solid state UHF receiver design, directed eventually to the mass production receiver industry. At the stage when this project was current, it was proposed that Australia adopt a frequency modulation broadcasting service in the UHF band. Several alternative systems of achieving stereophonic or even quadraphonic transmissions were under study, with a view to taking advantage of the increased bandwidth available in the UHF spectrum to overcome the severe signal to noise disadvantage of the pilot tone system in the stereophonic mode.

CONCLUSION

A successful UHF FM receiver was produced, of relatively simple construction. It was stable in operation, and as easy to tune as a conventional VHF pilot tone FM receiver.

In the demonstration, good quality stereo reception was obtained with a signal input to the receiver as low as 2 uV.

Under demonstration conditions receiving music, most listeners found the only detectable difference between a direct connection and the UHF FM system was a slight increase in noise caused by the phase lock loop integrated circuits.

RECEIVER DESCRIPTION

See Fig. 1 for the block diagram.

A wideband front end and IF amplifier are used, wide enough to accomodate the two carriers, which are treated as one signal. For example, with a peak deviation of 75 kHz, the bandwidth required to include the 1% sidebands for one carrier is 240 kHz, making the minimum bandwidth for two carriers with no overlap above 1% sidebands 480 kHz. If 400 kHz carrier spacing is used this gives a 160 kHz central

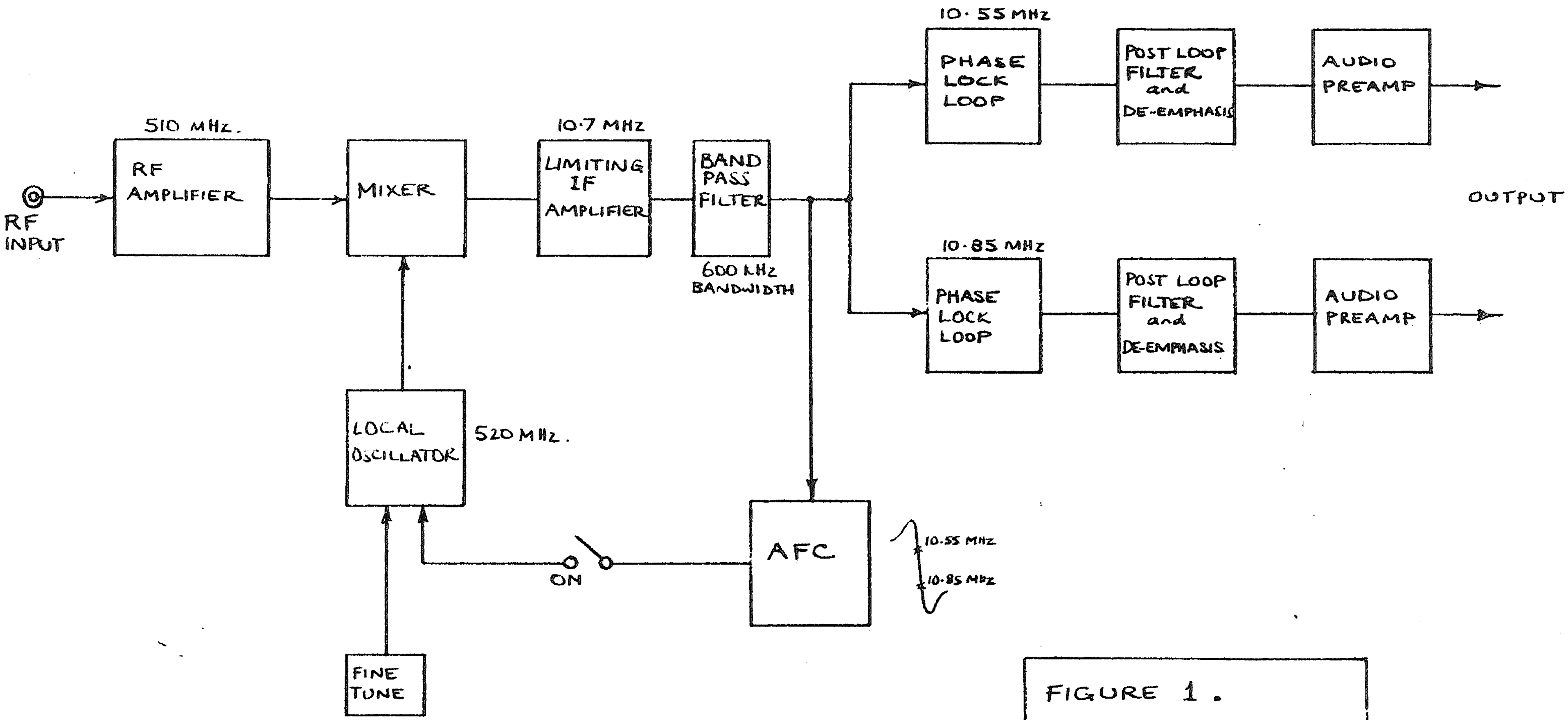


FIGURE 1.
BLOCK DIAGRAM — TWO CARRIER UHF FM RECEIVER.

guard band and an overall bandwidth requirement of 640 kHz.

The front end is fixed tuned to approximately 510 MHz. The local oscillator operates on the high side of the received signal, and the IF is 10.7 MHz. A wideband IF amplifier and limiter provides about 80 dB gain and is followed by a bandpass filter. A signal to operate the AFC loop is taken from the IF output, and a tuning meter is provided at this point.

The two phase lock loops are fed in parallel through a resistive attenuator to define the input signal level and hence the tracking range, assuming that the IF amplifier is limiting. One phase lock loop is tuned to a centre frequency of 10.55 MHz, the other to 10.85 MHz, and these provide the selectivity to separate the two channels. (300 kHz separation)

The phase lock loops are followed by post detection filters and 75 uS. de-emphasis networks, which feed audio preamplifiers.

Varicap fine tuning is provided on the local oscillator, and this can be controlled manually or by the AFC loop.

The AFC operating characteristic is broadened to allow both carriers to assume a symmetrical position within the linear portion of the loop discriminator S curve.

RADIO FREQUENCY AMPLIFIER

Two common base BF180 transistor RF stages are used, with trough line tuning. Trough transmission lines consist of a conductive bar enclosed in a shielding metal trough. The lines are less than a quarter wavelength long, and are earthed at one end, giving an inductive impedance at the far end which is tuned with a small trimmer capacitor.

Various impedances can be obtained by tapping along the bar. This sort of interstage network offers high unloaded Q, stability, ease of construction and adjustment, and lends itself to empirical selection of matching tap points.

This latter feature was taken advantage of to quickly obtain some satisfactory front end hardware.

