Advanced data communication techniques for Sub-Sea applications

by

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ABSTRACT

This thesis details research carried out in the through-water data communication field. An overview of the phenomena that prohibit acoustic communication in long-range shallow-water channels is constructed. Background research found that robust communications has not been achieved using single receiver reception in this environment. This work investigates the modulation technique itself and aims to improve on existing schemes (that have been applied to this environment). This is achieved with innovative techniques, based on multiple-frequency-shift-keying (MFSK) and space-frequency-shift-keying (SFSK). A number of industrial specified restrictions are placed on this work, including bandwidth restriction. Novel ways of intrinsically transmitting synchronisation information are therefore implemented. The development of appropriate systems is covered with general and platform specific implementation strategies being covered. A single modulation scheme (the three-chip four-frequency-shift-keying 3C4FSK scheme) has been put forward for consideration in any future research. Practical lab based tests and the mathematical analysis is detailed. Conclusions recommend further funding of long-range shallow sea-water trails of the 3C4FSK scheme and for the industrial scope of this work to allow investigation into multiple receiver systems that allow spatial processing of the signal as these schemes have been shown lately to have potentially in long-range channels.

CONTENTS

ABBREVIATIONS

Modulation Schemes and techniques:

OOK	On-Off Keying.
MFK	Minimum Shift Keying.
ASK	Amplitude Shift Keying.
FSK	Frequency Shift Keying.
BFSK	Binary Frequency Shift Keying
PSK	Phase Shift Keying.
BPSK (DPSK)	Binary (Double) Phase Shift Keying.
MFSK	Multiple Frequency Shift Keying.
MPSK	Multiple Phase Shift Keying.
QPSK	Quadrature Phase Shift Keying.
QAM	Quadrature Amplitude Modulation.
SFSK	Space Frequency Shift Keying.
DSSS	Direct Sequence Spread Spectrum
CDMA	Code Division Multiple Access.
TDMA	Time Division Multiple Access.
2C4FSK	Two-Coded Four Frequency-Shift-Keying.
5F4SFSK	Five-Frequency Four-Space Frequency-Shift-Keying modulation technique.
3C4MFSK	Three-Coded Four multiple-Frequency-Shift-Keying.
DSFMFSK	Dual Space-Frequency Multiple-Frequency-Shift-Keying.
PLL	Phase locked loop
LMS	Least mean Square.
DFE	Decision Feedback Equalisation.

Equipment:

Tx	Transmitter
Rx	Receiver
DSP	Digital Signal Processor.
VCO	Voltage Controlled Oscillator.
CCO	Current Controlled Oscillator.
DAC	Digital to Analogue Converter.
ADC	Analogue to Digital Converter.
AWG	Analogue Wave Generation.

LPF	Low Pass Filter.
BPF	Band Pass Filter.
fla, flb f2, f3, f4a, f4	Envelope detectors

Noise and Rate:

N	Noise
ISI	Inter-Symbol-Interference.
IBI	Inter-Band-Interference.
FSF	Frequency Specific Fading.
PN	*Pseudo* Noise.
SNR	Signal-to-Noise Ratio.
Rn	*Pseudo* Noise rate
Rs	signal rate
bps	bits per-second
BER	Bit Error Rate.
CNR	Channel to Noise Ratio

Signals and symbols:

S_i	The i^{th} symbol
C_{Tx}	The carrier at the transmitter
C_{Rx}	The carrier at the receiver
$Sp(t)$	Spread Spectrum Signal
F1, F2, F3, F4, F5, F6	Symbol frequencies.
Fs	Synchronisation frequency.

Period definitions:

T_c	Chip period
T_s	Symbol period
PT_s	Parallel symbol period
TS_s	Serial symbol period
T_b	Bit period
T_{bd}	The baud duration
T_r	The reverberation time
T_{cy}	Cycle period
Pp	Past 'perceived' bit period.

Ap	Active 'perceived' bit period.
Np	Next 'perceived' bit period.
F_s	Orthogonal frequency separation
F_{smin}	Minimum orthogonal frequency separation

Measurements:

Hz	Hertz.
Km	Kilometres.
kHz	kilohertz.
mw	milliwatts.
ms	milliseconds.
dB	decibels

Probability:

P_e	Probability of an error
P_c	Probability of a correct decision

Funding:

BG plc	British Gas plc.
EPSRC	Engineering and Physical Science Research Council.

Chapter 1. Introduction

1. The research described in this thesis covers data communication techniques for sub-sea applications, with particular emphasis on the modulation schemes ability to overcome the communication prohibitive acoustic propagation characteristics of the channel. The applications that require a sub-sea communication channel to be established are numerous but this work is focused on overcoming the specific problems of 'telemetering' data between the shore and an offshore platform or an unmanned sub-sea installation.

2. There are a number of different sub-sea environments that could be encountered in such applications, dependent on the channel topology; the trajectory, horizontal or vertical; the depth, shallow or deep; covering long, medium or short ranges. The acoustic characteristics of an individual channel can fluctuate with, time, temperature, weather, seabed type, surface state, salinity, sediment quantity and size, fish population and type, noise pollution and man-made obstacles (Fig.1.1). The instabilities of the channel's numerous characteristics and associated phenomena provide a medium that is not completely understood, and there is therefore scope for investigation. The environment provides an extremely complex multipath propagation characteristic[1, 2]. The acoustic signal's interaction with the surrounding environment can cause destructive interference, phase instability, inter-symbol interference (ISI) and frequency specific fading (FSF). Even with these harsh characteristics attributed to acoustic signal propagation through the environment, underwater acoustic signalling is still the best candidate for autonomous communications. With so many variables the problem of interest has to be clearly defined before investigation can be advanced.

3. This research covers five main stages that form the basis for the layout of Thesis.

Stage 1:

A. Background research carried out into the sub-sea environment and associated acoustical phenomenon.

B. Background research carried out into modulation techniques and their suitability; including techniques used in similar multipath environments and previous implementations specific to the sub-sea environment.

C. Formulation of ideas, preparation of design methodology and initial Matlab Simulink-based design assessment.

Stage 2:

A. Preparation of generalised designs, taking account of all the system specifications and project aims.

B. Assessment of the viability of platform specific implementation, including the choice of development platform.

C. Establishing digital signal processor (DSP) specific designs.

Stage 3:

A. Digital Signal Processor (DSP) implementation, testing and assessment.

B. Re-implementation

Stage 4:

A. Test design and preparation

B. Evaluation and assessment

Stage 5:

A. The assessment of the results

B. Thesis preparation and formulation of conclusions

4. This chapter of thesis sets out the aims of the research; stating the industrial involvement, providing an overview of the thesis and provides an introduction to the modulation schemes that have been developed. A definition of the contribution to knowledge is also given. Chapters 2 and 3 include background research, focusing on the sub-sea communications environment and modulation techniques, respectively. Chapter 2, looks at channel characteristics and associated underwater phenomena. Chapter 3 is an overview of recently implemented modulation principles and techniques. Chapter 4 is a general overview of the system implementations, providing an overview, of the transmission and receiver strategies, the synchronisation approaches and the system hardware. Chapter 5 details the first of the modulation schemes, the Three-Chip Coded Four-Frequency-Shift-Keying modulation technique (3C4FSK). Chapter 6 details the Five-Frequency Four-Space Frequency-Shift-Keying modulation technique (5F4SFSK). Chapter 7 details the Dual Space-Frequency Multiple-Frequency-Shift-Keying modulation technique (DSFMFSK). Chapter 8 Evaluation of the schemes, assessment and a compilation of conclusions from the research, recommendations for further work and a reiteration of the definition of the contribution to knowledge contained in this thesis.

1.1. Aims and Requirements of Research

5. This research is focused on developing a modulation scheme appropriate for the acoustically harsh conditions found in sub-sea environments, in particular transmission through long-range, shallow channels. The modulation schemes developed here are aimed at implementation in point-to-point links that are limited to the use of one omni-directional transmitter transducer (the projector) and one omni-directional receiver transducer (the hydrophone). It was believed at the outset of this work that a modulation scheme could be found that could improve the reliability of communications (telemetry) between submersed gas installations. The idea of this research was to study the capability to sustain a communications link of a variety of schemes before any implementation of additional processing, i.e. to find a scheme that is intrinsically resilient to the environment. This should enable a communication link to be set up with 'existing' British Gas equipment for the control of unmanned sub-sea installations.

6. The research aims can be condensed into a short statement:
To develop a modulation scheme that works reliably in long-range (10Km), shallow (10-100m) sub-sea channels using a pair of omni-directional transducers at a bit rate of between 120 and 150 bits per secound. This was to be carried out without the use

of multiple receiver techniques (multi-channel combining) or additional processing techniques (data coding, equalisation, adaptive equalisation and directive antennas).

7. A more in-depth account of the research aims and requirements follows. To assess the performance of the modulation techniques in the through-water environment, an accurate assessment of the effectiveness of each of the modulation schemes is required independently of other techniques that could influence the performance. A modulation scheme that works in the long-range, shallow water environment is required as this represents the worst communications scenario. The stipulation was; to use only one transmitter and one receiver which means having a resilient scheme that does not need the additional complexity or expense of multiple receivers, arrays, directional antennas.

1. The data requirement is to achieve continuous data transmission of at least 120-125 bits a second.

2. Sustained transmission must be maintain without reverting to burst transmission protocols, that are unable to maintain the data rate throughout transmission.

3. The data link should be able to be established using existing omni-directional transducers that have a three-kilohertz bandwidth that centre at a resonant carrier frequency of ten-kilohertz.

The requirement for a bandwidth limitation falls in line with other media such as radio frequency applications, as some bandwidth allocation may be a requirement in future sub-sea communications. Bandwidth usage within this environment is a concern[3]; the development of efficient modulation techniques and coding[4] will become more of a concern in the future as usage increases.

The modulation schemes are aimed primarily at two types of channels specified by BG plc:

1. A channel with one end in shallow water and one end in deep water as shown in Fig.1.1 (a).
2. A channel with both ends being in shallow waters Fig.1.1 (b).

1.2. Introduction to the modulation techniques developed

8. The choice of modulation schemes developed in this research all use a single omni-directional projector and a single omni-directional hydrophone. This hardware stipulation is important as it negates any use of multiple receivers and associated processing, which has been show to be the most effective form of communications through long-range through-water channels to date [5-7]. Other potentially useful techniques are also restricted with this choice including the use of highly directive antennas. The long-range, shallow sub-sea environment is extremely hostile particularly when using single 'non-directional' transducers.

9. The instabilities inherent in the medium combine to affect the signal characteristics predominately in the form of fluctuating signal phase and signal amplitude; thus phase and amplitude modulation schemes are particularly inhibited in the ocean environment [8-11]. A number of sub-sea schemes have recently been implemented that use Phase Shift Keying (PSK) modulation techniques [5-7, 12-15]. These schemes have used multiple widely spaced receivers (or arrays) and/or directive antennas to allow the techniques to be viable [10, 11, 16-22]. Without the ability to use widely spaced transducers (or arrays elements) at the receiver or directional antennas, some other way of overcoming the signal fluctuations is required, if phase or amplitude based modulation techniques are going to be used.

10. In this research the bit rates required mean that frequency-based modulation techniques can be looked on as viable means of limiting the problems associated with variable signal phase and amplitude produced by interaction with the environment.

Frequency based modulation techniques are relatively less susceptible to interaction with the environment compared to phase and amplitude based modulation schemes. This is mainly due to the signal phase and amplitude randomisation found with interaction with the environment. The modulation schemes developed during this research are thus based on Frequency Shift Keying (FSK) methods, which have traditionally been the modulation techniques most appropriate to long range, shallow-water communication.

11. Modulation techniques have been developed that use redundancy to combat interference, in the form of multiple frequency transmissions, whilst taking into account the bandwidth limitations imposed on the system. The transmission symbol rate restricts the choice of orthogonal frequencies that are available in the bandwidth allocated. The actual frequency spacing between symbols is defined essentially by two factors, (1) the requirement for orthogonal symbol frequencies when implementing matched filter detection (Appendix A), and (2) the need to limit filter tap sizes for real-time DSP processing. The orthogonal restriction becomes critical as the number of frequencies needed is increased for a particular symbol rate (T_s). There is a limit to the number of orthogonal frequencies available; with the number of frequencies decreasing as the symbol rate is increased. The implementation of receiver filters for signal detection is a restricting factor if the spacing between frequencies is small, as the implementation of matched filters can become highly tap-intensive, inhibiting the DSP's ability to perform real-time operation.

12. The design of the sub-sea modulation schemes has been carried out in two stages, a 'conceptual design' and a DSP-'specific' design. The conceptual design for implementing the modulation schemes was developed with the aim of providing a non-platform-specific structure from which specific designs could be developed. This provides an easily transferable generalised design. DSP platforms were chosen to enable a number of separate implementations to be developed using the same hardware and to allow quick transfer from one scheme to another while developing and testing the schemes.

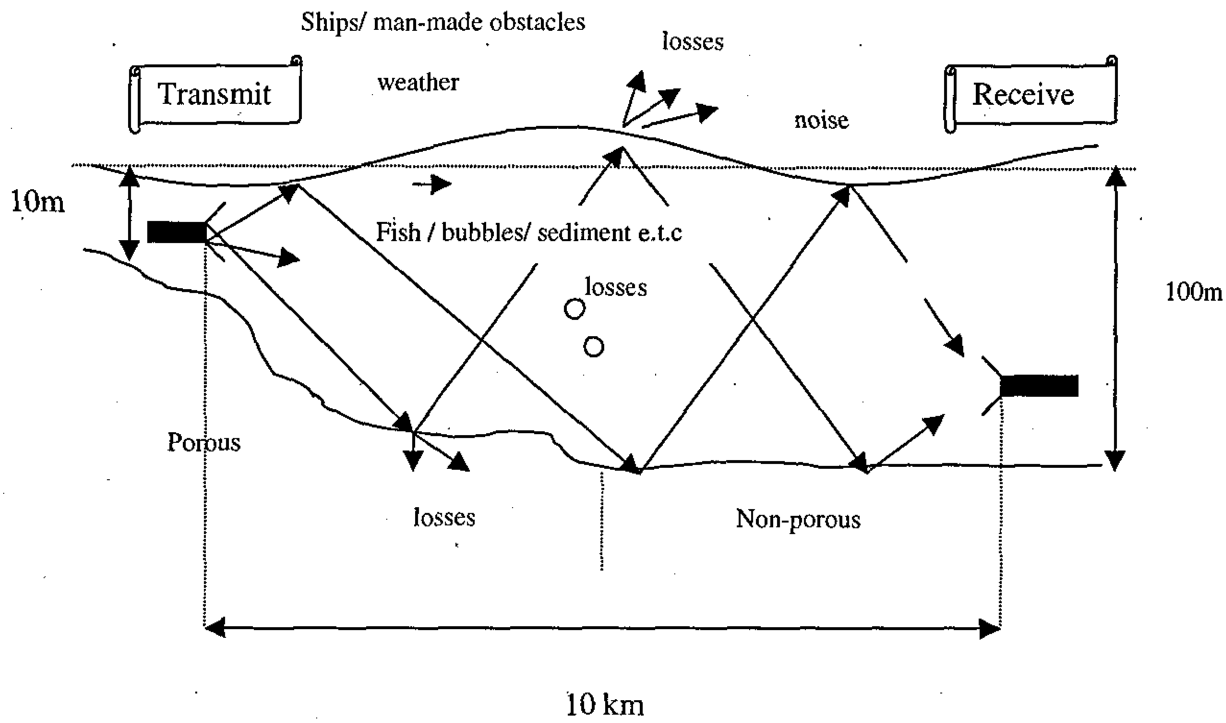


Fig.1.1 (a) Long-range communication channel, shallow to deep water

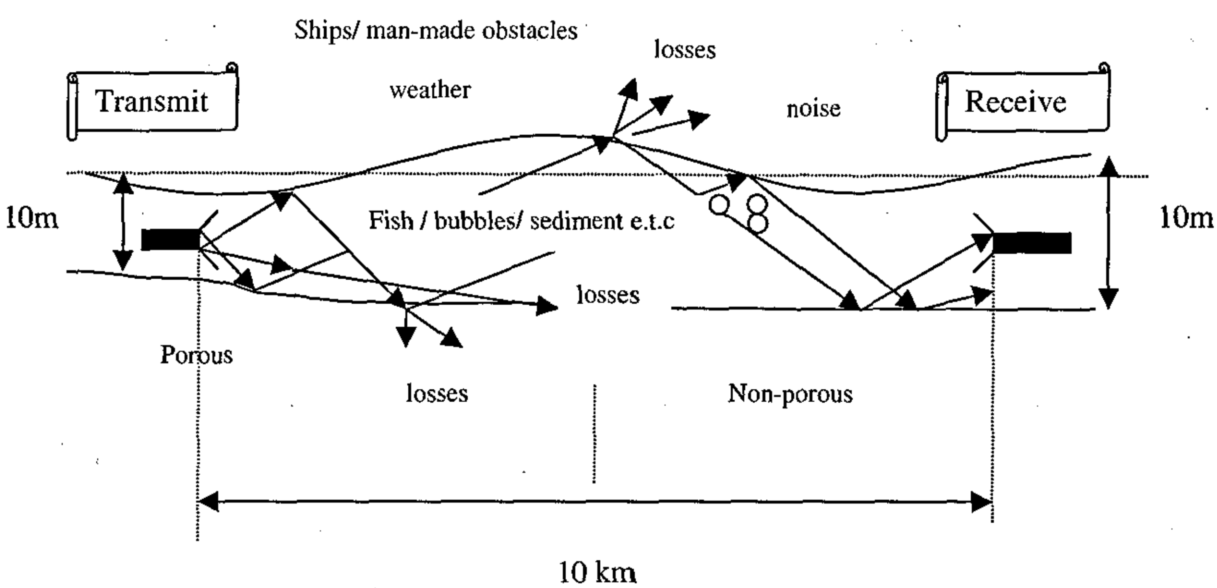


Fig.1.1 (b) Long-range communication channel, shallow water throughout

13. The initial designs ('conceptual designs') where developed using the MATLAB simulation package Simulink. The workings of the systems were initially tested by simulation with this package. The DSP-specific designs were implemented using DSPlay for overall system development, analysis and code generation; QEDesign software for filter coefficient generation; and DSWorks development software for filter analysis. The 'conceptual design' and the DSP 'specific design' are included in the system descriptions in the appropriate chapters of this thesis. The transmitter and receiver systems are explained, each being divided into a number of sub-systems that represent the individual processes required for generation and detection of the modulated signals.

1.3 Definition of the contribution to knowledge contained in this thesis

The contribution to knowledge contained in this thesis includes the; design; development; implementation and evaluation of three modulation schemes specifically for use within the shallow-water, long-range sub-sea environment. Frequency diversity, serial (time) diversity and parallel diversity are all utilised. The design and application of a number of innovative implementation techniques is also undertaken including a form of intrinsic synchronisation and coincident point envelope detection.

Concluding Remarks

This chapter provides an introduction to the research carried out, outlining the content and structure of this thesis. A description of the project planning and components of research are included. Implementation platforms and industrial requirements are covered. This leads onto the following Chapters, which cover background research with emphasis on relevant environmental phenomenon and modulation techniques.

Chapter 2. The Sub-Sea Channel and Associated Characteristics

1. The underwater environment provides a relatively hostile host for communications compared to more widely used media (atmosphere, coaxial cables, fibre optics) [23-25]. Problems with maintaining communications within the sub-sea medium are generally attributed to the intrinsic complexity of the acoustic signal's propagation found in highly fluctuating channel [26-29]. Evaluation of the relative performance of particular modulation techniques requires an understanding of the intricacies of the medium [27, 30]. An underwater channel is an environment that can be diverse in many ways, since there are numerous differences between channels, including straightforward structural differences such as depths and ranges.

2. The problems associated with underwater acoustic communication are complicated in particular when the medium to be traversed is the sea. A sub-sea channel has its own individual characteristics and processes that differ with time, location and the frequency of transmitted signal (Fig.2.1). The fluctuating sea surface (waves) and volume flow (internal waves) combined with temperature, pressure and salinity variations, cause a continuous change in the channel characteristics. Environment specific factors such as sound velocity profiles, amount of absorption, scattering and reverberation levels are all factors. These parameters all play a big part in defining the behaviour of acoustic waves as they transverse the medium. With all the different environmental parameters possible, a natural sea environment is never duplicated. Signal propagation can become extremely complex, especially near land. Near-shore transmission provides extra complexity when the signal interacts with the channel boundaries as they merge at the seashore [31]. Seasonal variation in the sound velocity profile, which is dependent on temperature and weather conditions (rain, wind etc.), only compounds the situation.

