DIRECT SEQUENCE SPREAD SPECTRUM TECHNIQUES FOR LAND MOBILE RADIO APPLICATIONS

by

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ABSTRACT

This thesis describes an investigation into the application and performance of direct sequence spread spectrum techniques for land mobile radio systems.

There is a brief description of the basic principles of operation of direct sequence systems.

The multiple user facility is analysed and values obtained for the maximum number of simultaneous system users in terms of system parameters. This clearly illustrates the need for power control. A possible method of providing power control is described. Comparison of user density is made against conventional narrowband modulation methods. There is some discussion of the effect of sequence cross-correlations on the number of system users. The system organisation is mentioned, showing possible application of a calling channel.

Consideration is given to the possibilities of bandsharing with narrowband modulation systems. Figures are derived for the resulting interference to existing systems which would be caused by such an arrangement.

A brief resume of the pertinent features of the land mobile radio channel is given. The effects of shadowing on the output quality and spectral efficiency of direct sequence systems is discussed. There is an analysis of the effects of shadowing on the user density in small cell schemes. An analysis shows the effects of multipath propagation on direct sequence performance by reference to a simple two path channel.

Details are given of a simple experimental direct sequence spread spectrum transmitter and receiver constructed. The measured results for the performance of the system against various forms of interference and channel degredation are compared with their theoretical values.

Finally ideas for future work are discussed.

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SYMBOLS and ABBREVIATIONS

a	Protection ratio	
B_D	Drift rate	
B_m	Modulated message bandwidth	
B_n	Narrowband receiver noise bandwidth	
B_{rf}	Direct sequence mainlobe bandwidth	
$c(\cdot)$	Spreading sequence	(time domain)
$C(\cdot)$	Spreading sequence	(frequency domain)
$\underline{C}(\cdot)$	Spreading sequence power spectral density	
d	Time delay	
d.s	Direct sequence	
D	Minimum propagation delay	
E	Propagation path gain	
f	Frequency	
f_o	Direct sequence centre frequency	
f_n	Narrowband system centre frequency	
F	Number of frequency bands	
G_p	Process gain	
$h(\cdot)$	Channel impulse response	
$H(\cdot)$	Channel transfer function	
$H_n(\cdot)$	Narrowband receiver transfer function	
$H_s(\cdot)$	Direct sequence receiver transfer function	
i	Spreading sequence length	
i.f	Intermediate frequency	
I	Interference power	
J	Speech modulation figure of merit	
K	Propagation constant	
$K_c K_1$	Sequence cross correlation factor	

L	Implementation loss	
m (·)	Message	(time domain)
M	Number of simultaneous users	
n(·)	Noise	(time domain)
N	Noise Power	
N (·)	Noise	(frequency domain)
N_o	Noise power spectral density	
p.d.m	Pulse duration modulation	
p.r.k	Phase reversal keying	
P_I	Received interference power	
P_{IT}	Total received interference power	
P_n	Narrowband power	
$\underline{P}_n(\cdot)$	Narrowband signal power spectral density	
P_{ns}	Narrowband power accepted by d.s receiver	
P_s	Direct sequence power	
P_{sn}	D.S power accepted by narrowband receiver	
P_u	Probability of unsatisfactory reception	
P_w	Received wanted signal power	
Q	$(i + 1)/_i$	
r	Distance between transmitters	
r_i	Inner radius	
r_o	Outer radius	
r_w	Distance of wanted mobile from cell centre	
$R_c(\cdot)$	Spreading sequence auto-correlation function	
s(·)	Signal	(time domain)
S	Signal Power	
S(·)	Signal	(frequency domain)
$\underline{S}(\cdot)$	Signal power spectral density	

s.c.p.d.m	Suppressed clock pulse duration modulation
s.i.k	Sequence inversion keying
s.i.r	Signal to interference ratio
$(s.i.r)_i$	s.i.r at input to d.s receiver
$(s.i.r)_o$	s.i.r at input to d.s message demodulator
$(S/I)_i$	$(s.i.r)_i$
$(S/I)_o$	$(s.i.r)_o$
s.n.r	Signal to noise ratio
$(s.n.r)_i$	s.n.r at input to d.s receiver
$(s.n.r)_o$	s.n.r at input to d.s message demodulator
$(S/N)_i$	$(s.n.r)_i$
$(S/N)_o$	$(s.n.r)_o$
$(S/N)_{omin}$	Minimum acceptable $(s.n.r)_o$
t	Time
T	Spreading sequence chip period
T_D	Drift time
T_m	Message bit period
$u(\cdot)$	Mobile transmitter spatial density
U	Mobile transmitter spatial density (constant)
V	Ratio of interference range to service range
x	Signal component
y	Signal component
Z	Resultant signal amplitude
β	Multiple user protection ratio
Δ	Excess path delay
σ	Standard deviation of log-normal shadowing

All other symbols and abbreviations have their usual meanings.

CHAPTER 1

Introduction

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Introduction

A difficult but valuable application for radio communication techniques is land mobile radio systems. Here the channel introduces many degredations making for difficulties in operating a high performance system. Other problems arise from the need to cater for many users and operate the system efficiently. Present day land mobile radio systems use conventional narrowband modulation methods operating in narrow bandwidth channels in an attempt to maximise spectral occupancy. There is little attempt to overcome the channel degredations or optimise the overall system performance. With increasing demands on the frequency spectrum and a desire for better system performance new ideas are required. It is therefore appropriate to consider the use of spread spectrum techniques which were originally developed for military applications.

It is difficult to find a definition of spread spectrum techniques, though one might run as follows: "A communications technique for which the transmitted bandwidth is far in excess of that of the information signal and independent of it, being determined by an auxiliary spreading signal". The most useful feature of systems using this technique is the ability to reduce the effects of a wide range of interfering signals. Hence in many circumstances the use of spread spectrum techniques can permit reliable communications where conventional narrowband techniques would be unusable.

Several implementations of spread spectrum techniques have been devised of which frequency hopping and direct sequence systems are the main types. A considerable amount of theoretical work has been carried out on frequency hopping systems for land mobile radio,[1] though no practical implementations have evolved. However little has been mentioned about direct sequence type systems for this application. These are in certain respects easier to implement than the frequency hopping type and should have a similar performance. Hence the reasons for investigating direct sequence techniques for possible application to land mobile radio systems.

An important area of investigation is the operation of many base station/mobile links together in a system. Using direct sequence techniques all links in a system operate on the same carrier frequency and the number of simultaneous system users is limited by self interference. Figures are therefore derived for the maximum number of simultaneous users of the system under various conditions. There is also a comparison of spectral efficiencies with conventional narrowband modulation methods.

In view of the wide bandwidths required it is unlikely that exclusive bands for spread spectrum operation could be found. Hence if such systems are to be used they will have to bandshare with existing narrowband systems. There is therefore an analysis of the compatability of direct sequence and narrowband modulation techniques.

If direct sequence spread spectrum techniques are to be useful in land mobile radio applications they must perform well under adverse propagation conditions. The main propagation degredations encountered in the land mobile channel are shadowing and multipath propagation. There is therefore an analysis of the effect of these degredations on direct sequence systems. This considers both the effects on the performance of a single link and also the effects on a complete system, especially on the spectral efficiency.

Finally details of an elementary direct sequence transmitter and receiver, which were constructed, are given. These were built to allow measurements to be made on the performance of a link under various conditions and the results compared with the theoretical ones.

In the analysis of system operation and performance it is convenient to study each topic in isolation as this simplifies the situation and aids understanding of the problem. However it will be realised that the topics are interrelated to some extent and to provide a complete analysis, all factors should be included. Hence the analysis sections of this thesis should not be regarded as isolated entities, rather as convenient interrelated sub-divisions.

The object of this work is essentially a systems study to investigate the possible application of direct sequence spread spectrum techniques to land mobile radio systems. Consequently the detailed aspects of implementation are not investigated in great depth as this is considered outside the scope of investigation. In particular little attention is given to the problems of spreading sequence synchronisation, as this topic is considered of such vital importance as to warrant a thorough study in its own right.

CHAPTER 2

Spread Spectrum Principles

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The main objective of any communications system is the faithful transfer of information from source to destination in an efficient manner. In attempting to achieve this the system has to contend with degredations introduced by the channel in the form of distortion and interference. Hence the receiver has the difficult task of recovering faithfully the original message from a mutilated version of the transmitted signal. Fortunately it can be aided in this task by a suitable choice of transmitted signal and the provision of apriori information about the intended signal. Therefore the transmitter serves to convert the message to a suitable form for transmission. The receiver uses its knowledge of the intended signal to operate on the received signal and recover the message. A similar philosophy applies to other elegant communications techniques, except that these do not have anywhere as large a ratio of transmitted to information bandwidths as do spread spectrum techniques.

2.1 Operation

The basic principles of spread spectrum operation have been described elsewhere.[2,3,4,6] Nevertheless a description of the direct sequence system is in order.

Fig 2.1 shows an elementary direct sequence system. At the transmitter a double modulation scheme is used to produce a wideband transmitted signal. The information signal is modulated onto a carrier using a standard modulation technique. This is further modulated by a wideband spreading signal to produce the transmitted spread spectrum signal.

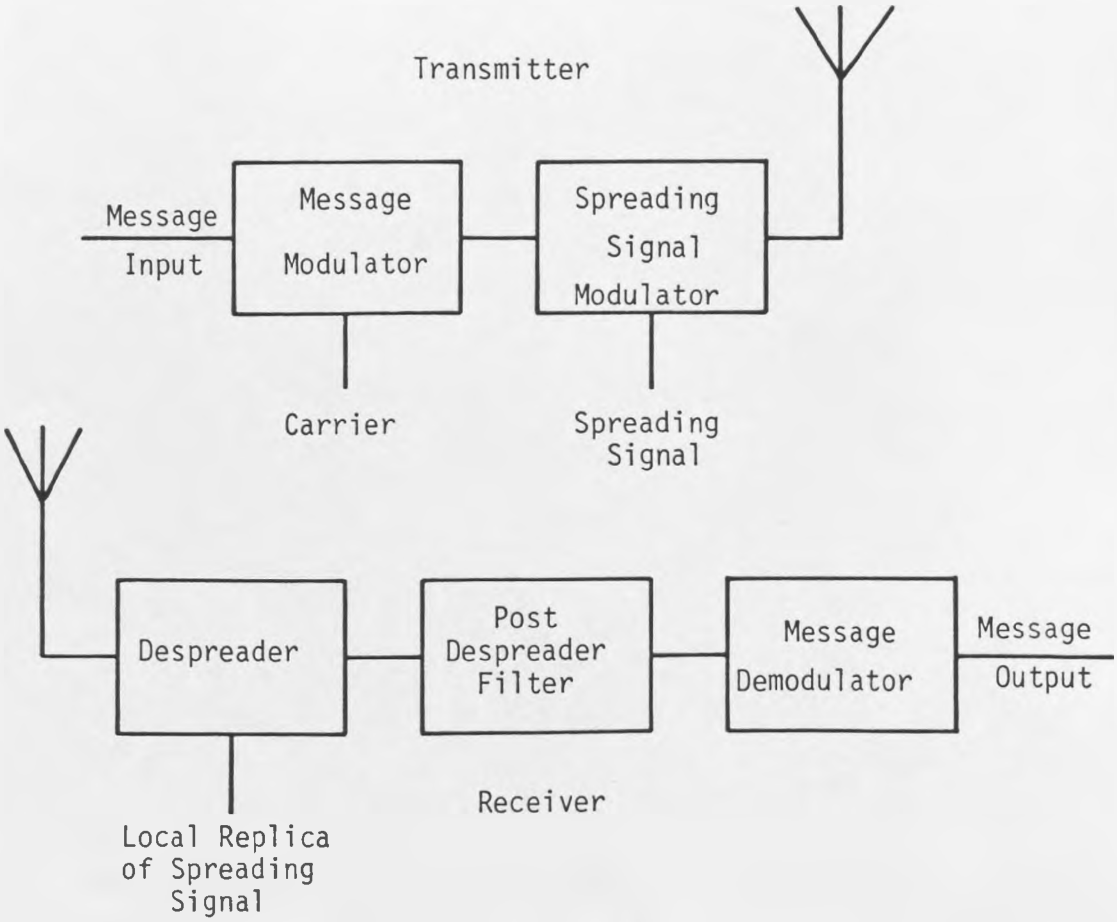

Fig. 2.1 Block diagram of elementary
Direct Sequence Spread Spectrum System

At the receiver the incoming signal is multiplied with a local replica of the spreading signal used at the transmitter. Providing the latter signal is correctly synchronised the incoming signal is despread to its original narrowband form. This can be filtered out and demodulated to recover the message.

Consider the reception of interference of fixed power which is uncorrelated with the spreading signal. In the despreading process this is multiplied with the spreading signal and consequently spread in frequency. By judicious choice of spreading signal the spread interference can be given noise like properties. Hence much of the interference energy will be rejected by the filter placed ahead of the message demodulator, that getting through appearing as noise. The system can therefore show an improvement in signal to interference ratio between input and output.

2.2. Process Gain

The performance of a direct sequence spread spectrum system is given by the process gain. This is a measure of the improvement in signal to interference ratio between input and output and is defined as :

$$G_p = \frac{(s.i.r)_o}{(s.i.r)_i} \qquad (2.1)$$

Here $(s.i.r)_i$ is the signal to interference ratio at the receiver input and $(s.i.r)_o$ is that at the input to the message demodulator. Several authors [3,5] show that for uncorrelated interference the process gain is theoretically given by:

$$G_p = \frac{B_{rf}}{B_m} \qquad (2.2)$$

Where B_{rf} is the spread bandwidth and B_m the bandwidth of the filter following the despreader. This value is exact for certain types of interference, though only approximate for others. In practice the process gain attainable will be less than that given by (2.2) due to various effects in the system. This is accounted for by including an extra term in (2.2), the implementation loss L. Thus the process gain becomes :

$$G_p = \frac{B_{rf}}{B_m L} \tag{2.3}$$

The definition of process gain only relates to the spread spectrum part of the system. The message signal to noise ratio is obtained from consideration of the signal to noise ratio at the input to the message demodulator.

Implementation loss arises from several causes, principally sequence tracking errors and the effects of channel filtering. At the receiver under conditions of low signal to noise ratio the local sequence generator is not perfectly synchronised to the incoming signal. This is due to the noise perturbing the sequence tracking loop, giving rise to a sequence tracking error. Hence the wanted signal output from the despreader is decreased, whilst extra noise is introduced resulting in a decreased signal to noise ratio at the message demodulator. The topic is covered by de Couvreur[23] who evaluates the signal and noise powers as a function of the variance of the assumed Gaussian jitter.

Thus the contribution to the implementation loss caused by sequence tracking errors is a variable quantity dependant on the signal to noise ratio at the receiver. The loss caused may be a small fraction of a dB at high input signal to noise ratios increasing to several dB at low signal to noise ratio's.

The system filtering contributes to the implementation loss as the following argument will explain. Due to the filters in the transmission path at transmitter and receiver normally only the main lobe of the transmitted signal reaches the despreader. The signal may also suffer amplitude and phase distortion in the filters. Consequently the incoming wanted signal is not an exact replica of the receiver local spreading sequence which is normally not filtered. Thus in the despreading process there is a slight reduction in the wanted signal level, generally referred to as correlation loss for true correlation receivers. Some loss of energy also occurs due to the filters restricting the signal spectrum. The contribution to the implementation loss caused by the system filtering is a fixed quantity which by careful design can be kept to less than 1dB. In system design a reasonable value to allow for implementation loss is 3dB.

Throughout the remainder of this thesis the effect of the implementation loss will for convenience be neglected. However this is not to deny its importance and in the overall analysis it should be included in considering the system performance.

It is necessary to point out at this stage, that spread spectrum systems are only useful against finite power interference. Consider a simple direct sequence system operating over a channel where the only interference is white noise having a single sided power spectral density of N_o, assuming the noise bandwidth of the receiver to equal the spread bandwidth B_{rf} then the signal to noise ratio $(s.n.r)_i$ at the receiver input is :

$$(s.n.r)_i = \frac{P_s}{N_o B_{rf}} \qquad (2.4)$$

where P_s is the received wanted signal power. This signal to noise ratio is improved by the system process gain to produce a signal to noise ratio $(s.n.r)_o$ at the message demodulator, where:

$$(s.n.r)_o = (s.n.r)_i \cdot G_p \qquad (2.5)$$

$$= \frac{P_s}{N_o B_{rf}} \cdot \frac{B_{rf}}{B_m}$$

$$= \frac{P_s}{N_o B_m} \qquad (2.6)$$

This is the same signal to noise ratio which could have been produced by the system without the spread spectrum section. Hence there is no overall improvement in performance over narrowband modulation methods.

2.3 Alternative Implementations

The system shown in Fig 2.1 is an elementary implementation of a direct sequence transmitter and receiver. There are a number of variations to this, particularly in the despreading circuitry at the receiver. Regardless of implementation the overall operation and performance remains virtually the same as for the system mentioned.

Instead of despreading using a mixer and filter some systems use a correlator comprising a mixer and integrator. This is operated on an integrate and dump basis. The output builds up to a peak which is sampled and sent for further processing. The contents of the integrator are dumped in preparation for the next information bit. Generally the incoming signal is converted to baseband before entering the correlator.

An alternative to the correlator is the matched filter, again often operating at baseband. The spread spectrum signal enters the matched filter where it builds up to a peak. This is sampled and further processed. Often a similar matched filter is used at the transmitter for sequence generation and information modulation combined. Matched filter systems are arranged so that the data bit period equals the spreading sequence period or some large fixed fraction of it. Practical constraints[7] on the matched filters tend to limit them to short spreading sequence lengths. Typically[41] s.a.w. type matched filters are limited to sequence lengths of less than 1000 chips. Both the matched filter and correlation type receivers are only suitable for fixed length chip binary signals.

2.4 Spreading Signal

The spreading signal is usually a binary sequence such as a Gold code or maximal length sequence. This is used to phase shift key modulate the carrier. Whilst any type of message modulation may be used, practical systems to date have used angle modulation. Digital data is generally combined with the spreading sequence prior to carrier modulation. This provides a simplification in implementation at the transmitter, being equivalent to phase shift key modulation of the carrier by both information and spreading signal. The simplicity of the system often results in analogue signals being converted to a digital format for transmission. Regardless of source, digital information is generally reclocked to make its transitions coincident with those of the spreading sequence. This ensures that the shape of the transmitted spectrum is independent of the information signal.

For convenience the thesis will concentrate on maximal length type sequences as these are well documented[31,43]. However most of the comments will apply to systems using other spreading sequences. It should be noted that the concentration on maximal length sequences does not imply that other sequences are unsuitable, in fact they may prove superior to the maximal length type.

2.4.1 Transmitted Spectrum

The transmitted spectrum can be determined by convolving the spectrum of the spreading sequence and information signal. Obviously the spreading signal has the widest spectrum. Hence providing the signals are unrelated the transmitted spectrum can be approximated by that of the spreading sequence. Perhaps of greater interest is the power spectral density, which can be obtained by application of the Wiener-Knichine theorem to the sequence auto-correlation function (a.c.f).

Now for a bipolar (± 1) maximal length sequence of length i clocked at $^1/T$ chips/s the normalised periodic a.c.f. is:

$$Rc(\tau) = \sum_{j=-\infty}^{\infty} \frac{(i+1)}{i} \Lambda\left(\frac{\tau - jiT}{T}\right) - \frac{1}{i} \qquad (2.7)$$

where

$$\Lambda\left(\frac{\tau}{T}\right) = \begin{cases} \left(1 - \frac{|\tau|}{T}\right) & -T \leq \tau \leq T \\ 0 & \text{elsewhere} \end{cases}$$

Hence the one sided power spectral density (p.s.d) for such a sequence modulated onto a carrier at f_0 Hz is :

$$\underline{S}(f) = \sum_{\substack{j=-\infty \\ j\neq o}}^{\infty} \frac{(i+1)}{i^2} \left[\frac{\sin \Pi\, T(f-f_0)}{\Pi\, T(f-f_0)} \right]^2 \delta\left(f-f_0-\frac{j}{iT} \right) + \frac{1}{i^2}\,\delta(f-f_0) \tag{2.8}$$

This is a line spectrum with a line spacing of the reciprocal of the spreading sequence period ($^1/iT$). It has a $(\sin x/x)^2$ type envelope with nulls at the reciprocal of the chip period ($^1/T$). As the sequence length i increases the lines come closer together and the p.s.d becomes a continuous function of the form:

$$\underline{S}\,(f) = T \left(\frac{\sin \Pi\, T(f-f_0)}{\Pi\, T(f-f_0)} \right) \qquad (i \rightarrow \infty) \tag{2.9}$$

Modulation by data will not alter the envelope, though the spectrum will tend more to a continuous form.

Generally the transmitted signal is filtered to prevent out of band interference. Typically the main lobe only is transmitted, giving a spread bandwidth of :

$$B_{rf} = \frac{2}{T} \tag{2.10}$$

2.5 Analogue Message Modulation

It was stated earlier that analogue signals are often converted to a digital format prior to transmission over a spread spectrum link. Whilst there are many methods of conversion to a digital format not all are suitable for this purpose. Generally for land mobile radio systems the main analogue signal to be transmitted is speech, therefore we shall concentrate on this.

The main requirement is to have an efficient method of speech conversion/modulation to optimise the spread spectrum operation. To maximise the process gain the bandwidth occupied by the speech conversion scheme should be small. Failure to ensure this causes the process gain to be small with a consequent reduction in overall system performance.

A closely related requirement is that the conversion scheme should not require a large input signal to noise ratio to produce an acceptable audio signal to noise ratio. This means that if possible the conversion scheme should exhibit a 'modulation improvement'. If this is not achieved then much of the advantage gained by the use of spread spectrum techniques will be wasted. A trade off is possible between'modulation improvement' and bandwidth without any loss in performance.

Finally the conversion scheme should have no thresholds or at least any threshold should be below the limits of intelligibility. Other criteria are involved in the selection of a speech digital conversion scheme but are not relevant here. Also whilst all the comments here are directed to methods of converting speech to a digital format, they apply equally as well to the complete process of message modulation.

Having outlined the desirable properties of speech converstion schemes it is useful to discuss the features of available systems. The main interests are in the i.f. bandwidth occupied by the converted signal and the ratio of audio to i.f. signal to noise ratios. As these can be 'traded off' without loss of performance a convenient method of comparing the available speech conversion schemes is by a figure of merit J. This is defined as:-

$$J = \frac{\text{Audio s.n.r./i.f. s.n.r.}}{\text{i.f. Bandwidth/Audio Bandwidth}}$$

Obviously the higher the figure of merit the more suitable the speech conversion scheme is for spread spectrum application.

Whilst it would be desirable to incorporate thresholds in this evaluation scheme it is difficult to give measure to their effect.

Table 2.1 shows figures at merit evaluated for a range of speech conversion schemes, the information being derived from the relevent sections of Carlson[40]. It should be noted that there is disagreement between authors[32,42] on the performance of pulse position and pulse duration modulation schemes as much depends upon the assumptions made. Fig. 3.5 at the end of chapter 3 presents the same information as table 2.1 though in a graphical form. Most of the speech conversion schemes mentioned here would benefit from speech processing technques such as companding, though no account has been taken of these in deriving the results.

TABLE 2.1

FIGURES OF MERIT FOR SPEECH CONVERSION SCHEMES

	Bandwidth Ratio	Modulation Improvement	Threshold s.n.r. dB	Figure of Merit	Notes
a) Pulse Code Modulation at Threshold					
No. of levels					
8 $(=2^3)$	12	6.82	8.47	.568	1,2,3,4
16 $(=2^4)$	16	20.09	9.78	1.255	
32 $(=2^5)$	20	62.66	10.83	3.13	
b) Pulse Code Modulation at 10dB above Threshold					
8 $(=2^3)$	12	1.36	8.47	.113	1,2,3,4
16 $(=2^4)$	16	4.02	9.78	.251	
32 $(=2^5)$	20	12.50	10.83	.625	
c) Pulse Position Modulation					
	4	.50	3	.125	1,2,5
	8	4.0	3	.50	
	12	13.5	3	1.125	
	16	32.0	3	2.0	
d) Pulse Duration Modulation					
	4	1	3	.25	1,2,5
	8	4	3	.50	
	12	9	3	.75	
	16	16	3	1	

NOTES:

1. Phase Reversal keying modulation of carrier assumed. Hence s.n.r. at output of carrier demodulator assumed to equal s.n.r. at input.
2. Sampling rate = 2 x Message Bandwidth
3. I.F. Bandwidth = 4 x Message Bandwidth x No. of bits/sample
4. Threshold s.n.r. = Carrier s.n.r. such that decoding errors equal quantisation noise.
5. I.F. Bandwidth = 4 x Message Bandwidth x Some Integer.

CHAPTER 3

Multiple User Facility

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Multiple User Facility

So far the discussion of spread spectrum systems has centred on a single link. From a system viewpoint it is necessary to simultaneously operate many links independently in the same geographical area. Consideration will now be given as to how this is achieved.

Obviously the use of separate channels, by assigning to each link a separate carrier frequency is impracticable because of the bandwidth required. Instead all links operate on the same carrier frequency and separation is achievedas follows. To each link is assigned a unique spreading sequence, from the set of available sequences, unrelated to the sequences used by other links. Thus at any receiver along with the wanted signal will be received signals from unwanted transmitters. The unwanted signals are treated as any other interference and their effect is reduced by the system process gain.

3.1 Multiple User Analysis

It is apparent that the output signal to noise ratio at any receiver is determined by the interference from other users. As each receiver requires a minimum signal to noise ratio for acceptable operation there is a limit to the number of simultaneous users of the system.

To evaluate this limit consider the situation shown in Fig. 3.1. Here M equi-power signals are present at the input to a direct sequence receiver along with noise of power N.