The local oscillator uses BF183 transistor, in a Colpitts circuit , with a trough line tank and varicap fine tuning. Careful filtering of the varicap voltage control lead is essential to provide stable operation. A small loop near the tank

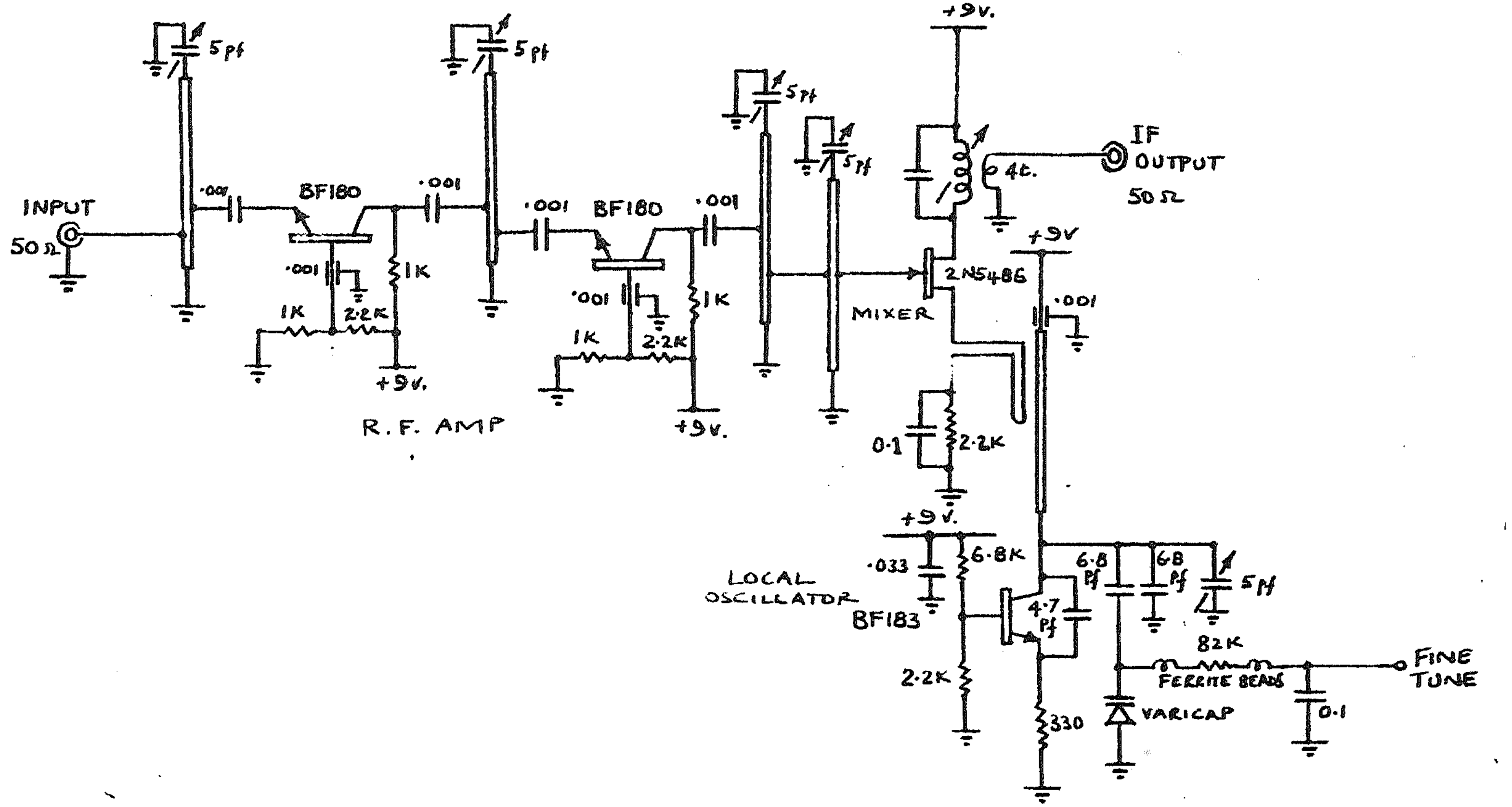


FIGURE 2.
R.F. FRONT END

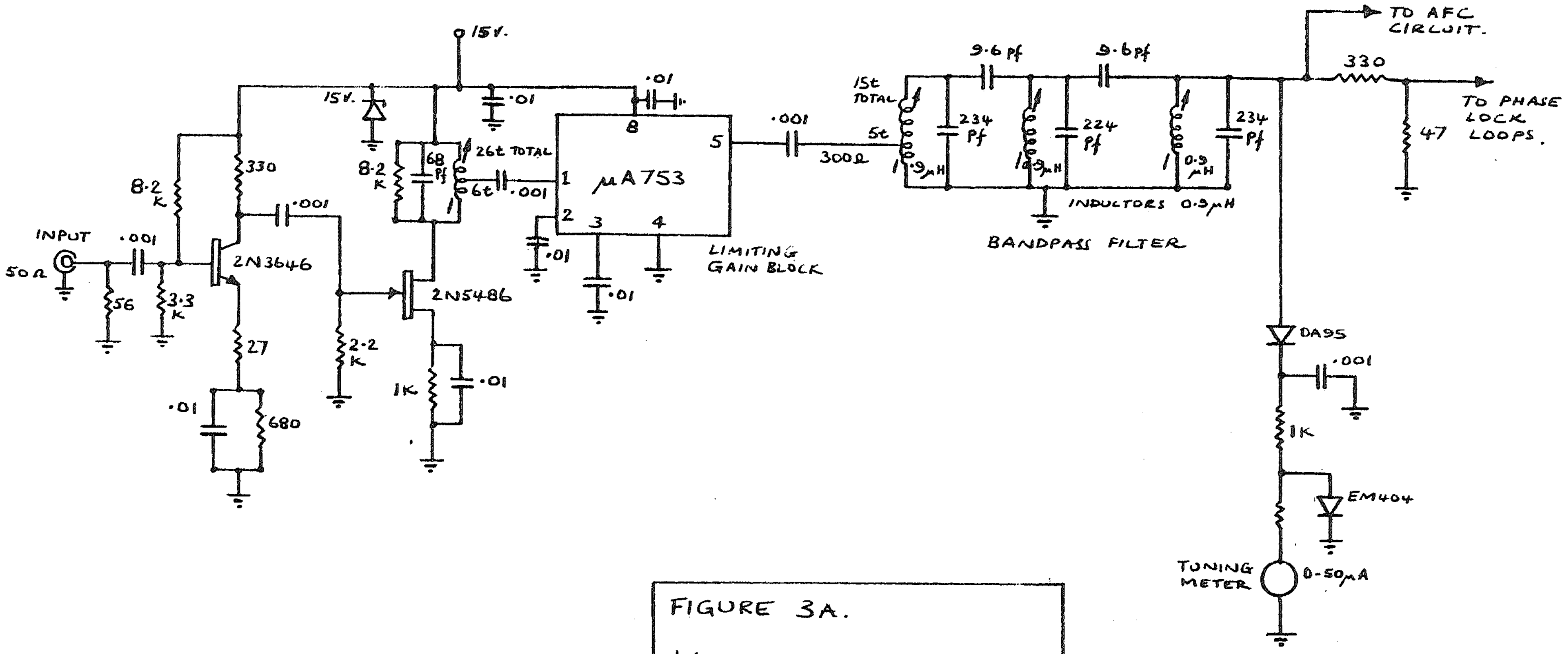


FIGURE 3A.