3. The underwater acoustics and communications communities have been unable to completely understand sub-sea propagation characteristics. Research has been carried out in the past aimed at establishing the effects of individual environmental parameters or combinations of parameters on sound propagation [32]. The idea was to understand the complete characteristics of these parameters to allow compensation for their effects on the acoustic signal to be applied.

4. The specific characteristics attributed to the sub-sea medium that have a significant bearing on the capability of the communications system, are numerous, varied and dependent on the type of application the system is to support. Despite the complexities, sound is the most suitable form of 'radiation' for communication because it is much less attenuated in water than other forms of energy, e.g. light. It therefore provides one of the best options for non-umbilical communications [33].

5. This chapter concentrates on the most influential characteristics of the sub-sea environment pertaining to the channels under investigation. In particular the multipath and frequency-specific effects that are considered particularly problematical to long-range shallow, sub-sea communications. Examples of experimental tests and verification of the effects found in the medium are referred to wherever necessary.

2.1. Sea Environment

6. The primary influence on acoustic propagation in a sub-sea channel is the environment itself. The environment that is encountered when undertaking sub-sea communications is continually changing and varied. Any channel can be influenced by a number of factors that may combine, establishing a particular channel characteristic.

The factors that can influence a sub-sea channel include:

1. Relatively stable components such as local seabed type; permanent man made objects; channel depth and channel range.

2. Slowly varying characteristics; including the sound velocity profile; pressure; temperature; salinity and foreign bodies in the channel.

3. Rapidly changing parameters like surface waves due to wind conditions; bubbles produced by rain and ship noise.

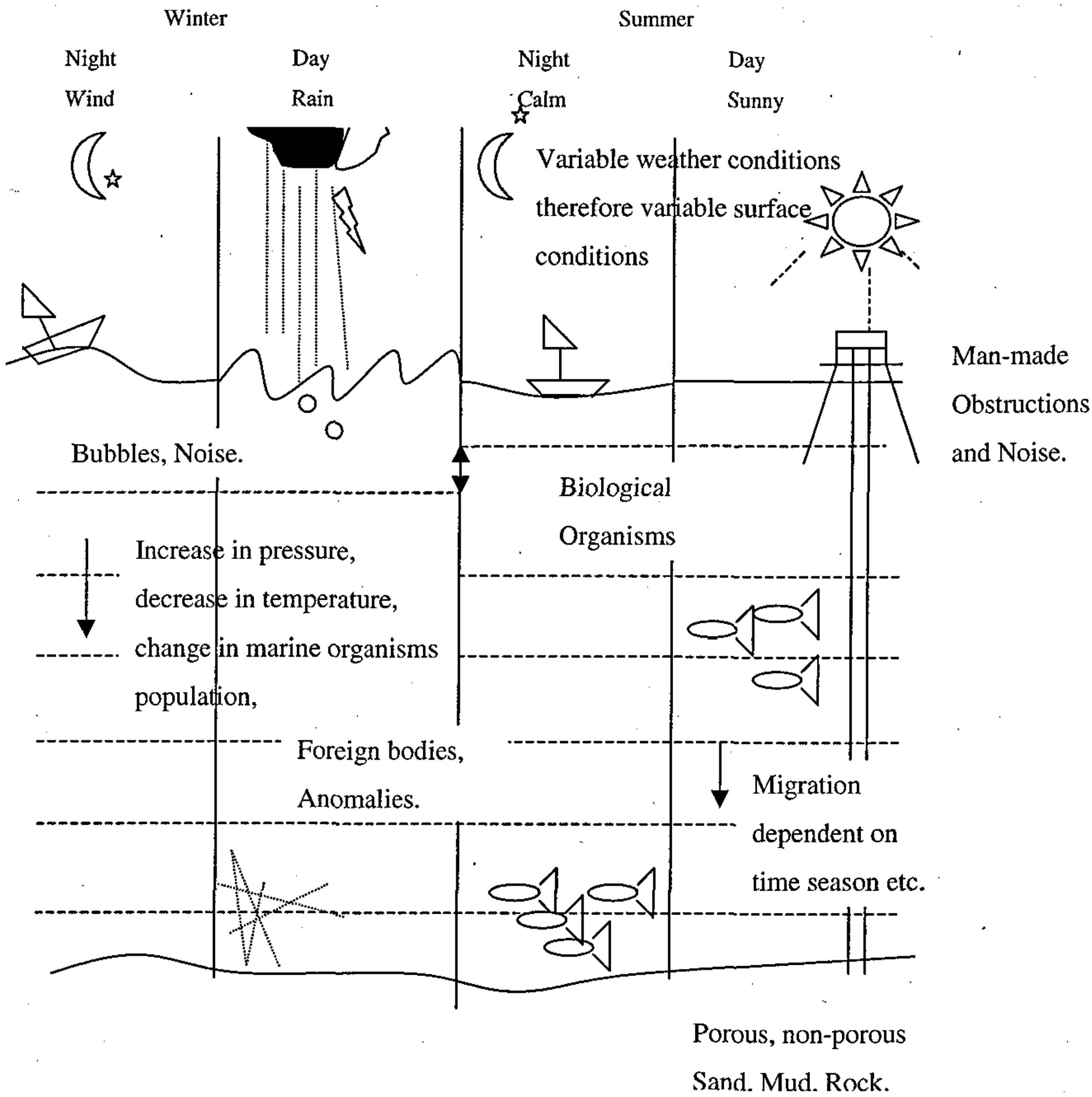


Fig.2.1 **The sub-sea variable characteristics**

2.1.1. Channel depth

7. There are many parameters that define an underwater channel, one of the most basic being the channel depth. A shallow medium can provide quite different problems and have different performance limitations than those found in a deep-water environment [34-41]. The depth parameter of the environment is relatively stable, compared to other environmental variables.

8. The depth can influence the propagation of the signal greatly. A shallow channel may have the same boundaries (Sea floor type, Surface condition) as a deep-water channel but they may have a different influence on communications in each case. Signal propagation is affected much more dramatically by the boundaries in a shallow water channel because of their proximity, whereas in a deep-water channel the boundaries have negligible affect in comparison. The increased signal degradation that is associated with a shallow channel can be combated using a variety of techniques [23, 24]. One of the main problems in the shallow environment, 'multipath interference', is generally attributed to interaction with the boundarys.

9. The sub-sea environment's depth can be characterised as layers of water each with slightly different parameters. The uppermost layer of the ocean has been found generally to be uniform in temperature and salinity during colder and windier periods accounting for the label 'mixed layer'. The mixed characteristic of the upper layer does not continue year round; it becomes less well 'mixed' in the calmer, warmer periods. The depth of the mixed layer is normally between 10m and 100m [42]. The thermocline produced at the bottom of the mixed layer is generally characterised by a discontinuous increase in the gradients of temperature and salinity. The mixed layer is deepened by strong winds, the gradients being re-established once the winds die down and there may also be a change in depth depending on the seasonal weather [43].

2.1.1.1 Sound velocity profile

10. Environmental parameters combine to affect communications in various ways; larger environmental processes result from these interactions and significantly influence acoustic signal propagation. A variable velocity of sound with depth in the sea is such a process [28]. The three parameters that influence the velocity profile in the ocean are temperature, salinity and pressure. The typical sound velocity profile is a result of the competition between the ocean temperature and salinity, which tends to give a sound-velocity decrease with depth, and the pressure gradient which tends to give an increase with depth [42]. Dominance in lower depths of the ocean is held by the pressure gradient with this being diminished in the upper levels of the ocean.

11. The sound velocity profile of a particular channel causes diffraction of the propagating sound [44] and can set up 'shadow zones' where minimal signal incidence is apparent. The small occurrences of the signal can be attributed to scattering from bubbles [59] and rough ocean surfaces. Placement of receiver hydrophones within a 'shadow zone', can severely limit the likelihood of sustainable communications.

2.1.2. Transmission Range

12. The range-dependent parameter can be defined in terms of how far the acoustic signals travel along the channel's horizontal axis. The length of a channel can significantly affect which channel characteristics inhibit effective communications [26, 45, 46].

13. Range categories used within modern communication systems vary, but channels are normally placed in one of three categories: short range, medium range, long-range. The exact definitions of 'range' varies, dependent on whom has implemented the scheme. Generally these categories can be considered as <1km for short-range transmission, 2km to 5km for medium range transmission and >5Km for long-range transmission [47]. The differences in distance between channels can bring into effect different sets of environmental phenomena, which may or may not be apparent in the other categories.

14. There is a particular interest within the sub-sea communications community; in the possibility of developing underwater acoustic communication systems, capable of operating at ranges which are long by comparison with depth. These long-range channels are known to have higher transmission losses [26, 45, 48]. The channels that are of interest to this research are within the long-range category (Fig.1.1).

2.1.3. Direction

15. Channel configurations that are relevant to the communications system depend not only on range and depth but also on the direction of operation (horizontal/vertical); thus numerous unique channel characteristics may be encountered. The propagation direction of interest to this research is horizontal, thus the transmitted signal encounters a much more fluctuating environment than generally found with vertical propagation. The increased interference found in a horizontal channel, means that the modulation techniques used have to be more robust.

2.1.3.1. Boundaries

16. When shallow, long-range, horizontal transmission is being implemented the boundaries of the channel have the potential greatly to influence the signal characteristics. The channel boundaries can affect an acoustic signal with the possibility of both surface and bottom reflected waves combining with the direct wave. The combination can be reinforced by producing greater pressure, or partially cancelled to produce a lower pressure, dependent on the phases. The phenomena that are prevalent at the boundaries when there is interaction with an acoustic signal include signal scattering, attenuation and absorption. The reflected signal is thus a depleted version of the incident signal.

17. Acoustic signal interaction with the surface causes scattering of the incident signal power and reflection of the propagation path [31, 45, 49]. The surface boundary can be approximated to a near-perfect reflector when smooth and as a scatter when rough. The surface path also changes to a specula reflection at long range from scattered reflection at short range. When reflection is primarily specula there is relatively little power loss but phase inversion occurs on reflection. In the case of scattered reflection the acoustic power is 'placed' in all directions thus higher losses are incurred, and the phase change is of a random nature. The scattering strength generally increases with frequency. The fluctuating nature of the sea surface can also cause 'frequency spreading' of the incident signal. Frequency components of the surface waves can 'smear' the signal envelope by producing upper and lower side bands in the spectrum. The moving sea surface 'superimposes' its 'Doppler motion' onto the frequency of the incident signal; which can have an effect on narrow band systems [50].

18. The seabed boundary interacts with an acoustic signal causing scattering, attenuation and absorption [31, 51-53]. The phenomena at this water-sediment interface are dependent on the seabed composition, the angle of incidence and the signal frequency. A wide band signal will experience frequency selective fading as different frequencies are attenuated to different degrees. The seabed composition is complex, as the seabed is normally a layered environment. The are many variable attributes that the seabed can exhibit including roughness, porosity and hardness [31].

19. For the kilohertz range of frequencies the scattering strength is highly variable for grazing angles up to about 60 degrees depending on seabed and sub-seabed roughness and composition. The surface roughness, which is mostly dependent on the wind, and the effects of surface roughness on surface reflected sound has been investigated in a number of papers [31, 43]. The Raleigh parameter has been used to quantify the relative state of the sea surface; it decreases with the grazing angle of the surface ray, as the range of the communication channel increases. The scattering strength generally increases with frequency. As grazing angles tend to zero degrees the scattering strength becomes negligible.

2.1.4. Volume characteristics

20. There are discontinuities in the volume of the ocean that causes signal attenuation, due to thermal conductivity and viscosity within the medium. There is also volume movement that can be attributed to 'internal waves' due to geographical mass flow [35, 54]. This is compounded by signal scattering and absorption; due to foreign bodies; discontinuities; bubbles and the increased absorption found with saline solutions [55-59].

21. A number of foreign bodies can be found in the ocean body, including salt, fish and man-made objects. Fine particles in the water have negligible effect on the acoustic signal but bubbles in the water can have strong scattering and absorber characteristics when their resonance frequency is related to the transmitted signal. Similar effects are found with certain populations of marine organisms such as the swim bladder fish [60].

2.1.5. Noise sources

22. There are a number of noise sources in the underwater environment that can have an effect on communications. Ambient noise can stem from a variety of sources and therefore can take a number of forms [61, 62]. The noise can be produced continuously, periodically or as a one-off occurrence, thus its effect on detection may vary. Noise sources include man-made noise, direct weather-induced noise, indirect weather-induced noise and biological sources to name but a few. Man-made interference generally stems from production machinery (petroleum or gas) [63] or shipping. Depending on the location of the communications channel; noise from man-made sources is likely to be either intermittent or sustained over longer periods.

23. Noise produced by the weather conditions, include direct phenomena like rainfall [64, 65], hail and snow breaking. Indirect influences such as wind induced waves produce noise and bubbles as they break [66]; noise is also produced in the volume of the medium, due to the thermal activity of the water molecules. The 'weather' noise occurrence is seasonal and intermittent. Biological organisms such as fish and mammals can also produce noise, which is usually of short duration and dependent on their movement, thus time dependent.

24. The frequencies that are used for the schemes in this work fall within the bandwidth where ambient noise production may be dominated by surface noise due to breaking waves and bubbles [67]. These results were gained in deep water, but intuitively it can be seen that in a shallow water environment the effects at the surface are likely to be even more pronounced, due to the proximity of the boundaries.

2.1.6. Channel variations

25. There are a number of channel characteristics, which do not stay constant. The 'types' of variation in the mediums characteristics can defined as:

1. Seasonal variations (weather dependent)
2. Variations influenced by the time of day or night
3. Temporary variations
4. Continuous variations

More than one of the channel characteristics may vary at one time they can vary continuously as well as exhibiting quantifiable seasonal differences. An example of this is the sound velocity profile, which can depend on the season and can vary over a much shorter period. A second example could be the population of marine organisms, which at certain depths may vary significantly between day and night and maybe from season to season. Weather has a temporary and continuous affect on some channel characteristics, with wind-induced waves and bubbles, rain induced bubbles and noise and other similar effects. Man-made objects can also cause continuous or temporary changes, or both, to the environment (ships, oil/gas installations etc.).

26. There have been studies that present evidence of variations in a number of the media characteristics and the effect on a communication system. Specifically, the influence of seasons has been shown in terms of the variation in propagation characteristics though out a year [47].

2.2. Underwater Sound Phenomena

27. An acoustic signal's interaction with the sub-sea environment produces phenomena that influence propagation of acoustic signals in a number of ways. The channel characteristics dominate the design of any acoustic data communications system, because these phenomena determine the characteristics of the acoustic signal at reception:

1. Reverberation due to multipath causes signal fading and delay spread at the receiver, which can lead to inter-symbol interference (ISI) as well as additive and destructive signal strength combining effects.
2. Frequency spreading due to signal and boundary interaction which can cause degrading of signal strength and in-correct symbol detection at the receiver [26].
3. Signal scattering due to signal interaction with the boundaries, suspended substances and gas bubbles, which lead to loss in signal strength and production of multipath (micro-multipath and macro-multipath).
4. Absorption due to interaction with a number of substances in the channel and with the seabed, which causes signal strength to degrade with range and can cause frequency dependent effects.
5. Ambient noise produced by man-made objects, breaking waves and rain can degrade the signal-to-noise characteristics at the receiver.
6. Refraction due to sound velocity differences in the medium, lengthening travel times and setting up internal sound propagation channels and shadow zones, which can make the performance of a receiver dependent on its placement within a channel (depth, range, etc.).
7. Random phase due to reflective paths, which can negate the use of a number of phase signalling modulation techniques without the use of spatial combining or some other compensatory technique.

8. Signal amplitude fading due to the multipath environments, which are variable and can cause the temporary loss of synchronisation or incorrect detection as the signal strength fluctuates.

9. 'Doppler shift' that is attributed to the relative motion of the source and receiver; sometimes needs to be compensated for [152, 154, 155, 156] (not applicable in this research as the source and receiver are both fixed).

28. The receiver signal is spread in time and frequency and has variable amplitude, visualised in Fig.2.2. The received signal also exhibits a randomised phase characteristic.

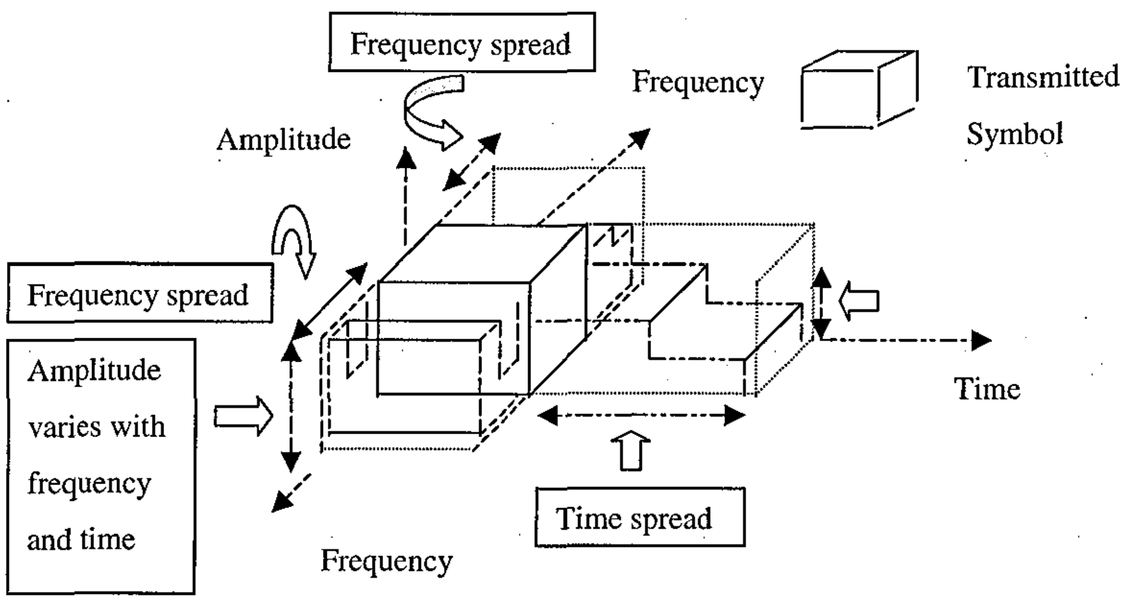


Fig.2.2 **Signal Visualisation**

2.2.1. Multipath

29. Boundary-induced (discrete) multipath includes components stemming from interaction with the sea surface, seabed and suspended objects within the body of the ocean. The occurrence of numerous transmission paths (multipath) between transmitter and receiver can be attributed to the interaction of sound rays with the channel boundaries (sea surface and seabed). Multiple reflected paths to the receiver (reverberation) are produced. Reverberation due to multipath is one of the most challenging phenomena of the underwater acoustic channel that a communications system has to overcome [27]. The existence of multipath causes signal fading and time spreading to occur, which combines with frequency spreading and attenuation inherently found in the environment to the detriment of the signal detection process at the receiver [68, 69].

30. Other macromultipath types include diffuse multipath that is attributed to the non-uniform body of the ocean. There are single path fluctuations (micromultipath); caused by volume scattering that is due to the movement of heterogeneity's in the body of the medium (suspended particles, marine organisms, bubbles and dissolved substances). Micromultipath is also attributed to 'internal waves' found within the volume of the channel which is generally attributed to geographical mass flow. The combined effects of the different multipath interfere with the coherence of a communication system, providing the need for multipath compensation to be used to enable effective communications to be maintained. The effect of the multipath propagation within the channel includes signal fading (SF) [1], and inter-symbol interference (ISI) [70], which can cause systems to exhibit substantial bit-error-rates (BER) [71].

2.2.1.1. Reverberation

31. The situation in a sea channel is a complex one with boundaries that are fluctuating and imperfect; time-variant multipath is a normal characteristic in all but a few propagation paths of interest. All configurations of sea communication channels whether the distance to be traversed is long, medium or short, and whether the water depth is deep or shallow, have effects that can be attributed to reverberation due to multipath.