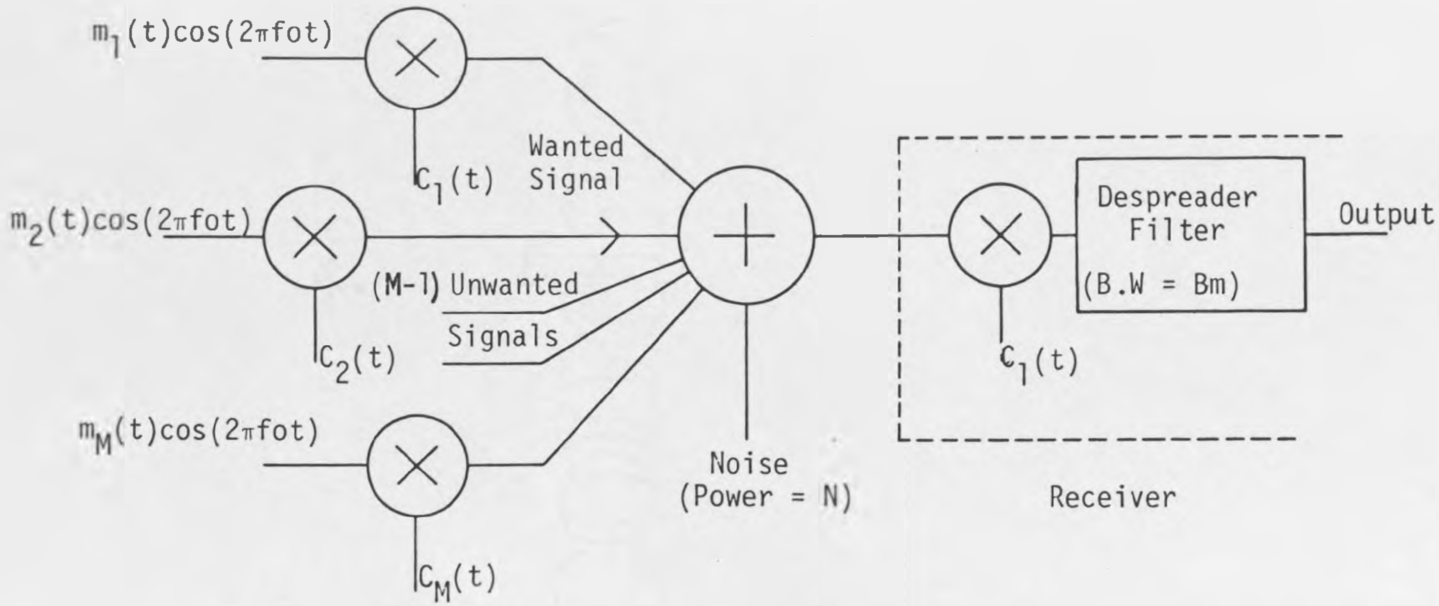


Fig. 3.1 Diagram showing Multiple User Facility

If one of the signals is the wanted one then the signal to interference ratio at the receiver input is :

$$\left(\frac{S}{I}\right)_i = \frac{S}{(M-1)\,S + N} \tag{3.1}$$

Assuming that the interference is reduced by the system process gain, then the signal to noise ratio at the message demodulator input is :

$$\left(\frac{S}{N}\right)_o = \frac{S\;G_p}{(M-1)S + N} \tag{3.2}$$

This can be re-arranged to give:

$$M = \frac{G_p}{(^S/N)_o} - \frac{N}{S} + 1 \tag{3.3}$$

Equation (3.3) shows that to obtain the largest number of simultaneous users in a given band the ratio of wanted signal to extraneous noise should be high. Hence if the only interference is that due to other users (3.3) can be written as :

$$M = \frac{G_p}{(^S/N)_o} + 1 \tag{3.4}$$

Or substituting for the process gain :

$$G_p = \frac{B_{rf}}{B_m}$$

we obtain

$$M = \frac{B_{rf}}{(^S/N)_o\;B_m} + 1 \tag{3.5}$$

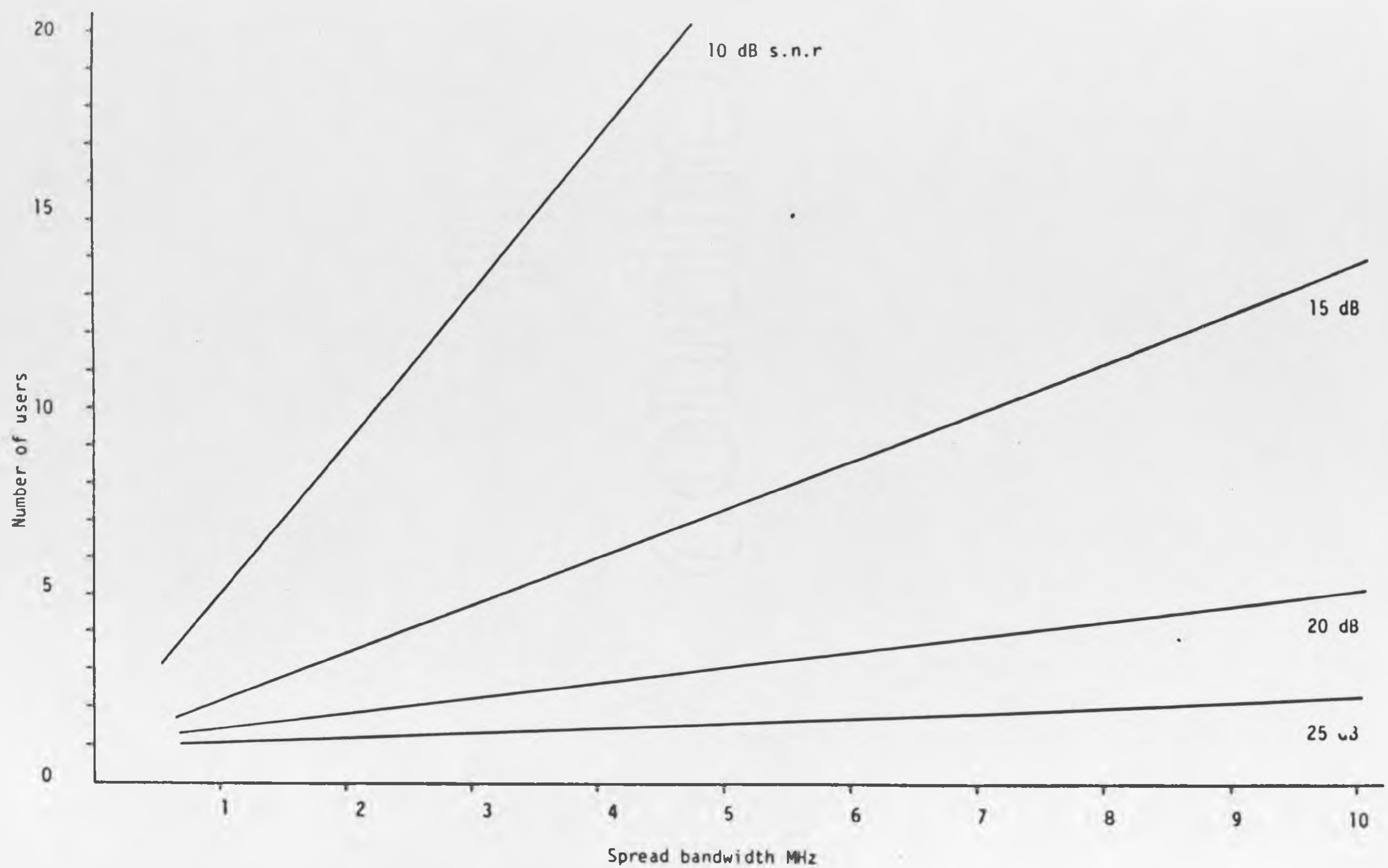


Fig 3.2 NUMBER OF USERS AGAINST SPREAD BANDWIDTH
Information bandwidth 25 kHz

Fig. 3 2 shows equation (3.5) plotted out, where the maximum number of users is evaluated against spread bandwidth for various signal to noise ratios at the input to the message demodulator.

As mentioned elsewhere the audio signal to noise ratio will generally be different to that at the input to the message demodulator. Examination of fig 3.2 shows that to obtain the maximum number of users in a given system the minimum required post-despreader signal to noise ratio must be small. This implies the use of an efficient message modulation method.

3.2 Graceful Degradation

A valuable feature of spread spectrum multiple user systems is that of graceful degradation. This allows the system to tolerate temporarily slight increases in the number of users above the limits evaluated without system collapse. To achieve this a message modulation technique having no sharp thresholds is required. Having satisfied this requirement a large system with many users can tolerate slight overloads with only small degradations in output signal to noise ratio.

Consider a large system using power control where the only interference at the message demodulator is caused by other users. Thus the signal to noise ratio at the message demodulator with M users operational is :

$$\left(\frac{S}{N}\right)_o = \frac{G_p}{(M-1)} \qquad (3.6)$$

If now the number of users increases by 10% this value becomes :

$$\left(\frac{S}{N}\right)_o' = \frac{G_p}{(M-1)\,(1+0.1)} \tag{3.7}$$

now $$\left(\frac{S}{N}\right)_o' \bigg/ \left(\frac{S}{N}\right)_o = \frac{1}{(1+0.1)} \tag{3.8}$$

Thus the signal to noise ratio is decreased from its original value by 0.4 dB. Providing the ratio of output to input signal to noise for the message demodulator is linear at this point such change is unlikely to be noticed.

Under conditions where the system is highly loaded priority users may find that communications are not reliable enough for their purposes. To overcome this such users may be allowed an increase in transmitter power or transmitted bandwidth. Either of these will provide an increase in signal to noise ratio, thus allowing more reliable communications. The use of increased transmitter power results in a decrease in signal to noise ratio for other users of the system. Hence use of this facility would have to be restricted. The use of increased bandwidth, providing an increase in process gain, carries no such penalty. However to obtain a 3dB improvement a doubling of the occupied bandwidth would be required.

3.3 Qualification of Multiple User Facility

At the beginning of this section it was assumed that the interference effect of unwanted spread spectrum users was reduced by the system process gain. Whilst this is a reasonable assumption it is nevertheless desirable to justify it.

A paper by Judge[9] considers multiplexing using maximal length sequences and correlation receivers. For M equal power received signals in the absence of noise this paper gives the output signal to noise ratio at any receiver as :

$$\left(\frac{S}{N}\right)_o = \frac{1}{(M-1)\left(k_1^2 + \frac{T}{T_m}\right)} \qquad (3.9)$$

The term k_1^2 accounts for the sequence cross-correlations and the ratio T/T_m is the ratio of chip to data bit period. This equation can be re-arranged as :

$$M = \frac{T_m}{T} \quad \frac{1}{(S/N)_o\left(1 + k_1^2 \frac{T_m}{T}\right)} + 1 \qquad (3.10)$$

Now to a reasonable approximation :

$$\frac{T_m}{T} = \frac{B_{rf}}{B_m} = G_p$$

Hence

$$M = \frac{G_p}{(S/N)_o} \quad \frac{1}{(1 + k_1^2 \; G_p)} + 1 \qquad (3.11)$$

Comparison with equation (3.4) shows that (3.11) is identical except for an extra factor. This extra factor of :

$$\frac{1}{1 + k_1^2 \; G_p}$$

reduces to unity for

$$k_1^2 \; G_p \ll 1$$

Judge evaluates the cross correlation factor k_1 for various maximal length sequences and indeed for moderate values of process gain

$$k_1^2 \ G_p << 1$$

3.3.1 Spreading Sequence Cross-Correlations

In general for the situation where all transmitters are received at equal levels the maximum number of users will be given by :

$$M = \frac{G_p}{(S/N)_o} \ k_c + 1 \qquad (3.12)$$

The factor k_c accounts for various effects, mainly though for the spreading sequence cross-correlations. For most situations k_c will be close to unity.

The significance of sequence cross-correlations in active despreader type direct sequence systems can be deduced from a qualitative argument. Consider such a system operating well above noise and having a single direct sequence interferer present at the input to the receiver. If the cross-correlation between locally generated and interfering sequences is low, then the signal at the despreader output will resemble a spreading sequence. It will therefore have a smooth spectrum with no concentrations of energy centred on the filter passband. Hence the interference energy will be reduced by the system process gain. If the cross-correlation between the sequences is high then the despreader output will not resemble a sequence. One could imagine the interfering signal having a sequence equal to the locally generated sequence with a few bits changed. Thus the filter input will have a spectrum with large energy concentrations falling on its passband. Hence most of the interference energy will appear at the input to the message demodulator.

The usefulness of sequences for spread spectrum applications is generally measured in terms of their auto and cross-correlation functions. However for a given set of sequences these functions can take different forms depending upon their definition.

Consider initially the sequence auto-correlation function (a.c.f), usually defined in general terms as:-

$$R(\tau) = \int_{-\infty}^{\infty} f(t)\, f(t-\tau)\, dt$$

For finite length sequences this gives rise to the periodic and aperiodic auto-correlation functions. The periodic auto-correlation function refers to the correlation of a cyclic shift of the sequence with itself taken over the complete length of the sequence. As the name implies the resulting correlation function is repetative with a period equal to the total sequence period. The aperiodic auto-correlation function refers to the correlation of only a small part of the sequence with itself. This is to say that only a small subsection of the sequence is correlated with that sequence in part, resulting in a nominally non repetitive auto-correlation function. Obviously as the length of sequence sub-section increased the aperiodic auto-correlation function approaches the periodic one for that sequence. Note that for infinite length sequences only an aperiodic type of auto-correlation function exists.

For many sequences the periodic auto-correlation function is fairly simple, only taking on a few values. The aperiodic auto-correlation function is however generally more complicated, taking on a wider range of values.

By analogy the periodic and aperiodic cross-correlation functions are similarly defined. Here of course the correlation is performed between different sequences rather than one sequence with itself. For spread spectrum applications the periodic auto and cross correlation functions are generally of interest. However it is not always necessary to have detailed knowledge of these functions, as values for upper and lower correlation bounds are usually sufficient.

A paper[43] by Sarwate and Pursley describes sequence properties in general and discusses maximal length and related sequences in some detail. Included is an evaluation of the peak periodic cross-correlation values for range of maximal length sequences. This shows that randomly selected sequences of this type may have high periodic cross-correlation peak values. However carefully chosen sub-sets can have quite small periodic cross-correlation peaks. Unfortunately the number of sequences contained in these (maximal connected) sets are rather low. For example a maximal length sequence of period 2047 contains 176 different sequences which can take peak periodic cross-correlation values of 287. However a maximally connected subset of just 4 sequences can provide a periodic cross-correlation peak of 65. The authors conclude that maximal length sequences are ideal where very small numbers of sequences with excellent auto and cross-correlation properties are required. They are inadequate in situations where large numbers of sequences are needed with good cross-correlation properties.

Under these latter circumstances Gold[14] sequences can provide a useful solution. These are sequences which form a large set having low bounded peak periodic cross-correlation values. This is obtained at the expense of increasing the peak periodic auto-correlation value. By way of example it is possible to obtain 2049 Gold sequences of period 2047 for which the peak periodic cross-correlation does not exceed a value of 17. Clearly such sequences would be of great value in multiple user systems having many potential users.

An alternative approach to sequence construction is that described by Milstem et al[44]. Here the emphasis is on sequences which can readily be synchronised in a short time, with only a few bits of the sequence received. These sequences are formed by combining 2 (or possibly more) short sequences to yield a long sequence, having less than ideal auto and cross-correlation properties. For a combination sequence of length 10^9 bits formed from 2 sequences of approximate length 3×10^4 bits Milstem states that 11 simultaneous system users can be tolerated for reliable operation. Obviously such sequences do not have immediate application in systems designed to operate with many users.

3.3.2 An Alternative Approach

An alternative approach to the calculation of the number of simultaneous allowable users of a spread spectrum system is that of Beale and Tozer[13]. They consider the problem of reliably synchronising a user in the presence of interference from existing users of the system. Using true correlation type receivers with synchronisation by searching for a correlation peak the results are pessimistic. For equi-power received signals a theoretical maximum of 10 users can be reliably synchronised for a 25 dB system process gain. The results for practical spreading sequences are shown to be slightly worse than this.

Examination of Fig. 3.2 shows that for a 25dB process gain (spread bandwidth 7.9 MHz) 11 simultaneous users can be active for a 15 dB despreader output signal to noise ratio. Whilst the values for the allowable number of simultaneous users obtained from Fig. 3.2 may be considered optimistic, the author feels that the values given by Beale and Tozer represent a lower limit. In practical systems synchronisation will be a sophisticated process and may not rely on searching for synchronisation peaks. In particular use of some reference for synchronisation, such as the base to mobile calling channel, would remove the restrictions imposed by Beale and Tozer. Consequently the number of allowable simultaneous users will increase to limits imposed by spreading sequence cross-correlations.

The results given in this paper do not alter the validity of equ (3.12) though the factor k_c may be quite small. At best this work indicates that considerable attention requires to be given to synchronising spread spectrum systems if maximum user density is to be achieved. At worst it shows that the user density is much lower than initially expected. However the topic is open to debate and suggests further work for sequence theorists.

3.3.3 Orthogonal Spreading Signals

It is interesting to consider the use of spreading sequences which are orthogonal; this is to say, a set of sequences for which the cross correlations between members of the set are zero. Thus the sequences are non-interfering and the number of links which could operate simultaneously over the same channel would only be bounded by the number of sequences available. Certainly this would be the case for true correlation type receivers if not for the active despreader type.

Ideally the sequences would retain their orthogonality regardless of any modulation imposed upon them. Furthermore there would be no requirement that the sequences have a common time epoch to ensure orthogonality. This implies that the periodic cross-correlations of the sequences would be zero, i.e. the sequence orthogonality is invarient to cyclic shifts. If sequences possessing these properties could be developed, having many members to a set, they would find widespread application in spread spectrum systems. In practice the conditions may be relaxed slightly to permit low cross correlations between members of the set, though this would limit the number of allowable simultaneous users of the system.

3.3.4 Sequence Lengths

It is appropriate at this point in the thesis to discuss briefly the topic of spreading sequence lengths for direct sequence spread spectrum systems. The topic is of relevance, as several properties of spreading sequences are related to sequence length, having an effect on the performance and operation of direct sequence systems.

In a large scheme with many individual users it is desirable to provide each user with a unique spreading sequence from a given set of available sequences. The number of unique spreading sequences available from a given set increases with increasing sequence length. Consequently direct sequence spread spectrum systems having many users each requiring a unique spreading sequence must use long length sequences. As an illustration of the numbers involved Table 3.i shows the number of unique maximal length spreading sequences available against length of sequence.[31]

As discussed in Section 3.3.1 it is necessary for all the members of a given set of available spreading sequences to have low cross-correlations in order to allow for the greatest number of simultaneous users of a system. For maximal length sequences Judge[9] shows that the sequence cross-correlations, as measured by the factor k_1, decrease for increasing sequence length. Representative values given by Judge are given in Table 3.i.

On a similar theme it was stated in Chapter 2 that received interference is given noiselike properties in the despreading process due to the properties of the spreading sequence. As useful spreading sequences are periodic they have line spectra, which to approximate to noise must have a small spacing between the spectral lines. It is difficult to quantify the line spacing required as this is generally a compromise between various factors, nevertheless it does effect the sequence length. As an example of the numbers involved consider a direct sequence system having a 10 MHz spread bandwidth for which a 100 Hz line spacing is required. From equation (2.8) the line spacing for maximal length sequences is :

$$\frac{1}{iT} = \frac{B_{rf}}{2i}$$

Hence rearranging:

$$i = \frac{10^7}{2 \times 10^2}$$

$$= \underline{\underline{5 \times 10^4 \text{ bits}}}$$

In practice a 65,535 bit maximal length sequence would be used as the nearest value.

Another consideration that enters into the topic of sequence length is that of synchronisation. Consider the initial synchronisation process, where the receiver local sequence replica is 'drifted' slowly past the incoming signal until they are aligned, a process generally referred to as sliding correlation. The 'drift' rate is determined by the post-despreader filter bandwidth and the time required to identify alignment between the sequences. As these factors are fixed the only variable affecting the synchronisation time is the sequence length. On the average the shorter the spreading sequence the shorter the initial synchronisation time, an important point in direct sequence communications systems.

For a sequence of length i 'drifted' at a rate B_D past the incoming signal the time T_D to drift through the complete sequence is:

$$T_D = \frac{i}{B_D} \tag{3.13}$$

Thus a 65,535 bit maximal length sequence 'drifted' at 10 k bits/sec would take 6.55 seconds to drift through the sequence. Assuming that initial synchronisation was completed when the received and local sequences first came into alignment the period T_D represent the maximum synchronisation time. On the average it would not be necessary to drift through the complete sequence and the synchronisation time would be correspondingly less. Other synchronisation schemes [38,] can be used which do not require the system to search through the spreading sequence and therefore have shorter synchronisation times.

The selection of a spreading sequence length is likely to be a compromise for many systems between multiple user and synchronisation considerations. For systems using matched filters constraints[7] on the filters will limit the sequence lengths usable and consequently the system performance. Using such systems the process gain will equal the sequence length unless recursive techniques are used. However for any direct sequence spread spectrum system it is apparent that the selection of the spreading sequence length requires careful consideration.

3.4 Near Far Problem

The analysis has so far assumed that each unwanted transmitter produces the same power at a receiver as the wanted signal. Failure to achieve this results in the classic "near far" problem. This arises where the signal from a wanted distant transmitter is swamped by local unwanted transmissions. Despite the system process gain it is impossible to achieve a satisfactory signal to noise ratio at the receiver output. In a fixed station situation control of power levels and the use of directional antennae may alleviate the problem.

The 'near far' problem in a mobile context is analysed in Appendix A for a circular coverage area and uniform distribution of mobiles. The problem also receives some attention in a paper by Matthews[8] et al. The analysis shows that in the mobile to base direction the penalty for not achieving equal received powers is a large decrease in the number of allowable simultaneous users of the system. Hence for this scheme power control is required in the mobile to base direction to ensure equality of received powers by overcoming the differing propagation losses. In consequence use of separate transmit and receive bands is necessary. Furthermore direct mobile to mobile communication is not possible in this scheme as the power control would be ineffective. This is not a serious restriction as the range would be limited.

It should be noted that power control is not required in the base to mobile direction of transmission. Here the ratio of wanted to unwanted signal levels will be constant throughout the coverage area for a central base station. However the absolute signal level will decrease with increasing distance from the transmitter.

To achieve power control, knowledge is required of the path loss between mobile and base station. The mobile transmitter power can then be adjusted to produce a constant received level at the base station. Assuming the propagation path to be reciprocal, power control can be achieved by adjusting the mobile transmitter power to follow the variations in received signal from the base station. The sensing circuit should be connected close to the message demodulator. This ensures that the system operates on the wanted signal rather than interference. A block diagram of this simple method of achieving power control is given in Fig 3.3.

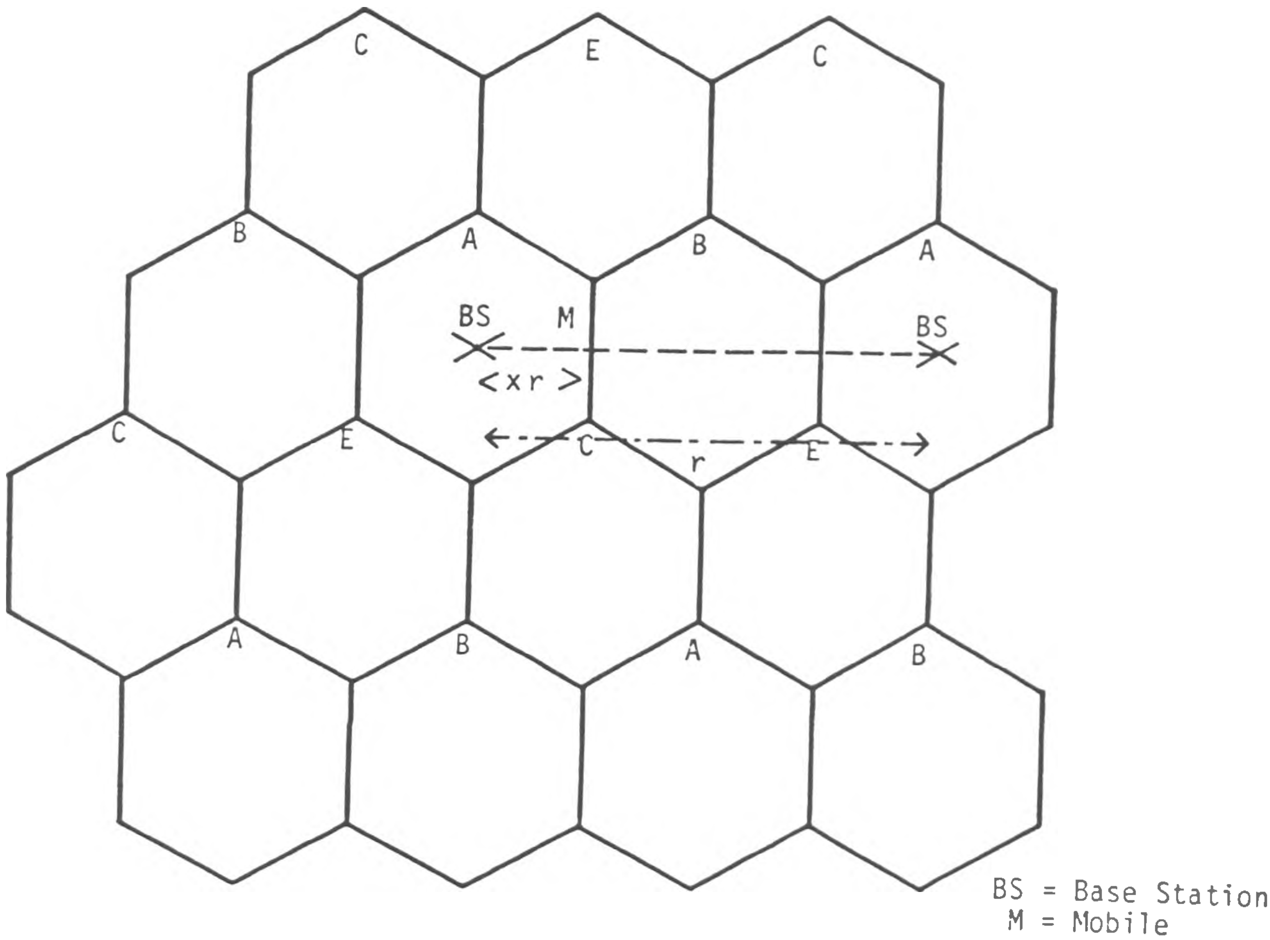


Fig 3.4 Layout for Co-Channel Interference Calculations

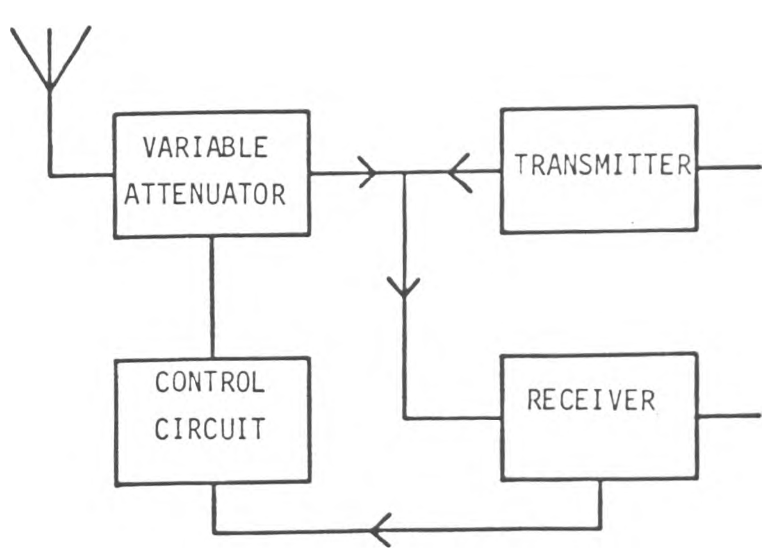


Fig 3.3 TRANSMITTER POWER CONTROL

3.5 Spectral Efficiency in Cellular Systems

It is useful to compare the spectral efficiency of direct sequence and existing mobile ratio systems. Generally frequency reuse is obtained by spatial separation, leading to mobile radio systems being interference limited. Thus for acceptable operation it is necessary to have a minimum protection ratio against co-channel interference. This leads to a minimum separation distance for frequency reuse. For narrowband modulation techniques a paper by Gosling[10] evaluates the spectral efficiencies as a function of protection ratio.

To make a meaningful comparison of spectral efficiencies it is necessary to compare like with like. Hence following Gosling consider a large coverage area divided into many equal area hexagonal cells, as shown in Fig 3.4. At the cell centre is a base station having M equal power transmitters for communication to mobiles in the cell. Base stations using the same frequency bands are spaced distance r apart. The mobiles are assumed to have omnidirectional antennae as are the base stations, whilst a fourth power propagation law is assumed. Thus a mobile located on the line joining a wanted and unwanted base station and a distance xr ($o<x<1$) from the wanted base station will receive power P_W from it, where

$$P_W \quad \alpha \quad \frac{1}{x^4 r^4} \tag{3.14}$$

The total received interference P_I will be due to M-1 other transmissions originating from the wanted base station and M interfering transmissions from the distant cell using the same frequency band.