LIMITING IF AMPLIFIER AND BANDPASS FILTER line couples the local oscillator signal into the source of the 2N5486 mixer FET, which mixes down to the intermediate frequency of 10.7 MHz.

Local oscillator frequency is above the received signal frequency, e.g. 521.10 MHz for a received signal frequency of 510.40 MHz.

See figure 2.

10.7 MHz INTERMEDIATE FREQUENCY AMPLIFIER

The IF amplifier is built around a Fairchild μA753 gain block-limiter with the bandwidth defined by a 3-stage bandpass filter following the limiter. See figure 3.

Charactistics of the IF amplifier are important to the overall receiver performance. The bandpass filter defines the predetection bandwidth, but must be wide enough to pass both modulated carriers.

Effective limiting action is important to feed a defined signal level to the phase look loops so the tracking range, which provides the selectivity for separating the two carriers, is defined.

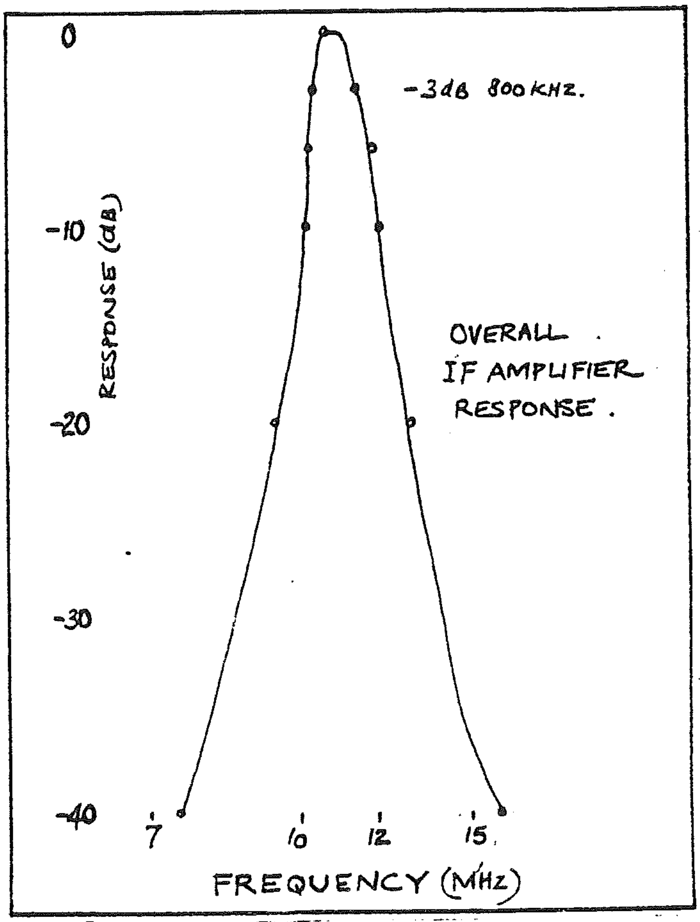


FIGURE 3B.

Used alone, the uA753 has insufficient gain (55dB). So a pre-amplifier incorporating a single low Q tuned circuit precedes the uA753, giving an overall gain of very approximately 80 dB.

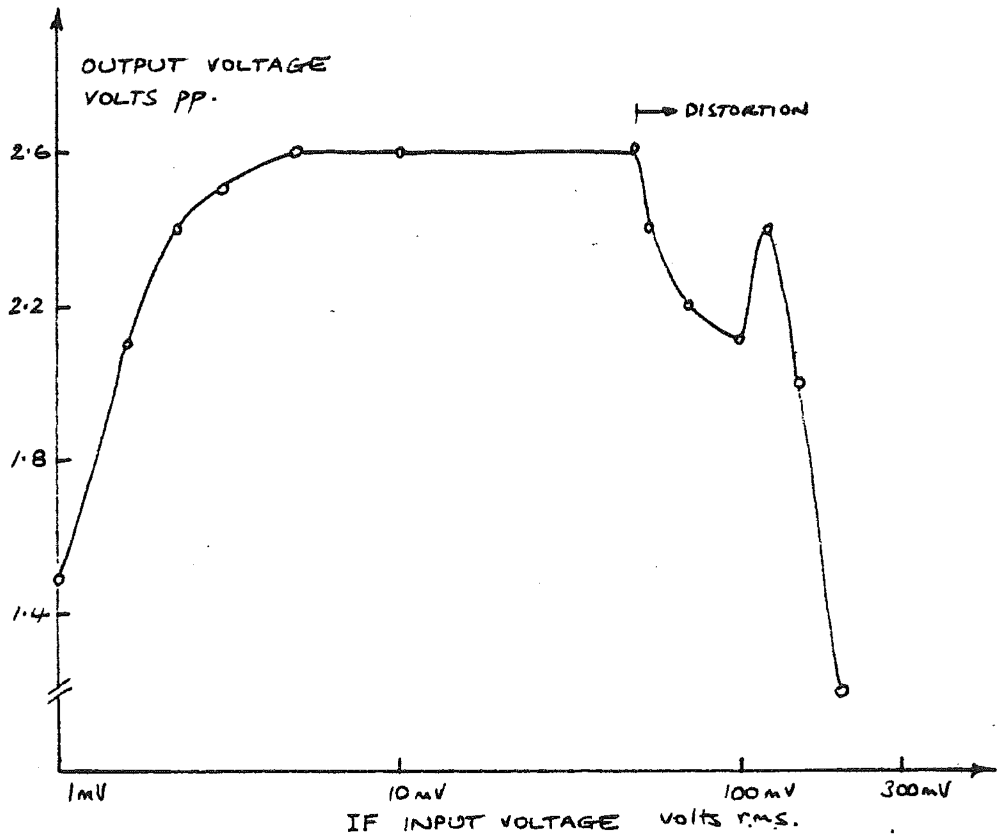


FIGURE 3C.
IF AMPLIFIER LIMITING CHARACTERISTIC

AUTOMATIC FREQUENCY CONTROL

An AFC loop is provided, using a TBS 261 amplifer/quadrature detector, followed by a single transistor DC amplifier to adjust levels and provide phase reversal. AFC may be switched on or off; when on it largely overrides manual fine tuning. The Q of the quadrature detector is lowerd by a 2.2 K ohm shunt, broadening the S curve operating range to about 1.5 MHz. Under normal operation the two carriers will position symmetrically on the central section of the S curve. The loop will lock in with one carrier somewhat over the S curve peak and the other still in the operating region, depending on the relative slopes of the two portions of the curve.