32. The implication of reverberation in the channel is the presence of multiple versions of the transmitted signal at the receiver each fluctuating in amplitude and phase and arriving with varying delays. Different underwater channels are affected by reverberation with differing degrees of interference to the transmitted signal. A short-range channel's communication performance is dominated by the signal's interaction with the channel boundaries and reflection from objects within the body of the water. In short-range shallow-water cases, the communication problem becomes extremely specific to the local environment with no one solution standing true for all cases [27].

33. A shallow water channel is highly susceptible to reverberation compared to a deep-water channel due to discrete multipath that is due to the proximity of its boundaries and their ability to create numerous paths between transmitter and receiver. In a shallow water channel, the small grazing angles with low reflection losses enable large multipath amplitudes to exist, which can combine constructively or destructively at the receiver. Constructive multipath techniques such as coherently combining multiple symbol arrivals are feasible under the right conditions [36, 72].

34. Far less interaction with the boundaries is found in deep-water channel than shallow-water. The sound velocity profiles found in the ocean, together with the distances travelled before interaction with channel boundaries, can provide the characteristics necessary to establish an internal sound channel. This can limit discrete multipath, as a large reduction in boundary interaction becomes evident. At depth a sound channel is produced by refraction, which enables a number of the sound rays to pass through the medium without encountering the boundaries, thus reducing the occurrence of discrete multipath.

35. It is difficult to distinguish between the different components of reverberation (produced from interaction with the seabed, surface and volume) found within long range, shallow water channels [52]. The receiver can not discriminate between the origins of the individual multipath and thus perceives all reverberation as the same. Reverberation has been measured in a 'control' environment where the effects of individual components can be discounted, showing there is a difference in the strengths exhibited by specific reverberation (surface, bottom, volume) [52].

2.2.2. Overcoming Multipath

36. There is a need for sub-sea communication systems to have the ability to dispel or minimise interference (multipath induced), in order to negate the negative effects it has on communications [69]. The phenomena of Inter-Symbol-Interference (ISI) and signal fading (SF) (including frequency dependent and channel dependent signal fading) which are inherent in a multipath environment require particular consideration.

37. Techniques that have been developed for modelling signal propagation and related phenomena enable approximate data to be obtained where experimental data is not available. The most appropriate modulation schemes for a particular environment can be chosen by an analysis of the simulated or experimentally collected data. A number of techniques that have been applied to the simulation of signal propagation characteristics in non-underwater media have been shown to be applicable in the sub-sea environment [23, 45, 63]. Ocean environment models have been used in the past to estimate the channel reverberation characteristics using adaptations of methods used for applications in similar multipath channels. Scattering function analysis is an example of a technique that can be applied that was first developed for electromagnetic communications. The usefulness of simulations is limited by how much information is available about the intricacies of acoustic effects in the underwater medium; therefore experimental data is preferable. Not only simulation techniques can be transferred from other media, modulation techniques may also be transferable. Modulation schemes that have been shown to improve the reliability of communication in one situation can usually be applied in others, as is the case with some of the dedicated interference combating methods (equalisation, directional antennas, etc.).

38. Any technique taken from implementations used in other communications media requires adjustment to account for specific sub-sea environment characteristics. Typical multipath spreads found in environments such as mobile radio channels cover two or three symbol intervals. The multipath spreads are multiplied in underwater channels increasing up to tens of the symbol interval [73], at symbol rates of hundreds of symbols per-second and above. Frequency hopping modulation was developed in other media as a security technique ('anti-jamming'), to facility the provision of secure communications [74]. The multipath avoidance properties of this method have been qualified in satellite communication systems where the scheme has been used to overcome macro-multipath interference. In the sub-sea, highly reverberent environment frequency hopping can also be used to combat the 'jammer-like' characteristics exhibited by multipath signals.

39. Other techniques available for interference compensation include equipment-based methods such as adaptive equalisation that compensate for the channel transfer function and highly directional antennas that suppress a proportion of the multipath signal. The multipath compensation schemes can sometimes complicate the situation, by introducing problems like slow equalisation adaptive convergence properties and alignment of antennas. Numerous papers have been produced in the past that describe a variety of methods aimed at establishing techniques for reducing interference within the communications channel [69, 72, 75-89]. The separate techniques available can be applied in varied channel conditions in attempts to reduce either SF or ISI. There is a need in many sub-sea environments to mitigate both these affects within a single channel. Hybrid techniques can be employed by implementing schemes in pairs (one to combat SF, and one to combat ISI) or by applying multiple techniques in conjunction to reduce the effects of each of the disturbances.

40. The four categories described below do not completely cover the full spectrum of techniques available, but summarise the main areas. The modulation schemes that have been investigated do not fall conveniently into any one category but can be seen as a combination of a number of the categories. The main techniques that can be used to combat ISI and SF can be placed into one of four categories: -

A. Symbol Period Manipulation Methods

41. Symbol period manipulation schemes are implemented by taking into account a channel's reverberation characteristics in terms of the duration and level of the signals [81]. The multipath deterioration rate within a specific channel is one of the factors to be considered when determining the modulation scheme's symbol period [23, 24]. Multipath signals stemming from previous symbol periods can be completely avoided or allowed to degenerate to an acceptable level by the implementation of symbol period manipulation methods.

B. Diversity Utilisation Techniques (spatial, frequency and time diversity)

42. Diversity can be found in the sub-sea environment in a number of guises, which can be manipulated to the benefit of communications. There is the spatial diversity (channel diversity) apparent in the medium in terms of channels with separate propagation characteristics [90, 91]. The deployment of 'widely' spaced receiver transducers can enable the separately fading channels to be detected. The spatial diversity found in the ocean propagation characteristics can be used to overcome channel dependent SF and phase randomisation. Frequency diversity is also a characteristic of the ocean medium with distinctive bands of frequency exhibiting various fading properties. Frequency-dependent SF can be minimised by transmitting separate frequencies. Other diversity techniques available include the use of differences in the propagation time (time diversity) of the separate multipath components. The reverberation components of multipath can be avoided in some environments where the reverberation time of the channel allows discrimination between transmissions that are spread far enough apart.

C. Channel Equalisation

43. Equalisation methods combat multipath disruption by cascading a model of the inverse of the channel transfer function. The ocean's, fluctuating nature points to adaptive techniques being employed if equalisation is to be a practical option. There have been a number of developments in adaptive equalisation, which have enabled these techniques to be used in situations where their use was previously limited by combining them with other interference combating methods [7, 17, 86]. The combination of techniques can allow the channel transfer function to be tracked more effectively by an adaptive equalisation algorithm.

D. Use of Angle of Arrival

44. The angle of arrival of the acoustic signal can be manipulated to reduce the multipath reverberation apparent at the receiver. The use of directional transmitters can reduce the received signal's interaction with the high reverberation inducing boundaries [6, 7, 21, 108]. Techniques for narrowing the beam-width of the transducers are available. They can reduce the multipath reverberation effects, and when employed together with a 'steering' technique the likelihood of the receiver being outside the signal's incidence can be limited.

2.2.2.1. Symbol Period Manipulation Methods

45. The length of the symbol period can have a direct influence on the reverberation's ability to degrade a communications system's performance. The employment of extended symbol periods (relative to the reverberation period of the channel) is effective at dispelling multipath components that stem from previous symbol periods. The channel characteristics determine the duration required for multipath components to degrade to an appropriate level for demodulation. Lengthening of the symbol period is viable, but not completely effective in situations where large reverberation times are evident. The channel's data rates are highly restricted in such a 'prolonged reverberation' environment.

46. The existence of Inter-Symbol Interference (ISI) may be apparent at the receiver if relatively short symbol periods are in use. When the symbol period is shorter than the reverberation duration the received signal is normally made up of the transmitted symbol and its multipath components as well as multipath components arising from previously transmitted data. The ISI corrupts a smaller proportion of an 'extended symbol period', as ISI dies down the signal that was transmitted (for that specific symbol period) becomes large in comparison to the multipath signal that arises from previous transmissions. Thus ISI can be combated in terms of frequency content by implementing extended symbol periods; the phase and amplitude of the signal are still inhibited by multipath.

47. Frequency-based modulation techniques can use the extended symbol period method for multipath suppression. Phase and amplitude modulation techniques still are affected by the random nature of the signals phase and amplitude, which is not dispelled by the extended symbol period method. A problem with the extended symbol period techniques is the inherently low data rates that are achievable. Inter-Symbol Interference is combated by this technique but signal fading is not overcome.

2.2.2.2. Diversity Utilisation Techniques

48. The frequency diversity found in the ocean can be exploited to combat fading that is apparent in this medium. Selective fading of frequencies in underwater channels can be used to transmit data simultaneously in separate frequency bands. A coherence bandwidth equal to the minimum separation of frequency bands has to be maintained so that the bands are independent in terms of their propagation characteristics. The coherence bandwidth may be different within different channels but it is always a significant constraint on frequency tone spacing (bandwidth allocation). A number of systems have been implemented using coherence bandwidths [78, 79, 82, 89, 92], enabling demodulation to take place separately for each frequency band. Channels that have previously been investigated have been found to have independently fading paths every 2kHz [75].

49. Frequency diversity techniques such as frequency hopping have been applied in radio communications [77]. Modifications of the basic frequency hopping techniques have been applied to the multipath environment [75, 76, 78, 79]. Frequency hopping in this case means the implementation of different frequency bands for the transmission of successive bits or symbols. Usually, the whole of the available bandwidth is used sequentially, as opposed to transmitting at the same time as in parallel transmission schemes. The modulation scheme used in the individual bands of a frequency hopping scheme is not restricted [79]; combinations can be used, such as Frequency Shift Keying (FSK) in some bands and Phase Shift Keying (PSK) in others.

50. The reliability of a frequency hopping system is maintainable as long as the channel reverberation duration is not too long. The period for the system to return to the same band should be longer than the channel reverberation time. The equation below can be used to estimate the value which channel reverberation time must not exceed to achieve sustained reliability. With a baud duration T_{bd} and a specified number of frequency bands n the limiting reverberation time T_r is defined:

$$T_r \approx (n-1)\, T_{bd} \qquad (2.1)$$

The number of frequency bands that can be used is restricted by the baud duration in combination with the time of arrival of the last reverberation component. Lots of frequency bands may help the situation but with more frequency bands the probability of Inter-Band Interference (IBI) is increased. If the baud duration is increased to overcome 'prolonged reverberation' the data rate can become severely restricted. There are advantages as well as disadvantage in implementing such a 'hopping' scheme. The hopping can partially clear the multipath signal and reduce signal saturation within a frequency band at the expense of additional decoder errors due to the tracking of the hop band sequence and extra bandwidth consumption [14]. Doppler shift that is evident in moving underwater systems can affect the frequencies, inducing interference between close frequency bands. There are circuits that provide Doppler correction [89] that can be applied to correct the shift at the cost of extra processing.

51. Another frequency diversity technique that could be used to limit the inherent signal fading is parallel transmission; each transmission conveys the same data. This can be achieved in the sub-sea media, because of the separate fading bands that are evident, as used in frequency hopping schemes. An overall value for the received signal is chosen, using some reliability measure to enable a choice of which of the frequency bands is more reliable. The reliability measure can be a measurable quantity, such as the level of received signal power in each frequency band.

52. Inter-symbol interference is not eradicated by these frequency diversity methods, so other schemes that combat ISI can be used in conjunction with the frequency diversity techniques. The high bandwidth utilisation of the frequency diversity techniques reduces their practicality in a lot of situations. The schemes are bandwidth inefficient but are flexible, so that different modulation schemes could be used in each of the bands or the same technique could be applied in all bands. The low data rate inherent in the extended symbol period technique is not found with these frequency diversity methods.

53. In the parallel transmission scheme the data can be different in each of its channels, allowing high data transfer rates to be implemented. If the parallel scheme is implemented to combat fading then the same data is transmitted over all the channels; consequently the data rate is not enhanced. The number of channels (bands) and their spacing is chosen to minimise Inter-Channel Interference (ICI) or Inter-Band Interference (IBI) [89]. In moving systems Doppler shift can cause ICI therefore the number of channels in a given bandwidth has to be chosen so as to maintain a frequency spacing between channels that is large enough to ensure that Doppler shift does not degrade performance.

54. The difference in arrival times exhibited by multipath signals (direct path and multipath components) can be used to manipulate the received signal and gain some advantage. The diverse time nature of multipath arrivals may even enable the direct path to be distinguished from the multipath signals. Methods that have previously been employed exploit knowledge of the multipath components arrival times, such as the burst transmission method (time gating) [81]. Systems operating under burst transmission [80, 81, 85] normally aim to avoid ISI altogether. Burst mode involves data being transmitted for short periods of time followed by a period with no transmission. The transmission period is implemented before the macro-multipath arrives, therefore only the direct path signal is detected and the multipath components are avoided.

55. High data rates are achievable for short periods of time (within the transmitted 'bursts') and for the rest of the time the data rate is zero. The system therefore provides a low overall data rate even though the data rate in the burst periods may be high:

$$Average-data-rate = (Burst-rate)\times\left(\frac{(Length-of-burst)}{((Length-of-burst)+(Length-of-reverberation))}\right) \quad (2.2)$$

The average data rate is sometimes even more highly restricted in a practical system as it may have to accommodate longer gaps in transmission than is theoretically required, as burst transmission implementations are influenced by the geometry of the channel.

The geometry dictates the difference between arrival times of the direct path and macro-multipath components. Therefore there must be adequate knowledge of a particular environment for the scheme to be implemented correctly.

56. The modulation scheme that is used in the 'burst' period is not restricted by the method itself and can be chosen independently. The burst transmission scheme addresses the problem of ISI but does not look at the problems brought about by the signal fading apparent in so many sub-sea channels. The phase wander due to macromultipath is avoided, but micro-multipath effects are still prevalent. The burst transmission scheme is inefficient and it becomes impractical where the reverberation time is large. The scheme is most effective where the reverberation is shorter and at least one symbol can be transmitted within a 'burst' transmission. The system mobility is limited, as the reverberation characteristics for each site need to be known before implementation. The channel-specific data has to be gathered by simulation or by experimental means prier to implementation.

57. Spatial diversity (channel diversity) can also be used to reduce the effects of signal fading and other multipath effects [93]. Multiple widely spaced hydrophones or array elements (several λ) can be used to take advantage of the spatial diversity. The use of a number of widely spaced hydrophones means that propagation-specific fading effects are likely to be independent for each hydrophone; thus the interference signals for each are incoherent. Spatially diverse channels are possible because widely spaced arrays are unlikely to have a flat wavefront incident across all elements due to ducting affects in the under-water medium [94]. Narrowly spaced arrays ($\lambda/2$) have interference signals that are almost coherent between the elements and consequently a coherent combination of the elements has no real improvement over a conventional array [94]. With widely spaced elements the same signal is effectively transmitted over the 'separate' channels and then can be coherently combined at the receiver.

58. Combination of the signals received can be achieved in a number of ways. A sum of all the received outputs or a selection of the channel with the most energy or SNR may be appropriate. In the sub-sea environment some form of adaptive combiner may be required to give weighting to each of the diverse channels. The weighting value for each channel can be found by the use of a known coded signal to estimate the error probability of each channel. Adaptive multi-channel combining uses an algorithm such as the Least Mean Square (LMS) algorithm to weight each channel. Receiver structures that coherently combine signals from a number of spatially diverse hydrophones or arrays have been shown to result in performance improvements [73, 95].

59. Another method that can be employed in sub-sea communications is the spread spectrum technique [96], which can suppress the effects of frequency-dependent signal fading [69]. Provision of 'security' against unauthorised interception of the signal [97] is an additional advantage of the scheme. The multipath signal, which can be likened to signals produced by a 'jamming signal', can be tackled using spread spectrum methods. Direct sequence spread spectrum (DSSS) systems have been implemented to discriminate against jamming by using bandwidth expansion [83]. In situations where bandwidth is not a problem a DSSS system can be effective in reducing multipath noise. The multipath is subjected to time delays by the direct spread spectrum technique, enabling the correct signal to be extracted at the receiver.

60. The system operates by *modulo*-2 addition of the data with a *pseudo*-noise sequence before modulation is undertaken. The signal produced by the *modulo*-2 addition has a frequency equal to that of the *pseudo*-noise rate (Rn). The original signal, with its lower rate Rs, is retrieved at the receiver by *modulo*-2 addition with a *pseudo*-noise sequence, which is frequency- and phase-locked to that at the transmitter. This technique works for multipath environments because when a *pseudo*-noise sequence is frequency- and phase-locked to the direct path, correlation with the multipath components is weak, so their bandwidth is not de-spread at the receiver. The two signals are thus 'bandwidth diverse'. A significant improvement can be achieved using direct spread spectrum for spreading ratios greater than 10. The signal-to-multipath ratio increases at the output of the data bandwidth filter by a ratio of Rn/Rs. With limited bandwidth situations the direct spread spectrum technique can cause a reduction in the overall data rate for the channel.

2.2.2.3. Channel Equalisation

61. Inter-Symbol-Interference has been compensated for in a number of less hostile channels by using an equaliser with a transfer function that, when cascaded with the channel's response, produces a flat magnitude and linear phase within the pass band. Multipath effects can be combated in some cases by applying the inverse of the channel transfer function at the receiver. The fluctuating nature of a sub-sea channel means that any static equalisation is impractical. Adaptive equalisation provides a way of tracking the channel transfer function whilst minimising the mean square error of the output at any time [98]. A reference signal is used by the equaliser, which minimises the error between the reference signal and received signal. The reference signal can be obtained by either periodically transmitting a known training sequence, or using a decision detection method.

62. The problems with implementing adaptive equalisation in a long-range shallow underwater environment are compounded by the combination of the rapid nature of the fluctuations in the channel and the extremely long reverberation times that are sometimes evident. The speed of variations in the transfer function will not allow the adaptive equaliser to converge in many situations and long reverberation times result in unrealistically long taps for filter implementation. Large filter implementations exhibit poor convergence properties and have increased noise enhancement as they increase in length [86]. Successful implementations of equalisation have been found [7, 17, 21, 86].

63. Implementation is normally in conjunction with techniques that can remove long multipath delays [84], e.g. the use of adaptive equalisation, together with spatial diversity (hybrid scheme), has been used [69]. The synergy of self-optimised receiver and spatial diversity has been shown to work in 'extremely shallow' channels with large range-to-depth ratios of the order of 10:1. In a non-minimum phase channel (dominant arrival not coming last), techniques such as the implementation of Decision Feedback Equalisation (DFE) can not be applied as the method requires past symbols for cancellation. Linear feedback implementations can be used in an attempt to minimise ISI but generally adaptive equalisation has been shown experimentally to be 'largely ineffective' in long range channels with non-minimum phase characteristics [94]. The use of multi-channel equalisation has been proposed and shown to produce better performance over long range channels than other equalisation methods [21, 93]. Multipath decision feedback equalisers can achieve near optimal performance but the high computational complexity, which commonly increases with the number of array elements, becomes a limiting factor at high symbol rates [47]. The long filter taps for each of the channels dramatically reduces the achievable symbol rate. The adaptation would have to be carried out continuously for rapidly changing channels if the technique were to work.

64. If the fluctuations are of a modest level then block adaptation can be implemented. Block adaptation involves the inclusion of 'training blocks' within the data so that the receiver parameters can be updated. One of the problems with equalisation stems from the fact that the channel response is not always known. In this case, implementation of an equaliser requires 'blind' equalisation. A number of methods for blind equalisation have been investigated [99-105]. Where the input symbols are of a finite alphabet the Viterbi algorithm can achieve the optimum estimate of the input signal [106, 107]. A number of the approaches are limited in the situation where the channel is time varying, as in the underwater case.

65. One recent approach applied to a Rayleigh fading channel goes about 'blind' equalisation by the direct examination of the input sequences [79]. By accepting a finite number of input sequences all can be looked upon as separate training sequences associated with a finite bank of filter models. By applying cost functions an approximate algorithm can be used to establish the most likely input sequence. Equalisation techniques have been found to be effective in some way in most channel configurations but a lot of the techniques available in equalisation are still in the development stage. However, they show promise even in such a fluctuating environment as the sub-sea medium.

2.2.2.4. Use of Angle of Arrival

66. Schemes that can be employed to reduce the multipath effects include the use of highly directional emitters and receivers, thus reducing the number of paths that are available. A directional antenna mitigates the effects of reverberation by increasing the directivity index and concentrating the power of the transmitted beam [69]. When the directivity of the transmitter and receiver are raised, pointing errors (due to offsets in alignment) can become an issue.