Hence $$P_I \quad \alpha \quad \underset{\text{Internal}}{\frac{M-1}{x^4 r^4}} \quad + \quad \underset{\text{External}}{\frac{M}{(1-x)^4 r^4}} \tag{3.15}$$

This can be rearranged to give the input signal to interference ratio as :

$$\left(\frac{S}{I}\right)_I = \frac{1}{(M-1) + \dfrac{M\,x^4}{(1-x)^4}} \tag{3.16a}$$

which can be rewritten as :

$$\left(\frac{S}{I}\right)_I = \frac{1}{(M-1) + \dfrac{M}{V^4}} \qquad V = \frac{1-x}{x} \tag{3.16b}$$

If the mobile is on the limit of its service area then V is the ratio of interference range to service range.

Assuming that the input signal to interference ratio is improved by the system process gain, then the signal to noise ratio at the input to the message demodulator is :

$$\left(\frac{S}{N}\right)_o = \frac{B_{rf}}{B_m\left[(M-1) + \dfrac{M}{V^4}\right]} \tag{3.17}$$

This can be rearranged to give the value for the maximum number of users of a cell as :

$$M = \frac{B_{rf}}{B_m\left(\frac{S}{N}\right)_o\left(1 + \dfrac{1}{V^4}\right)} + \frac{1}{\left(1 + \dfrac{1}{V^4}\right)} \tag{3.18}$$

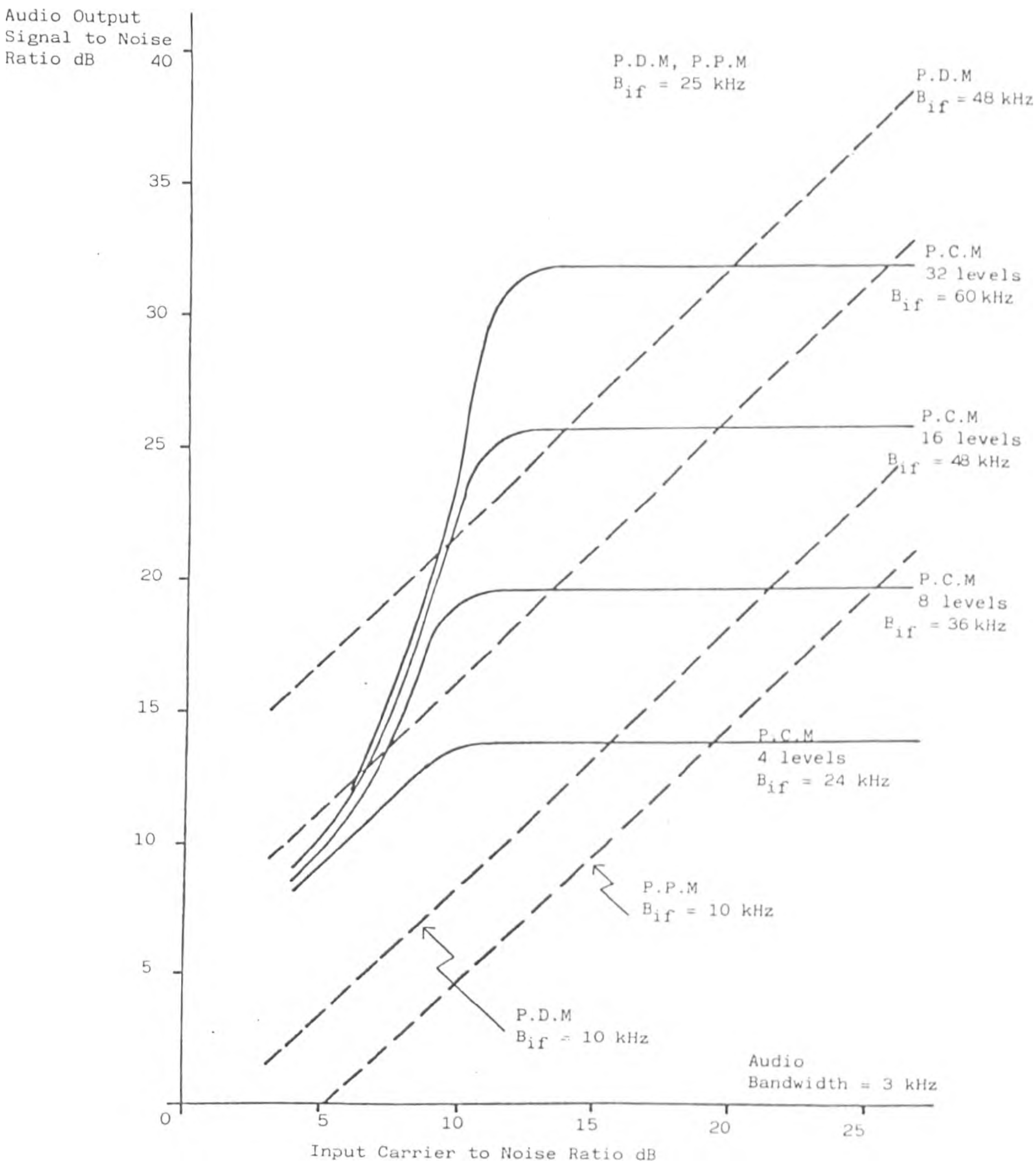


Fig 3.5 Graph of Audio Output Signal to Noise Ratio against Carrier to Noise Ratio for Various Modulation Schemes.

Thus if F frequency bands have to be used to provide frequency reuse, the spectral efficiency in Users/MHz is :

$$\frac{10^6 \ M}{F\ B_{rf}} = \frac{10^6}{F\ B_m \left(\frac{S}{N}\right)_o \left(1 + \frac{1}{V^4}\right)} + \frac{10^6}{F\ B_{rf} \left(1 + \frac{1}{V^4}\right)} \qquad (3.19)$$

(All Bandwidths in Hz)

For cellular schemes of this type there is generally some fixed relationship between the number F of frequency bands and the parameter V.

Table 3.ii shows the user densities obtainable in a direct sequence system for a range of post despread bandwidths and signal to noise ratios. As might be expected high spectral efficiencies only occur for low signal to noise ratios and small bandwidths at the despreader output. However the spectral efficiency falls as the position of the interfering cells moves outwards. Under these conditions the interference is predominantly from the users own cell, rather than from distant cells. The decrease in interference allows more simultaneous users in a cell, though not proportionally to the number of frequency bands needed to achieve this. Hence the number of users per unit bandwidth decreases. The user density is largest for the single band case with interference from immediately adjacent cells and is half of the maximum obtainable if co-channel interference is absent. ($V = \infty$)

For convenience Fig. 3.5 shows the relationship between audio output signal to noise ratio and despreader output signal to noise for a range of speech conversion schemes. The graphs were plotted from the same information used to derive table 2.1 in section 2.5, and assume a 3 KHz speech bandwidth.

Thus using Fig. 3.5 and Table 3.ii it is possible to obtain the spectral utilisation for a system using any of the speech conversion schemes shown. Consider a system for which a minimum audio signal to noise ratio of 12 dB is required and only a single frequency band is to be used. Using a 10 KHz p.d.m. speech conversion scheme Fig. 3.5 shows that a 13.5 dB despreader output s.n.r. is required, which interpolating from Table 3.11 permits a spectral utilisation of 2.28 Users / MHz. However a 25 KHz p.d.m. scheme requiring only a 5 dB despreader s.n.r. can achieve 6.37 Users / MHz.

For comparison user densities for narrowband modulation techniques are given in Table 3.iii as a function of the required protection ratio. This analysis included the effects of Rayleigh fading due to multipath propagation. However no account of multipath propagation was taken in calculating the user density for direct sequence systems. This topic is discussed elsewhere in this thesis as are the effects of shadowing, both of which lower the allowable user density of direct sequence systems. Hence the figures for the spectral efficiency of direct sequence systems evaluated here should be taken as upper bounds, only attainable under ideal conditions.

The wideband and narrowband analyses only considered the interference arising from the most proximate co-channel cell. However interference is present from other co-channel cells though at a lower level. Nevertheless this may be significant at certain positions of the mobile on the boundary of the cell.

User density figures for narrowband mobile radio systems have been calculated by other authors[11,15] and show reasonable agreement with those of Gosling. The values in Table 3.iii compare well with those for the user density of a frequency hopping scheme described by Cooper and Nettleton.[12]

3.5.1 Problems of Comparison

It is difficult to make accurate comparisons of the spectral efficiencies of mobile radio systems because of the many variables involved. Obviously as a start the analysis must be performed on systems operating under identical conditions. Even then the results can vary from analysis to analysis depending on the initial assumptions made.

An area where problems can arise is in measuring the system performance. Generally for speech communications the system quality is measured by the audio signal to noise ratio. This may be unsatisfactory where the audio interference is not noise but some other form of degredation, perhaps of an intelligible nature. The subjective effects will probably be different to that of true noise of an equivalent power. Connected with this is the minimum required performance for reliable communication. This is important in direct sequence systems where the spectral efficiency is highly dependent on the required signal to noise ratio.

It is apparent that direct sequence systems can achieve user densities of the same order of magnitude as narrowband systems, the actual user density obtainable varying with the system parameters chosen. It should be noted that the signal to noise ratios given in Table 3.ii are those at the input to the message demodulator. The audio signal to noise ratio will generally equal or exceed these values. On this basis direct sequence systems can be considered as spectrally efficient as conventional modulation techniques. Furthermore, spread spectrum techniques show several advantages over conventional systems.

In practical systems most users are only likely to be active for a small fraction of the time. As all direct sequence users are identical, except for spreading sequence, users can replace each other with no difference to the system. Hence providing the activity is evenly distributed in time it is possible to have many more potential users than active users in the system. Free access to the channel is available at all times, though communication may be difficult when many users are active. Thus the system will tend to be self regulating. When the loading is high users with non-urgent messages will find communications difficult due to the low signal to noise ratio. They may then cease communicating until conditions are more favourble. Users with urgent messages will probably develop more precise methods of communicating and will be able to operate under all conditions. These latter points are concerned with operating technique and are therefore matters of conjecture.

This situation should be contrasted with that for schemes using conventional narrowband modulation techniques. Again advantage can be taken of the low user activity factor to allow many more potential users than active users in the system. To do this entails channel sharing between users, implying that access to the channel may be restricted. In large schemes some form of channel assignment technique is necessary requiring a central control and considerable extra circuitry at the mobile. Such techniques may not adapt easily to changed circumstances and do not provide unrestricted access to the channel.

3.6 Aspects of System Organisation

It has been shown that direct sequence spread spectrum techniques allow several links to communicate simultaneously over the same channel without undue mutual interference. To make effective use of this facility requires some system organisation as the following discussion will illustrate. For simplicity a large area coverage scheme will be considered in which a central base station serves a small fleet of mobiles. The use of separate bands for transmission to and reception from the mobiles is assumed as is the use of power control at the mobile.

If simultaneous communication with several mobiles is required then a corresponding number of links must be set up. On this basis it is reasonable to assign to each mobile its own spreading sequence which acts as a unique address and allows selective calling thereby. Thus the base station only requires a small number of transmitters and receivers which can be programmed with any of the spreading sequences in use. When contact with a particular mobile is required the appropriate spreading sequence is programmed into an available transmitter/receiver and the mobile called.

When a mobile wishes to initiate a contact problems arise as the base station will not know which mobile is calling. Consequently the base station will not have a transmitter/receiver programmed with the appropriate spreading sequence. This problem can be overcome in a number of ways, the most apparent being scanning or the use of a calling channel.

In scanning, as the name implies, a base station receiver is regularly programmed with all the inoperative spreading sequences one at a time. The presence of a signal at any instant identifies a calling mobile allowing a transmitter/receiver to be programmed for communication.

An alternative is the use of a dedicated calling channel or more appropriately calling sequence. Mobiles wishing to initiate a contact would transmit an identifying signal using a common spreading sequence. This signal is received at the base station and causes equipment for communication to be set up as before. This system is perhaps more complicated than the previous one and problems would arise if several mobiles tried to initiate contacts simultaneously. Also the calling channel would reduce the spectral efficiency of the system slightly, though being low data rate it could be low power.

In the base station to mobile direction a form of calling channel/sequence is particularly valuable. It would provide an acknowledgement to a calling mobile that its call had been successful. If operated continuously it would provide a suitable signal for controlling mobile transmitter power control circuits. Finally the calling sequence could be used as a broadcast facility and might also find use for synchronisation purposes.

Table 3.i

Number of unique sequences and cross-correlation factor against sequence length for maximal length sequences

Number of Shift Register Stages	Sequence Length	Number of Sequences[31]	Sequence Cross-correlation Factor[9] K_1
2	3	1	
3	7	2	0.714
4	15	2	0.466
5	31	6	0.291
6	63	6	
7	127	18	0.134
8	255	16	
9	511	48	
10	1023	60	
11	2047	176	0.047
12	4095	144	0.078
13	8191	630	0.016
14	16383	756	
15	32767	1800	
16	65535	2048	
17	131071	7710	
18	262143	8064	

Table 3.ii

Number of Users/MHz for Direct Sequence Cellular System

Despreader Output s.n.r. (Minimum dB)	V = 1 Bm 25kHz	F = 1 Bm 10kHz	V = 2 Bm 25kHz	F = 3 Bm 10kHz	V = 3 Bm 25kHz	F = 4 Bm 10kHz
0	20.05	50.05	12.58	31.40	9.90	24.72
5	6.37	15.86	3.99	9.95	3.17	7.83
10	2.05	5.05	1.28	3.17	1.02	2.49
15	0.68	1.63	0.42	1.02	0.33	0.80
20	0.25	0.55	0.15	0.34	0.12	0.27
25	0.11	0.20	0.07	0.13		

(Assumes 10 MHz spread bandwidth)

V = Ratio of interference range to service range on boundary

F = Number of frequency bands required

Bm = Post despreader filter bandwidth

Table 3.iii

Number of Users/MHz for frequency modulation (from Gosling [10])

Modulation	Protection ratio dB 3<a<9.5	9.5<a<18.5	18.5<a
25 kHz F.M.	10	5.7	4.4
12.5 kHz F.M.	20	11.4	8.8

a = Protection ratio

(C.C.I.R.[37] recommends a minimum of 8 dB protection ratio for f.m. systems, other sources[27] suggest higher values of 12-14 dB).

CHAPTER 4

BANDSHARING

CHAPTER 4

BANDSHARING

Up to the present point in the thesis consideration has been given to spread spectrum systems sharing a common wideband channel. In view of the wide bandwidth required it is unlikely that channels could be allocated for exclusive use by spread spectrum systems. Hence the possibility of bandsharing with existing narrowband systems requires investigation. The onus falls on the spread spectrum system to prove that negligible interference would be caused to existing systems. Having established this then it remains to show that spread spectrum systems could operate satisfactorily under these conditions.

4.1 Interference to Narrowband Systems

For the narrowband system the concern is in the level and nature of the interference resulting from the spread spectrum overlay. To evaluate these consider the situation of a narrowband system operating in a band in which a single direct sequence transmitter is operational. The frequency domain representation is shown in Fig 4.1, where the effects of receiver noise and other interference are neglected. The direct sequence power P_{sn} accepted by the narrowband receiver is dependent on the power spectral density $\underline{S}(f)$ and the receiver frequency response $|H_n(f)|^2$. Thus:

$$P_{sn} = \int_{-\infty}^{\infty} |H_n(f)|^2 \underline{S}(f)\, df \qquad (4.1)$$

Considering the single sided spectral density, which for a long length maximal sequence is :

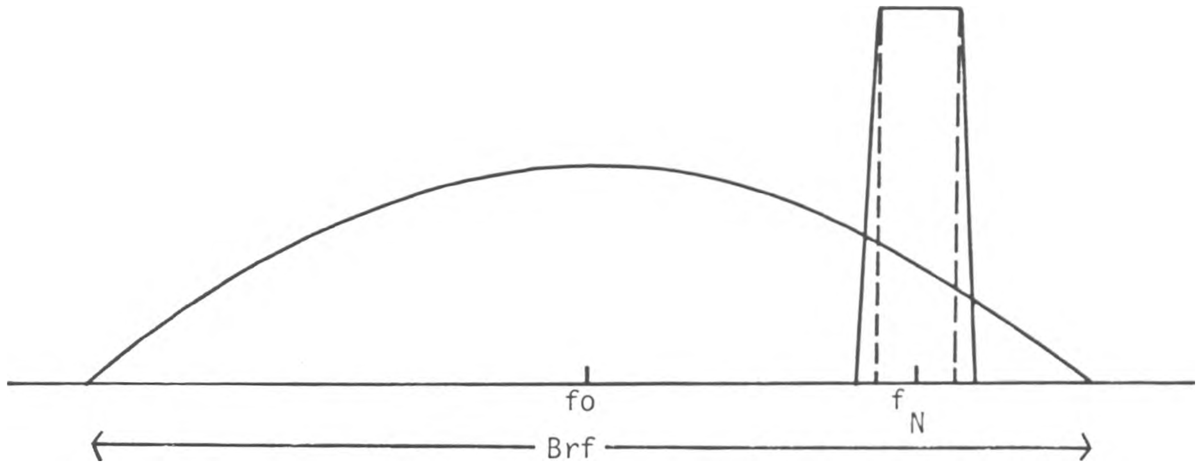


Fig. 4.1 Narrowband receiver tuned over Direct Sequence Transmission

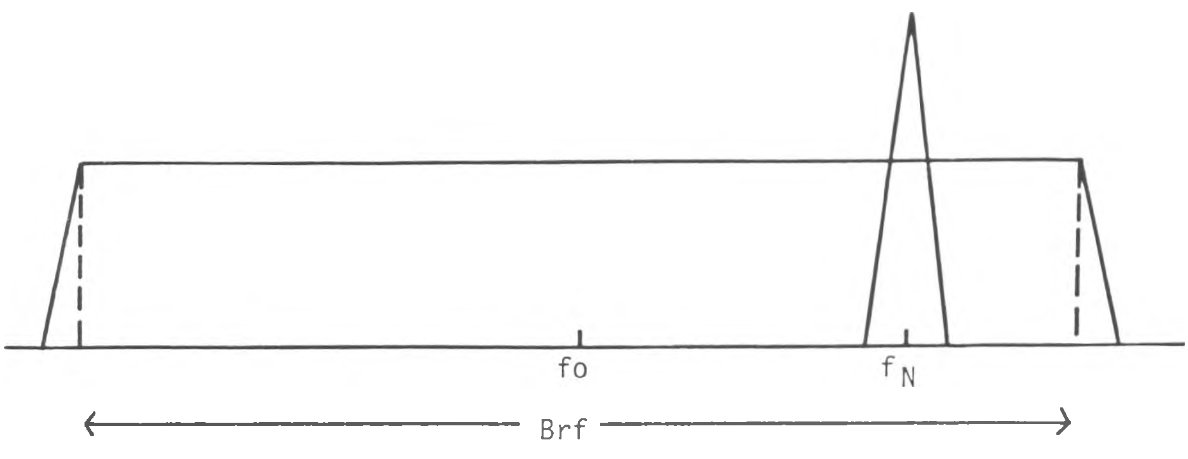


Fig. 4.3 Direct sequence receiver tuned over narrowband transmission

$$\underline{S}(f) = P_s T \left[\frac{\sin \Pi T (f-fo)}{\Pi T (f-fo)}\right]^2$$

equation (4.1) becomes :

$$P_{sn} = \int_{-\infty}^{\infty} |H_n(f)|^2 \quad P_s T \left[\frac{\sin \Pi T (f-fo)}{\Pi T (f-fo)}\right]^2 df \qquad (4.2)$$

If the receiver passband is assumed to have the form:

$$H_n(f) = \begin{cases} 1 & f_n - \frac{B_n}{2} \leqslant f \leqslant f_n + \frac{B_n}{2} \\ 0 & \text{elsewhere} \end{cases}$$

i.e. a rectangular passband of width B_n centred on f_n, the direct sequence power P_{sn} appearing at the output becomes :

$$P_{sn} = \int_{f_n - \frac{B_n}{2}}^{f_n + \frac{B_n}{2}} P_s T \left[\frac{\sin \Pi T (f-fo)}{\Pi T (f-fo)}\right]^2 df \qquad (4.3)$$

This can be rearranged to give :

$$P_{sn} = \int_{T(f_n - \frac{B_n}{2} - f_o)}^{T(f_n + \frac{B_n}{2} - fo)} P_s \left(\frac{\sin \Pi x}{\Pi x}\right)^2 dx \qquad (4.4)$$

The worst case situation occurs when the narrowband receiver has its passband centred on the direct sequence carrier ($f_n = f_o$), thus:

$$P_{sn} = 2\int_{0}^{\frac{TB_n}{2}} P_s \left(\frac{\sin \Pi x}{\Pi x}\right)^2 dx \qquad (4.5)$$

$$\simeq T B_n P_s \simeq \frac{2P_s B_n}{B_{rf}} \qquad (B_n \ll B_{rf})$$

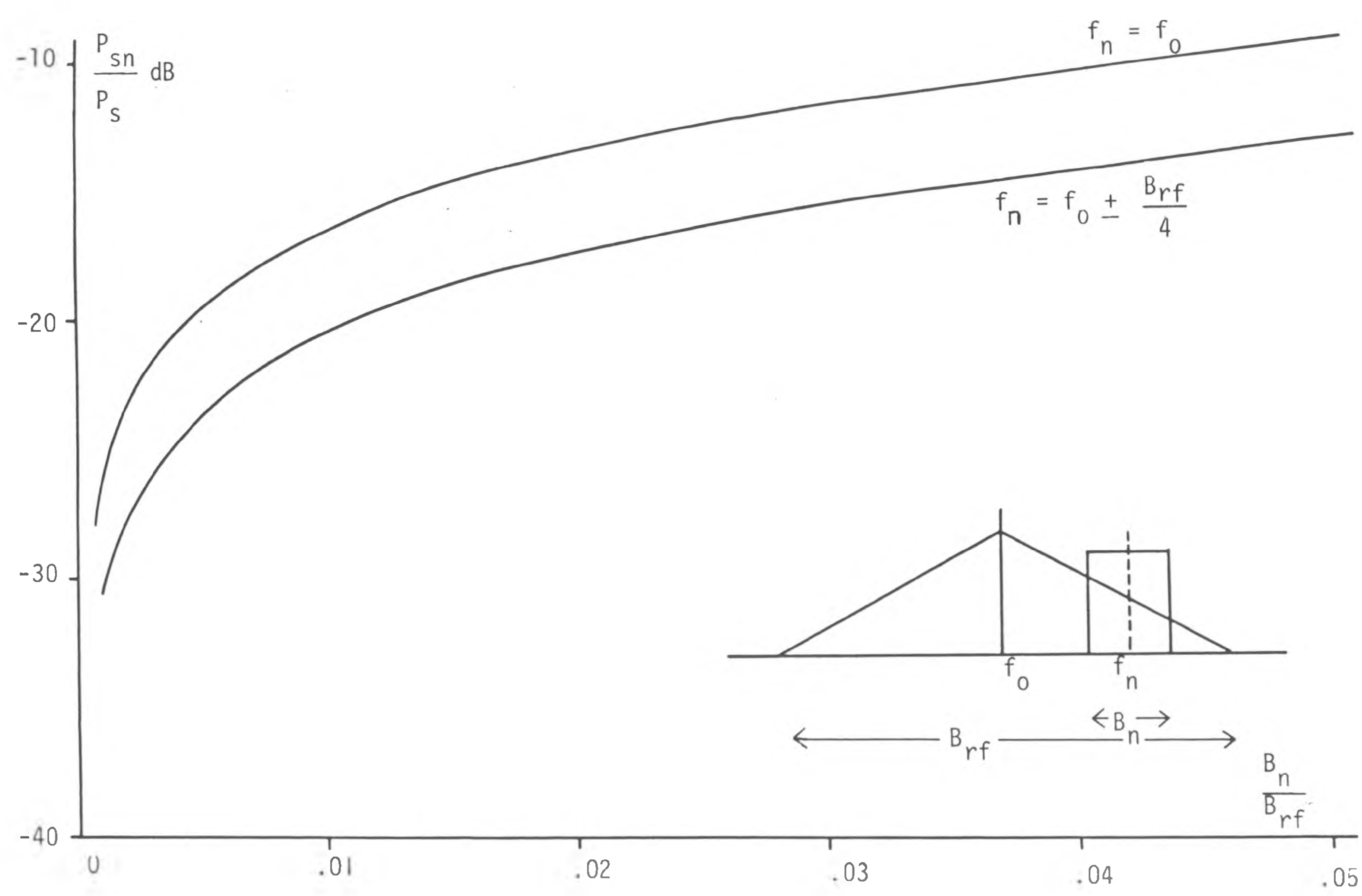


Fig 4.2 Graph showing ratio of direct sequence power at output of narrowband receiver to total direct sequence power against bandwidth ratios

Equation (4.5) has been plotted in Fig 4.2 for various ratios of narrowband to spread spectrum main lobe bandwidths. The vertical axis shows the ratio of power accepted by the narrowband receiver to the total direct sequence power. The results can be extended to the situation of many interfering direct sequence transmitters by superposition. It is apparent that for small ratios of narrowband to spread spectrum bandwidths the direct sequence power entering a narrowband receiver can be small. Knowing the narrow band signal power P_n Fig 4.2 can be used to evaluate the signal to interference ratio at the narrowband receiver.

For example :

$P_s = P_n$ $\qquad B_{rf} = 10$ MHz $\qquad B_n = 10^4$ Hz

$f_n = f_o$

$$\frac{P_{sn}}{P_s} = -26\text{dB}$$

The level of interference having been evaluated it remains to discuss its nature. The above analysis assumed the direct sequence signal to have a continuous spectrum, implying a close line spacing. Thus to a narrowband receiver the direct sequence signal will be indistinguishable from Gaussien noise. This is the case, for filtered maximal length sequences are often used as white Gaussien noise sources. As the sequence length decreases the direct sequence spectrum becomes a line spectrum. Providing the line spacing remains fairly small, with many lines in the narrowband passband, the same conclusion holds. However as the stage is reached where the receiver passband contains only a few lines the interference cannot be considered noise like.

4.2 Interference to Direct Sequence Systems

The previous section examined the effect of direct sequence transmissions on narrowband reception. Hence it remains to evaluate the effect of narrowband transmissions on direct sequence reception. Intuitively it is expected that the narrowband signals would be reduced in their effect by the spread spectrum process gain. To illustrate this consider initially the situation of a single narrowband transmission located in the passband of a direct sequence receiver, as shown in Fig 4.3. Assuming that the narrowband signal and spreading sequence are unrelated then the interference power at the direct sequence message demodulator is:

$$P_{ns} = \int_{-\infty}^{\infty} |H_s(f)|^2 \left[\underline{C}(f) * \underline{P_n}(f)\right] df \qquad (4.6)$$

Using the single sided spectral representation the spreading sequence can be taken to have the spectral density :

$$\underline{C}(f) = T \left(\frac{\sin \Pi Tf}{\Pi Tf}\right)^2$$

Thus the interference power becomes :

$$P_{ns} = \int_{-\infty}^{\infty} |H_s(f)|^2 \int_{-\infty}^{\infty} T \left(\frac{\sin \Pi T(f-g)}{\Pi T(f-g)}\right) \underline{P_n}(g)\, dg\, df \quad (4.7)$$

(where the convolution of power spectral densities has been expanded).

Furthermore if the pre-demodulator filter has a rectangular passband of bandwidth B_m centred on f_o, then :

$$P_{ns} = \int_{f_o - \frac{B_m}{2}}^{f_o + \frac{B_m}{2}} \int_{-\infty}^{\infty} T \left(\frac{\sin \Pi T\,(f-g)}{\Pi T\,(f-g)}\right)^2 \underline{P_n}(g)\, dg\, df \quad (4.8)$$

Knowing the form of the narrowband spectral density the resultant interference to the direct sequence message demodulator can be evaluated.