In practice the receiver was found to have sufficient local oscillator stability to operate for periods of up to about an hour without AFC and without noticable deterioration of audio quality.

See figure 5.

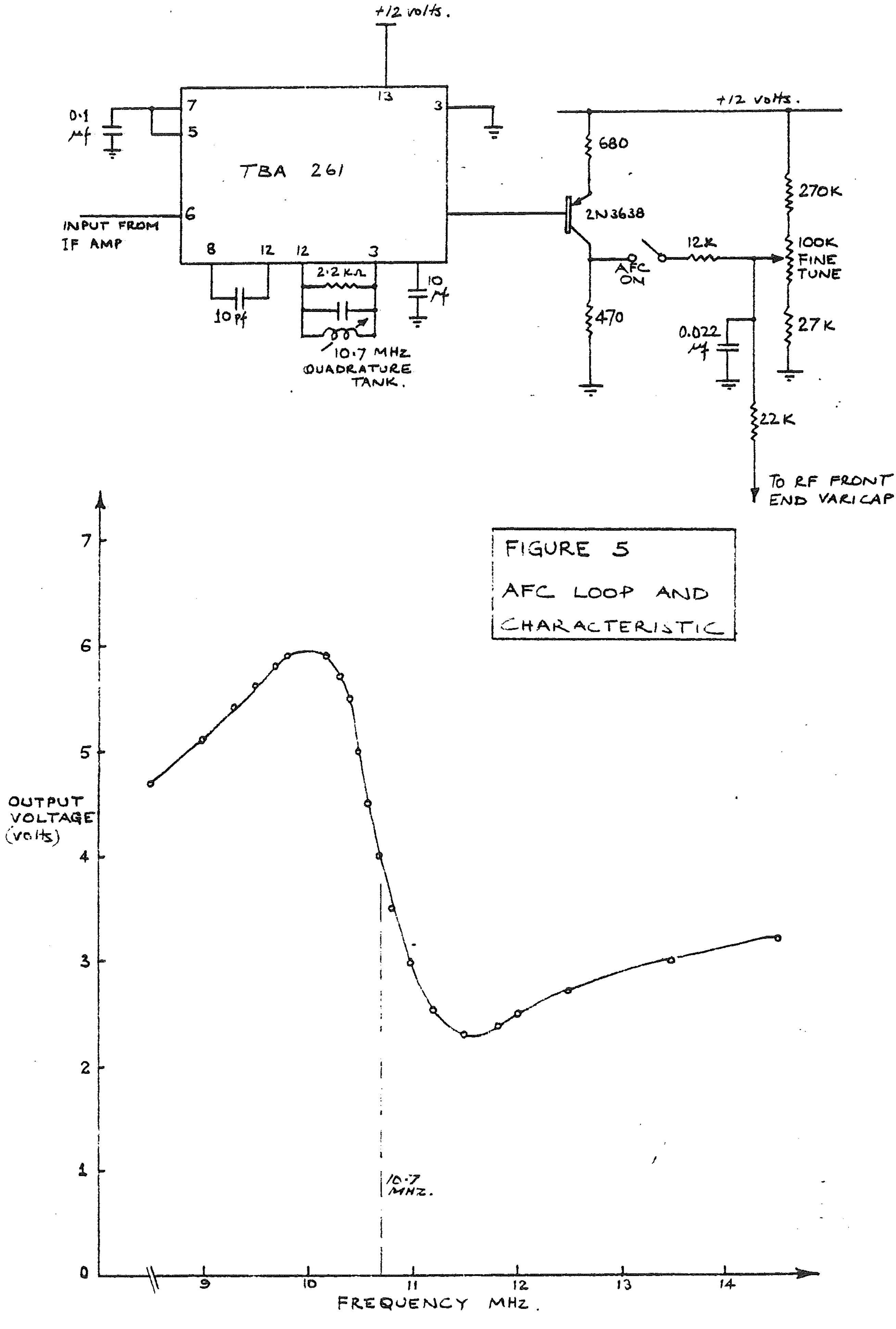


FIGURE 5

AFC LOOP AND CHARACTERISTIC

PHASE LOCK LOOP DESIGN

Integrated circuit phase lock loops are used, leaving the choice of the loop gain and loop filter parameters to the designer. An imperfect second order loop with a finite zero is chosen, as it is unconditionally stable, and allows an almost independent choice of loop bandwidth and damping factor. This is realised with a passive RC lead-lag filter, which initiates the filter pole away from the origin; if the loop filter was a perfect integrator the pole would initiate at the origin.

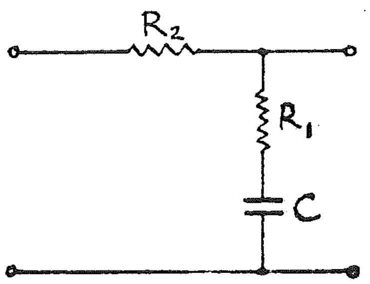
Total tracking range (or hold in range) equals twice the loop gain, and because of the characteristics of the phase detector, loop gain is a function of input signal level.

As the selectivity of the phase lock loops provides the only mechanism for separating the two carriers, the tracking range must be restricted to a value sufficient only to follow the modulation, plus a safety factor to allow for drift in the PLL centre frequency. Once the input signal amplitude to the PLL is defined by the IF limiting amplifier, tracking range can be set using an external resistor which adjusts overall loop gain. On the basis of an initial experiment, this was chosen to give a tracking range of about 250 kHz for 20 millivolts peak to peak input, the value required being 820 ohms.

Having defined the loop gain, which sets the static performance of the loop, design of the loop filter which governs the dynamic loop performance, can procede once values are chosen for the damping factor δ and natural frequency ω_n. These terms have the same significance as for a second order servomechanism. The loop noise bandwidth is a function of δ and ω_n, and is a minimum for $\delta = 0.5$. As δ reduces, noise bandwidth tends to increase rapidly, and phase error following an input pertubation increases. A damping factor of $\delta = 1/\sqrt{2}$ is traditionally chosen, and for this value the noise bandwidth is only 6% greater than the minimum value. We choose $\delta = 0.6$ which is very close to the minimum noise bandwidth. Within the range of value 0.5 to 1.0, the choice of δ is not particularly critical.

For demodulating sinusoidal FM, the steady state phase error at the highest modulating frequency must be $\leq \frac{\pi}{2}$ otherwise the loop will loose lock. In practise a margin to allow for noise is provided, so the steady state phase error is kept $\leq \frac{\pi}{4}$.

From Fig. 4-1 (Ref. 1), with $\delta = 0.6$, if fn = 40 kHz and the maximum modulating frequency fm = 15 KHz, $\omega_m / \omega_n = 0.38$ giving a steady state phase error of 0.75 radians. This allows a 0.8 radian safety margin for noise.