67. Narrow beam transducers such as parametric arrays use the underwater channel's non-linear behaviour to produce the beam [21, 87]. Two 'primary' frequencies are applied to the array. Combination of the two frequencies occurs a long the transmission axis in an 'end-fire' mode due to the non - linear properties of the water. The resultant is made up of sum and difference signals, the difference (or 'secondary') signal being capable of transmission over a greater range than the sum frequency or the primary frequencies. The non-linear action of the medium therefore generates a highly directive virtual end-fire array at the difference frequency.

68. Efficiency of such a transducer is low, typically only 1%, but this is overcome by the high transmitter directivity and the reduced absorption loss when transmitting at a lower frequency. In some situations excellent results can be achieved when using a fixed transmitter and receiver transducer system. Problems with the parametric system become evident when the transducers are not fixed, including complications with pointing errors, thus requiring some form of mechanical alignment [100]. There are also problems with the practicalities of the parametric source in terms of the large power requirements [21, 23].

69. Narrowing the active beamwidth of a transducer can be achieved by beamforming. This can be combined with some method of beam steering to enable communications to be maintained when using narrow beamwidths. Beem streering can be carried out at the receiver (nullifying unwanted signals) or at the transmitter (reducing boundary interaction). The beamforming technique requires a number of transmitter hydrophones or array elements to achieve narrow beamwidths. The main lobe of the beamforming hydrophone can be steered to any angle required. Beam steering can be achieved mechanically, which is a practical solution for fixed systems, but problematic for moving systems. Problems are encountered in terms of alignment and tracking of the moving system; extra data is required so the system can 'navigate'. There are also methods that use electronic beam steering, which involves the use of electronic delay elements to steer the beam to achieve the maximum signal level. The beamforming technique has been used over short distances in the underwater environment [108].

70. Problems have been encountered with aspects of the technique, including the level of angular resolution that is achievable and the ability to measure the angle of direct path arrivals. Adaptive beamforming can provide a higher level of angular resolution than basic beamforming [71, 107, 109]. The angular resolution required and the number of degrees of freedom (peaks and nulls that can be steered) dictate the number of array elements required. Long transmission ranges provide a situation where some multipath components may arrive from directions that differ from the direction of arrival of the direct path only slightly outside the resolution of the adaptive beamformer.

71. Adaptive equalisation techniques can be used in conjunction with adaptive beamforming, with the former compensating for the time diversity of arrivals. Beamforming may be effective for small range-to-depth ratios, but in a long-range channel equalisation is required because of small angular multipath separation [21]. Techniques for combined beamforming and equalisation (hybrid systems) have been investigated within other media [110] and there have also been proposals to apply these methods to the underwater channel [111]. Other proposals have included spatial diversity, combining multiple-channel decision feedback equalisation simultaneously with beamforming [21, 73].

72. When high rates of angular fluctuations are found, as in numerous underwater channels, a beamformer's adaptation ability to track the optimum angular response is severally restricted [94]. Training of the beamformer can be carried out periodically in some cases where the angular fluctuations are not too rapid and stay constant over numerous symbol periods. Over short-range channels, periodic training of a beamformer has been applied [71, 112] and also block adaptation, as used in equalisation. The general implementation for beamformer training involves the response being held for duration (reception period). As long as the incident multipath arrival angles stay virtually constant over several thousand symbols the technique works [94]. This method of adaptive beamforming is limited by the symbol rate of the system; higher symbol rates allow reception of data while the response of the channel is relatively stable. In channels where the response is rapidly changing over a few symbol periods a method of continual adaptation during the reception period is required.

73. A lot of the schemes that have been developed to reduce multipath pay the price of high bandwidth usage. The use of an adaptive beamformer to spatially filter the required direct path signal from the multipath can provide bandwidth savings in comparison with most other schemes that can be employed [71]. The large bandwidth wastage exhibited by many of the other adaptive antenna methods is unworkable in restricted bandwidth communications [113].

Concluding Remarks

The sub-sea environment is made up of many complex phenomena that combine to produce an extremely restrictive environment for through-water communication. In particular the long-range shallow-water channel can exhibit characteristics that combine to negate robust reception. Multipath and frequency specific fading are just some of the most restrictive of the phenomena that inhibit acoustic signal propagation. A number of techniques have been used in the past to try and overcome the problems found in this environment but most have been found to be lacking. A combination of different techniques is the mostly likely to provide a solution in such an environment. One of the areas that needs to be considered in more depth is the modulation technique itself and how it can be made more robust in such a restrictive environment. The following chapters describe the modulation techniques that have been developed through this research, starting with some background.

Chapter 3. Background to Modulation Schemes

1. The modulation techniques available to a communications engineer whilst implementing an acoustic though-water data telemetry system are limited, especially when a long-range shallow sub-sea channel is being considered. This is due to the complex nature of the acoustic signal's interaction with the environment. Signal instability that is inherent to such channels takes the form of amplitude and phase fluctuations. These fluctuations can mostly be attributed to the acoustic signal's multipath propagation characteristics, which limit the effectiveness of a number of modulation techniques that have been used reliably in other channels. Use of phase- and amplitude-based modulation schemes in shallow long-range channels, is limited by the 'extreme' nature of the signal propagation, in particular the phase and amplitude randomisation [94].

2. The choice of which modulation scheme to use is restricted by the type of hardware being designed, e.g. the number of receiver transducers, antenna directivity, etc. The occurrence of phase and amplitude instabilities particularly renders the phase- and amplitude-based modulation techniques ineffective when a single omni-directional receiver system is envisaged, as is the case with this work. Techniques have not been developed to date that effectively compensate for the complex channel transfer functions found in shallow long-range environments without the use of widely spaced transducers (or array elements) and/or highly directional antennas. The modulation techniques developed for this research are aimed for use with a single omni-directional transmission projector and a single omni-directional receiver hydrophone. This precludes the use of spatial diversity techniques (which require multiple widely spaced receiver elements) and the use of directional antennas.

3. The use of phase-based modulation schemes in long-range, sub-sea environments has increased recently [5-7, 13-15], which can mainly be attributed to the developments in spatial diversity processing algorithms which now can be used to overcome problems associated with phase signalling in these extreme situations [12]. The difficulties experienced with phase modulation and amplitude modulation in this reverberant environment, when using single omni-directional transducers, means that the most viable option for this acoustic data telemetry system is the use of some form of frequency signalling (modulation), based on frequency shift keying modulation [23]. Frequency-based signalling can take place effectively as long as the transmitted frequency is dominant at the receiver and the signal does not fade to a point that it can not be detected. The frequency-based schemes are primarily restricted by bandwidth limitations, frequency specific fading (FSF) and Inter Symbol Interference (ISI).

4. The choice between the implementation of coherent or non-coherent operation becomes a straightforward one when the channel to be transversed is a long-range, shallow sub-sea channel, and a single omni-directional receiver transducer is to be used. Coherent systems are generally able to provide greater data throughput compared to non-coherent signalling, but a coherent system's sensitivity to the non-stationary behaviour of a signal extracted from a shallow, underwater channel restricts its viability. The strong medium fluctuations negate coherent carrier-phase tracking; thus non-coherent signalling becomes the more viable alternative [27]. In channels where fluctuations are low enough not to exceed maximum levels for coherent signalling, coherent carrier acquisition and tracking is obtainable. Sub-sea environments that can accommodate coherent systems; include near-vertical deep-water channels [114]. Throughout this research non-coherent implementations are used.

3.1. Modulation Techniques Overview

5. A description of a number of the different modulation techniques that have recently been applied to the sub-sea environment follows together with an overview of most of the techniques that underpin their operation. The majority of the schemes are based on the Frequency Shift Keying (FSK) technique; there are also a number of techniques that utilise Phase Shift Keying (PSK)-based modulation protocols. The operation of the modulation techniques that use multiple symbols is covered (MFSK, MPSK); including variations that are appropriate for sub-sea through-water communications (SFSK).

3.1.1. Frequency Shift Keying (FSK)

6. Frequency shift keying (FSK) is a modulation technique that has often been utilised in the underwater medium. In its most basic form, the symbol alphabet consists of one frequency tone representing a bit one and a second frequency tone representing a bit zero. There is therefore generally a higher bandwidth requirement (to accommodate multiple symbol frequencies) when using frequency signalling modulation schemes than for phase or amplitude signalling schemes. This is particularly apparent in modulation schemes that require large numbers of symbols in their alphabets. The FSK technique is particularly useful in lower rate communications that transverse noisy channels (phase and amplitude instabilities); at high bit rates FSK becomes 'bandwidth expensive'.

7. One form of the FSK technique is Minimum Shift Keying (MSK); this involves transmitting the FSK signal with continuous phase transition between bit periods. The MSK technique is sometimes referred to as continuous phase FSK. The MSK technique eliminates the production of some of the unwanted high frequency components that can be found in FSK transmissions. Basic FSK is an adequate method in itself if the relatively large spectral side lobes caused by the phase discontinuities at the switching instants do not disrupt correct reception.

8. FSK symbol alphabets are not complex to generate; the basic tones can be produced by a group of separate oscillators tuned to the desired symbol frequencies. Division of a single oscillator frequency down to the appropriate symbol frequencies can be used to produce continuous phase FSK transmissions; the data sequence can be used to define which symbol is to be transmitted at any instant. The performance attainable with a single oscillator system is equivalent to that of a system that uses a group of separate oscillators producing an FSK signal with phase discontinuities, except for the benefit of a narrower spectrum. The narrower spectrum is possible because of the absence of abrupt transitions in the single oscillator system.

9. A FSK receiver that utilises an orthogonal symbol alphabet has to make a choice from the set of orthogonal signals over the symbol period; the symbol transmitted is dependent on the data. An *a priori* knowledge of each of the symbols that can be transmitted enables qualitative analysis of the received signal (transmitted signal plus channel noise) to take place at the receiver; in this way the transmitted sequence of symbols can be detected at the receiver. A FSK signal allows modulation onto a carrier waveform before transmission; thus it can be placed in a specific bandwidth. A FSK scheme can use coherent or non- - coherent demodulation, as the data is carried in the frequency component of the signal not the phase; phase continuity is not required between the transmitter and the receiver for symbol evaluations.

3.1.1.1 Multiple Frequency Shift Keying (MFSK)

10. A Multiple Frequency Shift Keying (MFSK) system involves the use of an arbitrary number of M frequencies (more than two) that can be used to transmit data. These frequencies are not used in parallel as with Space Frequency Shift Keying (SFSK); instead, one frequency element (tone) is communicated at a time. For optimum system operation the energy per element is kept constant by keeping the amplitude and time duration constant for each element. The transmitted signal is normally modulated onto a carrier before transmission. The relative performance of systems with differing numbers of M frequencies is characterised generally with improvements an increase in the number of frequencies implemented, rapidly at first but then becoming negligible for large numbers.

11. Four frequency tones are used to transmit data in a 4FSK modulation scheme. Each frequency can be assigned a two-bit code-word as shown in Table 3.1. In this case the systems alphabet are assigned one frequency tone each. An individual symbol is representative of a distinctive two bit code-word, the data rate is thus increased as each symbol transmission represents two bits.

BITS	SYMBOL	FREQUENCY
00	*S1*	*F1*
01	*S2*	*F2*
10	*S3*	*F3*
11	*S4*	*F4*

Table 3.1 Each frequency is assigned a two bit code-word

12. A Coded 4FSK (C4FSK) system can be implemented providing a more robust alternative to basic 4FSK. When implementing a scheme that uses *'character to frequency coding'*, each symbol of the input alphabet is represented by a sequence of frequencies instead of a single frequency as in 4-FSK. This can be achieved by one of two methods:

A. Retaining the same number of tones as in an 'uncoded' alphabet, and placing more than one element in each bit period. The number of codewords is reduced as well as the number of bits in each code word. Therefore there is a reduction in the system bit rate.

B. Increasing the number of tones in the symbol alphabet compared to an 'uncoded' scheme, each extra tone is used to make a sequence of tones in each bit period. This retains the same number of codewords in the code-word alphabet but this requires an increased bandwidth allocation, as there are large number of frequency tones used.

This means there is a sub-bit level of coding (sequence of symbols within each bit period).

13. Either of the two methods provides sequences of frequencies that can be sent within each bit period, dependent on the bit value, or transmitting 'checking' information. For a single element scheme such as FSK to be made into a 'coded' scheme, method B above is appropriate to do this an extra two extra frequencies is required, *F3* and *F4* in Table 3.2. A sequence of frequency tones can thus be assigned to each bit value.

BIT	SYMBOL	FREQUENCIES
0	***S1***	***F1, F3***
1	***S2***	***F2, F4***

Table 3.2 Bits are assigned two frequencies

14. A MFSK modulated scheme can use either coherent or non-coherent demodulation techniques at the receiver. Hybrid techniques can also be utilised, such a case is possible where the signal can be coherently tracked until a time when fluctuations are 'to high', at which point the demodulation is switched to non-coherent[75].

3.1.1.2. Space Frequency Shift Keying (SFSK)

15. Space Frequency Shift Keying (SFSK) is a modulation technique that uses multiple frequency tone combinations to convey data; the tones that are transmitted in parallel do not convey any additional data. The SFSK principle aims to utilise the fact that there are many separate bands of fading (frequency diversity) in a sub-sea medium. At any one instant different frequency bands fade more than others do, so that when data is transmitted over more than one frequency band the probability of correct reception is increased.

16. If all the combinations of parallel frequencies need to be detected at the receiver, this would provide no better performance than a simple FSK scheme, as one fading frequency would produce an error. The concept benefits from using a 'soft decision algorithm' receiver; where by the nearest ('most likely') combination of the discrete set of frequency combinations is chosen. With an, *a priori* knowledge of the possible combinations a decision can be made even if some of the frequency bands have encountered destructive fading.

17. As long as each frequency set (combination) is unique (none of its frequencies are part of another set) then a number of separate frequencies would need to be corrupted to sustain an error at the output. Performance is enhanced if the coding is chosen to be unique, so that the signal combinations used have the greatest 'distance' between any two signal combinations.

3.1.1.2.1 Eight-Space frequency shift keying (8-SFSK)

18. An 8-SFSK modulation scheme would use combinations of the eight frequency tones to represent each of the code-words in the alphabet. A soft decision reception algorithm could be used, the 'most likely' of the possible combinations being picked for each receiver decision. The assessment of which frequencies are present at the receiver can be made by pairing off each frequency assigned to 0 with one assigned to 1, then treating each pair as a BFSK signal. This is appropriate, as one of each pair of frequencies should be sent at any one time. The assessment then becomes how many of the pairs of frequency's point to a zero bit transmission or a one bit transmission (Fig. 3.1).

19. Variations on this theme are described below:

A. The assignment of all eight frequency tones (symbols), to one of the bit values can be carried out. Transmission of bit 1 can thus be carried out by sending all the frequencies and bit 0 by sending no symbols. The received frequencies would have to be compared against some 'measurement' above which a tone is considered as being detected, and below which a tone is considered as not being detected.

B. Two different frequency assignments from that in Fig. 3.1 can be used, one frequency set being sent for a bit 1 and the other sent for a bit 0 (Fig. 3.2). An, *a priori* knowledge of the possible frequency sets allows the received to assesses the 'most likely' bit value.

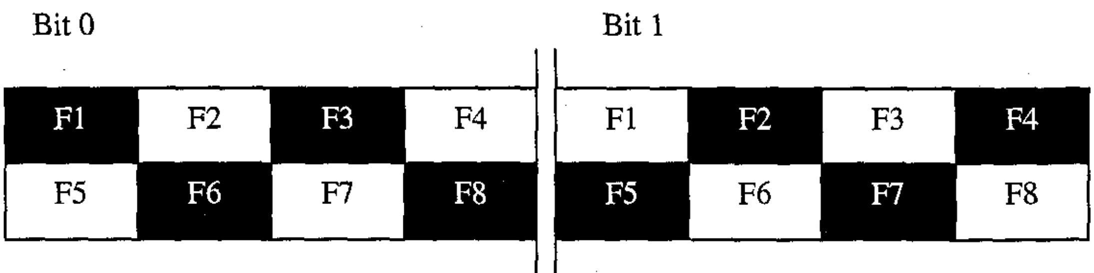
C. A scheme that may be more robust could use one of the two frequency sets in each symbol period, but it could also send another frequency set as a check in the next symbol period. One of the frequency sets followed by its 'check set' is sent for a bit 1 and the other frequency set followed by its 'check set' for bit 0. Bit 0 could be assigned F1 and F3 followed by F6 and F8 and bit 1 could be defined by F5 and F7 followed by F2 and F4.

D. Four frequency sets could also be used so that two bits can be transmitted at one time. One frequency set can be transmitted for each pair of bits: 00, 01, 10, and 11.

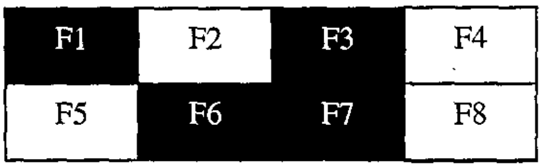
3.1.2. Phase Shift Keying (PSK)

20. Phase shift keying (PSK)-based modulation techniques have recently been shown to be applicable in long-range sub-sea communications channels [5-7, 12-14]. The randomisation of phase that is associated with transmission through this environment has limited the effectiveness of PSK in the passed. Without the ability to use spatial combining and directional antennas, PSK techniques are still severely restricted in long-range, shallow sub-sea channels. Techniques used to compensate for these channel effects in other less harsh media have been unsuccessful to date in this environment unless they are combined with spatial processing.

Frequencies are assigned to each bit value:



The frequencies are paired off for detection purposes: F1& F2, F3 & F4, F5 & F6, F7 & F8. Received frequencies: (F1, F3, F6, F7)



Three out of the four pairs of frequencies point to a bit 0 transmission.

Fig. 3.1 Soft receiver decision

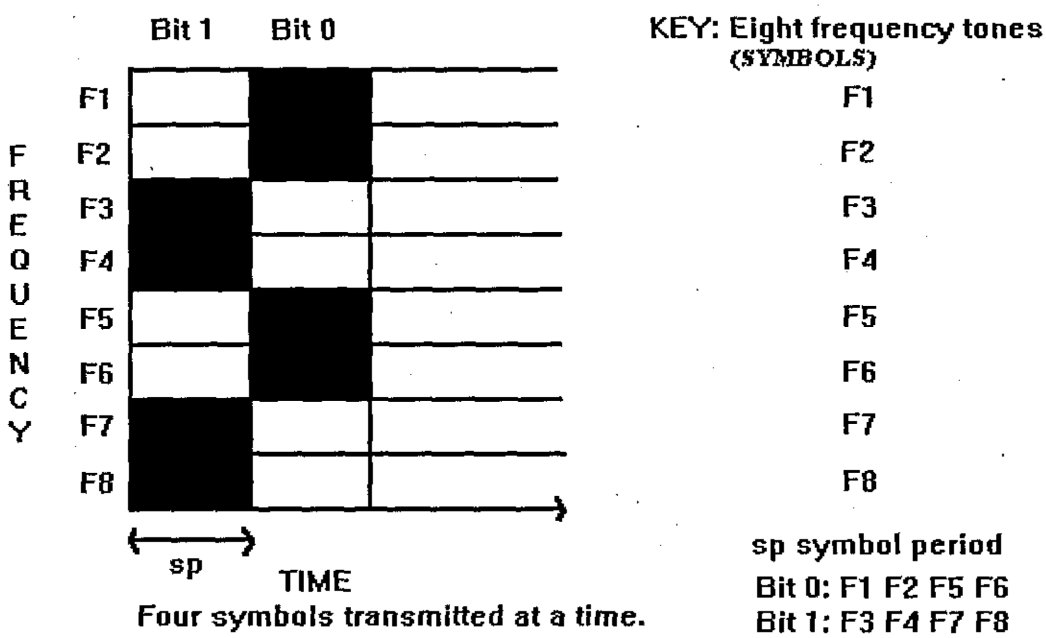


Fig. 3.2 SFSK Using two code-words

21. PSK requires the use of only one frequency tone because the phase component of the signal carries the information. PSK in its most basic form involves the assignment of a constant phase to each of the two possible bit values i.e. Binary Phase Shift Keying (BPSK); the frequency and amplitude are kept constant. The symbol alphabet therefore consists of a frequency tone with one phase representing a bit one and a second phase representing a bit zero. There is generally a lower bandwidth requirement when using PSK-based modulation schemes than when using frequency signalling. This bandwidth advantage is particularly apparent in modulation schemes that utilise large numbers of symbols. The PSK technique is particularly useful in higher rate communication channels where phase noise can be overcome; an increase in the size of the symbol alphabet is not directly related to an increase in the bandwidth requirements as is the case with FSK-based schemes.