For the situation where the narrowband signal has a small bandwidth compared to the direct sequence signal it is reasonable to write:

$$\underline{P_n}(f) = P_n \quad \delta(f-f_n) \tag{4.9}$$

This approximates the narrowband signal to a carrier.

Thus :

$$P_{ns} = \int_{f_o - \frac{B_m}{2}}^{f_o + \frac{B_m}{2}} \int_{-\infty}^{\infty} P_n T \left(\frac{\sin \Pi T(f-g)}{\Pi T(f-g)} \right)^2 \delta(g-f_n) \, dg \, df \tag{4.10}$$

which can be rearranged to give :

$$P_{ns} = \int_{f_o - \frac{B_m}{2}}^{f_o + \frac{B_m}{2}} P_n T \left(\frac{\sin \Pi T(f-f_n)}{\Pi T(f-f_n)} \right)^2 df \tag{4.11}$$

This equation is in an identical form to equation (4.3) dealing with the interference output from narrowband receivers produced by direct sequence transmissions. Hence similar results and conclusions apply as reached earlier. The worst case situation is when the interfering signal is centred on the direct sequence carrier.

In these circumstances :

$$P_{ns} = \frac{2 \, P_n \, B_m}{B_{rf}} \tag{4.12}$$

As expected the narrowband interference is reduced by the direct sequence process gain. Also interferers located away from the carrier produce less interference at the message demodulator.

The interference output for the situation of many interfering transmitters can be calculated by an extension of this analysis.

4.2.1 Problems with Multiple Narrowband Transmissions

Some care is required in interpreting the situation. Consider a direct sequence system operating over a band shared with many narrowband systems. Due to the number of signals the signal to interference ratio at the message demodulator may be low. If in an attempt to improve this the process gain is increased by increasing the occupied bandwidth extra interference enters the receiver from narrowband transmissions which now fall in the direct sequence passband. Hence the expected improvement in signal to interference ratio is not achieved. If in some defined sense the average total narrowband power per unit bandwidth is constant then the signal to interference ratio at the direct sequence message demodulator will be fixed regardless of spread bandwidth. The situation is akin to that of white noise interference where the interference level at the message demodulator is independent of spread bandwidth.

The above analysis has investigated the compatability of direct sequence spread spectrum and narrowband systems. It shows that the interference caused to one system by the other is fairly innocuous providing the interference power is not excessive. Hence the possibility of bandsharing by both systems arises. Whilst bandsharing is possible it would be difficult for the two systems to simultaneously share a common service area. The interference levels would be too high for satisfactory operation of either system.

However bandsharing would be practicable where the systems used adjacent service areas. The interference levels would be acceptable, except perhaps on the boundaries where the signal to interference ratio would be at its lowest.

To minimise interference the available bandwidth should be split into two. Both spread spectrum and narrowband systems covering adjacent service areas would use the same band for mobile to base transmission and the remaining half for base to mobile communication.

Whilst bandsharing with existing mobile radio systems is possible there are considerable organisational problems. However one part of the spectrum where bandsharing appears feasible is the U.H.F. television band (U.K.) In any given geographical area only a part of the band is used for local television reception, the remainder is left fallow to prevent co-channel interference between transmitters serving different areas. Hence spread spectrum systems could operate in the fallow band, where they would only cause interference to receivers in adjacent service areas. The television receivers in these areas might gain some added protection due to their antennae rejecting the direct sequence transmissions. Also the location and coverage area of the television transmitters is known, enabling the system to be planned for least overall interference. If required,techniques[16,17] could be used to prevent spread spectrum operation in the channels used for broadcasting in the immediate adjacent areas. The idea is discussed by Ormondroyd [18,19] who provides information on the protection ratio's required for minimal television interference.

CHAPTER 5

PROPAGATION EFFECTS

CHAPTER 5

PROPAGATION EFFECTS

The land mobile radio environment provides possibly the most difficult channel over which to provide a radio communications link. The transmitted signal is degraded by the channel making it difficult for the receiver to produce a useable output. The degradations are caused by the methods of propagation which are closely related to the environment in which the system is operating. It is thus important to investigate the effects of the channel on direct sequence spread spectrum systems to evaluate their suitability for land mobile radio use. This task will be undertaken here and where relevant comparison made with narrowband systems. Before doing this a brief discussion of the channel characteristics is worthwhile.

5.1 The Land Mobile Radio Channel

It is generally accepted [20] that the land mobile radio channel has the following characteristics :

a) Mean signal strength related to distance
b) Doppler shifts
c) Shadowing
d) Multiple propagation paths

The list is not complete, though these are the pertinent characteristics. Also the characteristics listed are not exclusive to the mobile radio channel.

The variation of signal strength with distance is not a channel degradation per se. However it is perhaps the most important factor to be considered in the design of a radio communication system.

In most land mobile radio systems the doppler shifts encountered are small (<100 Hz) and are in consequence not troublesome. However in those systems where phase coherence is important the tracking loops can be designed to compensate for doppler shifts of this magnitude. Finally doppler shifts in the received signal only occur when there is relative motion between transmitter and receiver.

The characteristics of shadowing and multiple propagation paths are almost unique to the land mobile radio channel. They are the characteristics which cause the degredation of signals sent over the channel and arise from the method of propagation.

In urban and suburban areas, radio waves to and from low aerials are propagated predominantly by reflection and diffraction.[21] Shadowing is caused by obstacles blocking the propagation path, resulting in a shadow area into which signals are not directly propagated. Consequently signals only reach the shadow area by diffraction, which results in a mean signal strength less than predicted directly from free space/plane earth propagation models. Furthermore the mean signal level varies over small areas in which it would otherwise be expected to remain sensibly constant. This spatial variation is only apparent over distances of many tens or hundreds of wavelengths. By its nature shadowing affects all frequencies equally in a given band and can be viewed as a kind of extra path loss.

Multipath propagation is caused by the reflection of the transmitted signal by obstacles in the locality. Thus the received signal is composed of many components of varying amplitudes and phases. At a fixed location these will interfere either constructively or destructively depending upon the frequency. Hence the channel has a frequency response comprising peaks and nulls. At a fixed frequency the channel response will be a function of position, having a spatial distribution of peaks and nulls. The spacing between nulls is of the order of a wavelength or less.

An assessment of the effects of shadowing and multipath propagation on a radio communications scheme can take several forms. An obvious approach is to consider a single link in isolation and investigate the effects of the channel on the output quality. An alternative approach is to consider a scheme in which many links are operating. Here the effects of the channel on output quality of a single link could be investigated along with any effects on the operation of the complete system. Both approaches have their merits though the latter is more valuable for systems design and operation.

For either approach it is simplest to investigate the effects of shadowing and multipath propagation separately. This is valid as the two modes of propagation are essentially independent. However in practical mobile radio channels both shadowing and multipath propagation are likely to be present simultaneously.

5.2 Effects of Shadowing in Area Coverage Schemes

To analyse the effects of shadowing on a direct sequence spread spectrum system consider initially a single area coverage scheme. This has a central base station attempting to provide complete coverage to the service area. The only interference is that arising from other users in the service area. Neglecting for the moment receiver noise we shall investigate the effect of shadowing on the spectral efficiency of the system.

Consider initially the base to mobile direction of transmission. The signals arriving at the mobile will have travelled over the same path. Thus it is reasonable to assume that the shadowing on each is totally correlated. Hence the ratio of wanted signal to other user signal power will be constant regardless of the position of the mobile and regardless of the shadowing. In the mobile to base direction of transmission signals reaching the base station will have travelled over different paths. Consequently the shadowing is likely to be uncorrelated between the incoming signals. However assuming each mobile/base station path to be reciprocal the mobile power control circuit will ensure equality of received signals at the base station. Hence the ratio of wanted signal to other user signal powers will be the same for all users. Thus for this scheme shadowing causes no reduction in the number of allowable simultaneous users.

The above statement requires qualification to indicate the tacit assumptions in the argument. The problems arise in those locations where the path loss is high, i.e. deep fades. It was assumed that the mobile power control was effective at all signal levels. However this may not be the case for low signal levels, where considerable mobile transmitter power would be required. Furthermore the analysis assumed that at all times the wanted signal was well above the noise level, predominantly receiver noise. Again this will not be true for deep fades.

* There is no work to support or dennounce the existance of reciprocal paths other than the Lorentz reciprocity theorem[39].

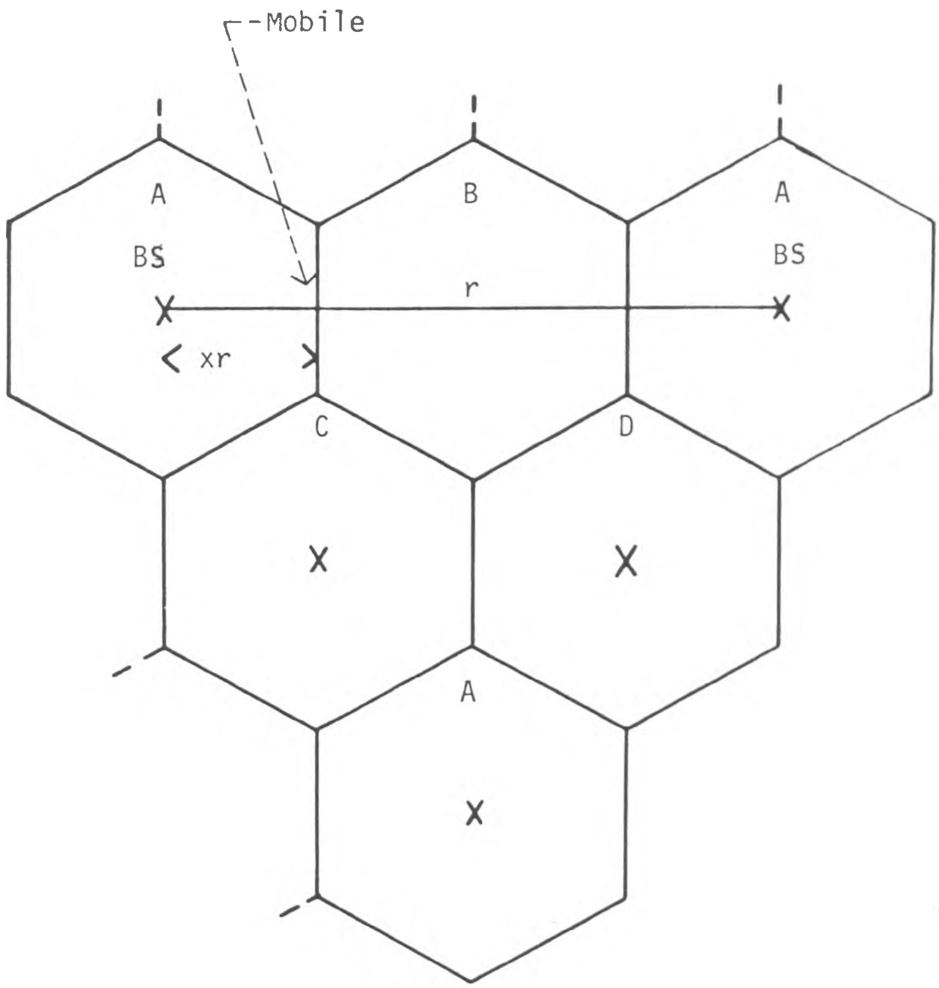
The situation is similar to that for narrowband transmission. For reasonable coverage of a given area high transmitter powers are required to overcome the extra path loss (over free space/ plane earth). Also unless repeaters or other schemes are used there will be locations to or from which communication is impossible.

5.3 Effects of Shadowing in Small Cell Schemes

Complications arise when shadowing in small cell schemes is considered. Here the shadow fading on signals to and from separate base stations may not always be correlated. Hence there will be times when the ratio of wanted signal to total unwanted signal power will be less than expected. This is due to the mobile being shielded from the wanted local base station and in 'full view' of an interfering base station. The situation for conventional narrowband modulation methodsis discussed by French,[15] who shows the deleterious effects of shadowing on the spectral efficiency of cellular narrowband systems.

This work can be used to gain a reasonable estimate of the effect of shadowing on the spectral efficiency of a cellular direct sequence mobile radio system. Consider the situation shown in Fig 5.1 for which the following are assumed :

a) Each cell has a central base station for communications with the mobiles in that cell.
b) Frequency reuse is obtained by spatial separation.
c) Each mobile has its own unique spreading sequence.
d) Only interference is from transmissions in the wanted mobiles cell and the nearest adjacent unwanted cell.



BS = BASE STATION

Fig 5.1 Layout for Co-channel Interference Calculations with Shadowing

e) No other interference or propagation effect is present.

Due to local obstacles there is shadowing on all signals giving rise to wideband fading. The shadowing on all signals from any base station is assumed correlated, whilst the shadowing between signals from adjacent base stations is independent, i.e., uncorrelated. The shadowing is assumed to have a log-normal probability density function.

Consider the base station to mobile direction of transmission only. For a given wanted mobile close to the cell boundary let the received wanted signal power from the wanted base station transmitter be P_S. Now if there are a total of M active users per cell, including the wanted mobile, the total received interference power from the base station in that cell will be :

$$(M - 1)\, P_S \tag{5.1}$$

From the nearest adjacent unwanted cell if each transmitter produces a received interference power of P_I at the mobile, then the total interference power received by the mobile from this cell is :

$$M\, P_I \tag{5.2}$$

Hence at the mobile the ratio of wanted signal power to total interference power $(S/I)_I$ is :

$$\left(\frac{S}{I}\right)_I = \frac{P_S}{(M-1)P_S + MP_I} \tag{5.3}$$

This is improved by the system process gain to produce a signal to interference ratio $(S/I)_o$ at the message demodulator.

input where :

$$\left(\frac{S}{I}\right)_o = \frac{P_s\ G_p}{(M-1)\ P_s + M\,P_I} \qquad (5.4)$$

Obviously there is a minimum acceptable value of signal to interference ratio $(S/I)_{omin}$ at the input to the message demodulator, below which system operation is unsatisfactory. This must correspond to a worst case situation when the maximum number of allowable simultaneous users are active.

Hence for the situation considered system operation is unsatisfactory if :

$$\frac{P_s\ \ G_p}{(M-1)P_s + M\,P_I} \leqslant \left(\frac{S}{I}\right)_{omin} \qquad (5.6)$$

which can be rearranged as :

$$P_s \leqslant P_I \quad \frac{M}{\left[\dfrac{G_p}{(S/I)_{omin}} - M + 1\right]} \qquad (5.7)$$

This can be rewritten as :

$$P_s \leqslant P_I \quad \beta \qquad (5.8)$$

where

$$\beta = \frac{M}{\left[\dfrac{G_p}{(S/I)_{omin}} - M + 1\right]}$$

Now fading due to shadowing is a statistical process and the interest is therefore in the probability of unsatisfactory reception. i.e. :

$$P\ (P_S \leqslant P_I\ \ \beta\ ) \qquad (5.9)$$

Thus Fig 4, Section V of French can be used to evaluate a value for β as this corresponds to the protection ratio required in narrowband modulation schemes.

Hence knowing β a value for the maximum allowable number of simultaneous users M can be evaluated. This of course depends on the mean received signal powers $\overline{P_s}$ and $\overline{P_I}$ and also on the desired probability of not being able to communicate.

$$\text{Hence } M = \frac{\dfrac{G_p}{(S/I)_{omin}}}{1 + \dfrac{1}{\beta}} + 1 \qquad (5.10)$$

Substituting for the process gain from (2.2) and rearranging gives a value for the spectral efficiency in users/MHz :

$$\frac{M}{B_{rf}} = \frac{10^6}{B_m \left(\frac{S}{I}\right)_{omin} \left(1 + \frac{1}{\beta}\right)} + \frac{10^6}{B_{rf}\left(1 + \frac{1}{\beta}\right)} \qquad (5.11)$$

(all bandwidths in Hz)

Consider the wanted and interfering base stations to be situated a distance r apart. Let the wanted mobile be on the boundary of its cell on the line joining the two base stations and distance xr (0<x<1) from the wanted base station. Assuming equal transmitter powers and the use of omnidirectional antennae, then for a fourth power propagation law the mean received wanted transmitter power $\overline{P}_s$ is :

$$\overline{P}_s \;\alpha\; \frac{1}{x^4 r^4} \qquad (5.12a)$$

Whilst the mean received interference power $\overline{P}_I$ from the interfering cell is :

$$\overline{P}_I \;\alpha\; \frac{1}{(1-x)^4 r^4} \qquad (5.12b)$$

Hence

$$\frac{\overline{P}_s}{\overline{P}_I} = \frac{(1 - x)^4}{x^4} = V^4 \tag{5.13}$$

Where $$V = \frac{1 - x}{x}$$

Hence from French

$$\beta = \frac{V^4}{10^{zd/10}} \tag{5.14}$$

As in Section 3.5 equations (5.11) and (5.14) can be used to obtain values for the user spectral density under various conditions. Typical figures are given in Table 5.i for the spectral efficiency in users/MHz for 10 MHz spread bandwidths. To permit a comparison with the results given in Table 3.ii account has been taken of the number of frequency bands F required to provide frequency reuse, following Section 3.5.

For the purposes of comparison table 5.ii shows the spectral utilisation of cellular f.m schemes operating under identical conditions. The values given are based on the work of Gosling and French. Examination of tables 5.i and 5.ii shows that an increase in the standard deviation of the shadowing or a decrease in the allowable outage time reduces the spectral utilisation of f.m and direct sequence modulation schemes.

Comparison of tables 5.i and 3.ii shows that shadow fading reduces the spectral utilisation of direct sequence systems, the reduction being greatest for small reuse distances. When the reuse distance is small the ratio of mean wanted signal power to mean co-channel interference power is small. Consequently a large protection margin is required to ensure reliable operation in the presence of shadow fading. To provide this protection margin a reduction in the number of active simultaneous users is necessary. Hence the spectral utilisation of the complete system decreases from the situation where shadowing is absent.

As the reuse distance increases the mean ratio of wanted signal power to co-channel interference power also increases. Thus the protection margin required against shadow fading decreases with a consequent increase in the spectral utilisation. As the spacing between co-channel cells increases the spectral utilisation will approach that for the situation where shadow fading is absent.

Thus for cellular direct sequence systems operating under conditions of shadowing several frequency bands are required to provide greatest spectrum utilisation. This contrasts with the analysis in Chapter 3, where shadowing was absent, for which greatest spectrum utilisation was achieved using a single frequency band.

The relationship between despreader output signal to noise ratio and audio signal to noise ratio is shown in Fig. 5.4 for a range of speech conversion schemes. This is to be used in conjunction with table 5.i in an identical manner to the use of Fig. 3.5 with table 3.ii in Chapter 3. As an example of the numbers involved consider a direct sequence system using a 25 KHz p.d.m. speech conversion scheme which is required to maintain a minimum 12 dB audio signal to noise ratio. Thus from Fig. 5.4 a minimum of a 5 dB despreader output signal to noise ratio is required. In the presence of log-normal shadowing having a standard deviation of 6 dB the spectral utilisation would be 2.77 Users/MHz for a 10% outage time. For a similar outage time with shadowing of 12 dB standard deviation the spectral utilisation falls to 1.12 Users/MHz.

Comparison with table 5.ii shows that this particular direct sequence system has a lower spectral utilisation than 12.5 and 25 kHz f.m. schemes operating under similar conditions and requiring a 10 dB protection ratio.

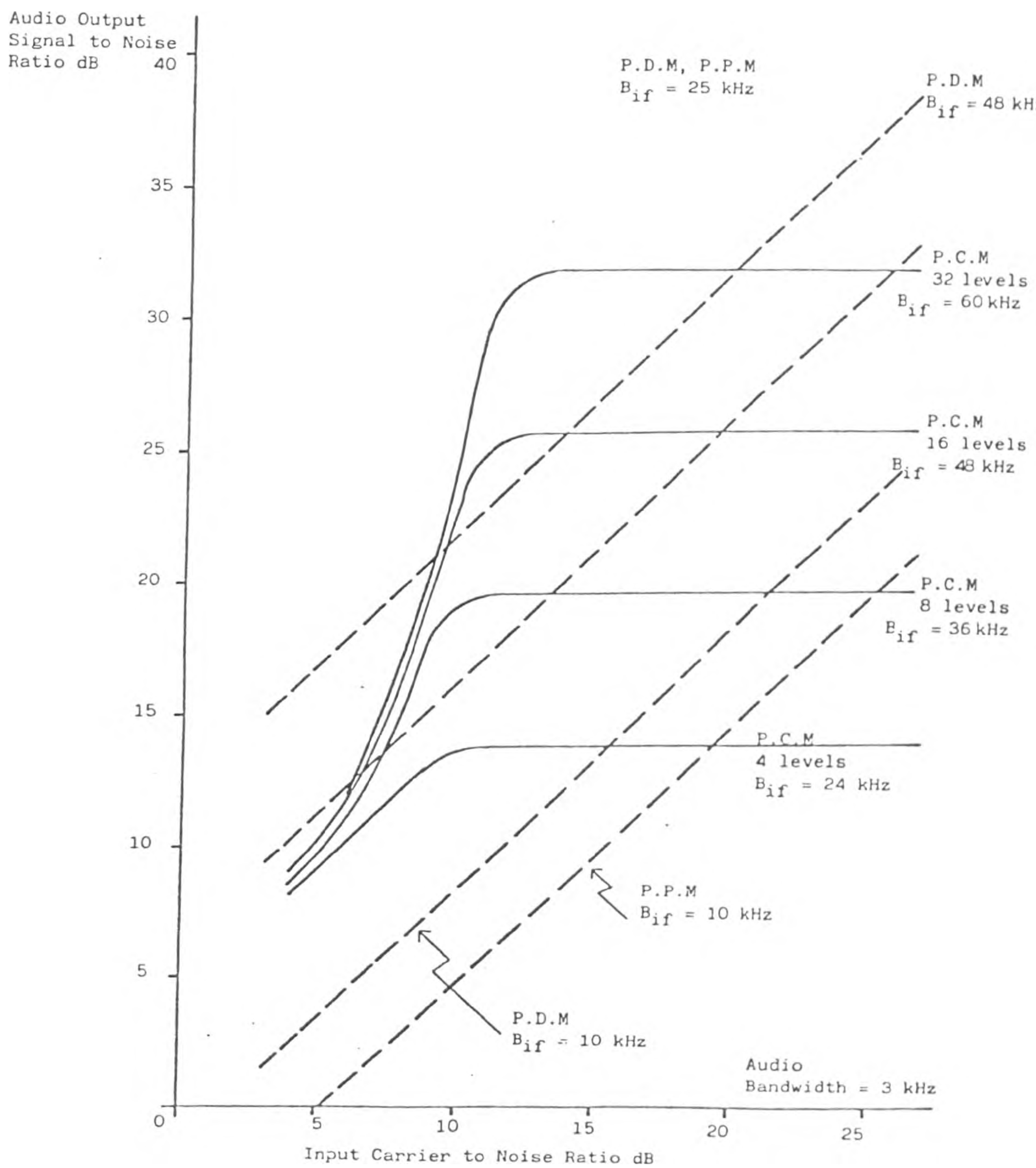


Fig [redacted] 5.4 Graph of Audio Output Signal to Noise Ratio against Carrier to Noise Ratio for Various Modulation Schemes.

5.4 Effects of Multipath Propagation

To investigate the effect of multipath propagation on a direct sequence spread spectrum system consider initially a single link. For convenience assume that no interference is present and that receiver noise etc., is negligible. Furthermore, assume the mobile is stationary.

For a multiple propagation path channel the impulse response is given by :

$$h(t) = \sum_{j=o}^{n} E_j \ \delta(t-\Delta_j-D) \tag{5.15}$$

Where E_j is the gain of the j^{th} path, Δ_j is the excess path delay and D is the minimum propagation delay. Without loss of generality the minimum propagation delay D can be neglected as only relative delay is important. Also let the minimum propagation delay path have j=o i.e., Δ_j = o. Thus the channel impulse response becomes :

$$h(t) = E_o \delta(t) + \sum_{j=1}^{n} E_j \ \delta(t - \Delta_j) \tag{5.16}$$

Consideration must now be given to the signal sent over the channel. For simplicity this will be a maximal length sequence p.s.k modulated onto the carrier, no message modulation being used. Any adjustments to the analysis for message modulation can be discussed later. Thus the transmitted signal is :

$$s(t) = c(t) \ \cos(w_o t + \theta) \tag{5.17}$$

$$\left[c(t) = \pm 1 ; \ w_o = 2\Pi f_o\right]$$

Where c(t) is the maximal length sequence, of bit period T.

After transmission over the channel the received signal is :

$$x(t) = s(t) * h(t)$$

$$= \int_{-\infty}^{\infty} c(t-\tau) \cos (w_o [t-\tau] + \theta) . \Big[E_o \, \delta(\tau)$$

$$+ \sum_{j=1}^{n} E_j \; \delta(\tau-\Delta_j) \Big] \; d\tau \qquad (5.18)$$

$$x(t) = E_o \, c(t) \cos (w_o t + \theta) + \sum_{j=1}^{n} E_j \; c(t-\Delta_j) \cos (w_o t + \theta - W_o \Delta_j) \qquad (5.19)$$

In the receiver this signal is multiplied with a locally generated replica of the spreading sequence to give :

$$y(t) = x(t) \, c(t-d)$$

$$= E_o \, c(t) \, c(t-d) \cos (w_o t + \theta)$$

$$+ \sum_{j=1}^{n} E_j \; c(t-\Delta_j) \, c(t-d) \cos (w_o t + \theta - w_o \Delta_j) \qquad (5.20)$$

The resulting signal is bandpass filtered prior to demodulation. As the system is linear it is convenient to consider only a single delayed component present at the input to the receiver along with the minimum delay signal. The complete system is as shown in Fig. 5.2. Hence :

$$y(t) = E_o \, c(t) \, c(t-d) \cos (w_o t + \theta) + E_1 \; c(t-\Delta 1) \, c(t-d) \cos (w_o t + \theta - w_o \Delta_1) \qquad (5.21)$$

$$= y_1(t) + y_2(t)$$

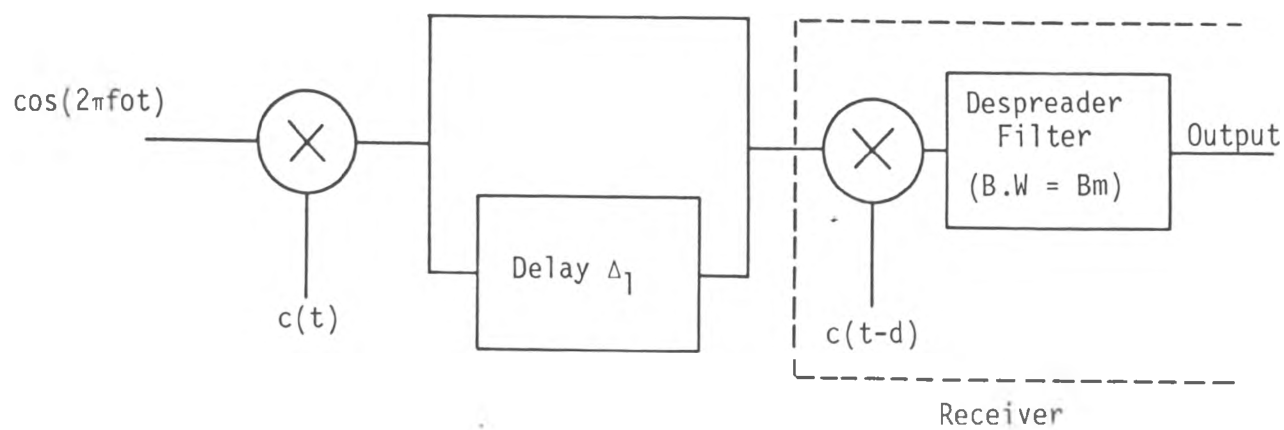


Fig.5.2 Diagram showing Multipath Channel for Direct Sequence Spread Spectrum System

Without loss of generality the receiver local spreading sequence will be assumed to be perfectly synchronised to the minimum delay (direct) component :

i.e. $d=o$

Thus : $y_1(t) = E_o \cos(w_o t + \theta) \quad [c(t)\ c(t) = 1]$

and $y_2(t) = E_1\ c(t-\Delta_1)\ c(t) \cos(w_o t + \theta - w_o \Delta_1)$ (5.22)

The first term $y_1(t)$ is the "wanted" signal, whilst the second $y_2(t)$ is the delayed path component. The effect of $y_2(t)$ depends upon the delay Δ_1 and two cases arise :

a) $\Delta_1 < T$ b) $\Delta_1 > T$

It is apparent that the performance of direct sequence spread spectrum systems under multipath propagation conditions is determined by the auto-correlation function of the spreading sequence. This is obviously the case for true correlation type receivers. Hence the results obtained would probably have been different for other spreading sequences. In this respect maximal length sequences are considered optimum as they have a two level auto-correlation function (periodic).