LOOP FILTER

$$\tau_1 = R_1 C$$
$$\tau_2 = R_2 C$$

The loop filter equations are

$$\omega_n = \left(\frac{K}{\tau_1 + \tau_2} \right)^{1/2}$$

$$\delta = \frac{1}{2} \omega_n \left(\tau_2 + \frac{1}{K} \right)$$

These are solved for τ_1 and τ_2

$$\tau_1 = \frac{K}{\omega_n^2} - \frac{2\delta}{\omega_n} + \frac{1}{K}$$

$$\tau_2 = \frac{2\delta}{\omega_n} - \frac{1}{K}$$

For a tracking range of $\pm$ 125 KHz,

loop gain K = $125 \times 2\pi \times 10^3$

natural frequency $\omega_n = 40 \times 2\pi \times 10^3$

damping factor $\delta = 0.6$

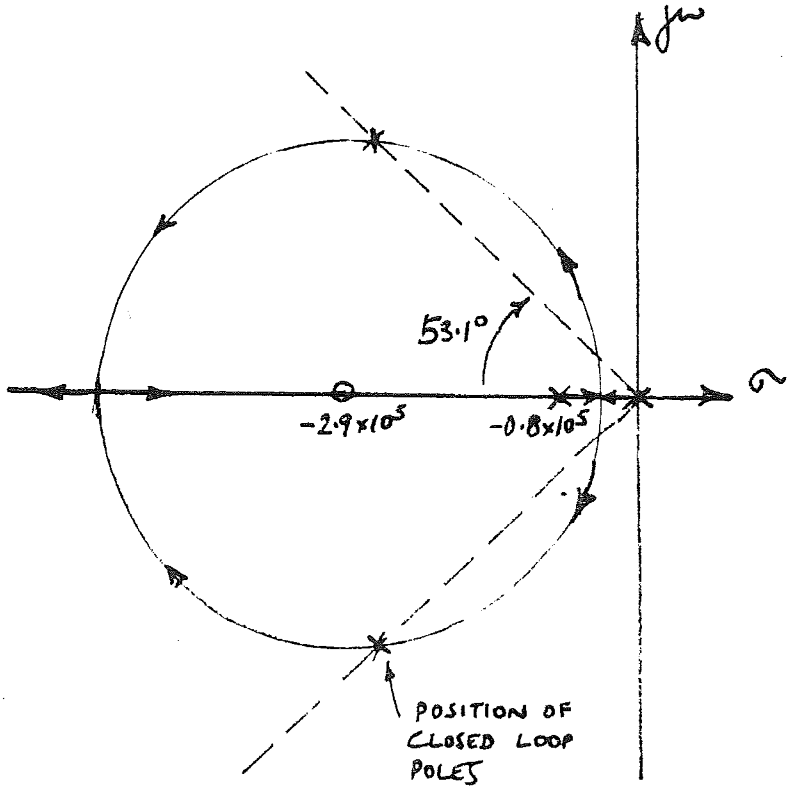
thus:

$$\tau_1 = 3.5 \times 10^{-6}$$
$$\tau_2 = 8.9 \times 10^{-6}$$

$$C = \frac{\tau_2}{R_2} = 1.5 \text{ nanofarads}$$

$$R_1 = \frac{\tau_1}{C} = 2.2 \text{ kilohms}$$

Note that $R_2 = 6$ Kohm is an internal component within the integrated circuit.



ROOT LOCUS PLOT

FIGURE 6

Closed loop poles lie on a line at angle Q from the negative real axis where

$$\text{artan}\ \theta = \frac{(1-\delta^2)^{1/2}}{\delta}$$

$$\delta = 0.6$$

So $\quad \text{artan}\ \theta = 1.33$

and $\quad \theta = 53.1$ degrees.

Reference 1: "Phaselock Techniques" F.M. Gardner, John Wiley and Sons, Inc. 1966

NOTE : PHASE LOCK LOOP :

(i) $820\,\Omega$ between pins 14, 15 to reduce loop gain.

(ii) $27\,k\Omega$ current injection resistor on pin 7.

POST DETECTION FILTER

The post detection filter is a lowpass filter following the phase lock loop demodulator, and serves to define the system noise bandwidth.

At the demodulator output considerable high frequency noise exists, much of it being at the loop voltage controlled oscillator frequency, and although this is inaudible it makes noise measurements difficult. The post detection filter cleans up the output considerably and makes noise measurement easier to perform.

An active filter is used, giving an ultimate slope of 18 dB per octave. The circuit is shown in Figure 7 and the response curve is shown in Figure 8 , which also gives the response of a 75 microsecond de-emphasis network. Note that the de-emphasis network overrides the effect of the post detection filter. However in a prototype model where measurements may be made without the de-emphasis network, the filter is useful.

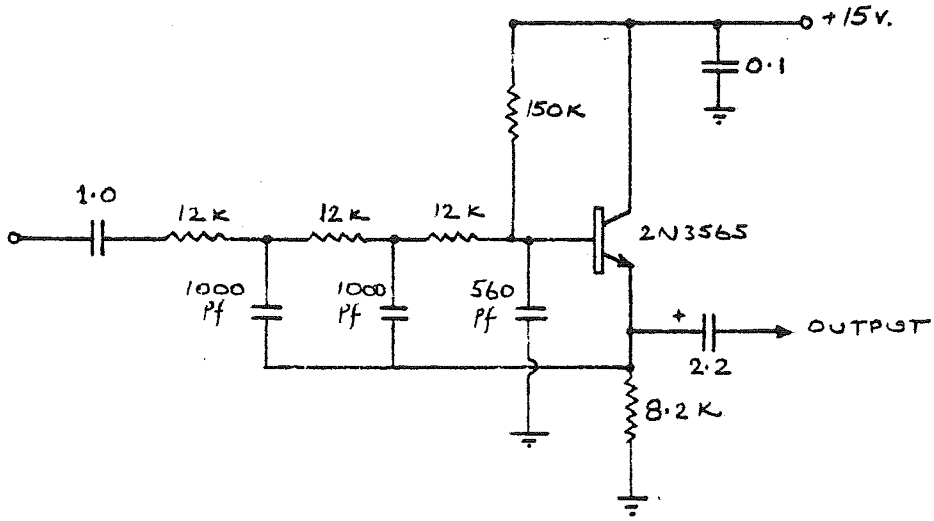


FIGURE 7.