22. The BPSK when the phase assignments are 180 degrees from each other or grey coded QPSK is most effective in terms of establishing minimum bit error rates. PSK symbol alphabets can be generated similarly to FSK signals; the PSK frequency tones can be produced by a group of separate oscillators generating the desired symbol phases. A single oscillator can also be used with a direct path and delayed paths; the data sequence can then be used to define which symbol is to be transmitted at any instant.

23. A PSK receiver utilising a discrete symbol alphabet is required to make a choice from the set of symbols in the symbol alphabet, over a symbol period; the symbol transmitted is dependent on the data. An *a priori* knowledge of each of the possible symbols that may be transmitted enables the analysis of the received signal (transmitted signal plus channel noise) to be simplified; the choice is between the symbols in the symbol alphabet. A PSK scheme uses coherent demodulation; as the data is carried with phase, and signal phase continuity is required between the transmitter and the receiver for symbol evaluation. .

24. When the absolute phase of the transmitted signal is not known, coherent techniques can be used to detect changes in phase (phase shifts); as is done with DPSK techniques. DPSK uses shifts in phase to represent data; if a one bit is transmitted the phase does not change; if a zero bit is transmitted a change of phase occurs. The receiver has to compare neighbouring intervals to determine which bit value has been sent; the absolute value of phase on its own is not adequate to establish the transmitted bit value.

3.1.2.1 Multiple Phase Shift Keying (MPSK)

25. A Multiple Phase Shift Keying (MPSK) system involves the use of an arbitrary number of M phase assignments (more than two) that can be used to transmit data through the channel. The most intuitive MPSK scheme is the Quadrature Phase Shift Keying (QPSK) technique where M=4. One form of QPSK is equivalent to the simultaneous transmission of two separate PSK signals. This is achieved using coherent techniques to distinguish between four possible orthogonal signals (Fig. 3.3). QPSK can also use the four symbols to represent one of four pairs of bits: 00; 01; 10; 11 (Table 3.3, Fig. 3.3), with one symbol transmission in each symbol period. The baud rate is half of that of a BPSK implementation but the error susceptibility is increased, as the thresholds between symbols are closer. The error performance of systems with differing numbers of M phase symbols is degraded as the number of symbols implemented increases, due to the separation between symbols being smaller.

26. Variations on the QPSK scheme includes orthogonal QPSK which negates the use of large phase changes between neighbouring symbol periods; phases can only change by 90 degrees, which suppresses the sidelobes produced by QPSK schemes that have abrupt changes at symbol transitions. The benefits produced by implementing orthogonal QPSK in terms of the reduction in phase transmission magnitudes can be further improved using a MSK continuous phase implementation.

27. The MPSK technique can involve the use of larger numbers of M symbols. Higher bit rates can be achieved with a larger number of symbol assignments; this is achieved at the expense of the error susceptibility because the symbols are placed closer together. The separation of the symbols, assuming uniform spacing, is defined by $2\pi/M$. A variation on the basic MPSK technique combines MPSK with MASK techniques to produce Quadrature Amplitude Modulation. QAM, which utilises a distinct distribution of symbols in the signal space (Fig. 3.4). Placement of symbols as far from each other as possible helps to establish better performance. In this scheme amplitude as well as phase modulation is used to distinguish between symbols.

BITS	SYMBOL	SIGNAL
00	*S1*	$+ \, cosine \, (2 \pi t) + sine \, (2 \pi t)$
01	S2	$+ \, cosine \, (2 \pi t) - sine \, (2 \pi t)$
10	*S3*	$- \, cosine \, (2 \pi t) + sine \, (2 \pi t)$
11	*S4*	$- \, cosine \, (2 \pi t) - sine \, (2 \pi t)$

Table 3.3 Each signal phase - Assigned a two bit code-word

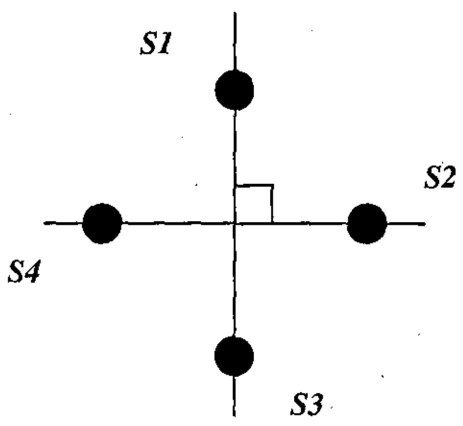


Fig. 3.3 QPSK Symbol Phase Assignments

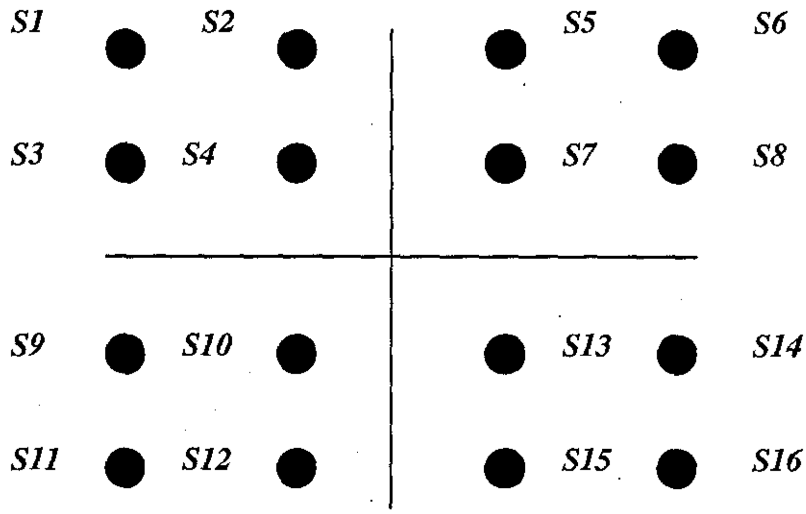


Fig. 3.4 16-QAM Symbol Amplitude Phase Assignments

3.1.3. Amplitude Shift Keying (ASK)

28. The use of modulation techniques based on Amplitude Shift Keying (ASK) are restricted in long-range sub-sea communications channels; this is due to the randomisation of the acoustic signals amplitude within this environment. A short overview of ASK based techniques is included here to complement the sections on FSK and PSK.

30. Amplitude Shift Keying (ASK) requires only one frequency tone, since it is the amplitude component of the signal that carries the information. ASK in its most basic form involves the assignment of a sinusoid of constant amplitude to a bit one and a zero amplitude to a bit zero; this form of ASK is defined as On-Off Keying (OOK). This is a form of Binary Amplitude Shift Keying (BASK), which involves the transmission of symbols consisting of two different amplitudes. A BASK symbol alphabet consists of a frequency tone with one amplitude representing a bit one and a second amplitude representing a bit zero. There is generally a lower bandwidth requirement when using ASK based modulation schemes compared to frequency signalling-based modulation schemes, as only one frequency is normally required in ASK schemes.

31. ASK symbol alphabets can be generated in a straightforward way; the frequency tones can be produced by a single oscillator. For OOK transmission an oscillator output can be passed or not passed by a switch, depending on if a one bit or zero bit is being transmitted. In the case of BASK schemes where both bit values are represented by signal amplitudes that are non-zero, a single oscillator can be used, the two required amplitudes can be obtained from the single oscillator output.

32. An ASK receiver that utilises a limited symbol alphabet is required to make a choice ever symbol period from the set of symbols in the symbol alphabet; the symbol transmitted depends on the data. An *a priori* knowledge of each of the possible symbols that may be transmitted enables the analysis of the received signal (transmitted signal plus channel noise) to be simplified to a choice between the symbols out of the set that can be transmitted.

3.1.3.1 Multiple Amplitude Shift Keying (MASK)

33. A Multiple Amplitude Shift Keying (MASK) system involves the use of an arbitrary number of M amplitude assignments (more than two), that can be used to transmit data through the channel. The simplest MASK scheme is where $M = 4$. Coherent techniques can be used to distinguish between four possible symbols that represent one of four pairs of bits: 00; 01; 10; 11 (Table 3.4). The bit rate is increased in this way without reducing the system symbol period. Further increases in bit rate can be achieved by increasing the number of M symbols; in a scheme with $M = 8$, each symbol transmission can be assigned a three-bit code-word (Table 3.5). The error susceptibility of symbol detection is increased as the thresholds between symbols are closer to one another. As the M symbol amplitudes implemented increase, the error susceptibility increases.

34. The MASK technique can involve the use of larger numbers of M symbols. Higher bit rates can be achieved with a larger number of symbol assignments; this is at the expense of the error susceptibility as the symbols spacing decreases. A variation on the basic MASK technique combines MPSK to produce Quadrature Amplitude Modulation (described earlier).

BITS	SYMBOL
00	**S1**
01	**S2**
10	**S3**
11	**S4**

Table 3.4 Each signal amplitude is assigned a two-bit code-word (M = *4*)

BITS	SYMBOL
000	**S1**
001	**S2**
010	**S3**
011	**S4**
100	**S5**
101	**S6**
110	**S7**
111	**S8**

Table 3.5 Each signal amplitude is assigned a three-bit code-word (M = *8*)

3.1.4 Multiplexing techniques, spread spectrum and frequency hopping

35. Multiplexing techniques can be used to set up multiple parallel data transfers over a single communication link. This is applicable in situations where one transmission is not using the full capabilities of the system, in terms of bit rate or bandwidth. The are a number of types of multiplexing available including:

1. Time Division Multiplexing (TDM)

2. Frequency Division Multiplexing (FDM)

3. Code-Division Multiple Access (CDMA), using direct spread spectrum techniques

36. Time division multiplexing facilitates the transfer of multiple transmissions sequentially within a channel. This can be achieved as long as the maximum bit period limitation of the link is larger or equal to the combined bit rates required for component transmissions. Fig. 3.5 shows a case where four separate transmissions of four bits are multiplexed into one transmission. Each of the four bits has the same bit rate requirement, which is one fourth of the bit-rate of the resultant multiplexed signal. Each 'bit period' of the multiplexed signal actually holds four sequential bits, one for each of the four separate transmissions.

37. Frequency division multiplexing facilitates the parallel transmission of multiple signals through the channel [115]. In situations where the bandwidth is not completely used, other communications can be established within the unused bandwidth. The additional frequencies used should not overlap with or fall within the matched receiver bandwidth of the other signals, when matched filter reception is used (Fig. 3.6). Orthogonal frequencies should be used to optimise performance (Appendix A).

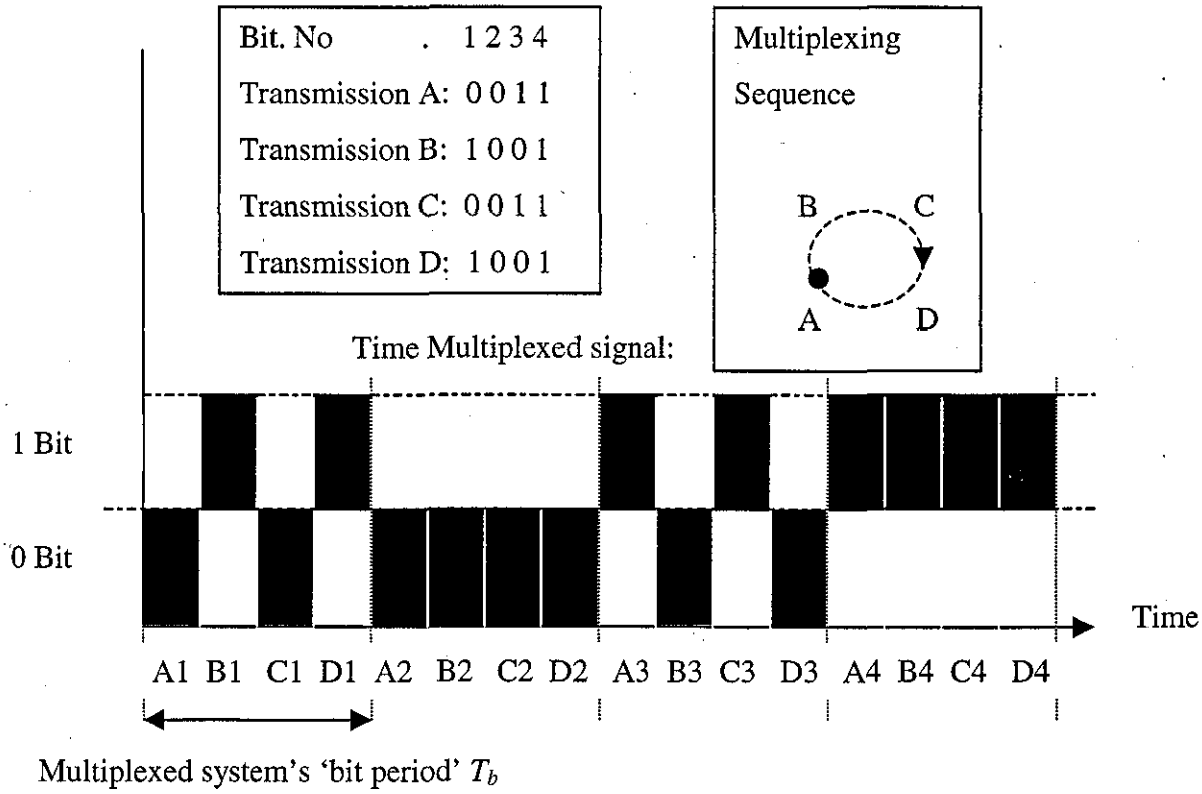


Fig. 3.5 Time Division Multiplexing

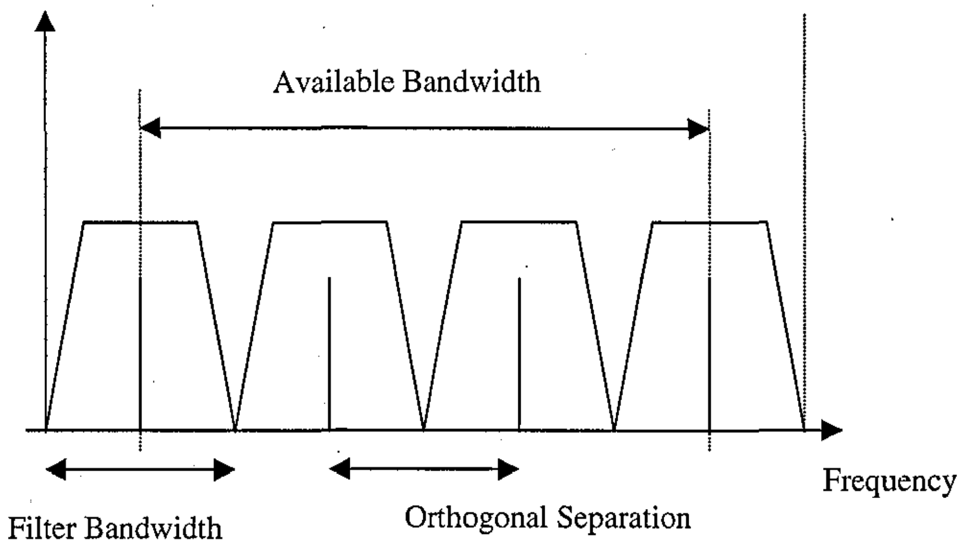


Fig. 3.6 Frequency Division Multiplexing

38. Spread spectrum is a wideband technique that generally requres a bandwidth of the order of 10 to 100 times that required for the information transfer [116]. In systems that are not bandwith limited the technique can be used to camouflage the signal and combat 'jamming' signals. The spread spectrum signal resembles wideband noise and therefore is less likely to be noticed by a uninitiated obserever. For a signal to be categorised as spread spectrum the signal must be dependent on a signal other than the base-band signal, thus wideband FM does not constitute a spread spertrum signal [116].

39. Direct sequence spread spectrum *pseudo noise* (*PN*) spreading is carried out by modulating a noise waveform with a data signal (binary) to produce a resultant signal that still looks random. The noise waveform is a binary *PN* sequence that approximates wideband noise whose spectrum is proportional to the bit rate and whose frequency is much higher than the data signals bit rate. The resultant signal is thus made up of blocks of signal equal to the data bit period and has characteristics that are equal to the *PN* sequence (noise signal) for a bit 0 and the inverse (*modulo* 2) of the *PN* sequence for a the bit 1 as shown in Fig. 3.7. The equation below defines the process, where $Sp(t)$ is the spread spectrum signal, $d(t)$ is the data, $n(t)$ is the noise and + is a *modulo* 2 addition:

$$Sp(t) = d(t) + n(t) \tag{3.1}$$

The *PN* sequence is produced at the receiver and used to de-spread the signal. The receiver has a noise advantage over narrow band schemes, which can be put down to the spreading of the noise that occurs when the *PN* sequence is introduced.

40. The third multiplexing technque mentioned earlier (code-division multiple access (CDMA)) uses spread spectrum techniques. In the case of CDMA, orthogonal *PN* sequences are used. Each *PN* signal is summed with one of the data signals (multiple transmissions), then the addition of the resultant signals takes place. The *PN* sequences are chosen to have an equal frequency that is a lot higher than the data signals bit rate. At the receiver the *PN* sequences orthogonality cames into play. When the received signal is summed with one of the *PN* sequences, the product of two different orthogonal (uncorrelated) components of the signal produces a noise waveform (at the frequency of the original noise waveforms used at the transmitter).

Filtering can therefore be used to reduce the effects of the product of the separate *PN* waveforms. Matched filtering techniques can be used to perform the detection process, limiting the crosstalk effects since the *PN* sequences are orthogonal [116] (Appendix A).

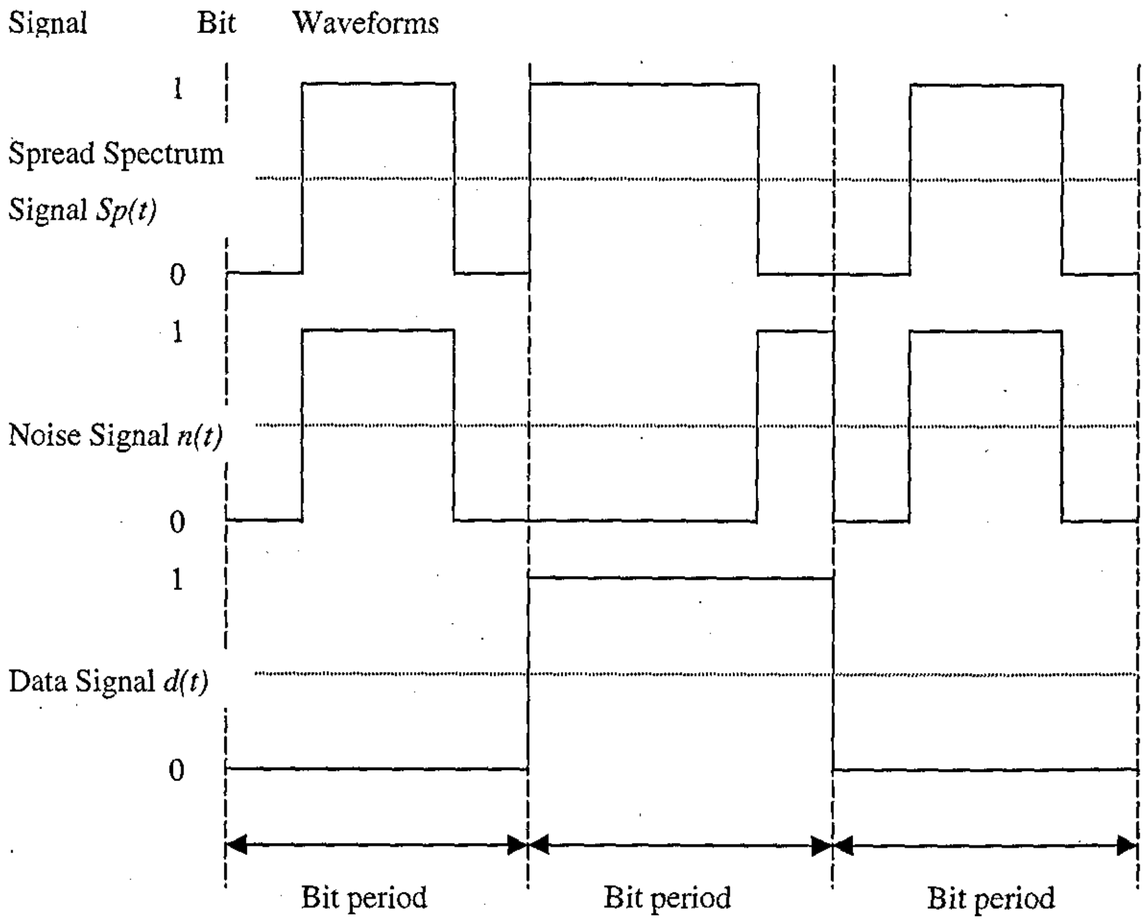


Fig. 3.7 Spread spectrum signal

41. Frequency hopping uses a number of different frequencies (symbols) to convey a single transmission. A *PN* sequence can be used to control a 'hop' sequence, which defines the active frequencies at any given time. The sequence is cyclic, with a number of frequencies (or sets of frequencies) being used before repetition (Fig. 3.8). The transmitter 'hop' is exactly mirrored at the receiver by using a locally generated version of the *PN* sequence. The active frequencies are evaluated at all times. Spacing of the frequencies in a frequency hopping scheme are generally orthogonal and equi-distant across the allocated bandwidth , so that matched filter reception can be used effectivly.