5.4.1 Excess Path Delay Less Than Chip Period

(a) $\Delta_1 < T$

Here the excess path delay is less than the spreading sequence chip period. With regard to the delayed component the situation is similar to the receiver having a small synchronisation error. Hence at the message demodulator the delayed signal produces a coherent component and a noiselike non-coherent component. Calculation of the power spectral density is difficult, though the coherent component due to the delayed signal is given by :

$$y_2(t) = E_1\left(1 - \frac{Q\Delta_1}{T}\right) \cos\,(w_o\Delta_1 + \theta - w_o\Delta_1) \qquad (5.23)$$

$$0<\Delta_1<T$$

At the message demodulator the total coherent component will be the sum of components from direct and delayed path signals :

$$= y_1(t) + y_2(t)$$

$$= E_o \cos(w_o t + \theta) + E_1\left(1 - \frac{Q\Delta_1}{T}\right) \cos\,(w_o\Delta_1 + \theta - w_o\Delta_1) \quad (5.24)$$

$$= Z \cos(w_o t + \gamma)$$

where $Z = \sqrt{E_o^2 + E_1^2\left(1 - \frac{Q\Delta_1}{T}\right)^2 + 2E_oE_1\left(1 - \frac{Q\Delta_1}{T}\right)\cos w_o\Delta_1}$

Obviously the resultant depends on the amplitude and phase of the delayed signal component. The coherent signal power can therefore be greater than or less than for the direct path signal only.

Calculation of the noise power at the message demodulator is not so easy. However the noise level will be greater than if the delayed signal component was absent. A paper by De Couvreur[23], on synchronisation errors in direct sequence systems, gives a similar result for the coherent delayed signal power as (5.23). Also given is an expression for the noise spectral density at the input to the filter due to the non coherent component. For an infinite length sequence the one sided noise spectral density is :

$$E_1^2 \;\; \frac{\Delta_1}{T}^2 \left[\frac{\sin \Pi\, \Delta_1(f - fo)}{\Pi\, \Delta_1(f - fo)}\right]^2 \qquad (5.25)$$

Thus at the output of a narrowband filter of bandwidth Bm centred on f_o the noise power will be approximately given by :

$$\frac{B_m E_1^2 \Delta_1^2}{T} \qquad (5.26a)$$

which can be rewritten as:

$$\frac{B_m E_1^2}{} \frac{\Delta_1^2 B_{rf}}{2} \qquad (5.26b)$$

As might be expected the extra noise is small for a small tracking error.

Thus from (5.24) and (5.26b) the resultant signal to noise ratio $(s.n.r)_o$ at the input to the message demodulator is :

$$(s.n.r)_o = \frac{Z^2}{B_m E_1^2 \Delta_1^2 B_{rf}} \qquad (5.27)$$

As an example of the figures involved let :

B_m = 25 kHz $\quad$ B_{rf} = 10 MHz $\quad$ $\Delta_1 = 0.1\ \mu s$

f_o = 100 MHz $\quad$ Q = 1 $\quad$ $E_1 = E_o$

i.e. an equal amplitude signal delayed by half a chip period :

Hence

$$(s.n.r)_o = \frac{2.25}{25 \times 10^{-4}}$$

$$= \underline{\underline{29.5 \quad dB}}$$

Still a fairly high signal to noise ratio, due to the two signal components interfering constructively. Had the signal phasing been such that the interference was destructive the resulting signal level would have been low, though the noise level would have altered but little.

5.4.2 Excess Path Delay exceeds Chip Period

(b) $\Delta_1 > T$

In this instance the excess path delay exceeds the spreading sequence chip period. Consequently the delayed signal component does not correlate well with the receiver local spreading sequence, producing non coherent signals at the input to the message demodulator. An exact calculation of the resultant spectral density and hence interference power is difficult. However an approximation can be made using the 'shift and add' property of maximal length sequences.[31] For application here this is taken as follows : 'A binary maximal length sequence multiplied with a version of itself delayed by an integer number of chip periods results in a shifted version of that sequence".

Hence for the case under discussion :

$$y_2(t) = E_1\ c(t)\ c(t-qT)\ \cos\ (w_0 t + \theta - w_0 \Delta_1)$$

which by the above statement can be written as :

$$y_2(t) = E_1\ c(t-pT)\ \cos(w_0 t + \theta - w_0 \Delta_1) \qquad (5.28)$$

From previous work (equn (2.9) this is known to have an approximate single sided power spectral density :

$$\underline{Y_2(f)} \simeq \frac{E_1^2}{2} T \left(\frac{\sin \Pi T(f - fo)}{\Pi T(f - fo)} \right)^2 \qquad (5.29)$$

This is a familiar spectral density which will produce a power P_{2N} at the output of a narrowband filter of bandwidth Bm centred on f_o where :

$$P_{2N} = \frac{E_1^2\ Bm}{B_{rf}} \qquad (5.30)$$

Furthermore the resulting signal is noiselike in form.

The wanted signal term is of course $y_1(t)$ which produces a power P_1 at the input to the message demodulator where :

$$P_1 = \frac{E_o^2}{2}$$

Thus the resulting signal to noise ratio is :

$$\left(\frac{S}{N}\right)_o = \frac{P_1}{P_{2N}} = \frac{E_o^2}{2\ E_1^2} \quad \frac{B_{rf}}{B_m} \qquad (5.31)$$

$$= \frac{E_o^2}{2E_1^2} \quad G_p$$

At the input to the receiver the ratio of direct to delayed path signal is :

$$= \frac{E_o^2}{E_1^2}$$

It is apparent that there has been an improvement in the ratio of direct path to delayed path signal powers by a factor related to the system process gain. The delayed signal in this case is treated as interference and its effect reduced by the system as for any other interference.

Results for the general case where the excess path delay takes any value exceeding the chip period are likely to be better. Generally for this case the spectrum at the output of the despreader will contain less energy close to the carrier. Thus there will be less pseudo noise power at the input to the demodulator as much of it will be rejected by the filter. In consequence the signal to noise ratio will be higher than given above for integer chip period delays.

The effect of a simple two path propagation model on a direct sequence system has been investigated. Whilst not complete the analysis shows that multipath propagation causes an increase in noise level at the message demodulator compared to the single propagation path case. This gives a corresponding decrease in signal to noise ratio for the case where the excess path delay exceeds the sequence chip period. An excess path delay of less than a sequence chip period can produce an enhancement or reduction in signal to noise ratio depending upon the path delay. For the direct sequence system described operating over a multipath channel, only signal components arriving within a chip period of the receiver local spreading sequence contribute to a coherent signal at the message demodulator.

The analysis here of the effects of multipath propagation on direct sequence systems was performed in the time domain, with the channel described by its impulse response. However, the analysis could, by Fourier transform theory, have been performed in the frequency domain, the channel being described by its transfer function. Providing the same models are used in both time and frequency domains the results will be the same regardless of which domain is used. For preference time domain analysis was chosen here as preliminary examination showed it would probably be computationally easier..

5.4.3 Effect of Multipath Propagation on Message Signals

Throughout the analysis of the effects of multipath propagation the transmitted signal carried no message modulation. It is therefore useful to briefly discuss the situation when digital modulation of messages is applied to the spreading sequence.

Generally the message bit period will be considerably greater than the maximum excess path delay of around 10 μS or less. Thus over most of each message bit the analysis will hold as the system will be in a quasi steady state. However when message transitions occur this will not be the case for a time duration equal to the maximum excess path delay. Therefore there will possibly be some brief transient effects at the beginning of message bits where transitions have just occurred. Due to the relatively short duration of these initial transients it is unlikely that they will be troublesome and can be neglected.

As mentioned elsewhere the message modulation of the spreading sequence will alter its properties from those given here. However as mentioned earlier this is not too serious a problem as the changes will be slight. Consequently for practical purposes the presence of message modulation can be neglected as it will have negligible effect on the final result.

5.4.4 Problems of Analysis Over Multipath Channels

In the analysis of the effects of multipath propagation it was assumed that there was no sequence tracking error at the receiver. This is to say that the receiver locally generated sequence was in exact alignment with the direct received signal component. This assumption will not always be valid,[3] as can be shown for a delay lock sequence tracking loop used at the receiver. Under certain conditions signal components arriving over paths with small excess delays interfere with the tracking loop operation producing tracking errors. Due to this the signal component is reduced and the noise increased giving an overall decrease in signal to noise ratio at the message demodulator

To evaluate the performance of the direct sequence system over real multipath channels requires knowledge of the excess path delays and their relative amplitudes. This information might be obtained from the work of Cox[25,26], who used a kind of maximal length sequence matched filter spread spectrum system to probe the mobile radio channel. Fig 5.3 is a typical envelope delay profile showing the amplitude of received signal components against excess path delay. For application to direct sequence communications it may be possible to reduce this to a small number of discrete paths of sufficient amplitude to be significant.

Such work would probably require considerable computational effort to be of value. Furthermore the measurements of Cox had a 0.1 μS resolution, implying a 10 MHz measurement bandwidth. Thus the results would not be useful for wide band spread spectrum systems having spread bandwidths in excess of this figure. For these reasons the work has not been pursued in this direction.

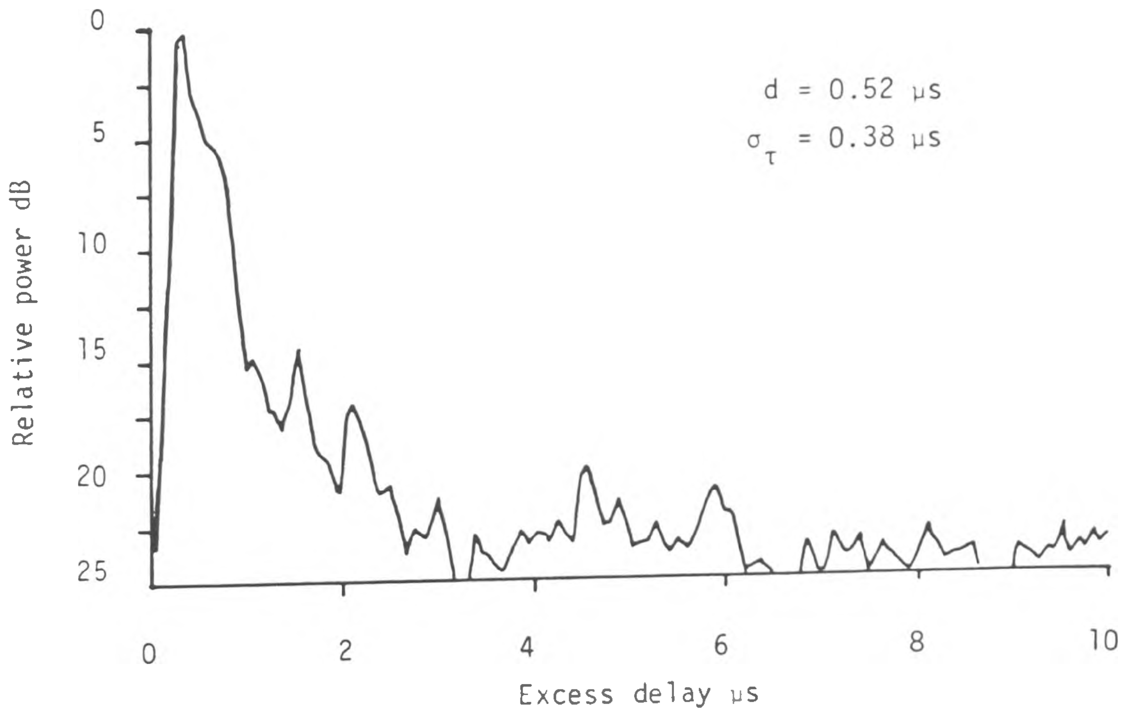


Fig 5.3 Typical Envelope Delay

Table 5.i

Number of Users/MHz for Direct Sequence Cellular System with shadowing

Despreader Output s.n.r $(S/N)_{omin}$ dB	V = 1 Bm 25kHz	F = 1 Bm 10kHz	V = 2 Bm 25kHz	F = 3 Bm 10kHz	V = 3 Bm 25kHz	F = 4 Bm 10kHz
	σ =6 dB		Pu = 10^{-1}			
0	3.04	7.59	7.58	18.94	8.71	21.75
5	0.97	2.40	2.41	6.00	2.77	6.89
10	0.31	0.76	0.77	1.91	0.89	2.19
15	0.10	0.25	0.26	0.62	0.29	0.70
	σ =6 dB		Pu = 10^{-2}			
0	0.42	1.05	1.94	4.85	4.64	11.58
5	0.13	0.33	0.62	1.54	1.47	3.67
10	0.04	0.10	0.12	0.49	0.47	1.17
15	0.01	0.03	0.06	0.16	0.15	0.37
	σ =12 dB		Pu = 10^{-1}			
0	0.26	0.67	1.29	3.24	3.53	8.82
5	0.08	0.21	0.41	1.03	1.12	2.79
10	0.02	0.06	0.13	0.33	0.36	0.89
15					0.12	0.28

(Assumes 10 MHz spread bandwidth)

V = Ratio of interference range to service range on boundary

F = Number of frequency bands required

Bm = Post despreader bandwidth

Pu = Probability of unsatisfactory reception

σ = Standard deviation of log-normal shadowing

Table 5.ii

Spectral efficiency in Users/MHz for cellular F.M. systems with shadowing

System	User/MHz		
	$\sigma = 6$dB	$P_u = 10^{-1}$	
Type	1.18<a<8.22	8.22<a<13.22	13.22<a<17.1
25 kHz f.m	10	5.71	4.44
12.5 kHz f.m	20	11.42	8.88
	$\sigma = 6$dB	$P_u = 10^{-2}$	
	0<a<4.35	4.35<a<8.23	8.23<a<11.39
25 kHz f.m	5.71	4.44	3.07
12.5 kHz f.m	11.42	8.88	6.14
	$\sigma = 12$ dB	$P_u = 10^{-1}$	
	2.35<a<6.23	6.23<a<9.39	9.39<a<12.07
25 kHz f.m	4.44	3.07	2.50
12.5 kHz f.m	8.88	6.14	5.00

a = Protection ratio dB

P_u = Probability of unsatisfactory reception

σ = Standard deviation of log-normal shadowing

(C.C.I.R[37] recommends a minimum of 8dB protection ratio for f.m systems, other sources[27] suggest higher values of 12-14 dB)

CHAPTER 6

Experimental Transmitter and Receiver

CHAPTER 6

Transmitter Receiver System

At the start of building a direct sequence spread spectrum communications link a plan was devised providing the broad outlines of the system. This would provide sufficient information to allow the circuits forming the system to be designed and built.

In the broadest terms the plan was to build a single transmitter and receiver capable of speech and data communication. The system would operate over a cable, though the equipment should be capable of conversion to a radio link if required. Finally the design and construction should be flexible to permit alterations to the system as found desirable. At the commencement of the project no details of similar equipments were known. Thus the details of the system were reached on an arbitrary basis, taking account of convenience and practicability. Hence the receiver would have an active despreader, whilst the transmitter would have an active spreading sequence generator. It was considered that passive despreaders were less flexible than active types.

One of the major considerations was the speech modulation method to be used. Little had been reported on analogue modulation methods for spread spectrum and this coupled with the ease of digital modulation led to the use of digital speech modulation. Of the many speech digitisation methods available suppressed clock pulse duration modulation was chosen. The paper by Jacobson et al[28] shows this to be a valuable technique, with many advantages for spread spectrum applications. At this stage the sampling rate was fixed at around 8 kHz, this being adequate for general speech.

The selection of spread bandwidth and carrier frequency were to some extent interdependent. As some useful equipment existed in the laboratory having a 30 MHz centre frequency this seemed a prudent carrier frequency to use. If required the frequency could be easily changed at some future time. The spread bandwidth was somewhat arbitrarily fixed at 4MHz. This was compatable with the 30 MHz carrier frequency and could again be readily changed if desired.

The use of a maximal length type spreading sequence was assumed, though the length still had to be fixed. The sequence generator would be programmable to generate a number of different length sequences, so a maximum length of $2^{18}-1$ (262, 143) was chosen. This would allow a 10 Hz line spacing to be achieved at a 2 Mb/s clock rate if desired.

6.1 Construction

The transmitter and receiver are built as two separate units, each comprising a card frame built into a small cabinet. In keeping with the theme of versatility a modular approach to construction was used. Each circuit function was assembled as a self contained entity, allowing easy replacement or alteration.

All digital and low frequency circuits were assembled on plug-in Veroboards, whilst r.f and analogue circuits were built on double sided printed circuit boards. The latter were made the same size as the plug in cards to allow them to slide in the card frame. Connections to the printed circuit boards are made using screw connectors and co-axial plugs and sockets, allowing for easy removal etc. All circuits incorporate adequate supply decoupling. Power is obtained from regulated mains power packs fitted to each cabinet.

6.2 General Outline of Transmitter

Following on from the system plan the transmitter design is as given in the block diagram Fig 6.1. The sequence generator produces the pseudo-random spreading sequence which is then modulated by the information signal. The information, either digitized speech or data, is reclocked prior to sequence inversion key modulating the spreading sequence. The output from the information modulator then phase reversal key modulates the r.f carrier for transmission over the channel.

Speech is converted to a binary format by the s.c.p.d.m modulator.

6.2.1 Sequence Generator

The sequence generator generates the wideband spreading sequence and modulates it with the information signal. The circuit consists of a shift register and feedback network forming the sequence generator, an information modulator and a modulator driver. The 18 stage shift register is constructed of 9 dual J-K flip-flops, arranged as shown in the circuit diagram Fig B 1. Feedback is provided by an 'exclusive or' gate which can be connected to any 2 outputs of the shift register. This allows a variety of sequences up to length 262,143 bits to be generated. The generator will operate at clock rates up to 15 Mb/s.

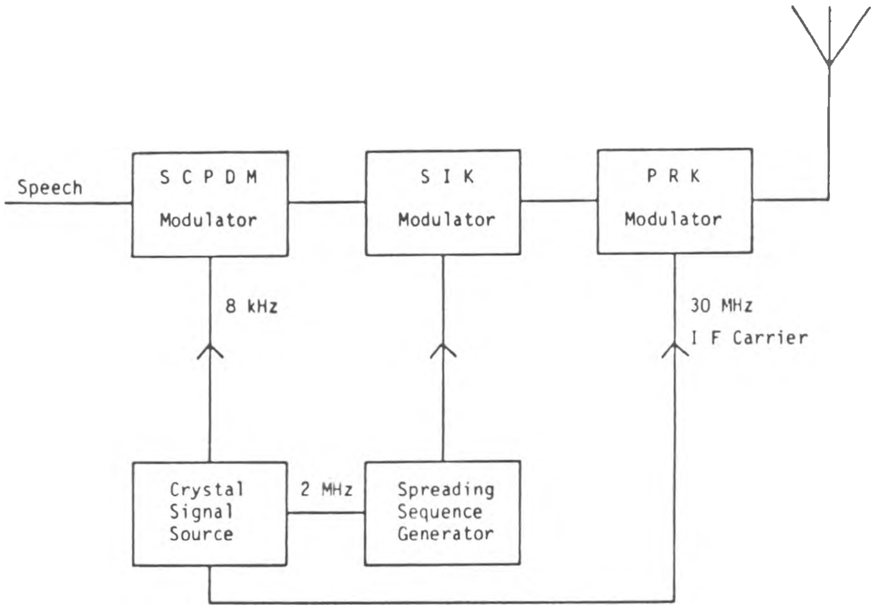


Fig. 6.1
Direct Sequence Transmitter

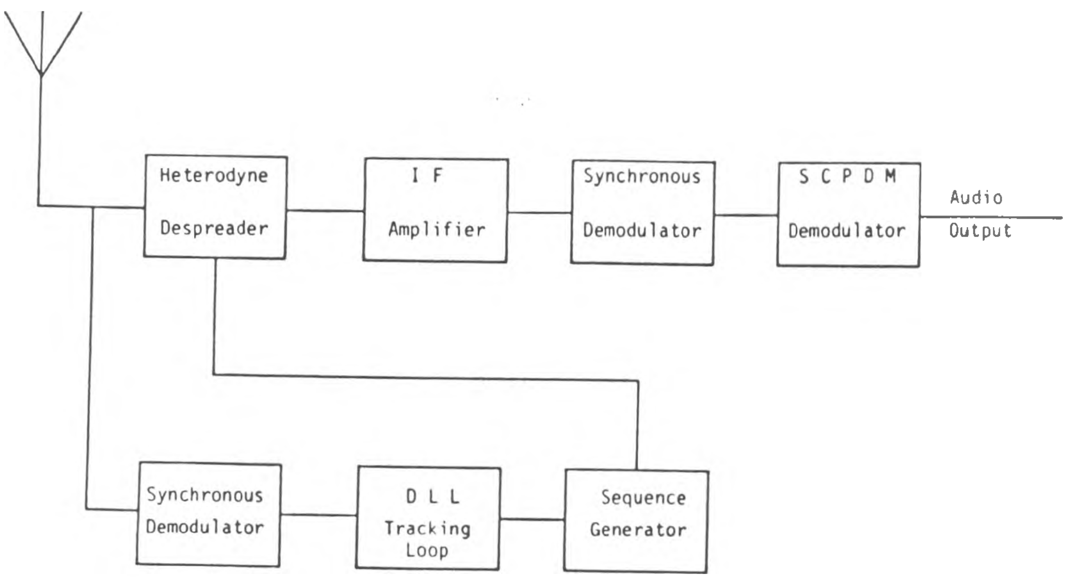


Fig. 6.2
Direct Sequence Receiver

An 'exclusive - or' gate forms the information modulator to provide sequence inversion keying modulation of the spreading sequence. A logic network selects either data or suppressed clock pulse duration modulated speech, which is reclocked by a D flip flop prior to modulation. The modulated spreading sequence goes via the driver to the carrier modulator where it phase reversal key modulates the r.f carrier.

6.2.2 Speech Modulator

The speech modulator converts speech to a suppressed clock pulse duration modulation (s.c.p.d.m) binary format. The circuit is similar to that described in (28), Fig B.2 shows the circuit used.

After initial amplification the speech is converted to pulse duration modulated speech using a circuit based on a 555 timer i.c. This drives an 'exclusive-or' gate driven by a half rate clock to produce s.c.p.d.m speech.

Speech compression can be applied from a 570 compandor i.c. switched in after the amplifier. In circuit this reduces the dynamic range and thereby allows a greater mean depth of modulation to be achieved.

6.2.3 Ancillary Circuits

All clock and carrier signals in the transmitter are derived from an integrated 10 MHz oscillator. A chain of dividers produce outputs at 2 MHz for the sequence generator, and 8 kHz and 4kHz for the speech modulator. The crystal

3rd harmonic at 30 MHz is extracted by a tuned amplifier to provide the r.f carrier. The carrier modulator is a double balanced diode modulator minicircuits type SRA-1.

6.3 Receiver

To the receiver falls the task of recovering the information signal from the incoming signal and interference. The block diagram Fig 6.2 shows that the receiver achieves this by performing the reverse process to the transmitter. The incoming signal is despread to a narrowband form, filtered out and demodulated. This produces the original information, either data or s.c.p.d.m speech. This latter signal is further processed by the s.c.p.d.m demodulator to reproduce analogue speech.

A heterodyne despreader is used to despread the wanted signal. It has the advantage of preventing interference leaking round the unit and entering the i.f amplifier. Furthermore it provides a downconversion of the wanted signal to a convenient frequency for filtering and amplification.

The i.f amplifier is connected to the despreader output. It filters out the wanted narrowband signal and rejects, as far as possible, interference. Thus it should have the smallest bandwidth necessary to pass the wanted signal along with good selectivity to provide maximum rejection of interference. Earlier work[29] shows that to avoid extra interference entering the i.f amplifier via image effects it should have a centre frequency of at least twice the spreading sequence clock rate. For acceptable reproduction s.c.p.d.m speech requires around 25 kHz i.f. bandwidth for an 8 kHz sampling rate. To provide adequate image rejection and good selectivity a double i.f scheme is used. The first i.f is at 10.7 MHz with down conversion to a second i.f. at 460 kHz. The gain is not critical, though for reasonable sensitivity a value in excess of 40 dB is required.

Following the i.f amplifier is the synchronous demodulator, used to produce a baseband output. As s.c.p.d.m speech has on the average no d.c component the incoming signal has no carrier component. Hence a carrier regeneration circuit is needed to produce a correctly phased carrier for demodulation. Demodulation occurs in a double balanced mixer with the output filtered to remove unwanted components. A squaring type carrier regeneration circuit is used to produce the carrier. The implementation of this circuit is simpler than the Costas and remodulation loops whilst all three have been shown [30] to have almost identical performances.

For speech signals the synchronous demodulator is followed by a s.c.p.d.m demodulator to provide recovery of the analogue speech. The s.c.p.d.m demodulator converts the incoming pulses to p.d.m pulses which are then processed to obtain a speech output. The classical method of p.d.m demodulation is conversion to p.a.m and low pass filtering. However a baseband output can be obtained by directly filtering the p.d.m signal. This latter method of baseband recovery is claimed[32] to cause less distortion with naturally sampled signals, whilst the former method is better for uniform sampling. Both methods of baseband recovery are used to allow a subjective evaluation of each to be made.

6.3.1 Heterodyne Despreader

At the front end of the receiver is the heterodyne despreader. It is made from 2 double balanced modulators, minicircuits type SRA-1. One modulator is used to modulate the receiver local spreading sequence onto a 40.7 MHz carrier. The output from this forms the local oscillator drive to the second modulator. The second modulator acts as a conventional receiver mixer. The input port is driven by the incoming signal and the output port drives the i.f amplifier.

6.3.2 I.F. Amplifier

As mentioned previously the i.f amplifier uses a dual conversion technique to obtain the required image rejection and selectivity. The circuit diagram is given in Fig 8.3.

The first stage at 10.7 MHz uses a f.e.t amplifier with single tuned circuits on input and output. Transformer coupling is used to provide low input and output impedances. An integrated circuit double balanced mixer type MC 1496[34] provides down conversion to 460 kHz. The local oscillator is crystal controlled at 11.16 MHz.

The main amplification and filtering is provided by the 460 kHz amplifier. This uses 3 double tuned transistor amplifier stages. A combination of damping and stagger tuning is used to obtain the required bandwidth. An emitter follower output stage provides a low impedance drive to the synchronous demodulator.

6.3.3 Synchronous Demodulator

The synchronous demodulator has the configuration shown in Fig 6.3. Incoming signals are split between the carrier regeneration circuit and the double balanced mixer used for demodulation.