POST DETECTION FILTER

ONE CHANNEL

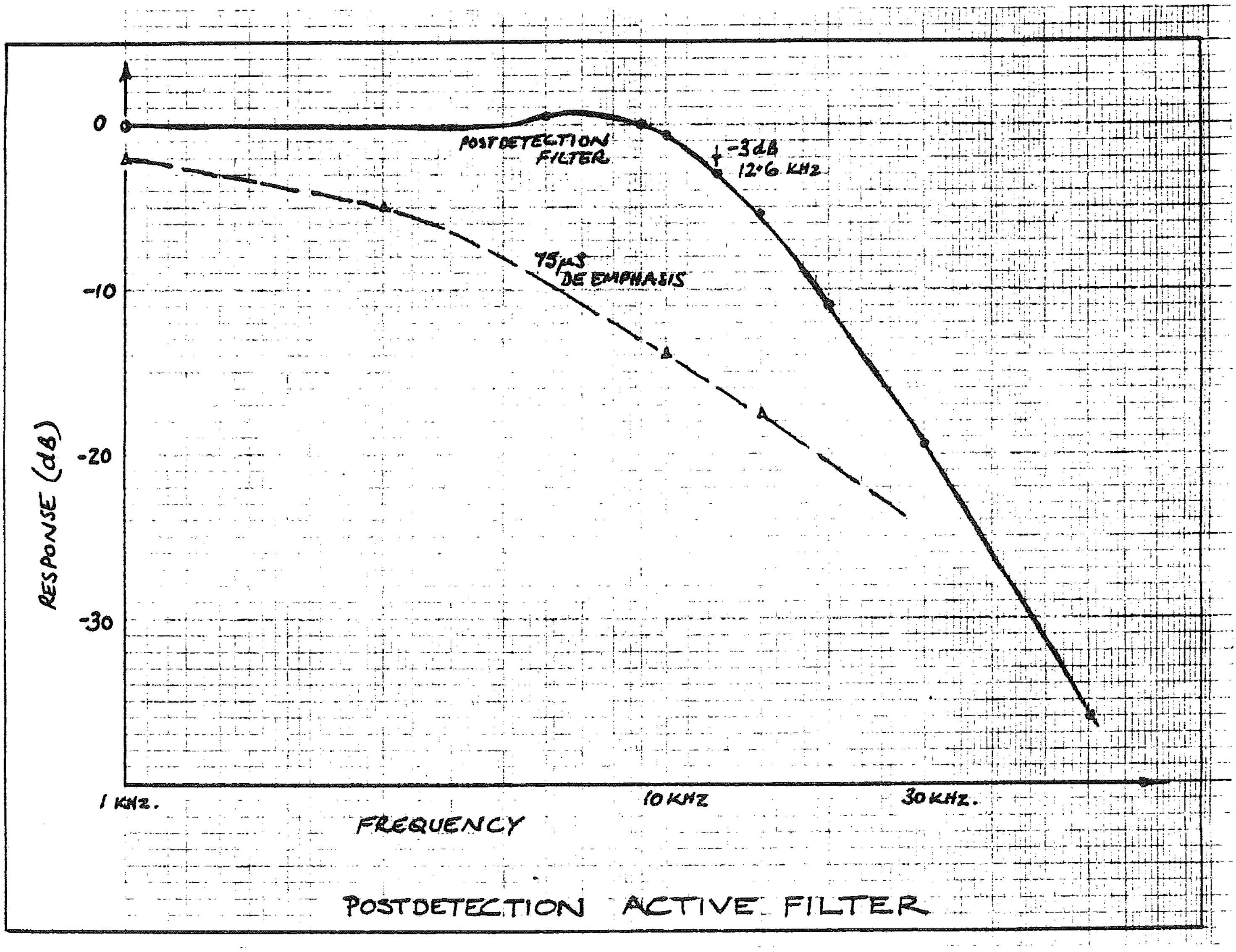


FIGURE 8.
POSTDETECTION FILTER
RESPONSE.

PHASE LOCK LOOP NOISE

Signetics type NE560B integrated phase lock loops are used. These are designed for communications systems use rather than high fidelity applications.

The output noise level is sufficiently high to make an unweighted full spectrum audio signal to noise ratio of better than 40 to 50dB unlikely to be achieved with a 10.7 MHz signal and 75 kHz peak deviaiton. This is a sufficient degradation to be noticeable when a direct comparison is made with reproduction from a good quality gramophone record.

A selection of the two quietest NE560B's from a total of 6 was made for the demonstration receiver.

The noise probably arises from the RC multivibrator type voltage controlled oscillator. An LC tuned VCO would have a much narrower noise bandwidth and would eliminate this problem. So ideally, a discrete phase lock loop, or a specially developed integrated version, should be used for this application. An alternate approach would be to use the same sort of PLL but double convert the IF to a lower frequency. If, for example, the IF was reduced to 1.07 MHz, the ratio of deviation to centre frequency is increased by 20dB and the audio signal to noise ratio would also increase by 20dB.