3.1.5. Modulation Schemes Recently Used In Sub-Sea Applications

42. This section describes a number of communication systems that have been used for numerous sub-sea applications. A brief account of the modulation schemes that have been implemented follows, along with relevant results and conclusions. There are numerous systems that have been developed for implementation in the underwater environment [8, 83, 87, 117-137, 145-148, 150, 151, 157, 159]. The developments in long-range, shallow–water sub-sea communications channels are of particular interest and are focussed on here [8, 83, 87, 117-137].

43. The requirement for control of, and data acquisition from, remote sub-sea installations began to increase in the early 1980's, providing the catalyst for more intense exploration into modulation techniques that would be able to stand up to the acoustically inhibitive environment. A number of techniques have been used, the choice of modulation scheme being predominantly dictated by a number of factors. These include the nature of the signal interaction with the environment; the techniques available for overcoming any inhibitive effects of signal environment interaction; the hardware available (e.g. number of receiver transducers); the bandwidth available; synchronisation requirements; and the required data rate.

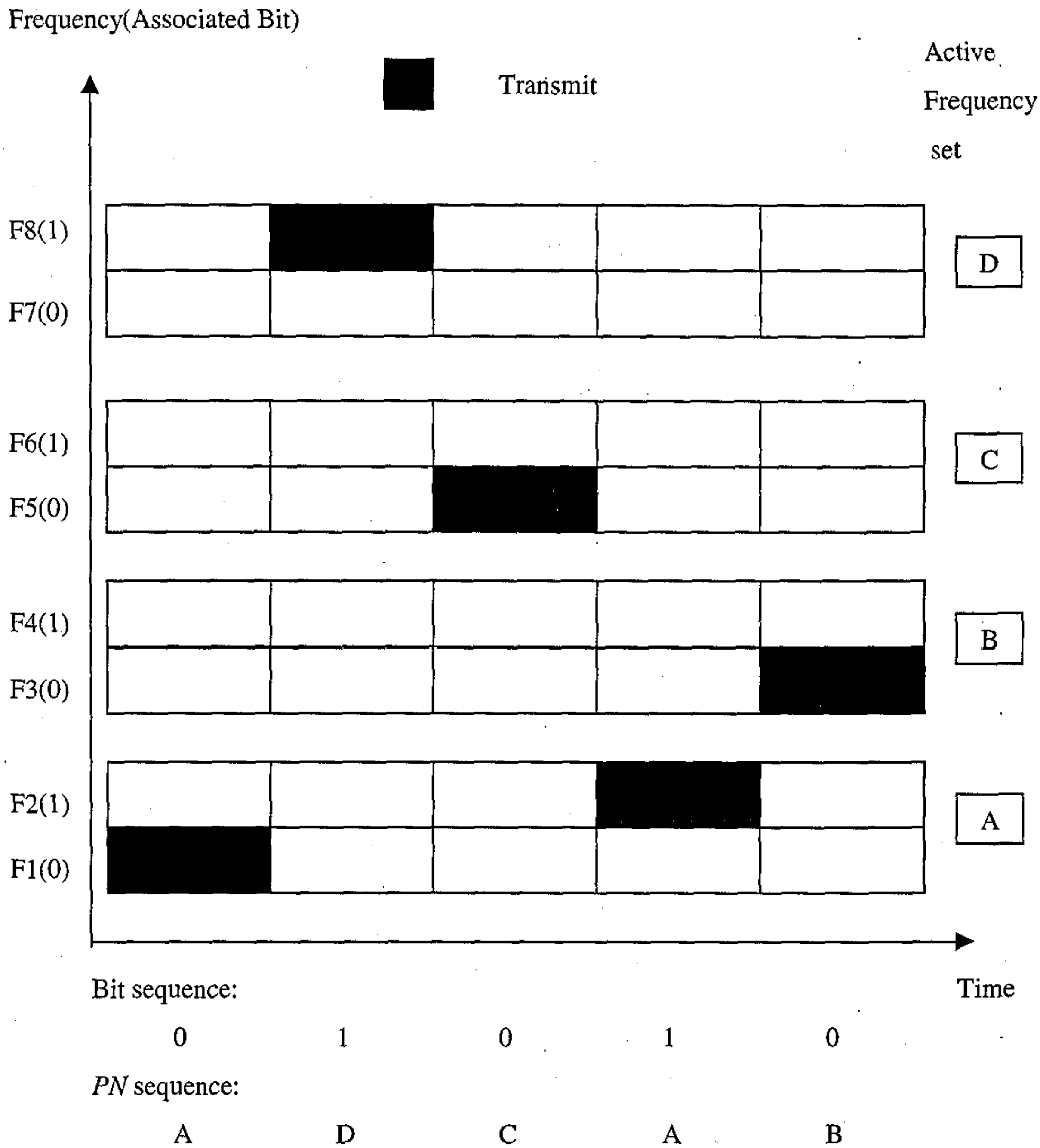


Fig. 3.8 **Frequency hopping**

3.1.5.1. Frequency-based Implementations

44. Multiple Frequency Shift Keying (MFSK) modulation has been seen as a possible means of providing reliable communications since the early 1980's. At the *Oceans '81 Conference* a number of papers were published dealing with MFSK systems, their applications, and the sub-sea environment, with an emphasis on robust acoustical communications [76, 138]. MFSK is considered one of the most promising techniques available to the sub-sea communications engineer, because it does not have a reliance on frequency and phase stability and there is potential to utilise/overcome frequency selective medium characteristics.

45. The MFSK experimental system envisaged in 1981 was designed for deployment over long-range (up to 10 kilo-yard) channels, and a predictions of possibly achieving 99% reception of messages under worst-case noise and multipath conditions was made. The system specifically used energy detection of frequency and time diverse MFSK, with four diverse frequency tones used for data transmission and two frequency tones for initialisation. Simulations were run with individual error sources being applied separately to determine their individual effect on communications. The worst-case situation was then mimicked, by combining all the error sources in one simulation. The largest single source of error was found to be the Doppler frequency shift (due to relative movement of transmitter and receiver), with none of the errors found to stem from any other noise source during testing. Doppler shift handling was improved by applying a Hamming window algorithm to the data, increasing the processing bandwidth by three at a cost of 50% more noise bandwidth [136]. An in-water one-way link was established in a lake (Washington Lake) for experimental purposes, covering ranges up to 5.7 nautical miles in 20 feet of water and sustaining a 99% probability of correct reception. The lake used being essentially an isothermal and low noise water environment [138].

46. A microprocessor based multi-frequency shift keying (MFSK) *Digital Acoustic Telemetry System (DATS)* [75, 117] was also presented. DATS had been used to transmit data over short ranges by 1984 and experimentation was carried out in Woods Hole

Harbour. The scheme is of interest to this research even though it was aimed at short ranges because the channel depths considered were shallow. Incoherent and coherent operation were both used with the data transmission system, which used up to sixteen tones (between 45kHz and 55kHz) for data transmission. A Doppler pilot tone was also used at 60kHz and a short synchronisation pulse at 30kHz. The Doppler pilot tone was used to compensate for the mean Doppler drift between transmitter and receiver, while the synchronisation pulse was used at the start of each data word. The system used coherent demodulation by tracking the Doppler pilot tone with a phase locked loop (PLL) until a situation arose where signal fluctuations were too high. Incoherent operation was then used. The system modulation scheme used frequency hopping to reduce effects of time spreading, by hopping to a new set of frequencies for the transmission of ever data word. The data transmission experiments which were carried out in the harbour produced in-water data errors consisting of transmission errors, and timing synchronisation errors due to errors in tracking the beginning of each data transmission. The synchronisation system performed acceptably only when deep signal fading was infrequent. The overall conclusions stated that there is a need for simpler systems compared to the high level of complexity required for DATS operation, but there is also a parallel need to incorporate sophisticated coding to combat the shallow water channels effects on the transmitted signal [75].

47. Multi-purpose communications systems have been envisaged that can be used in a number of different channels; one such system is the *Multi-modulation Acoustic Transmission System*, which uses different modulation schemes for different channels [139]. Phase Shift Keying (PSK) is used for vertical links and either chirp or frequency hopping modulation for horizontal channels. Non-coherent frequency hopping modulation is used for shallower longer range communications, with signal demodulated being carried out by a maximum likelihood receiver. Tests carried out in a pool to simulate horizontal transmission produced encouraging results. A frequency band spanning the bandwidth from 10kHz to 14kHz was used for the chirp modulation scheme for experimentation in a channel in the Bay of Breast; no errors were detected for ranges up to 4000m at 20 baud. Further trials carried out in other areas yielded low error rates, which were attributed to vessel noise and specific multipath structures.

48. The *Birmingham Acoustic Signalling Systems (BASS) 600 FSK* non-coherent scheme employed continuously operating FSK [140]. The field trials on this communications link included FSK and DPSK systems; thus allowing a comparison between the modulation schemes to be undertaken. The tests were aimed at short-range (25 to 250m) communications with a transmission frequency of 600kHz. The effects of multipath, weather conditions, and platform movement, were all considered. Small beam angles were used to eliminate the multipath at the modest range-depth ratios that were under analysis. Trials were carried out at sea-water sites, with virtually error-free transmission being established at ranges up to 100 metes in smooth seas. When the tide was stronger the link broke down into an unstable state. Loss of performance was attributed to a combination of the increasing attenuation losses with range and an exponential fall of error probability with increasing signal-to noise-ratio [140]. This effect was found at about 250m range with a low transmission power of 100mW. Results showed that a four-fold increase in power is needed to increase the range by two.

49. A multi-carrier system based on an Orthogonal Frequency Division Multiplexing (OFDM) principle was presented in the *Oceans '94 Conference*, for use in horizontal underwater channels [141]. The system was presented together with the principles of using an OFDM modulator and demodulator by means of Discrete Fourier Transform (DFT), overcoming the need for multiple identical modulators and demodulators. Simulation results of a fading and non-stationary channel showed that the system would be robust with the appropriate guard interval of approximately 30ms to counteract channel spread. Multi-carrier transmission over a multipath channel could also be carried out using a guard interval to combat symbol interference [141]. Sea trials of a OFDM system with incoherent demodulation and convolution coding produced results with very few errors for a signal-to-noise ratio of greater than 16dB.

3.1.5.2. Phase-based Implementations

50. Phase shift keying (PSK) techniques have been used in sub-sea communication systems; for this to be viable, problems due to phase instabilities have to be at a level that can be overcome by processing or low enough not inhibit reception in the first place. A number of the systems that deploy PSK-based modulation schemes are hybrid modulation systems or are aimed at vertical deployment. Other schemes that use PSK have been implemented by making use of widely spaced receivers and techniques that exploit the environment's spatial diversity [7, 12, 13]; this allows PSK to be viable over long-range channels.

51. The *MAST2* project *"LORACOM"* used phase shift keying techniques over long-range sub-sea channels, producing good results. Investigations into the use of coherent Binary Phase shift keying (BPSK) was carried out over a 50 km channel. Initial simulations and implementations carried out pointed to the potential of using widely spaced hydrophones and adaptive combining [5].

52. Investigations were carried out in deep waters (over 300m deep) using 1.7kHz frequency carrier and a bandwidth of 400Hz [6, 7]. The receiver was designed in an attempt to overcome the problems brought about by the poor signal-to-noise ratio (SNR), the time-varying channel characteristics, and multipath-induced phase fluctuations. The system was implemented using adaptive equalisation to reduce the effects of ISI. This was combined with adaptive beamforming to decrease the angular spread of the transmission of the acoustic signal, thus reducing the number of multipath signals received. Further signal manipulation was carried out by adaptive multi-channel combining, in an attempt to coherently combine the signals exhibiting spatial diversity from separate channels. The channel's spatial diversity was exploited by deploying a number of widely spread receiver hydrophones for the tests; in a separate experiment a short, vertically aligned transducer array was deployed. Adaptive equalisation was found to be largely ineffective in the channels encountered, as was the adaptive beamformers ability to track the changes of the incident signal's magnitude and angle of arrival at low symbol rates. But consistent success was achieved with the widely spaced hydrophones when coherently combining arrivals from the spatially diverse channels [6, 7].

53. Spatial diversity techniques are not an all-encompassing solution. There are situations when their use may not be feasible (cost, equipment available) or may require other techniques to be used in conjunction

54. In multiple user channels wide band spread spectrum techniques and multiplexing are of interest. Spread spectrum is relatively robust against fading compared to PSK. CDMA is able to discriminate between a number of signals at the receiver and therefore can be useful in multi-user implementations. TDMA techniques use is restricted by the long propagation delays found in many sub-sea channels.

55. In some sub-sea channels PSK techniques have been shown to be a viable modulation techniques without the need to incorporate spatial diversity. A number of communication links have been investigated, pointing to the use of PSK-based schemes in these less hostile scenarios [15, 23, 24]. Short range and vertical communication links provide particularly viable mediums for PSK-based communications due to the reduced interaction between the acoustic signal and the channel boundaries as well as the reduced channel losses.

56. A vertical link was established as part of a 'New Innovative Multimodulation Acoustic Communication System. This was a deep water vertical link that used differential coded BPSK to transmit over ranges up to 2000m with a carrier frequency of 53kHz [142]. Transmissions were made without coding (raw data transmissions) at 100 - 200 bits a second to establish comparable results.

57. A communications link implemented by Birmingham University included FSK and DPSK systems; and included comparison between the two modulation schemes. The tests were aimed at short-range (25 to 250m) horizontal communications; with a transmission frequency of 600kHz. The *BASS 600 DPSK* scheme [140] used small beam angles to eliminate the multipath at the modest range-depth ratios that were under analysis. Trials were carried out at sea-water sites, with effective transmission being established at ranges up to 100 metes in smooth seas. Conclusions from this work were that H.F. communications is possible over ranges of hundreds of metes.

Concluding Remarks

This chapter sets out the background to the modulation techniques described in later chapters, including a look at the underlying modulation schemes and techniques that have been implemented in the past. This leads intuitively onto the following four chapters that present an overview of the communication systems adopted and the modulation techniques employed. Chapter four provides an overview. The modulation techniques in chapters five and six use some of the techniques developed for the scheme presented in chapter seven but in a more refined state. The modulation techniques are presented in an appropriate sequence in order to help the readers understanding.

4. Communications System Overview

1. The incoherent reception techniques that have been adopted lend themselves to relatively non-complex implementation, compared to that required for coherent reception [127]. This is because the signal detection is independent of phase considerations. Coherent demodulation requires knowledge of the transmitted signal phase; this is not the case with incoherent demodulation. Incoherent frequency-based schemes have a higher bandwidth requirement than their coherent counterparts; with the same number of modulating frequencies. In a system alphabet twice as much frequency separation is required for non-coherent demodulation as compared with coherent demodulation. Coherent reception would provide the desirable solution in an accommodating environment as optimum demodulation requires a local carrier at the receiver, whose frequency and phase are the same as the transmitted signal. In a long-range, shallow sub-sea media the harsh acoustic signal propagation characteristics, especially the randomisation of signal phase, negate the implementation of coherent techniques unless processing to overcome the phase fluctuations can be provided. Incoherent reception methods do not require phase stability and therefore are appropriate for applications in long-range, shallow water environments when using omni-directional transducers.

2. Within this work a bit period represents one bit value. The bit period can be made up of one symbol period equal to the bit period. A bit period can also be made up of a number of symbol periods of equal size; in this case the symbol periods are called chip periods. A receiver's function is to achieve the minimum probability of error ($P_{e\min}$) in some 'optimum sense' when evaluating which symbol has been transmitted. This is carried out with the assumption that the symbols are equally likely. Maximum likelihood techniques lend themselves to reception under this assumption, i.e. the maximum likelihood (ML) receiver is a method of achieving minimum error probability (Appendix A). With coherent reception the ML process can be implemented either with a correlation receiver or a matched filter receiver. Coherent correlation reception requires the multiplication of two time functions followed by integration of the product for each symbol (Fig.4.1 (a)); a decision is made once every symbol period (T_s). The matched filter detector is an equivalent system to the correlation receiver, providing a measurement of the received symbol by maximising the output signal to noise ratio. A matched filter bank is required with one matched filter for each symbol in the symbol alphabet (Fig 4.1(b)).

3. Incoherent ML reception is implemented either as a modified correlation receiver or a matched filter envelope demodulator. Modified correlation reception involves separate multiplication of the input by a sine version and cosine version of each symbol frequency. The operations required for each symbol are integration, square root and summation (Fig 4.1(c)). The matched filter envelope detector technique involves a bank of matched filters (each matched to one of the symbol frequencies), the outputs of which are passed onto envelope detectors (Fig 4.1(d)).