The mixer is constructed from the MC 1496 integrated circuit double balanced modulator. This is followed by a low pass filter to remove unwanted products produced in the demodulation process. The filter was designed[33] for a second order Butterworth response with a 3dB cut-off frequency of 15 kHz. A 741 operational amplifier used in an active filter circuit provides a convenient method of implementing this function.

The circuit diagram, Fig B.4, shows the squaring type carrier regeneration circuit in more detail. A full wave rectifier and tuned amplifier produce a twice carrier frequency component from the incoming signal. This is tracked by a NE 561 integrated circuit p.l.l to obtain a relatively noise free output. The p.l.l loop filter bandwidth was set at 120 Hz to achieve a compromise between low output jitter and rapid response to changes in the input frequency. The p.l.l output drives a J-K flip-flop via a fet buffer. This produces a component at the carrier frequency which is phase shifted by an R.C. network connected to the Q. and $\overline{Q}$ outputs of the flip flop. This is adjusted to compensate for the $\pi/_2$ phase shift inherent in the p.l.l operation as well as other phase shifts in the circuit. A tuned amplifier and emitter follower provide a sine-wave drive to the double balanced modulator.

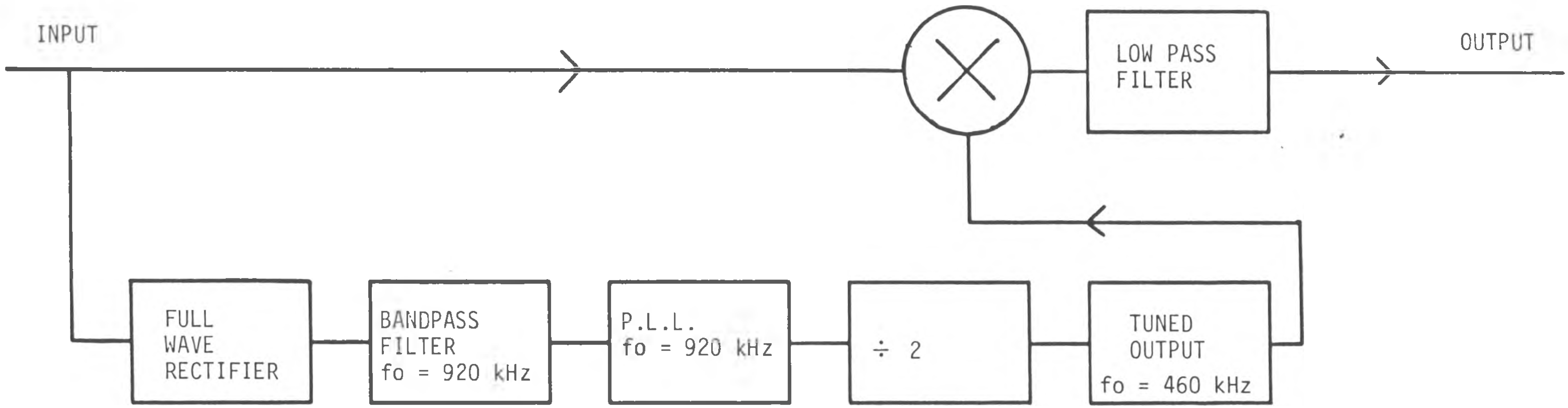


Fig. 6.3 Block Diagram Synchronous Demodulator

6.3.4 S.C.P.D.M. Demodulator

The s.c.p.d.m demodulator has the circuit given in Fig. B.5. Incoming signals are fed to a 710 integrated circuit comparator which converts them to TTL levels. This also allows the Unit to operate with a wide range of input levels. The comparator output is then multiplied by a half rate sampling clock (4kHz) in an 'exclusive or' gate. Providing the recovery clock is correctly phased the output is a p.d.m waveform. The clock is generated by a 74 LS 124 v.c.o running at 8 kHz, a 7474 flip flop providing the 4 kHz output. The clock generator acts as a p.l.l, the control signal being obtained by low pass filtering the recovered p.d.m signal. The loop bandwidth is set by the filter bandwidth at 20 Hz.

For the filter method of baseband recovery the p.d.m signal is routed to an active low pass filter. This was designed[33] as a fourth order low pass Chebychev type with ½ dB ripple and a 3 dB cut-off at 3 kHz. It is built around two 741 operational amplifiers and has a gain of 4. Audio output is obtained from a small amplifier and speaker.

The second method of baseband recovery uses conversion of the p.d.m signals to p.a.m followed by low pass filtering of this signal. The conversion to p.a.m is performed by an integrate and dump technique. The p.d.m pulses from the 'exclusive or' output drive a constant current source charging a capacitor. This produces a potential across the capacitor proportional to the pulse duration. At the end of the sampling period the capacitor potential is sampled by a LF 398 sample and hold and the capacitor discharged ready for the cycle to be repeated. The output from the sample and hold is a staircase approximation to the modulating signal and is low pass filtered by the filter described earlier. All the control timing is performed by a set of 74123 monostables.

6.3.5 Sequence Tracking

An important requirement for a direct sequence spread spectrum system is a local sequence replica at the receiver. Initially in the system described this was obtained from the transmitter sequence generator. A version of the spreading sequence unmodulated by information was fed directly into the despreader at the receiver by a coaxial link, the path delay on this link being adjusted to compensate for the delay on the r.f path.

Whilst this technique was adequate for many purposes it was obviously unsuitable for comprehensive measurements on the system. Hence to complete the system a sequence generator and tracking unit were constructed.

The complete sequence tracking unit is as shown in Fig 6.4 synchronised to the incoming signal by being incorporated into a delay lock tracking loop. The tracking loop operates at baseband as the implementation is simpler than for a system operating with signals modulated on a carrier.[35] A baseband signal is therefore obtained by synchronously demodulating the incoming signal. To obtain a 30 MHz coherent carrier to drive the demodulator a mixing process is used on the two local oscillator signals and regenerated carrier.

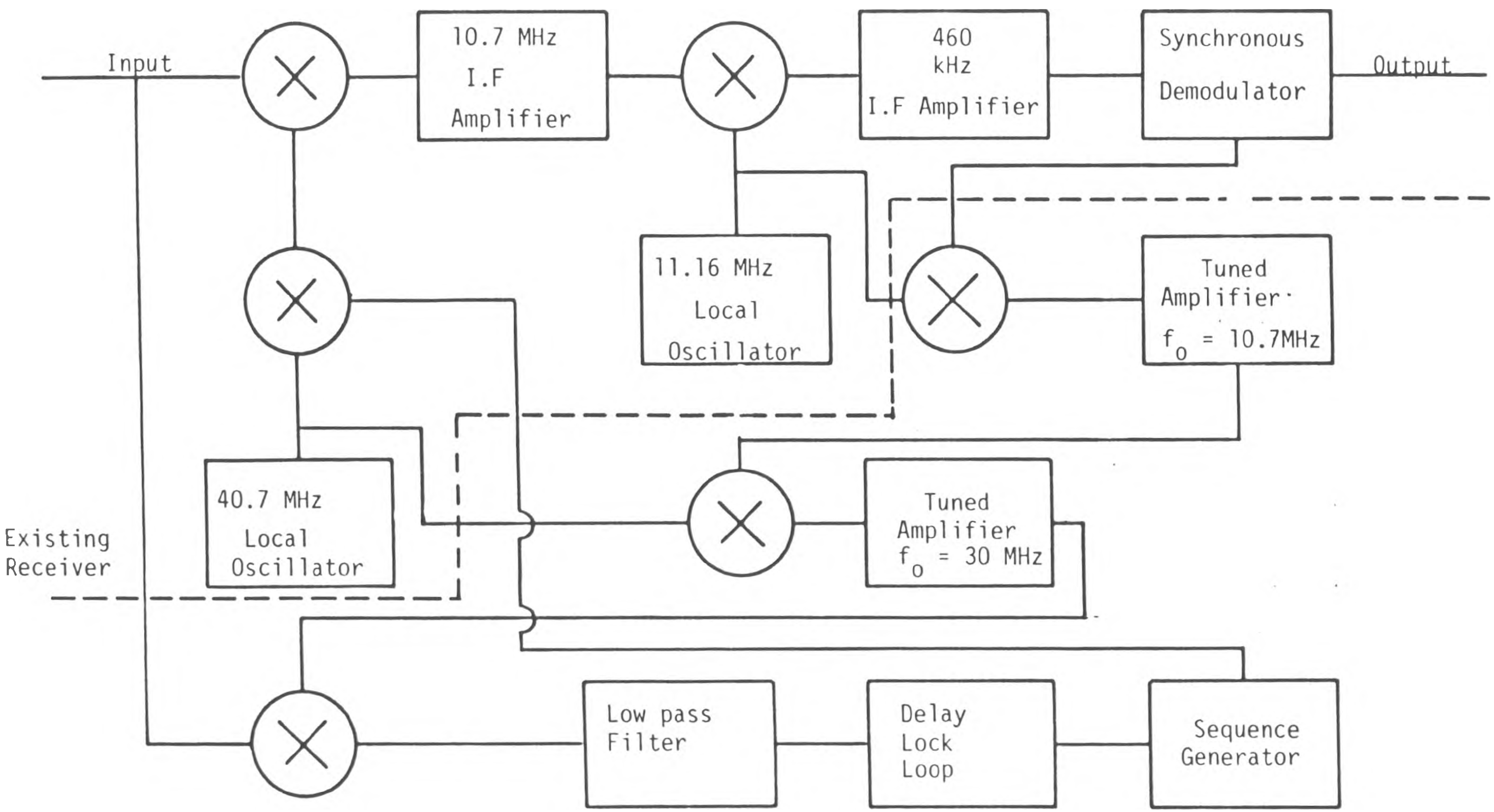


Fig 6.4 Block Diagram of Direct Sequence Receiver with Additions for Sequence Tracking

An obvious problem is that the system is not self starting, for unless the sequence generator is synchronised there is no signal for the carrier regeneration loop to track. This in turn deprives the sequence tracking loop of a signal and the condition is perpetuated. For lab tests it is acceptable to initially inject the baseband spreading sequence directly from the transmitter into the tracking loop, the system being synchronised by sliding correlation action and the injected signal removed as soon as correct tracking is maintained.

The technique is plainly unsuitable for elaborate experiments where a totally self contained synchronisation method would be needed at the receiver.

6.3.5.1 Delay Lock Loop

The delay lock tracking loop follows the general outline given by Ward [36] and serves to track the incoming maximal length sequence. It operates in conjunction with an appropriate sequence generator. This sequence generator is identical to the one used at the transmitter except that the message modulator and associated circuitry are omitted.

A block diagram of the delay lock loop is shown in Fig. 6.5 The incoming signal is split between two double balanced mixers using the MC 1496 integrated circuits. One of these is driven by the output of the 17th stage of the sequence generator shift register whilst the other is driven by the output of the 18th stage. The balanced outputs from the mixers are of an unsuitable d.c level and have an insufficient amplitude to drive the diode rectifiers. The signals are therefore amplified using 741 op amps

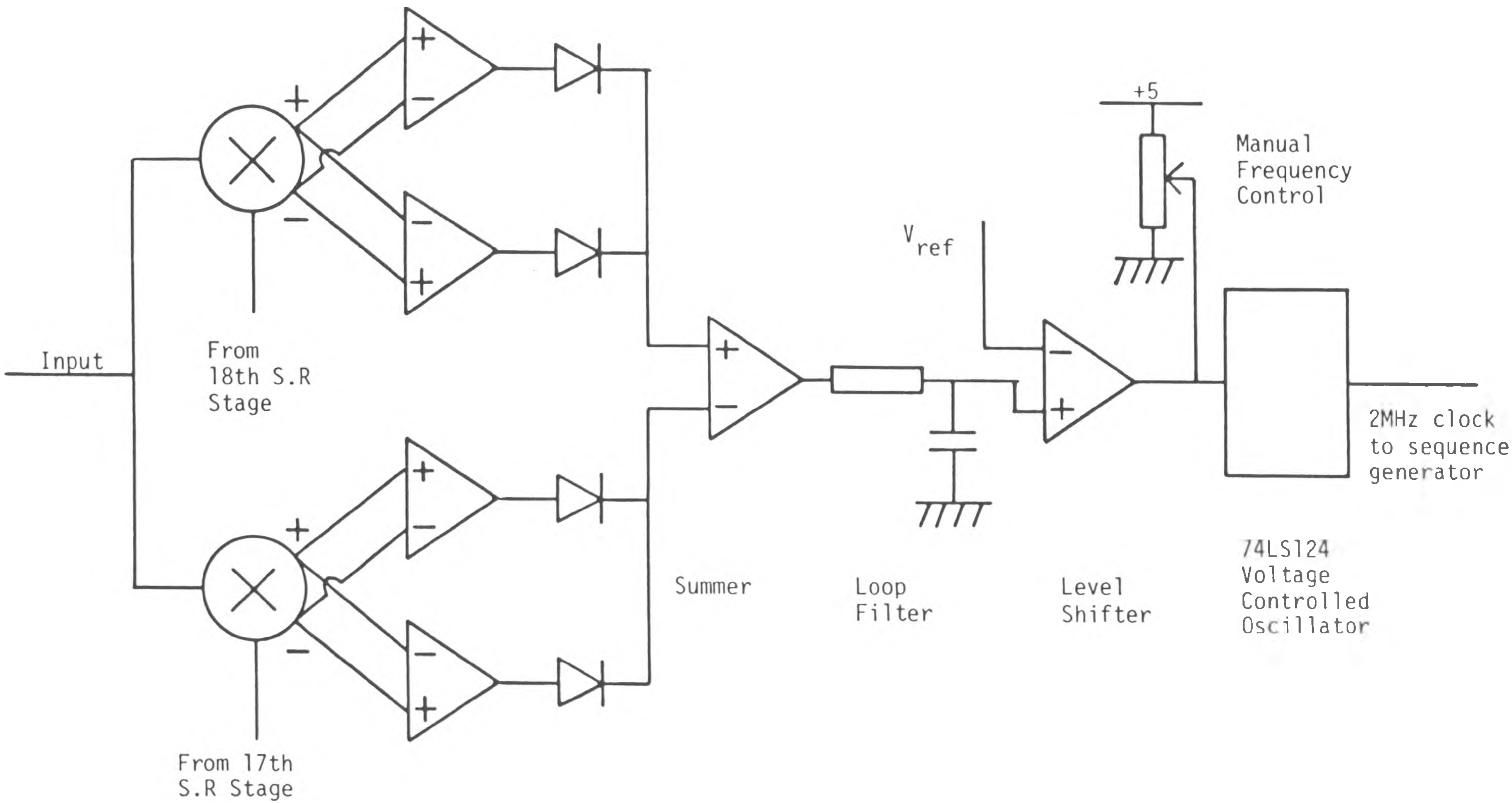


Fig. 6.5 Block Diagram of Delay Lock Tracking Loop

connected as differential amplifiers. Their outputs are connected to diode rectifiers arranged to form two full wave rectification circuits. The rectifier outputs are subtracted by a 741 op amp to form the raw error signal.

This is filtered by a simple RC type low pass filter having a 1.2 kHz 3dB bandwidth. This value was chosen as a compromise between rapid tracking and acquisition and low jitter. After level shifting with a 741 op amp the error signal is fed to the control input of the voltage controlled oscillator. This is a 74 LS 124 TTL voltage controlled oscillator having a free running frequency of 2 MHz. The clock signal drives the sequence generator and also a 7474 D flip flop controlled by the 17th stage of the shift register. The latter circuit provides a spreading sequence of the correct phase to operate the despreader.

A manual control of the voltage controlled oscillator frequency provides a small frequency offset for acquisition purposes.

6.3.5.2 Carrier Reconstruction Circuitry

The circuit used to obtain a 30 MHz coherent carrier is as given in Fig 6.4. The first mixer accepts signals at 11.16 MHz from the i.f amplifier local oscillator and 460 kHz from the carrier regeneration loop. The output at 10.7 MHz is fed to the second mixer along with a 40.7 MHz signal from the receiver main local oscillator. The resulting output at 30 MHz drives a Minicircuits SRA-1 double balanced mixer, to demodulate the incoming signal to baseband.

Both mixers and their associated output filters have almost identical circuits, differing only in the filter components. Each mixer is constructed from an MC 1496 integrated circuit double balanced modulator. This is coupled to a single stage transistor amplifier having single tuned circuits on input and output to remove unwanted mixer products.

6.3.5.3 Delay Lock Loop Problems

The delay lock tracking loop described failed to operate satisfactorily when connected in the receiver. With no information modulation on the spreading sequence the loop tracked well after initial acquisition by manual adjustment of the voltage controlled oscillator frequency. However when the sequence was modulated by information the loop would not track.

Investigation of the discriminator characteristic showed the central section over which the loop normally operates to consist of two parts, instead of one. Thus the discriminator curve was dependent on the sign of the incoming sequence, as shown in Fig 6.6, the shift being around 0.05 of a chip. This shift in the delay discriminator characterstic was sufficient to prevent the voltage control oscillator from tracking. Each time the information bit changed the transient small error forced the oscillator out of lock and this was maintained over the data bit until the next transition.

The reason for the shift was not immediately apparent. Considerable investigation of the circuit showed some imbalance in the integrated circuit mixers and in the rectifier diodes. Despite careful selection and balancing of the mixers some imbalance remained here. Hence it was concluded that a complete rebuild of the circuit would be required using better balanced components.

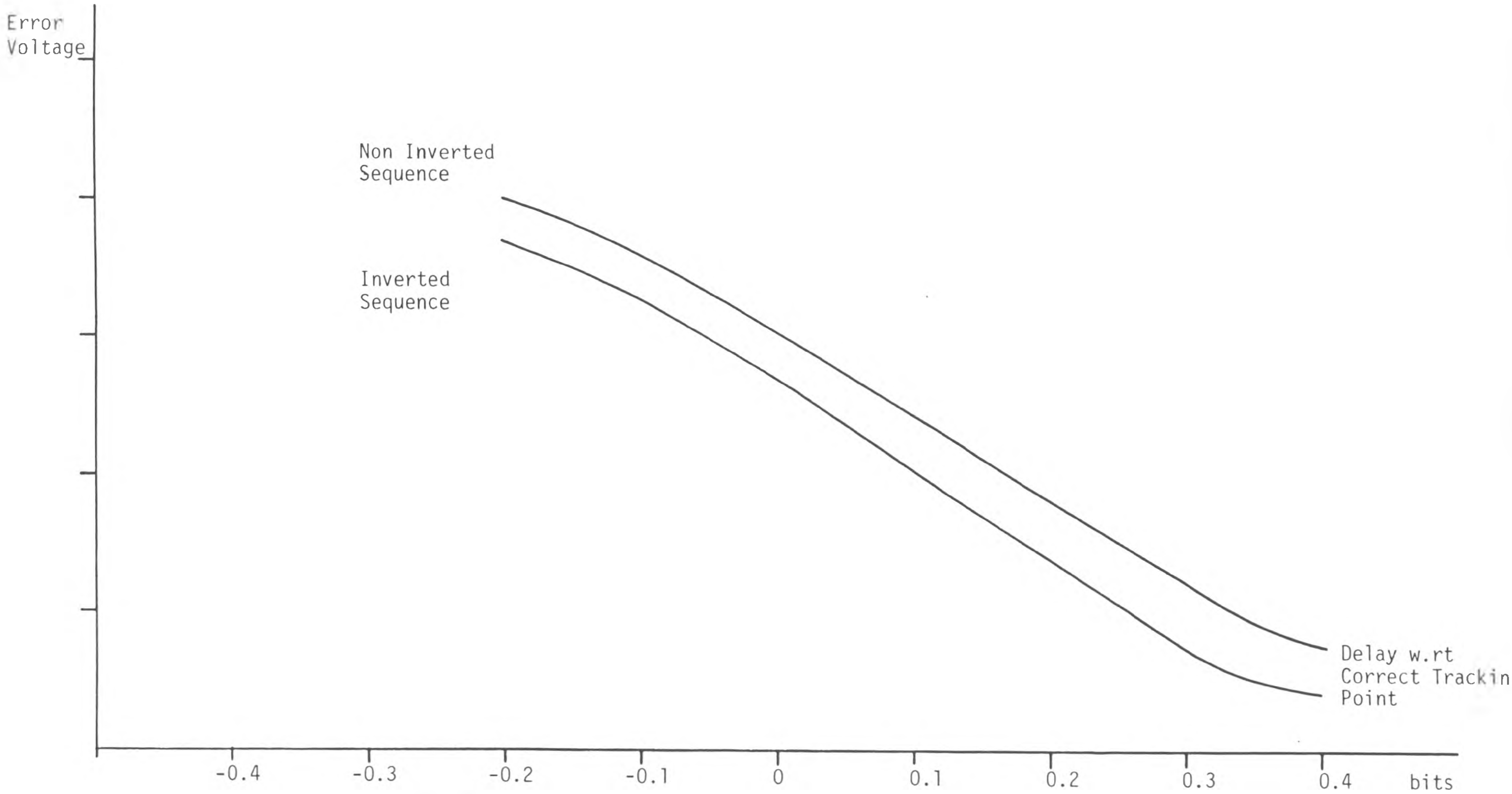


Fig 6.6 Delay Lock Loop Characteristic

In retrospect it is possible that the problem was compounded by an incorrect choice of loop filter bandwidth. Had this been higher then the loop transient response would have been much improved, possibly leading to acceptable tracking under the existing conditions. Doubtless this could have been confirmed by simple experiment.

6.4 Ancillary Circuits

The receiver requires two local oscillators. One at 11.16 MHz provides down conversion from 10.7 MHz to 460 kHz in the i.f. amplifier. The second of 40.7 MHz provides the drive to the heterodyne despreader. Both oscillators use similar circuits comprising a crystal controlled transistor oscillator coupled to a tuned buffer/driver stage.

CHAPTER 7

Measurements

CHAPTER 7

Measurements

The purpose of this section is to give details of the measurements made on the direct sequence system described previously. The main interest is in the performance of the complete system when operated with various channel degradations. However some results are also given of the measured performance of the sub assemblies forming the system.

7.1 Equipment Performance

The measurement of equipment performance was straightforward, the pertinant results being given in Table 7.i. Consideration of the performance of the i.f amplifier shows that there is perhaps scope for improvement here. Certainly some small improvement in selectivity might have been obtained, though this is not significant. The rest of the equipment operated satisfactorily, except for the synchronisation circuitry as discussed previously.

7.2 System Performance

Before discussing the overall performance of the system when subjected to various channel degrodations a number of points should be clarified. Of considerable importance is the method of system synchronisation. During the measurements this was achieved by feeding part of the signal from the transmitter sequence generator directly to the despreader at the receiver. The length of this link was adjusted to provide a time delay equivalent to that of the direct r.f path. This method of synchronisation ensured that at all times the transmitter and receiver were in perfect synchronisation. Whilst this technique is adequate here it is obviously unsuitable for larger scale tests.

The majority of measurements were made using a spectrum analyser, which could be connected to various points in the system. For qualitative measurements, such as the taking of representative spectra, unmodulated s.c.p.d.m. signals were sent over the link. However to facilitate quantitative measurements the transmitted signal consisted of a carrier modulated by the spreading sequence only. Thus at the output of the i.f amplifier in the receiver only an unmodulated carrier was present, save for any interference. Hence the carrier level and noise spectral density could be easily measured and converted into an equivalent carrier to noise ratio. As the system is considered linear this action is perfectly valid.

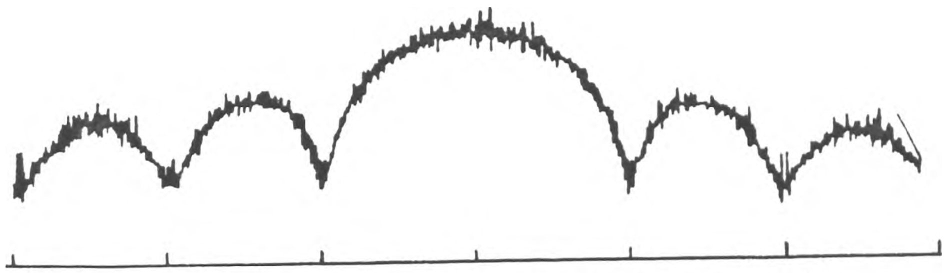
To preserve impedance matching and prevent variation of signal levels channel degradations were introduced using matched tees. Furthermore an adjustable attenuator at the receiver input was used to set the signal to a constant level. This provided compensation for the varying path losses and ensured the receiver was operating under known conditions.

For all measurements the spreading sequence was a maximal length type of length 2^{15}-1 bits.

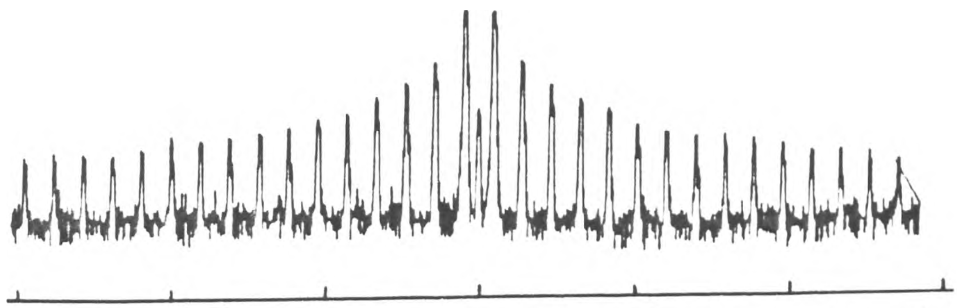
7.2.1 Representative Spectra

Fig 7.1 shows the spectrum of signals at various points in the system with no channel degradations. These are given to illustrate the type of spectra which might be typically seen

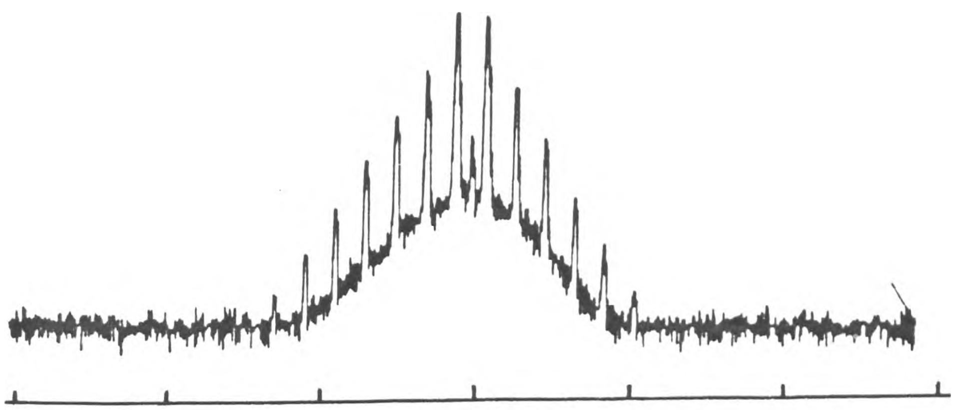
Fig. 7.1
Typical Spectra Direct Sequence
Spread Spectrum System



a) R.F Input, fo = 30MHz, Horizontal 1div = 2MHz



b) Despreader Output, fo = 10.7MHz, Horizontal 1div = 40kHz



c) I.F output, fo = 460kHz, Horizontal 1div = 40kHz in this type of equipment. The many sidelebes in the transmitted spectrum are due to the absence of r.f filtering in the equipment.