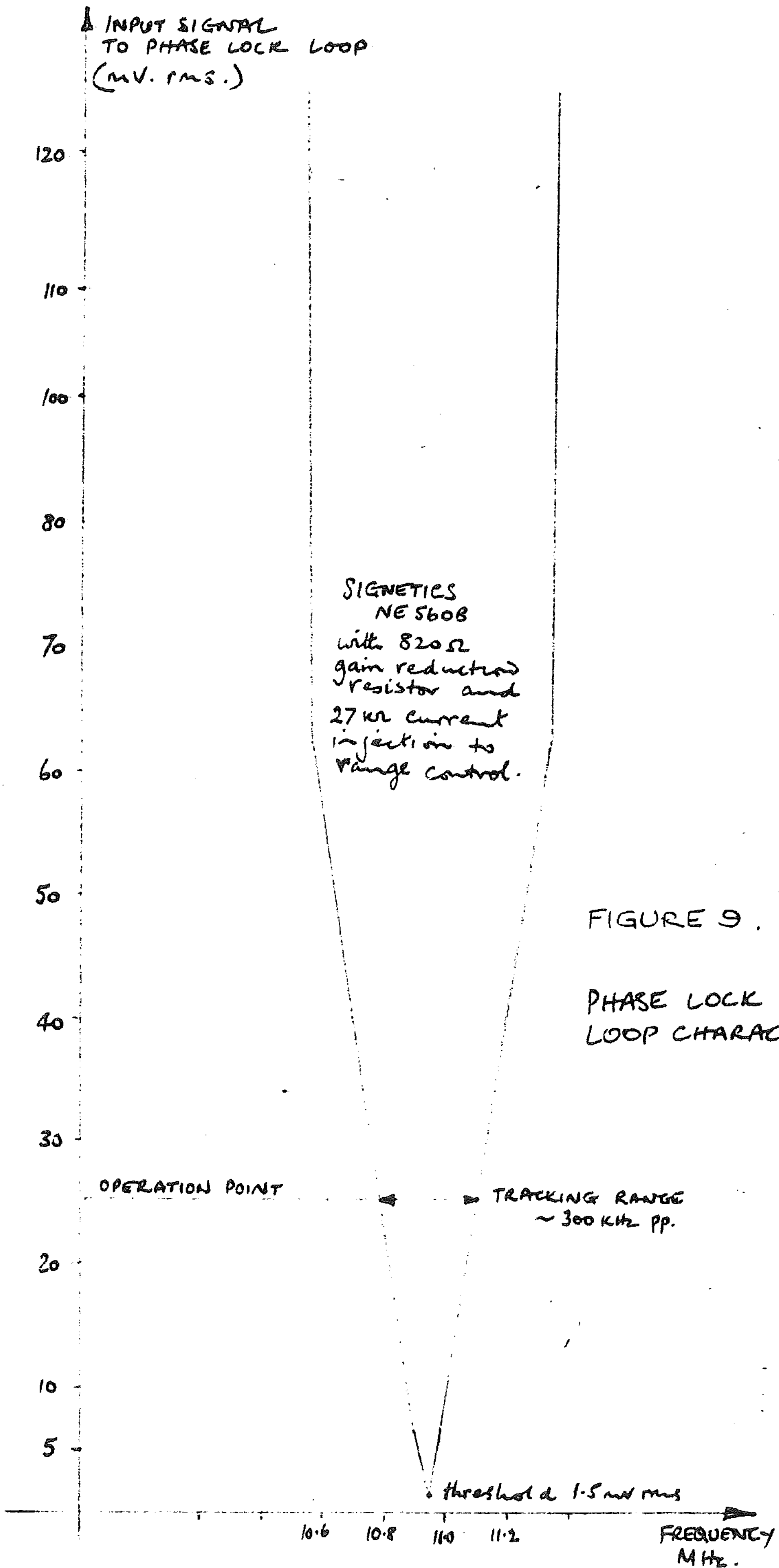


FIGURE 9.

PHASE LOCK
LOOP CHARACTERISTICS

MEASUREMENT OF RECEIVER NOISE, CROSSTALK AND DISTORTION:

With the receiver set up in the theatrette and the two signal generators acting as a transmitter on the opposite side of the room, a series of noise, crosstalk and distortion measurements were made on both channels as a function of channel spacing over the range 500 kHz to 50 kHz.

For these tests the transmitter signal generators were set to an output level of 30 mV, giving an adequate signal input of approximately 30 uV to the receiver.

The left channel signal generator was modulated with a 400 Hz sinusoidal tone, and the right with a 1300 Hz tone, both to a peak deviation of 75 kHz. An HP Noise and Distortion meter was connected to the receiver audio outputs, to each channel in turn.

The measurement procedure is as follows:

The noise and distortion meter is connected to the Right channel, and the 400 Hz tone is removed. The 1300 Hz fundamental is filtered out, and noise plus distortion is measured.

Then the 1400 Hz tone is removed, and noise is measured, both straight and via a 1000 Hz high pass filter included in the noise and distortion meter, removing the hum component of noise.

To measure crosstalk, both tones are applied, and the 400 Hz signal is nulled from the left channel signal, leaving crosstalk plus noise plus distortion.

From these measurements the separate contributions of noise, crosstalk, and distortion can be calculated.

Channel Frequencies (MHz)		Spacing
Left	Right	kHz
509.95	510.45	500
510.00	510.40	400
510.05	510.35	300
510.10	510.30	200
510.15	510.25	100
510.20	510.25	50

At the close spacings of 100 kHz and 50 kHz an interference effect is noticed with tone modulation, consisting of large amplitude noise spikes caused by phase lock loop cycle slipping on deviation peaks. This can easily be heard on tone modulation, but is not evident on music. Furthermore, because these noise spikes are narrow they do not increase RMS noise readings very much.

Results are present in graphical form in Figures 10 and 11.