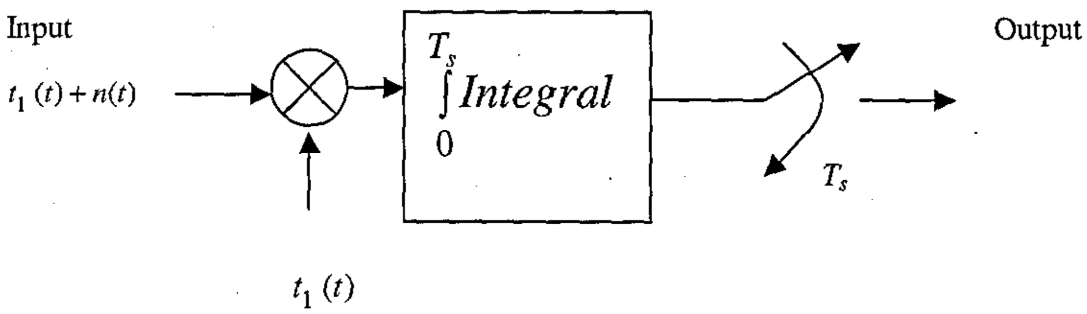


Fig.4.1 (a) Coherent Reception (Correlation Receiver Block)

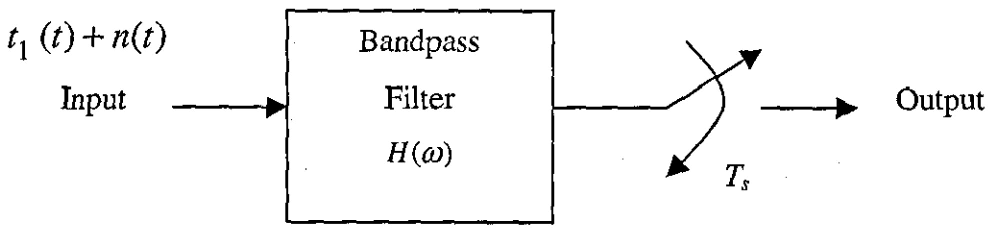


Fig.4.1 (b) Coherent Reception (Matched Filter Detector Block)

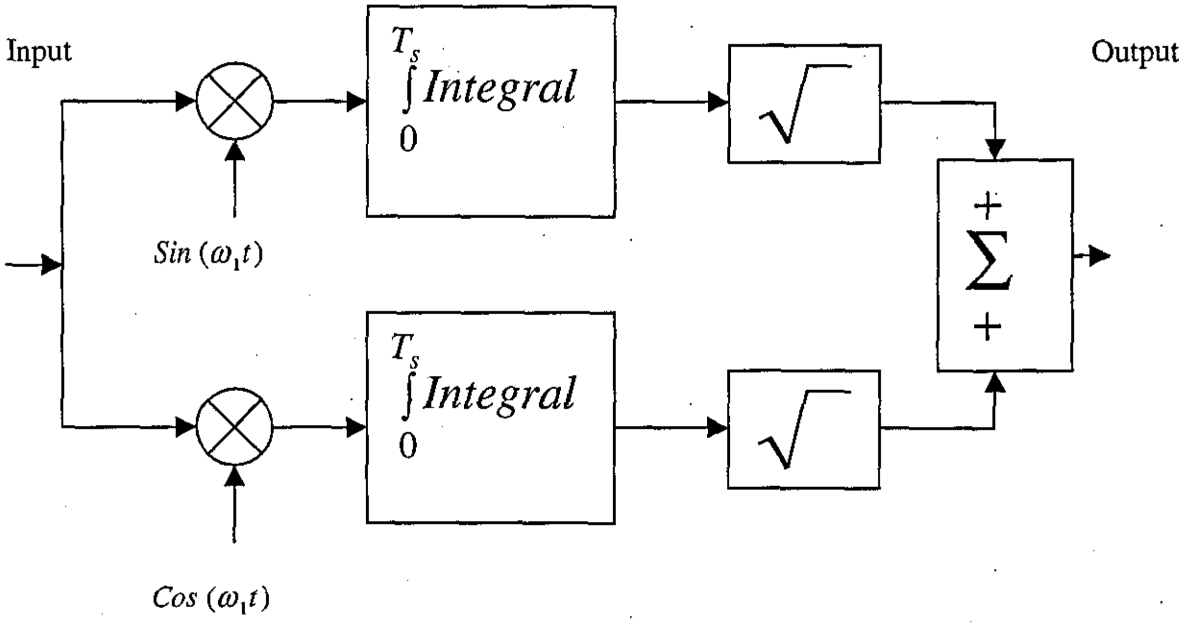


Fig.4.1 (c) Incoherent Reception (Modified Correlation Receiver Block)

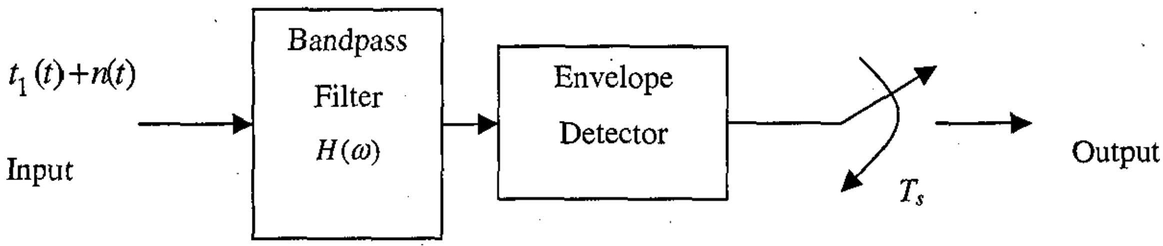


Fig.4.1 (d) Incoherent Reception (Matched Filter Envelope Detector Block)

4. The receivers produced in this research are based on incoherent matched filter envelope detection. Orthogonal spacing of the symbols (tones) in the system alphabet enables the appropriate matched filter for a particular transmission, to show a maximum output while the outputs of all other filters are minimised. So the detection of the transmitted tone can be made with 'maximum confidence' at the appropriate instant. Performance of the matched filters when evaluated is independent of the frequencies chosen, only on the frequencies being mutually orthogonal (Appendix A). Frequency-specific Fading (FSF) and Inter-Symbol Interference (ISI) exhibited by the acoustic signal interaction with the channel can interfere with the detection process [1, 165]. Modulation-based strategies are used in this research to increase the likelihood of detecting the 'correct' bit sequence.

4.1. Overview of Transmission Strategy

5. The transmitter's main function is to generate the symbols (frequency tones) that represent the data as well as the synchronisation information and to transfer them through the channel. Further functions include the ability to generate a carrier and the modulation of the signal onto the carrier. The processes can be defined as five operations which work in synergy to produce the signal that is transmitted (Table 4.1, Fig.4.2).

A.	Symbol alphabet generation
B.	Generation of the symbol period
C.	Assigning symbols to symbol periods
D.	Generation of the carrier
E.	Modulation onto the carrier

Table 4.1 General Transmitter Operations

6. Each transmitter implementation has specific requirements; therefore the operations in Table 4.1 and Fig.4.2 are given only as an overview of the transmission operations. In each chapter of this thesis that is associated with a modulation scheme the processes are described in more depth and sub-processes are included to aid understanding.

7. A symbol alphabet generator is used to generate the modulating frequencies. Within the 3kHz system bandwidth the symbol frequencies are all chosen to be orthogonal to each other, thus exploiting the optimum matched filter response at the receiver.

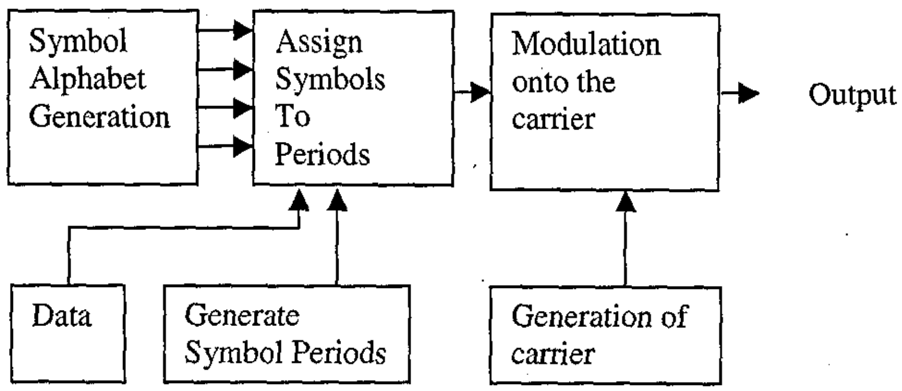


Fig.4.2 General Transmitter Structure

8. Generation of symbol periods and the allocation of symbols are described together here. These processes are specific to each modulation scheme. In general, the two processes work in synergy to produce the symbol configuration required within each bit period. A 'cyclic code' may be used to keep track of the exact symbol period being transmitted; alternatively in modulation schemes where the symbol period is equal to the bit period, the data is used. The multiple symbol periods that make up a single bit period are called 'chip' periods (Fig.4.5). The modulation techniques that require the 'cyclic code' use a 'chip selection' to picks the symbols for each chip period dependent on the cyclic code and the data (Fig.4.3). The chip period is equivalent to the symbol period when there are two or more symbol periods of equal length in a single bit period. This is a form of modulation coding called character-to-frequency coding.

9. Modulation onto a carrier involves two operations, these being carrier frequency generation and the actual modulation process, both of which are carried out within the transducer unit electronics. The multiplication performed by the modulation process shifts the modulating alphabet frequencies onto the carrier frequency (10kHz).

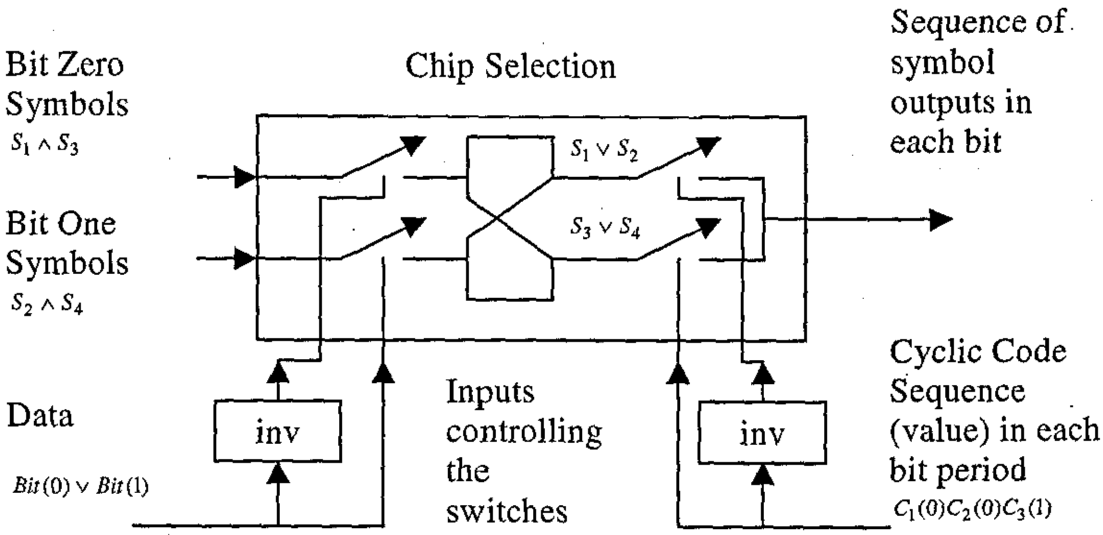


Fig.4.3 Chip selection process

1.	Demodulation
2.	Matched Filtering
3.	Envelope Detection
4.	Bit Decision
5.	Equality Decision (Energy Decision)
6.	Synchronisation
7.	Bit evaluation

Table 4.2 Reception Operations

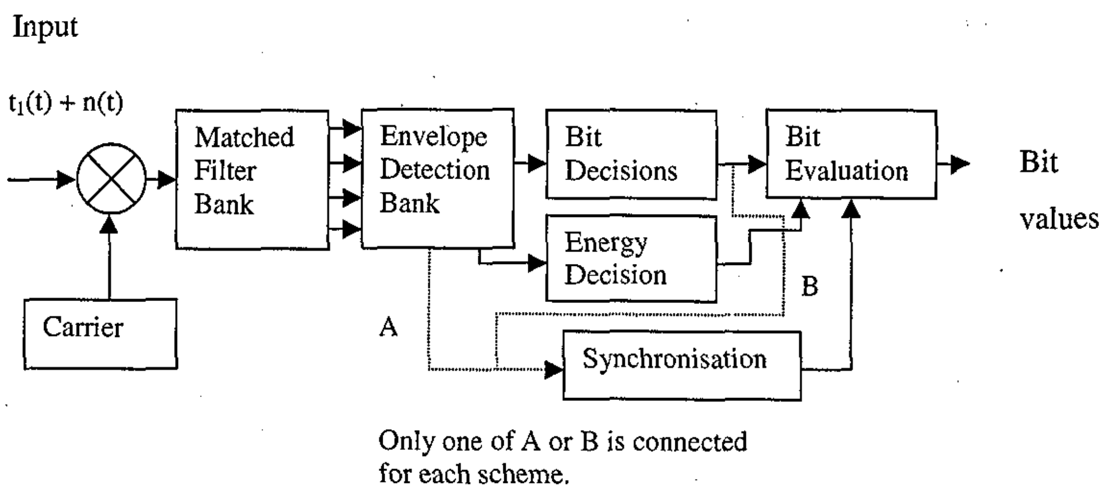


Fig.4.4 Reception Processes

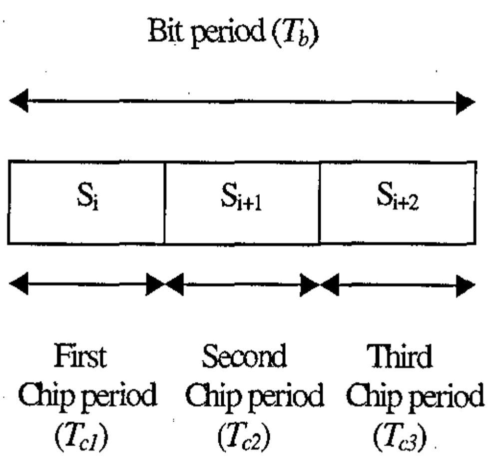


Fig.4.5 Definition of 'chip' period for a three-chip coded scheme

4.2. Overview of Reception Strategy

10. The reception strategy adopted is significantly more complex than the transmission strategy. This is due to the well-known communications problem of multipath interference and the noise, that is found in long-range, shallow sub-sea environments. The receiver's purpose is to produce an estimate of the transmitted sequence of symbols and evaluate the individual bits. The receiver has to be capable of:

1. Performing demodulation; distinguishing between the symbol frequencies
2. Quantifying each symbol
3. Evaluating which symbol is present in each symbol period
4. Providing a bit decision and a secondary evaluation of that decision ('check mechanism')
5. Providing a methodology for dealing with a 'tie' in the bit decision
6. Maintaining synchronisation with the transmitter (at bit or sub-bit level).

11. The reception strategy chosen involves separating the receiver process into a number of logical stages performed in the evaluation of each transmitted symbol. The concept of matched filtering forms the basis of the receiver structures. The matched filter reception technique is believed to be appropriate as a finite symbol alphabet is used, allowing the receiver to have an *a priori* knowledge of the possible symbol transmissions and as symbol frequencies have all been chosen to be orthogonal to one another. The matched filter method also lends itself to Digital Signal Processor implementation, which is used primarily to facilitate the development of multiple receivers.

12. The reception processes work in synergy to evaluate the bit value for each symbol period. The stages of processing are listed in Table 4.2 and shown graphically in Fig.4.4. The listing is in a logical order, starting from the extraction of the modulating signal from the carrier signal (demodulation), including the evaluation of the frequency of the received signal (matched filtering) and finishing with the bit evaluation process. The demodulation, matched filtering and envelope detection processes can be looked upon as working with the received signal sequentially. The bit decision and energy decision can be viewed as two parallel processes that utilise the output of the initial sequential processing blocks. Conceptually, the synchronisation is best looked at as a process that works in parallel with all the other processes, to keep the receiver synchronised with the transmitter. The final bit choice block uses the output of the two parallel blocks (bit decision and energy decision), together with the synchronisation signal, to establish the bit values for each bit period.

4.2.1. Demodulation

13. Demodulation of the modulating signal from the carrier is carried out as appropriate to non-coherent reception throughout the work documented in this thesis. The acoustically hostile nature of the environment prohibits the implementation of coherent detection without some form of compensation for the received signal's 'randomised' phase. Numerous techniques have been used in the past to attempt to overcome the channel's prohibitive effects but have generally been found to be effective only in other more hospitable sub-sea channels. Some of the techniques have been found to work adequately over short ranges, in deep water and in a vertical trajectory; or with the use of multiple receiver hydrophones or widely spaced arrays [2, 3]. These techniques generally attempt to overcome the phase fluctuations produced by the signal's interaction with the environment in one of a number of ways, by attempting to reduce signal-boundary interaction, compensating for the channel's transfer function, utilising the spatial diversity inherent in the environment in some way. In channels in which these techniques work competently, the use of coherent detection is not necessarily prohibited.

14. In the channels of interest here, the implementation of non-coherent reception, without the use of the techniques mentioned above, is appropriate. The non-coherent demodulation strategy involves the multiplication of the received signal with a locally generated version of the carrier frequency. The demodulation operation places the modulating signal back in the original bandwidth that was occupied before modulation onto the carrier was performed at the transmitter. This produces an output signal that is made up of the symbol frequency that was transmitted and the 'noise' picked up in the channel.

4.2.2. Matched filtering

15. Central to the implementation of the receiver structures described in this thesis is the concept of matched filtering (Appendix A). Filters are matched to each of the symbol frequencies in a particular scheme's symbol alphabet. The matched filters provide a quantifiable measure of the frequency content of the received signal in terms of the symbol frequencies. This is possible because the receiver has an *a priori* knowledge of the finite symbol alphabet being utilised. The symbols are all transmitted with equal energy and the frequency tones are orthogonal. The probability of symbol detection may be presented in terms of the i^{th} symbol S_i, and the n^{th} symbol's *a priori* probability of transmission P_n. For a finite alphabet made up of, S_i equal energy symbols where $i = 1,2,......M$, there is an *a priori* probability $P_n = P_i$ where n is any integer between 1 and M. There is an underlying assumption that each symbol has an equal *a priori* probability when making each symbol decision.

16. The matched filter outputs; are used by envelope detectors to produce a measurement of the frequency content in each symbol period. The highest output for each symbol period is assumed to be the transmitted symbol. The other matched filter outputs should be at a minimum, not matched to the frequency. Thus a comparison of outputs provides a measurement of which symbol is present. This, of course, is the idealised case; interference produced by the transmitted signal interaction with the surrounding environment complicates the situation. There may be signal components at the filter outputs produced by noise or inter-symbol interference. The occurrence of fading or frequency drift is also a possibility, causing lower than maximum 'correct' matched filter output. Therefore at some 'noise level' the wrong matched filter could produce the highest output and an incorrect symbol decision may be produced. With the symbol frequencies chosen as an orthogonal set, potential interference between them is limited.

17. In theory the bandwidth of each filter should be equal to that of the signal that is 'matched'. If the bandwidth of the filter is wider than the signal bandwidth, then an increase in the noise content of the filtered signal is inevitable; the signal strength is unaffected by the increase in filter bandwidth. A narrow bandwidth results in a reduction in signal strength as well as a reduction in noise, but the reduction in signal strength is not overcome by the reduction in noise as it is not enough to make up for the degraded signal strength. Thus excessively wide or small bandwidth allocations will reduce signal-to-noise ratio (SNR).

18. Minimum orthogonal frequency separation (F_{smin}) can be defined in terms of the symbol period (T_s):

$$F_{s\min} = \frac{1}{T_s} \qquad \textbf{(4.4)}$$

The maximum orthogonal spacing possible that is able to accommodate the alphabet within the limited bandwidth allocation is used, not the minimum. This is done to share out the available bandwidth efficiently. The signals are thus placed at a 'maximum range' in terms of their inter-interference potential, and are provided with the largest possible bandwidth for each matched filter implementation. A general definition for orthogonal incoherent FSK frequency separation is used to establish the appropriate separation:

$$F_s = \frac{m}{T_s} \qquad \textbf{(4.5)}$$

where F_s is the frequency of separation, T_s is the symbol period and m is any integer. A tight bandwidth for each matched filter is desirable to allow only the matched frequency through and to reduce noise. The practicalities of real-time operation notably DSP power and memory limitations negate the utilisation of filter tap-intensive narrow bandwidths. The limited system bandwidth and the real-time practicalities mean each matched filter bandwidth implementation needs to be maximised to limit the number of taps required. When all the filters are matched to one of a finite set of modulating symbols and are allocated the same bandwidth, a comparable measure of the relative signal strengths can be made.

19. The taps of a digital filter are a time domain representation. Matched filtering in the time domain requires the impulse response of the filter to be the time reversed implementation of the signal to which it is to be matched. This is achieved using the Kaiser technique. In practice there is a restriction on the length of taps, dependent on the DSP sampling frequency and the length of the symbol period. Long filter tap implementations may incorporate the frequency content of more than one symbol period. The matched filters are required to measure the frequency content of the received signal, so the frequency content within each symbol period can be established through envelope detection.

4.2.3. Envelope Detection

20. The matched filter outputs provide a quantifiable measure of the frequency content of the received signal. The 'measurements' are converted from the 'instantaneous' values of the filter outputs to a representation of the signal content within each symbol period. This procedure is carried out by a process called envelope detection.

21. A concept of coincident point envelope detection was initially considered; this is applicable when the portions of the symbol period used are not uniform. The detection period used for each signal is dependent on the number of cycles required for the detection process. Here, each matched filter's signal detection process is performed over half a cycle of each matched frequency. Symbol detection periods are thus intersected by a 'guard' period, for all matched frequencies that have more than half a cycle in a symbol period. This can be seen as artificially extending the length of the symbol periods, thus leaving them less susceptible to Inter-Symbol Interference (ISI). The envelope detection is carried out by magnitude integration, which is performed with an equal number of each signal's samples. In coincident detection an equal number of samples are used for each matched filter symbol decision over an equal number of cycles of each frequency. The samples are spaced equi-distantly across the detection cycles of each frequency (Fig.4.7 (a)). A comparable value is produced for each symbol frequency, in every symbol period.

22. The coincident point envelope detection requires high sampling frequencies to provide high numbers of coincident points over a small number of cycles of each symbol frequency. At high sampling rates, decimation techniques provide one way of conserving the DSP power reducing the number of redundant samples that are processed. When utilising these techniques, reconstruction of the sampling frequency is necessary for synchronisation purposes.