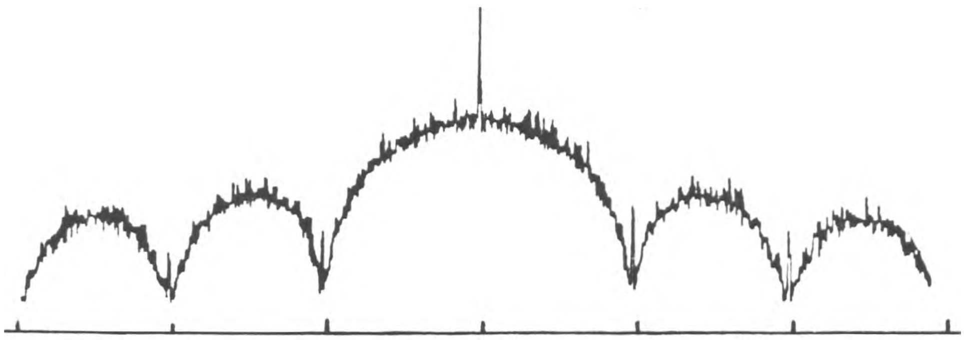
The long sequence length used ensured that the transmitted spectrum was continuous, the individual lines only being resolvable at the highest resolution of the spectrum analyser. Furthermore, no detectable change in the transmitted spectrum was observed when s.c.p.d.m. signals were sent over the link.

7.2.2 Tone Interference

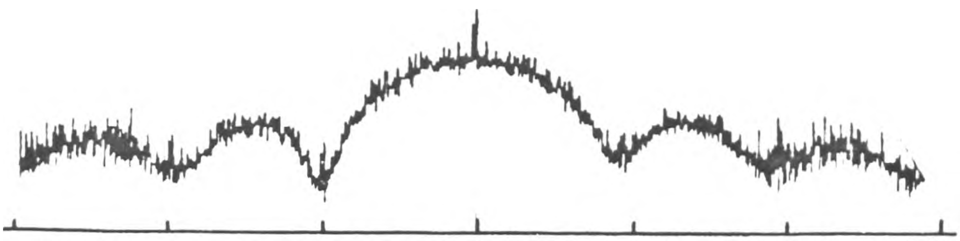
The first type of channel degradation investigated was tone interference. A single unmodulated tone, nominally on the system carrier frequency, was fed into the receiver along with the wanted signal. By varying the tone level a wide range of signal to interference ratios could be produced at the input to the receiver.

Fig 7.2 shows typical spectra at various points in the system. The spectrum at the output of the despreader shows that the interfering tone has been spread in frequency, as expected. Furthermore it has been endowed with the properties of the spreading sequence. Using the spectrum analyser the carrier level and (pseudo) noise spectral density were measured. As the noise density is essentially flat near the centre frequency the noise power in a small bandwidth at this frequency is easily calculated. Thus the signal to noise ratio in the i.f bandwidth can be evaluated. Fig 7.3 shows the output signal to noise ratio plotted against the input signal to interference ratio. Also shown is the theoretical result obtained from

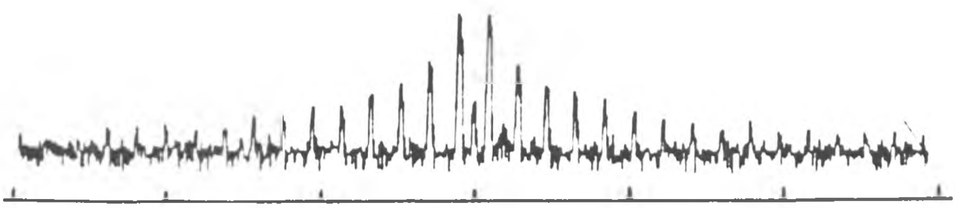
Fig. 7.2
Spectra at various points in the system
for Tone Interference



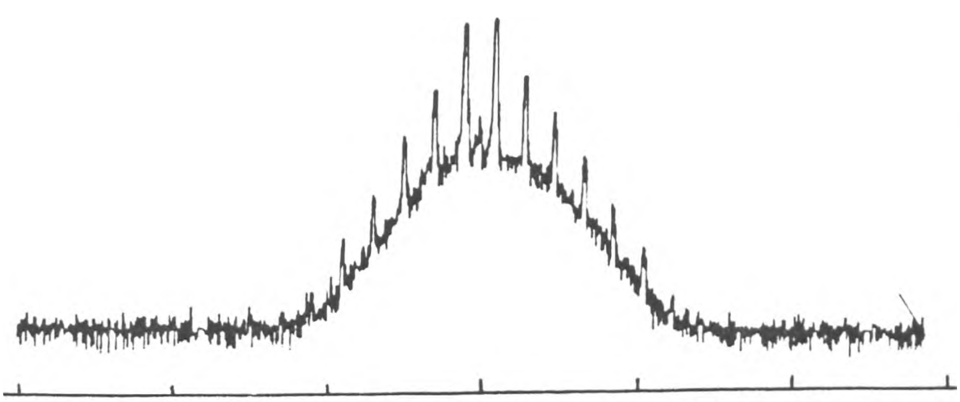
a) R.F Input, fo = 30MHz, Horizontal 1div = 2MHz



b) Despreader Output, fo = 10.7MHz, Horizontal 1div = 2MHz



c) Despreader Output, fo = 10.7MHz, Horizontal 1div = 40kHz



d) I.F Output, fo = 460kHz, Horizontal 1div = 40kHz

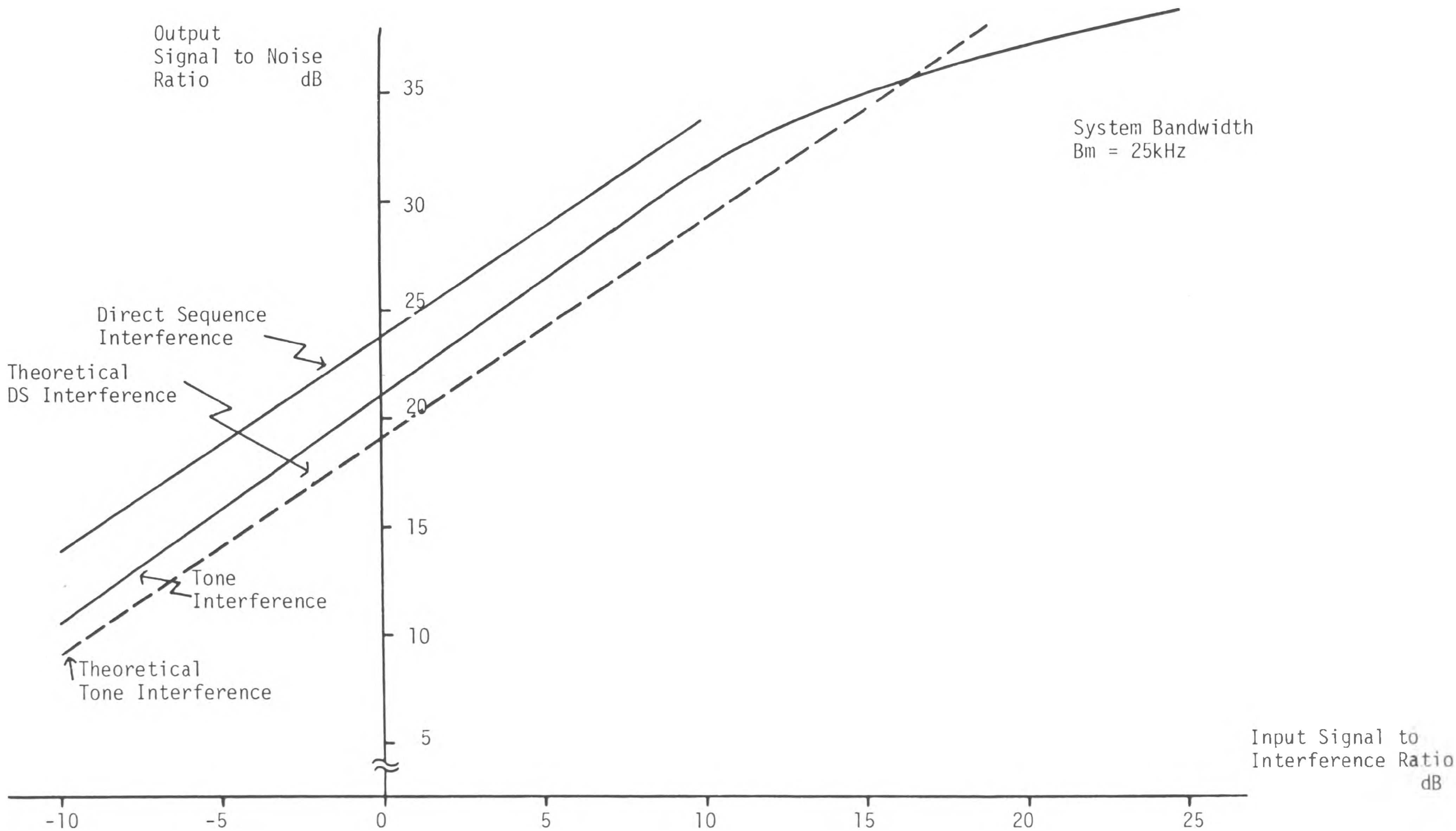


Fig. 7.3 Graph of Output Signal to Noise Ratio against Input Signal to Interference Ratio for Direct Sequence System

equation (4.12). Both curves assume a 25 kHz i.f bandwidth, though the actual value is not critical as it will identically effect both results.

Interestingly the measured result is slightly better than the theoretical one by around 1 to 2 dB. The reasons for this are not apparent, though this may be due to inaccuracies in measurement. At high input signal to interference ratios the measured curve falls below the theoretical and rises less steeply. This is almost certainly due to receiver noise which obviously only becomes noticable as the interference related noise decreases. Nevertheless the two curves are in close agreement, showing the system to provide the expected 20 dB improvement in signal to interference ratio. Whilst no measurements were made, the effect of altering the frequency of the interfering tone was observed. As predicted the noise was greatest when the tone was on the direct sequence carrier frequency and decreased as moved from this .

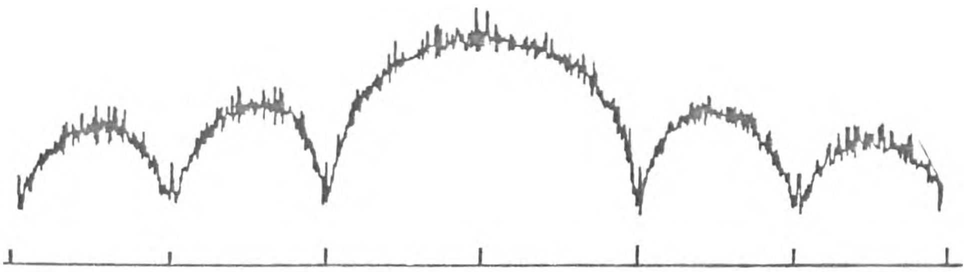
7.2.3 Spread Spectrum Interference

The second type of interference investigated was that from another spread spectrum source. This was provided by a 2^{10}-1 bit maximal length sequence generator clocked at 2 Mb/s, the output phase modulating a nominal 30 MHz carrier. Operation of this second source was asynchronous of the main system.

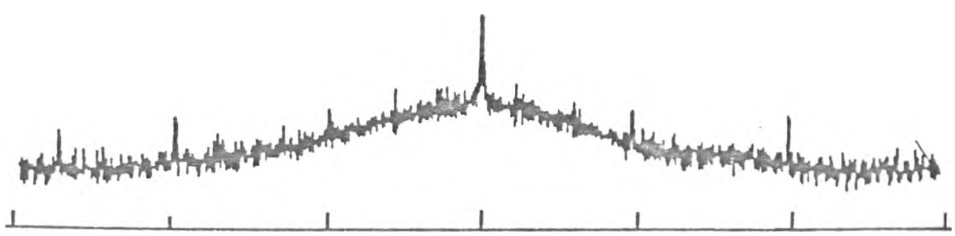
The resultant spectra before and after despreading are shown in Fig 7.4. At the output of the despreader there is an increase in noise compared with the situation of no interference. It is therefore concluded that this extra noise arises from the

Fig. 7.4

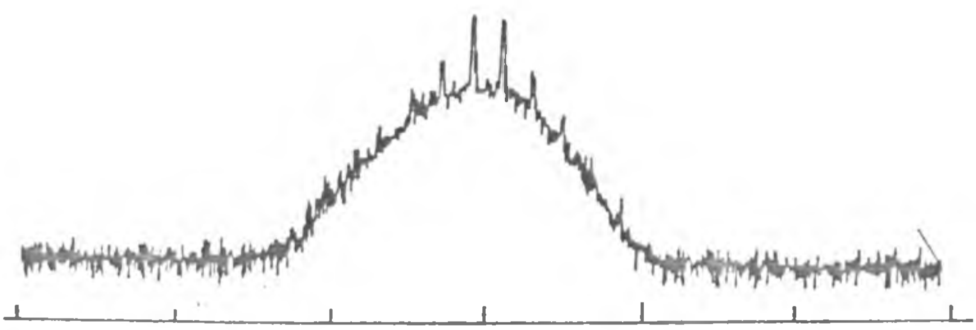
Spectra at Various Points in the System for one Spread Spectrum Interferer



a) R.F. Input, fo = 30 MHz, Hor 1 div = 2 MHz



b) Despreader Output, fo = 10.7 MHz, Hor 1 div = 2 MHz



c) I.F Output, fo = 460 kHz, Hor 1 div = 40 KHz unwanted signal. However the overall spectrum of the noise was difficult to determine and no firm conclusions can be drawn about the changes to the interference. As previously the carrier level and noise spectral density were measured and converted to a signal to noise ratio in a specified bandwidth. The results are shown in Fig 7.3 to allow a comparison to be made with those obtained for tone interference.

The improvement in signal to interference ratio is better than expected on the basis of the interference being reduced by the system process gain. Indeed the output carrier to noise ratio is around 3dB better than that measured for equivalent tone interference.

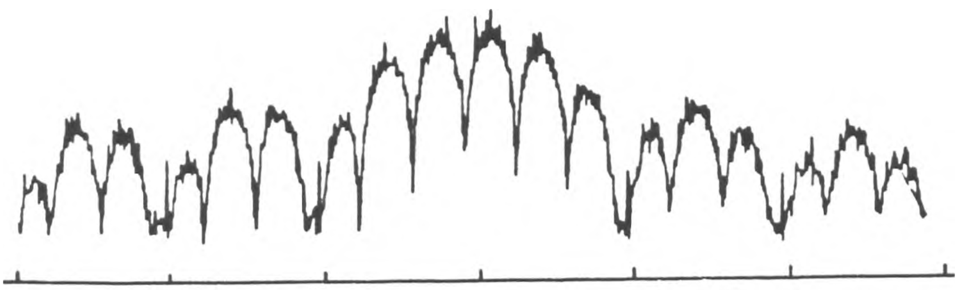
7.2.4 Multipath Interference

The final type of interference investigated was that caused by multipath propagation. For ease of implementation and subsequent analysis only a simple two path model was considered. Matched tees were used to split the signal from the transmitter into two paths and recombine them at the receiver. The shortest path was a direct path and consisted of an attenuator to set the ratio of direct to delayed path signal levels. The delayed path was formed from lengths of coaxial cable to provide delays of .53 and 1.5 μS.

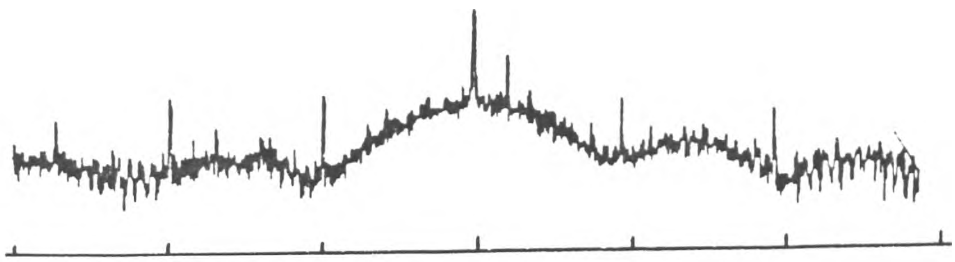
Representative spectra of input and output signals are given in Fig 7.5. The received signal is badly distorted by the channel which places several peaks and nulls in the transmitted spectrum. At the output of the despreader the delayed path signal is present as (pseudo) noise with the familiar $(\sin x/x)^2$

Fig. 7.5

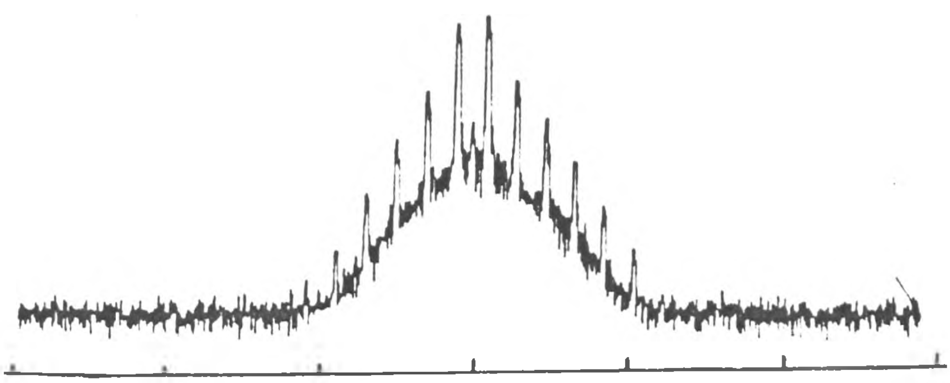
Spectra at Various Points in System
for Simple Multipath Channel



a) R.F Input, fo = 30 MHz, Delay = 1.5μs
Horizontal 1div = 2MHz



b) Despreader Output, fo = 10·7 MHz, Horizontal 1div = 2MHz



c) I.F Output, fo = 460kHz, Horizontal 1div = 40kHz

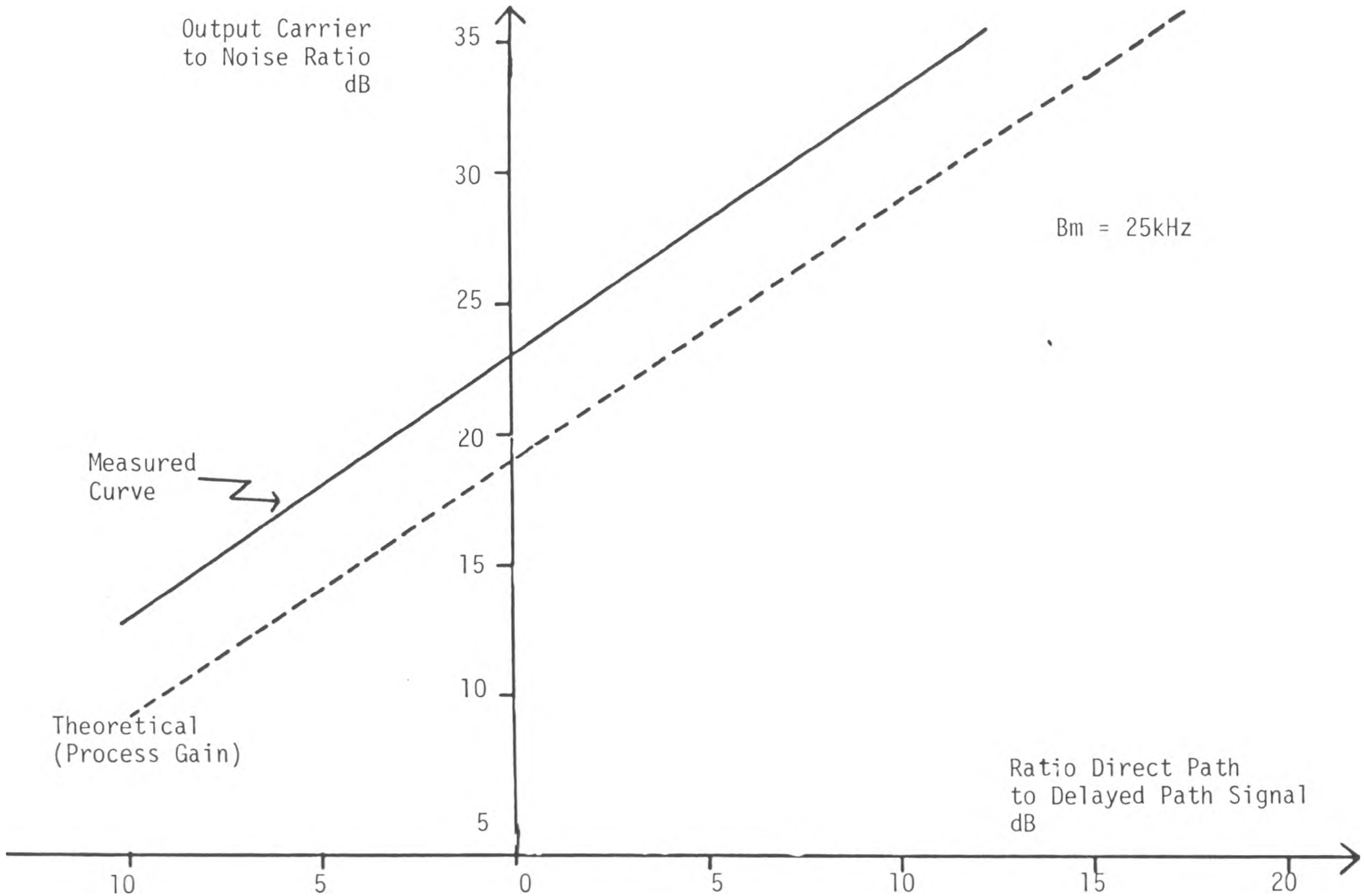


Fig 7.6 Graph Showing Output Carrier to Noise Ratio for 2 Path Propagation of 1.5μs and 0.53μs Excess Delays

type power spectrum. This and the subsequent spectra were identical regardless of the excess path length. Thus the results are as expected for this type of channel degradation. Again the carrier to noise ratio was evaluated and this is plotted in Fig 7.6 against the ratio of direct path to delayed path signal level.

The measured carrier to noise ratio at the output of the despreader was identical for both excess path delays. Furthermore the measured result is around 1 dB better than expected on the basis of the delayed path signal being reduced by the system process gain.

It should be noted that both excess path delays exceeded the spreading sequence chip period. In accordance with section 5.4 different results would have been expected for excess path delays of less than a sequence chip period. Unfortunately no measurements were performed for path lengths of less than 0.5 μS due to an oversight by the author. This is a regrettable omission as the results of such measurements would have aided the understanding of system performance.

7.3 Qualitative Observations of Performance

Several qualitative observations of the performance of the s.c.p.d.m speech modulation scheme were made and these are recorded here.

A simple measurement with a power supply and oscilloscope showed the p.d.m modulator to be non linear. Thus the pulse duration was not exactly proportional to the signal amplitude at the sampling instant. The effect was somewhat similar to a form of compression and compared to a small signal was obviously greatest for large amplitudes. The problem was in the p.d.m modulator itself, the timer integrated circuit not being specifically designed for this application had a non-linearity on its control input. The only solution would have been a complete redesign of the p.d.m modulator, which was not considered worthwhile. The resultant distortion did not appear to be objectionable and anyhow the compression was probably beneficial.

At the receiving end of the link aural observation of the system performance was made, particularly as to the preferred method of baseband recovery. The conversion to p.a.m and the filtering method of audio recovery both gave similar performance and there was little to decide between them. However if asked to state a preference most listeners liked the audio from the filtering method slightly better.

The use of a speech compressor at the receiver gave an improvement in speech intelligability even though no expansion was provided at the receiver. The reduction even further in dynamic range by the compressor allowed a greater depth of modulation without detracting from the overall quality.

Addendum

It is interesting that for all the measurements the measured values exceed the theoretical ones by a few decibels. Furthermore for any given set of measurements the ratio of measured to theoretical results is substantially constant. As a check all measurements were repeated several times, with no appreciable variation in results.

One of the main problems of performing measurements of this type is that of accurately measuring the noise. The spectrum analyser used was claimed to give accurate noise level readings, once the noise bandwidth was accounted for. However in many spectrum analysers the displayed noise is below its true value and a correction factor of a few decibels must be applied to obtain the true level. In retrospect it is difficult to ascertain if this applies here, nevertheless there were more than likely inaccuracies in the noise measurements. Whilst the possibility of errors in the signal measurement can not be discounted the errors here should be small compared to those for the noise level measurements.

Table 7.i

Equipment Performance

I. F Amplifier

Gain = 40dB (f = 10.7 MHz)

Bandwidth		Image Rejection
		> 40 dB
- 3dB	26 kHz	
- 6dB	37 kHz	
- 60dB	153 kHz	

Synchronous Demodulator

Minimum Input = 5mV (To track signal)

Output Filter Gain 12 dB

Output Filter 3 dB Bandwidth = 10 kHz

S.C.P.D.M. Demodulator

Low Pass Filter

Bandwidth 3 dB = 3 kHz

Gain = 20 dB

Response at 4kHz > 20 dB down

CONCLUSIONS

Conclusion

This thesis has examined many aspects of applying direct sequence spread spectrum techniques to land mobile radio systems. Most of the topics have been covered in isolation and it remains to bring all the major points together.

The main trend to emerge from the analyses is the low number of allowable direct sequence users per unit bandwidth. The early analysis in chapter 3 showed that a large area coverage scheme with no channel degredations permitted the greatest number of simultaneous users. The inclusion in the analysis of channel degredations due to propagation effects reduces this figure, whilst incorporation into a cellular scheme reduces it even further. In calculating the number of allowable simultaneous system users allowance must be made for implementation loss and interference, both of which lower the spectral utilisation. Finally account must be taken of the reduction in spectral utilisation caused by sequence cross-correlations. Overall the various values of spectral utilisation obtained are upper bounds which will not be achieved by practical systems.

As an example consider a simple cellular scheme in which 4 frequency bands are used. Assume the log-normal shadowing has a 6 dB standard deviation and that a 10% outage time is acceptable. Using pulse duration modulation in a 25 kHz bandwidth fig 5.4 shows that a minimum 5 dB despreader signal to noise ratio is required for an audio signal to noise ratio of 10 dB. From table 5.1 this corresponds to a maximum of 2.77 users/MHz. Already the direct sequence scheme looks poor compared to the 4.44 users/MHz for a 25 kHz f.m. scheme with 14 dB protection ratio operating under similar conditions. When for the direct sequence scheme account is taken of implementation loss, multipath propagation, sequence cross-correlations and interference the spectral utilisation will fall below the present low value. Allowing a generous underestimation for the above effects of 3 dB would reduce the direct sequence spectral utilisation to 1.38 Users/MHz. Even this represents an upper limit unlikely to be achieved in practice.

A problem in the assessment of direct sequence spectral utilisation is the extreme sensitivity of such systems to the message. By this is meant the trade off between message output signal to noise ratio, bandwidth and number of allowable simultaneous system users. Hence if message modulation techniques having high figures of merit and low thresholds can be developed then an increase in direct sequence spectral utilisation is possible. Consider the same situation as before except that at the despreader output a 5 dB signal to noise ratio is required in a 10 kHz bandwidth. This immediately leads to a maximum 6.89 users/MHz before implementation loss etc. is taken into account. Making the same estimation as previously for these latter effects results in a direct sequence spectral utilisation of 3.44 users/MHz. At this stage direct sequence techniques look comparably attractive with 25 kHz f.m. schemes, though still poor compared to 12.5 kHz f.m. schemes. Also at this sort of spectral occupancy the spread spectrum audio output signal to noise ratio will be just acceptable. However there will be improvement here as the number of simultaneous system users decrease.

In making comparisons with conventional narrowband modulation methods only 25 and 12.5 kHz channel spaced f.m. systems were considered. There is however a considerable amount of information[10,11,15] on s.s.b. systems which shows these to have higher spectral utilisations than f.m. schemes. These figures were not used in the comparison as most authors take f.m. systems as the reference for comparison. It is no surprise therefore that the spectral utilisation of direct sequence schemes is well below that of s.s.b. systems

When viewed as stand alone systems direct sequence spread spectrum techniques do not appear useful. However when the idea of bandsharing is considered the situation is slightly more encouraging. Direct sequence systems are probably the only, modulation technique which can bandshare successfully at small reuse distances. However the number of allowable simultaneous direct sequence users has to be reduced slightly to cope with the extra interference from bandsharing. Whilst the idea required careful consideration bandsharing by direct sequence and conventional narrowband modulation systems could provide a small but useful increase in overall spectral occupancy.