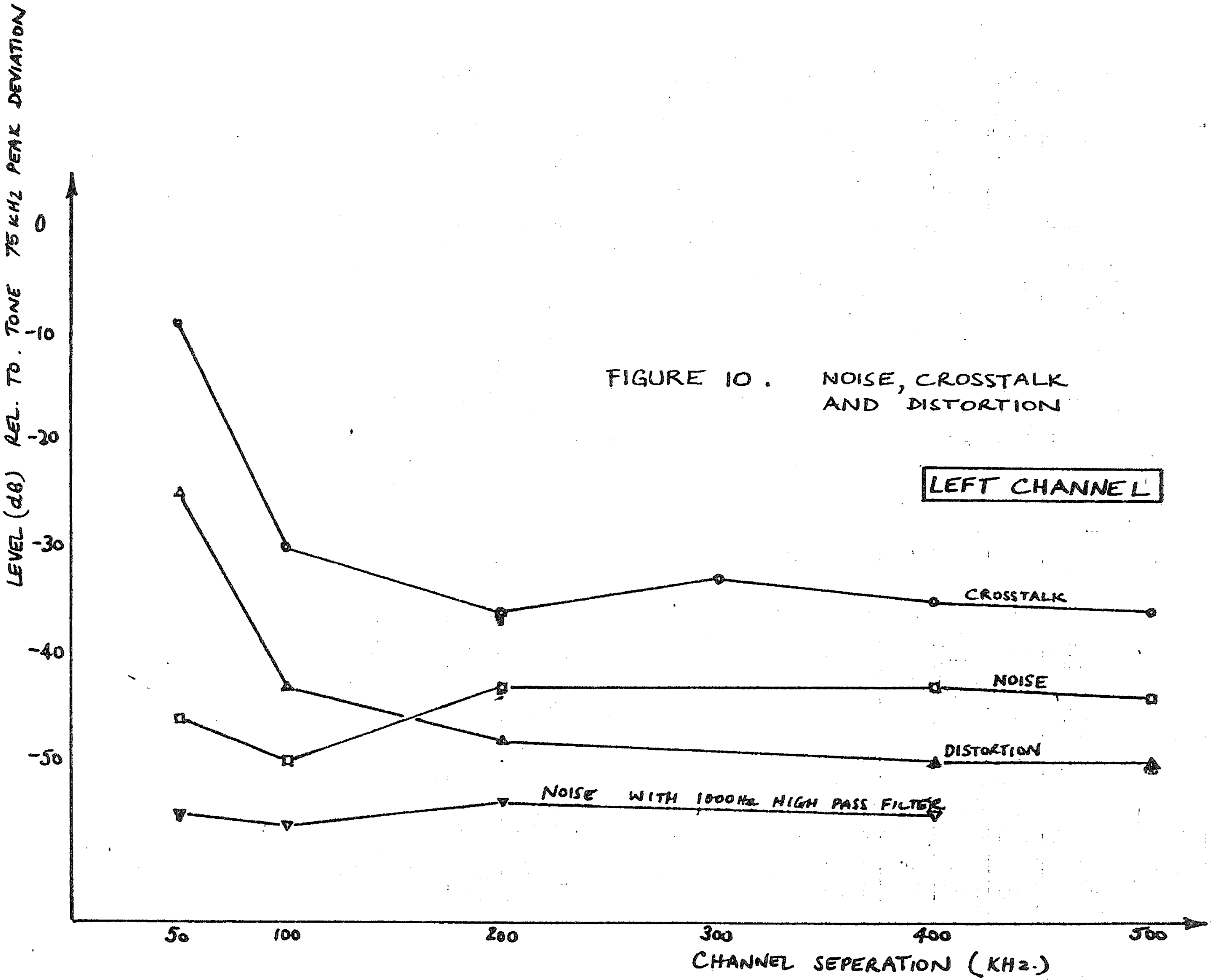


FIGURE 10. NOISE, CROSSTALK AND DISTORTION

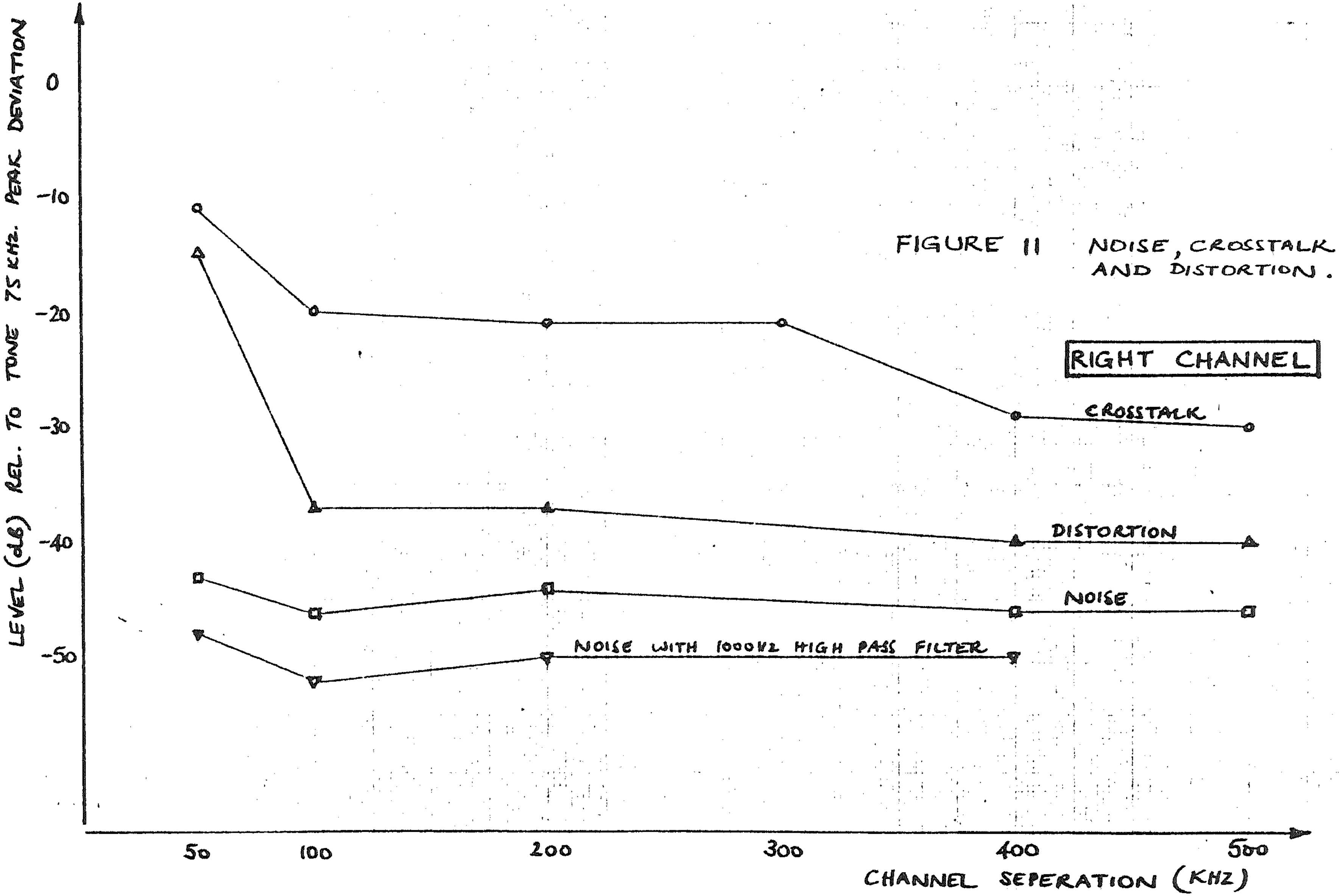


FIGURE 11 NOISE, CROSSTALK AND DISTORTION.

Receiver Demonstration

A demonstration for the Independent FM Inquiry was arranged in the ABCB theatrette. Frequency modulated carriers were transmitted across the room from a simple quarter wave vertical radiator, using two Singer model SG1000 FM signal generators as the source. These instruments, with a stability of and digital frequency read out, allowed the close spacing required between carriers to be easily set up and maintained.

A record player (Pioneer turntable model PL-41 Audio Technica cartridge, model VM 35F) provided the program source, and after pre-amplification in the front section of a Kenwood model KA-4002A amplifier and pre-emphasis (75 microsecond) the signal frequency modulated the Singer generators to a peak deviation of 75 KHz, one signal generator being used for the left channel and the other for the right.

At the opposite end of the demonstration room the experimental receiver was arranged with a quarter wave vertical antenna. A Kenwood model KA-4002A stereo amplifier and a pair of Leak loudspeakers (sandwich) completed the receiver.

A de-emphasis network was incorporated in the receiver.

A direct connection between the program source and receiver amplifier was provided, via a shielded cable and separate de-emphasis and gain adjustment network, set to give the same output level as the receiver. So by switching the amplifier input selector, program via the radio path and the receiver could be compared with program over a direct wired connection.

FIELD STRENGTHS IN DEMONSTRATION ROOM

Transmitter Sig. Gen output setting (each channel)	Field strength at Receiver Antenna (approx.)	Signal level to Receiver from Receiver antenna	Comments
22 millivolts	64 dBu	20 uV	
2.2 mV	48 dBu	2 uV	good quality music reception
0.7 mV	approx. 43 dBu noise significant part of F.I. set indication.	less than 1 uV	marginal reception

RECEIVER PERFORMANCE AT LOW SIGNAL LEVELS:

With one signal generator (simulating one carrier) feeding directly to the receiver input, and with tone modulation:

Receiver Input	Total noise and distortion
1 uV rms	5.6%
1.5 uV rms	2.8%

For these low signal level tests the IF amplifier was below the limiting threshold, and consequently feeding a reduced signal to the phase lock loops, resulting in reduced tracking range. To stay within this reduced tracking range, peak deviation was reduced to 40 kHz for these tests.

ACKNOWLEDGEMENTS

This project was instigated and supervised by Mr. J.M. Dixon. Much of the receiver design and construction, particularly the RF front end, was undertaken by Mr. A. Slamin.