The construction of a transmitter and receiver proved valuable and showed this not be a task undertaken lightly. Despite the problems with the synchronising circuitry the equipment performed as expected. Despite being somewhat optimistic the results show reasonable agreement with those expected, supporting the use of the process gain as a measure of the systems performance. Following on from here it would be useful to adapt the transmitter and receiver for operation over a radio transmitting channel. This would permit realistic measurements of performance to be made over a representative mobile radio channel. Further assessment of multiple user performance could be obtained by adding several transmitters to the system, having different but related spreading sequences.

Almost the entire analysis of spectrum utilisation for direct sequence schemes has been based on the effects of unwanted signals being reduced by the system process gain. In chapter 3 there was mention of the importance of using spreading sequences with low cross-correlations to obtain greatest spectral utilisation. The reason for doing this was explained using a qualitative approach. However no relationship between sequence cross-correlations, process gain and spectral occupancy was established, except to mention that sequence cross-correlations effect the implementation loss. Despite papers by Judge and others there appears to be a lack of information in this area. Consideration of the topic provides some interesting problems, particularly when orthogonal or nearly orthogonal spreading sequences are used. Whilst present multiplexing theory is well established the process gain agrument is perhaps somewhat weak. Consequently the topic would provide a valuable area for further investigation.

Another topic worthier of further investigation is multiple user performance under conditions of multipath propagation. This is of considerable interest for practical land mobile radio systems, yet did not receive a lot of attention in the thesis. This was mainly due to the difficulty of obtaining useful wideband channel models, also the analytical complexity. Most of the latter problem could be solved by computer simulation, assuming accurate models of the system and channel can be obtained. Indeed a good computer simulation would be useful for further study of many aspects of direct sequence techniques.

As mentioned previously the overall performance of direct sequence systems is particularly sensative to the message modulation used. Consequently work on this topic would be of great benefit if message modulation techniques having high figures of merit and low thresholds could be devised. Perhaps vocoding or linear predictive encoding techniques may be of interest here. As stated at the beginning of the thesis spreading sequence synchronisation is a valuable area for further investigation. Both the acquisition and tracking processes require study, especially the rapid acquisition of long sequences.

This thesis has provided an investigation into the possible application of direct sequence spread spectrum techniques to land mobile radio systems. Due to the low spectral utilisation of these techniques they are unlikely to find widespread application. However they have obvious use in hazardous radio environments where reliable communications are necessary. Furthermore the possiblities of bandsharing and the unrestricted multiple user facility show that the value of direct sequence spread spectrum techniques should not be underestimated.

ACKNOWLEDGEMENTS

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APPENDIX A

Near Far Problem

APPENDIX A

Near Far Problem

Consider the situation shown in Fig A1 where a simple circular area coverage scheme is served by a central base station. Distributed in some manner throughout the coverage area are several simultaneously active direct sequence spread spectrum mobile transmitters all operating on the same centre frequency. It is required to evaluate at the base station the ratio of received power from a wanted transmitter. In order to simplify the problem the following assumptions are made :

All antennae are omnidirectional

All transmitters radiate identical power

Plane earth propagation

Propagation effects and other interference neglected

Hence for the wanted mobile transmitter located distance r_w from the base station the received power P_w is :

$$P_w = \frac{K}{r_w^{\,4}} \qquad \text{(A 1)}$$

Now let the distribution of active mobiles be a function of distance r from the base station. To describe this we use the mobile density u(r), which gives the average number of active mobiles per unit area.

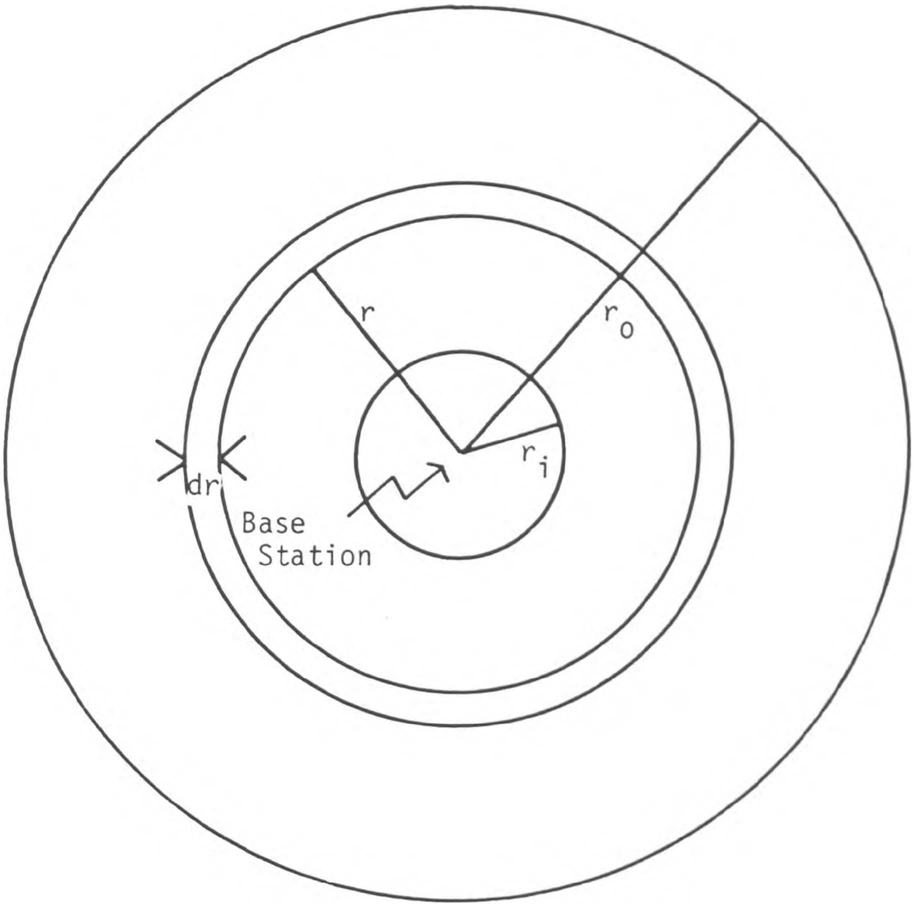

Fig. A.1 Layout for 'Near Far' problem Calculations

To calculate the total received power at the base station the procedure is as follows. The total received power from an incremental annular ring of width dr will be calculated and this integrated to give the total power.

For an annular ring of incremental width dr situated a distance r from the base station the area is :

$$= 2\Pi r \ dr \qquad \text{(A 2)}$$

assuming second order terms involving small quantities can be neglected.

Thus the total power received at the base station from this ring is:

$$= \frac{2\Pi K}{r^4} \ r\, u(r) \ dr \qquad \text{(A 3)}$$

Hence the total power P_T received at the base station from all transmitters within an area bounded by the outer perimeter r_o of the coverage area and an inner guard radius r_i is :

$$P_T = 2\Pi K \int_{r_i}^{r_o} \frac{u(r)}{r^3} \ dr \qquad \text{(A 4)}$$

An inner guard ring is generally necessary mathematically, for with u(r) represented by certain types of function a value r_i = o would make the integral undefined or limitless. Physically it represents the closest distance which a mobile may approach the base station. In reality a mobile too close to a receiver would de-sensatise it unless it had a very large dynamic range.

We can now evaluate the ratio of wanted signal power to total interference power at the base station as :

$$\frac{P_W}{P_T} = \frac{1}{r_W^{\,4}} \cdot \frac{1}{2\Pi \int_{r_i}^{r_o} \frac{u(r)}{r^3}\, dr} \qquad \text{(A 5)}$$

This considers the wanted mobile as extra to those existing in the system, though for all purposes the difference is negligably small.

Consider the simple situation where the mobiles are uniformly distributed over the coverage area with a density of U per unit area.

Hence the total received power at the base station is:

$$P_T = 2\Pi K \int_{r_i}^{r_o} \frac{U}{r^3}\, dr \qquad \text{(A 6)}$$

$$= U \Pi K \left[\frac{r_o^{\,2} - r_i^{\,2}}{r_o^{\,2}\, r_i^{\,2}} \right] \qquad \text{(A 7)}$$

and if the mobile is on the boundary of the coverage area ($r_W = r_o$) then the ratio of wanted signal to interference is :

$$\frac{P_W}{P_T} = \frac{1}{r_o^{\,2} \Pi\, U \left[\frac{r_o}{r_i}^{\,2} - 1 \right]} \qquad \text{(A 8)}$$

Obviously for large values of outer to inner radii and large values of U the signal to interference is small.

To evaluate the maximum allowable number of simultaneous users under these circumstances a value for the mobile density can be found for the minimum acceptable signal to interference ratio :

$$U = \frac{1}{r_o^2 \Pi \left[\frac{r_o^2}{r_i^2} - 1\right]\left(\frac{P_w}{P_T}\right)_{min}} \qquad (A\ 9)$$

Now the number of users M in the coverage area is :

$$M = U \Pi (r_o^2 - r_i^2) \qquad (A\ 10)$$

Or substituting for U from A 9 :

$$M = \frac{1}{\left(\frac{r_o^2}{r_i^2}\right)\left(\frac{P_w}{P_T}\right)_{min}} \qquad (A\ 11)$$

This can be compared to the case where power control is used for which the maximum number of simultaneous users M_c is given by :

$$M_c = \frac{1}{\left(\frac{P_w}{P_T}\right)_{min}} \qquad (A\ 12)$$

Obviously the maximum number of allowable simultaneous users is far less without power control.

As an example of the numbers involved assume :

$$\left(\frac{P_w}{P_T}\right)_{min} = 0.01 \qquad r_o = 10 \text{ km} \qquad r_i = 1 \text{ km}$$

Hence $M_c = \frac{1}{0.01}$

$= \underline{\underline{100}}$

Whilst $M = \frac{1}{\frac{10^2}{1} \times 0.01}$

$= \underline{\underline{1}}$

APPENDIX B

Equipment Circuit Diagrams

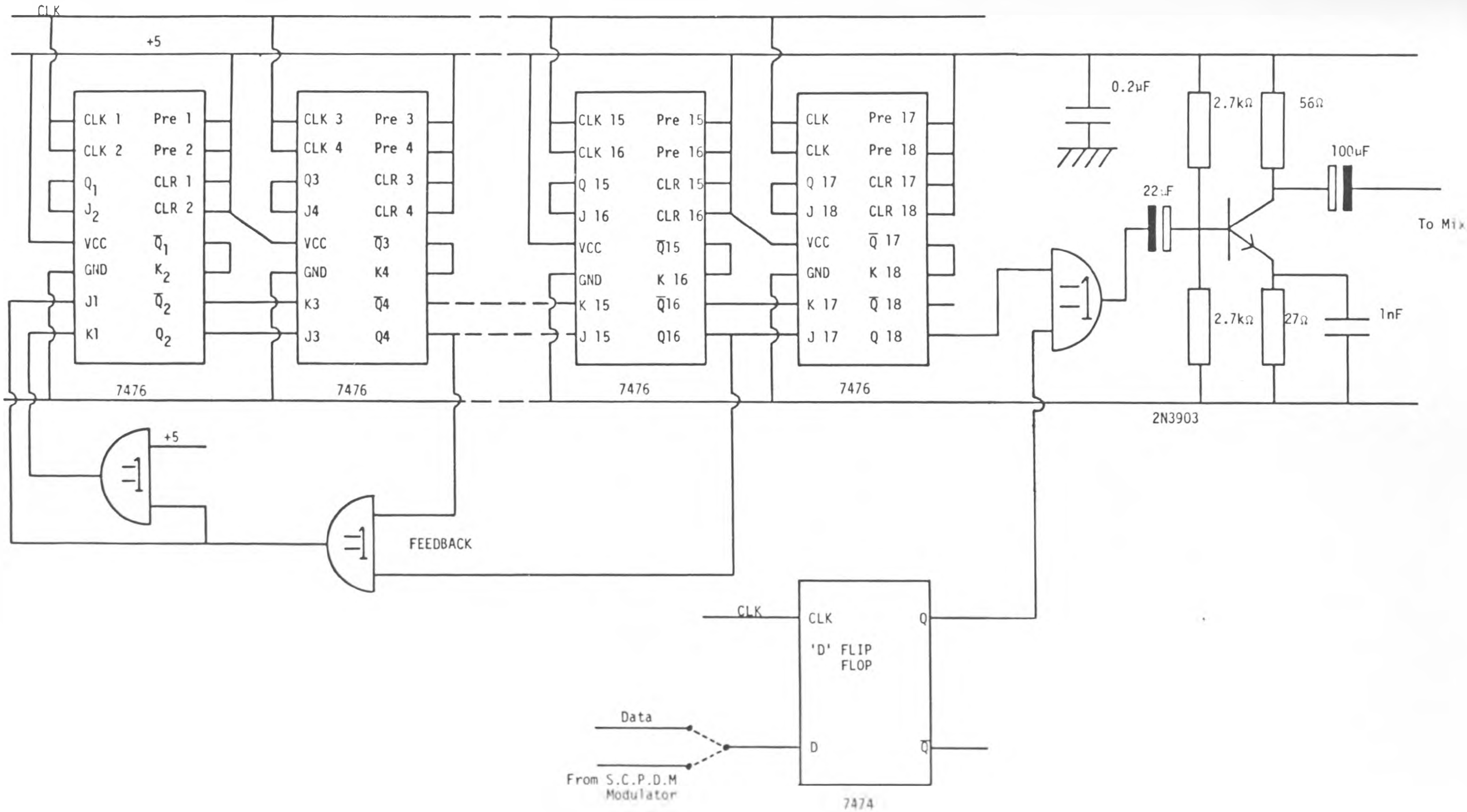

Fig. B.1 Diagram of Sequence Generator and Message Modulator

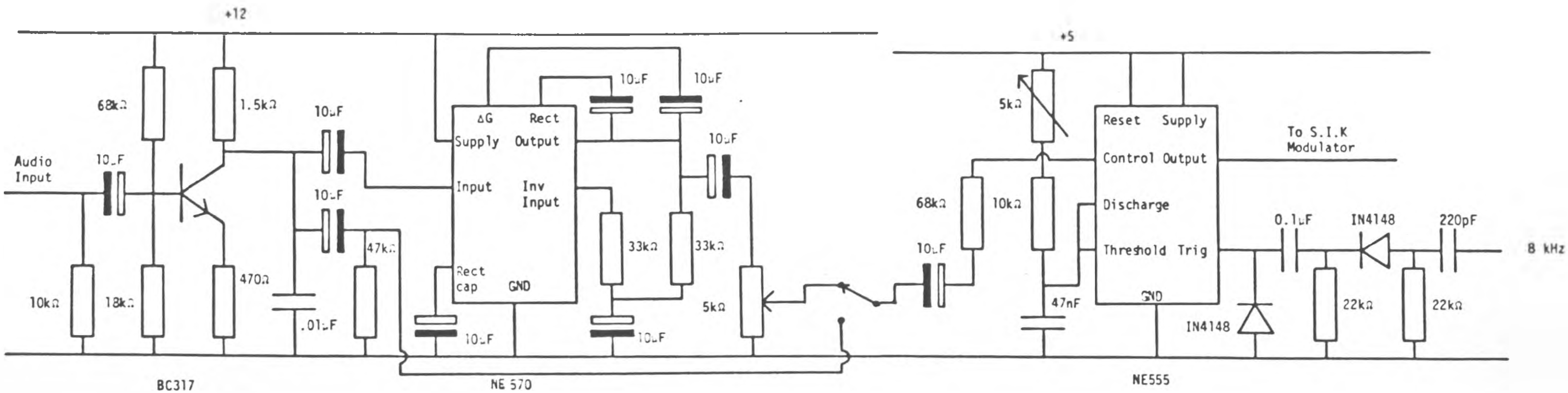

Fig. B.2 Suppressed Clock Pulse Duration Modulation Modulator

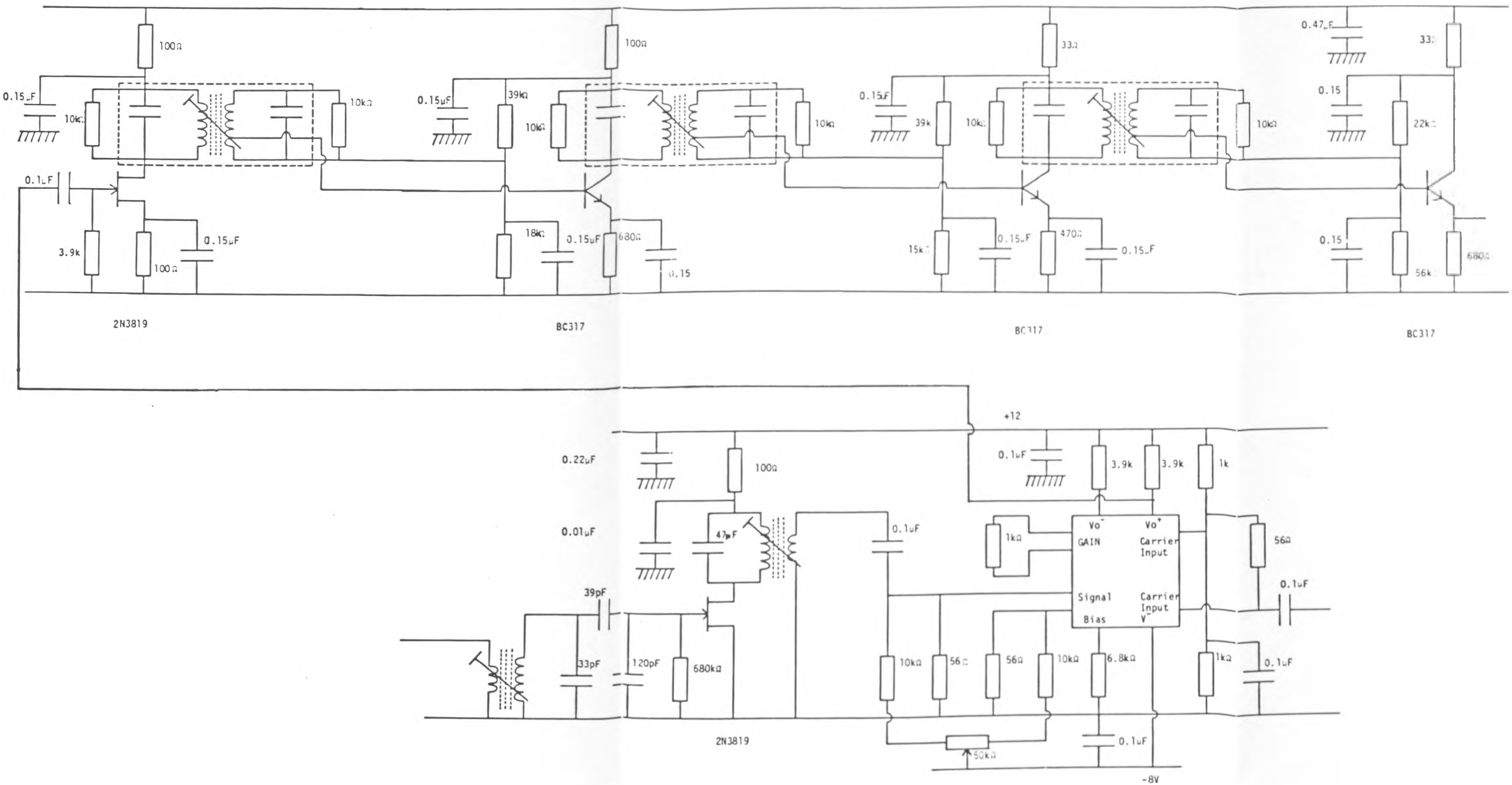

Fig. B.3 Circuit Diagram of

I.f. Amplifier

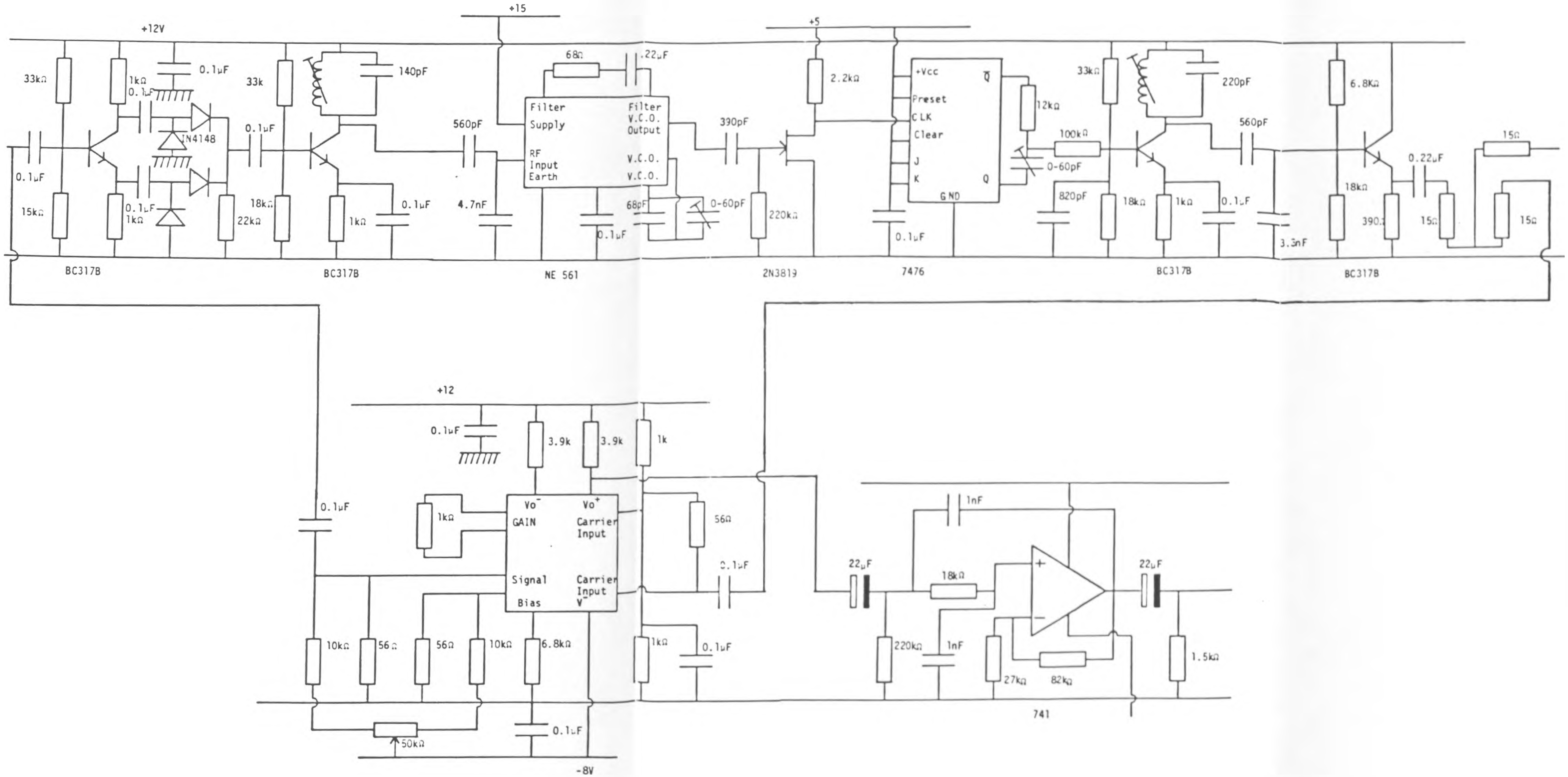

Fig. B.4 Circuit Diagram of Synchronous Demodulator

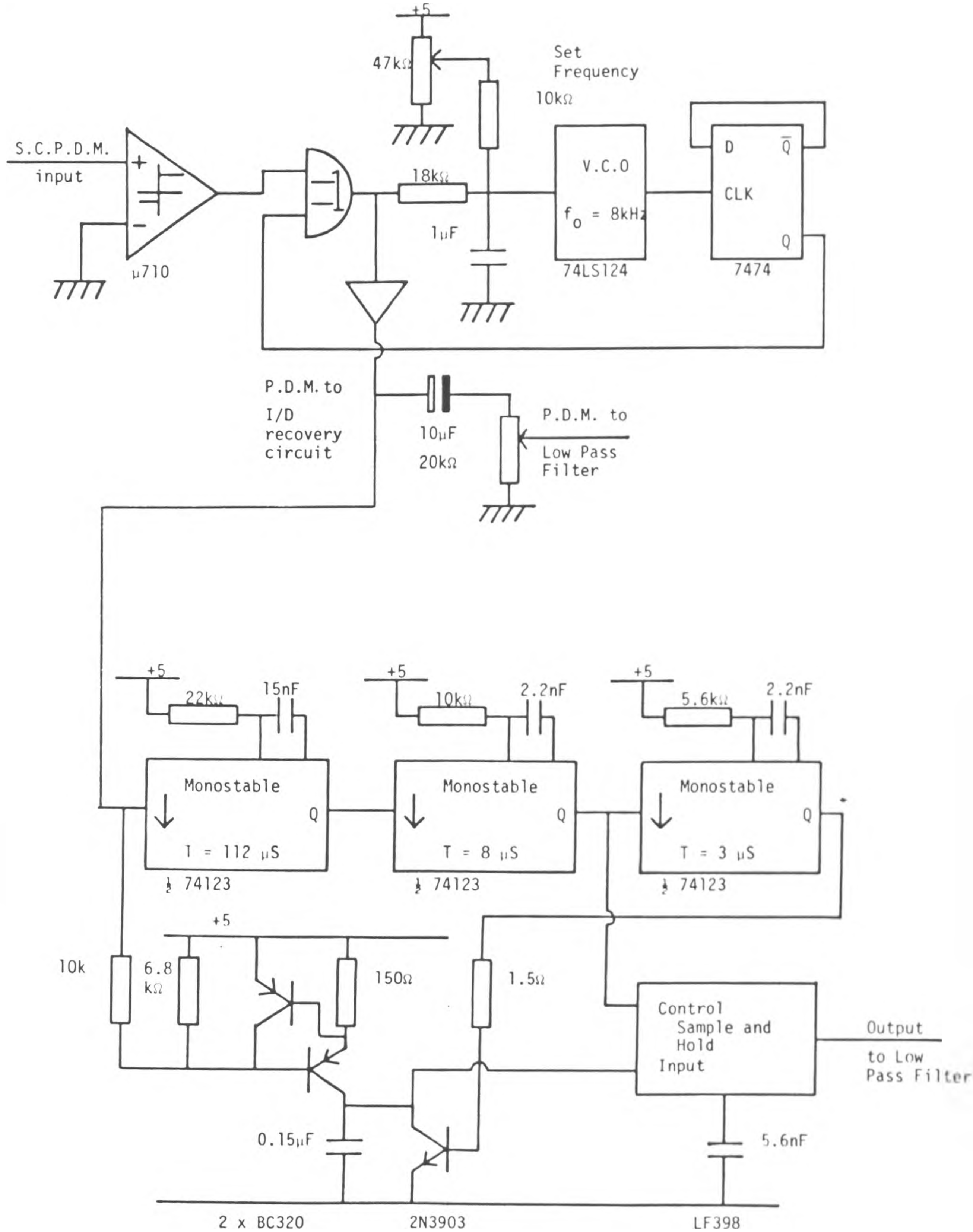

Fig. B.5 Suppressed Clock Pulse Duration Modulation

Demodulator

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