23. Another envelope detection technique that was considered performs the straightforward integration of the signal magnitude over the full symbol period. The technique does not include any guard periods and performing the detection using all the samples, not just coincident ones (Fig.4.7 (b)). This 'straight forward' magnitude envelope detection technique is provided with adequate samples even at low sampling rates.

24. A third tèchnique was designed to replace the coincident point envelope detection, to allow low sampling frequency implementation, which is required as real-time operation means all the processing has to be carried out between sampling instants. This guard envelope detection's design is influenced by the other two methods. Direct integration of the signal is performed as in 'magnitude envelope detection' (which is also used in some of the real-time implementations). This is carried out over a portion of the symbol period, not the full period (Fig.4.7 (c)), therefore a guard period is inserted. A fourth technique (cycle technique) was also developed that also provides guard periods, but in this case the same number of cycles are used in each envelope detector as with coincident point envelope detection but all samples are used in the evaluation. The output signal of each envelope detector is adjusted dependent on the number of samples used in its processing.

4.2.4. Bit evaluation methodologies

25. The bit decision strategy adopted throughout this work uses 'redundant' information that is incorporated in the transmissions. The information is used to limit the likelihood of detecting the wrong bit, even when individual symbol decisions are in error.

26. Multiple symbol assignments associated with each bit value provide a facility to 'question' each bit decision. The possibility of checking a bit choice with some other, separate information (uninfluenced by the first decision) reduces the likelihood of detecting an incorrect bit value. This is because decisions have to agree with each other for an outright bit decision to be made. A third decision, the energy decision, is defaulted to in a case of disagreement. Each of these decisions comprises up to three symbol decisions.

27. One of two techniques is used to perform this 'bit checking' process, in each of the modulation schemes developed here. The first technique is utilised in the modulation schemes that use sequential transmission of a number of symbols, within each bit period. For a bit decision to be made there is a requirement for the sequence of symbols to be detected at the receiver. A separate bit decision is made for a zero bit and a one bit, utilising completely different symbols. If there is an incorrect symbol detection within a bit period then an assessment of which bit's sequence is 'most correct' is made (zero-bit sequence or one-bit sequence), in terms of the number of correct symbols in each sequence. If a tie occurs an energy decision is implemented.

28. The second 'bit checking' method is used in schemes where the symbol period is equal to the bit period. Separate decisions are made using additional symbols that are transmitted in parallel with the first decision symbols (Fig.4.6). The outputs of all the decisions are compared. If the additional decisions do not agree, then a tie is called and the 'energy decision' is used to evaluate the bit value.

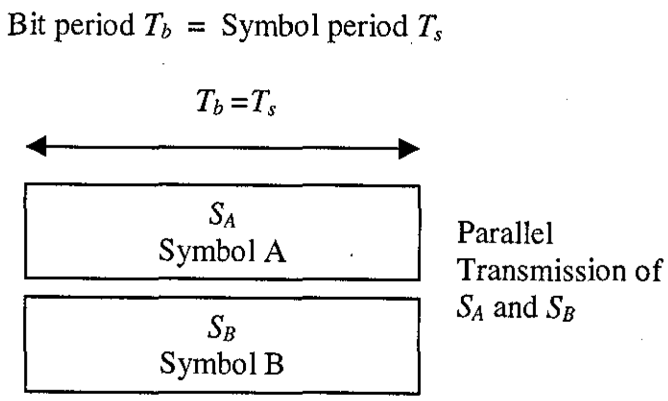


Fig.4.6 **Parallel transmission of symbols over a bit period**

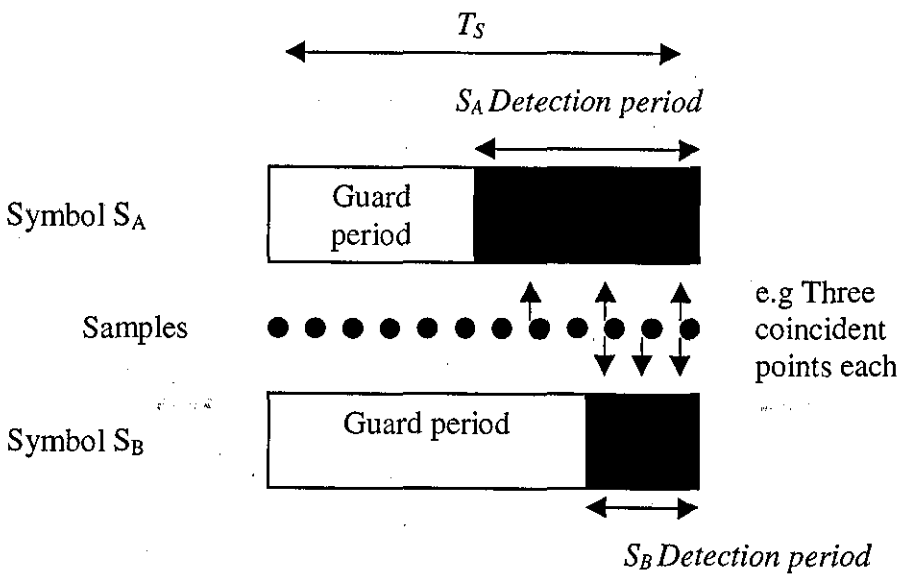


Fig.4.7 (a) **Coincident Point Envelope Detection**

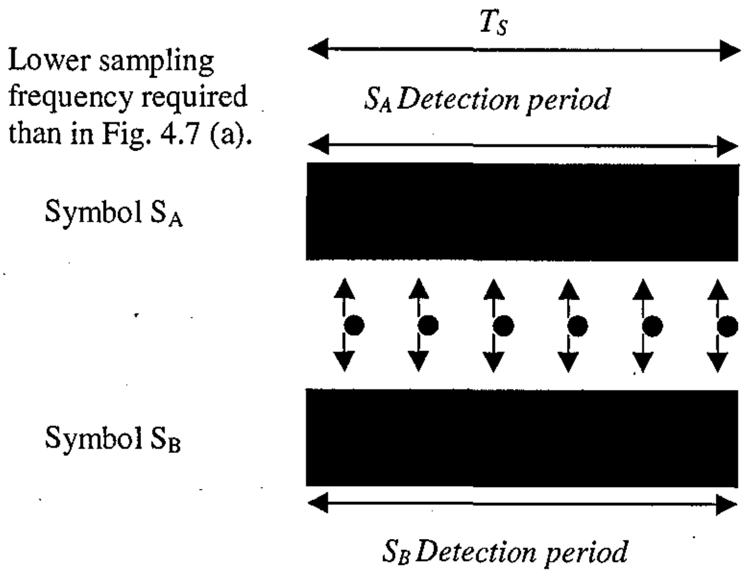


Fig.4.7 (b) Magnitude Envelope Detection

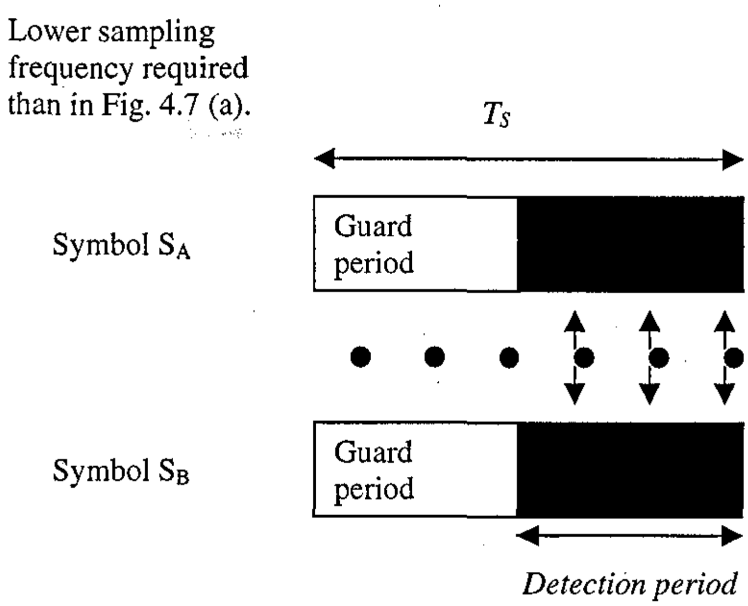


Fig.4.7 (c) Guard Envelope Detection

29. In the serial transmission schemes the two bit decisions (and the energy decision) are made up of three separate symbol decisions. Therefore each individual bit evaluation comprises up to nine symbol evaluations (Fig.4.8 (a)). The energy decision uses the same inputs as both of the bit decisions and combines the signals before performing the energy evaluation. The three energy decision evaluations carried out in each bit period are not completely independent of the six bit decision evaluations. The multiple evaluations achieve a relatively robust way of determining the bit decision compared to single tone assignment schemes, where only one bit decision is made, as there is no redundant information to make additional evaluations, e.g. basic FSK. In a 4FSK scheme each symbol normally represents two bits; thus there is only one decision for every two bits compared to up to eighteen decisions for two bits in one of the serial decision systems described here. In the 'parallel' systems described in this thesis there are two separate bit decisions (equivalent to the number of symbol decisions in this case) and a further energy decision that combines the two. Therefore up to three decisions are made for each single bit decision with the 'parallel' systems (Fig.4.8 (b)).

4.2.5. Bit decision evaluation (Energy decision)

30. In the case of a 'tie' between the bit decisions (when the bit decisions are not conclusive), a random choice may be made. The maximum likelihood detection principle allows a random choice, as this does not affect the average probability of error (Appendix A). A check of the initial decision is carried out by the bit decision; ties (inconclusive results) are only overcome randomly if the energy decision also yields a non-definitive result.

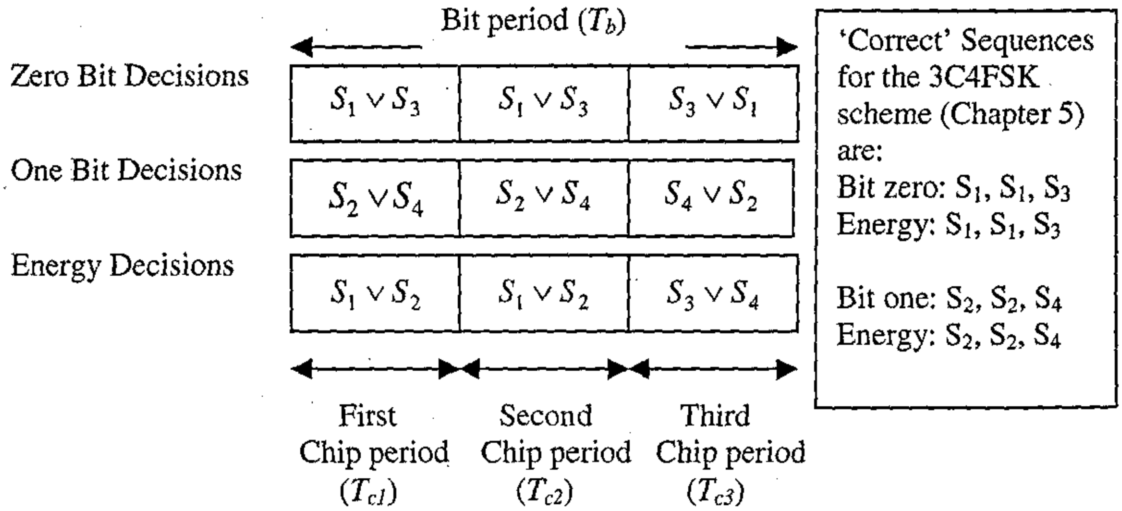


Fig.4.8 (a) Serial System's Nine Decision Bit Evaluation

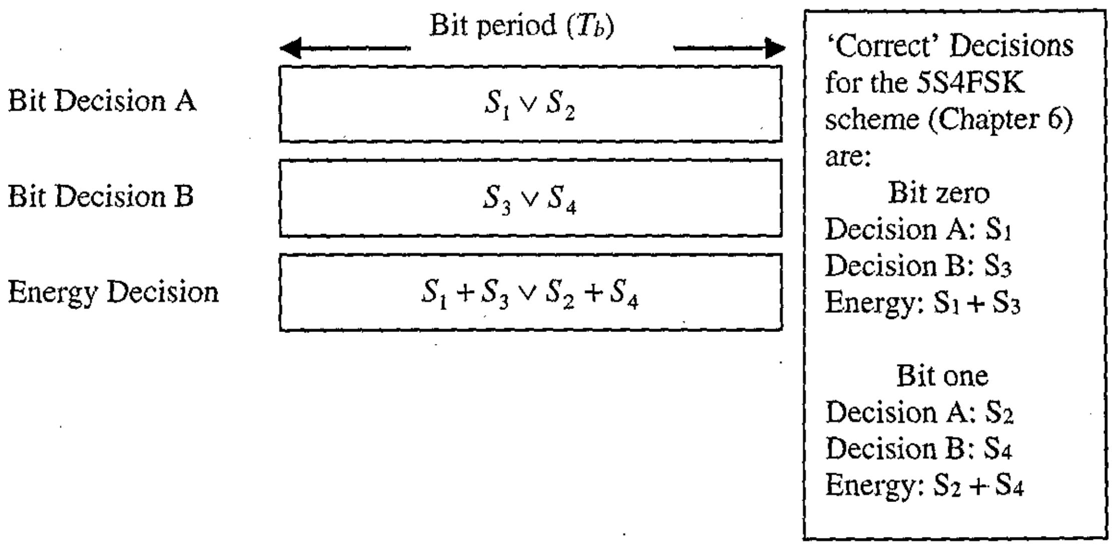


Fig.4.8 (b) Parallel Systems Three Decision Bit Evaluation

31. The energy decision is thus a way of deciding between the two possibilities, when the bit decisions are inconclusive. A measure of the energy content appropriate to a zero bit and a one bit is made by combining all the outputs from envelope detectors associated with each bit. The bit one and bit zero values are then compared to establish the bit received when a contention has occurred.

32. One of two 'energy' decisions is used in each system. The first technique extracts the 'energy' from a number of serial decisions. This technique is used in modulation schemes where the symbol periods are smaller than the bit periods. The energy is measured and the appropriate bit value (the one with most energy) is chosen. Thus, inconclusive bit decisions cause the evaluation to be made over the full bit period instead of by the sub-division of the bit period into its separate symbol decisions, as is carried out in the bit decision process.

33. The second method combines the 'energy' from a number of parallel decisions. In this case the symbol period is equal to the bit period; when a 'tie' occurs the 'energy decision' is made by taking the combination of the decision energy (for each bit value) and comparing the results. The highest combined content decides which bit is being received.

34. The energy decision output is used in the bit evaluation process instead of the bit decision(s) in the case of an unresolved bit transmission. The energy decision makes use of the 'extra' information provided by multiple symbol assignments to make the bit evaluation more robust. A random choice is not made unless the energy decision is also inconclusive.

4.2.6. Decision Combing and synchronisation utilisation

35. The decision 'combing' block takes the bit decision(s) and energy decision outputs to determine the transmitted bit. The decision combination is made only once in each bit period using the synchronisation signal. In contentious cases the energy decision output is used to provide the output; if this is inconclusive the choice is random.

36. The modulation techniques transfer the synchronisation information together with the data either intrinsically or explicitly. In this way some mechanism for transfer of the bit rate to the receiver is undertaken. A PLL is utilised to lock a reference frequency to the frequency of change of a regular sequence of chips (Fig.4.9 (a)), or a dedicated frequency (symbol) transmission dependent on the modulation scheme used (Fig.4.9 (b)). The synchronisation signal is used to define the period over which envelope detection should be carried out as well as defining the point at which the decision 'combing' output is relevant.

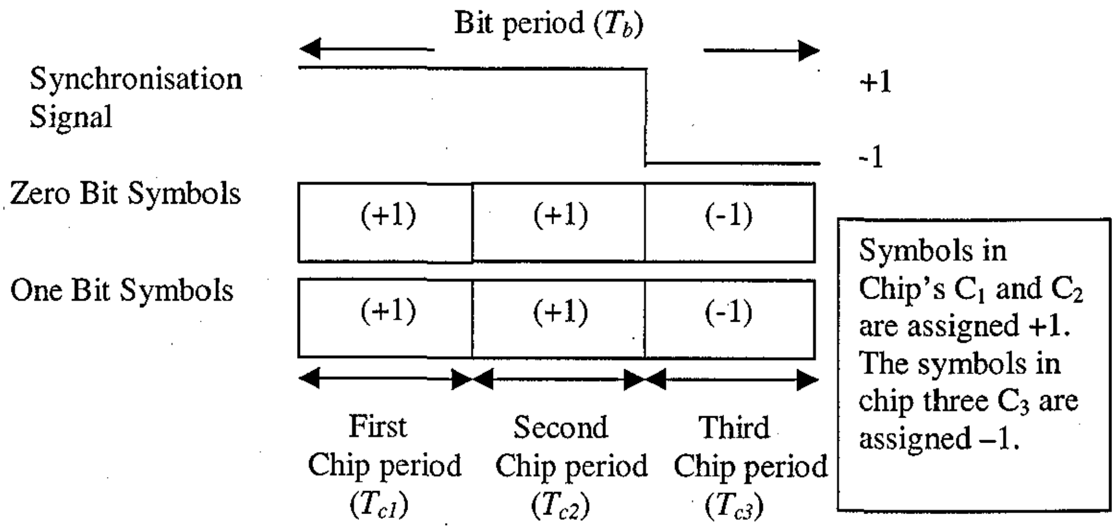


Fig.4.9 (a) The intrinsic transfer of synchronisation information

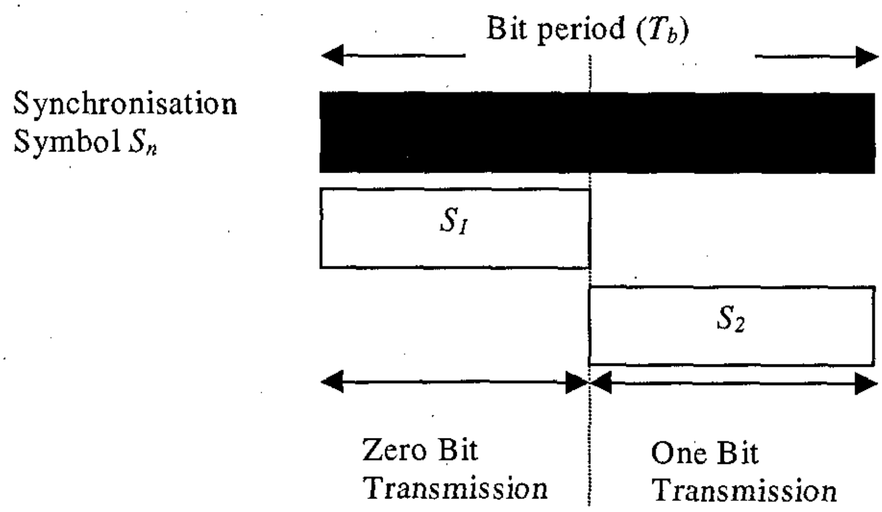


Fig.4.9 (b) **The explicit transfer of synchronisation information**

4.3. Transmission-Reception Synchronisation Requirement

37. For communication between the transmitter and receiver to be established, the transmission and reception protocols at either end of a channel must be compatible. When designing a modulation technique, for a specific environment and a particular purpose, one must not forget the underlying problem: *The two ends of a link have to work in synergy to produce the intended transfer of data.*

38. This is not straightforward in the acoustically harsh, long-range, shallow sub-sea environment; there is no certainty that the transmission and reception processes are working in unison. There are inherent propagation difficulties found in the environment including multipath propagation (varying delays, phase randomisation) and frequency dependent fading. There are also hardware associated characteristics that tend to preclude the two ends of the link from staying synchronised, such as the relative drift of the transmission and reception oscillators. This oscillator drift can mean bit decisions are made over the wrong portions of the input signal, as the reception bit rate becomes 'miss-aligned' from the transmission bit rate. A number of codes can be used to overcome bit errors that are inevitable, including the simplest parity checks, block codes and convolution codes.

39. The receiver requires knowledge of the transmission rate so that all decisions can be made at appropriate instants. Bit level decisions require the bit rate to be established (utilising transmitted synchronisation information), allowing each bit choice to be made over the correct period of the received signal. For transmission and reception techniques to work in unison there must be some way of transferring the synchronisation information together with the data signal. This should be carried out with minimum interference between the two types of information. The synchronisation signal is independent of the data; it therefore does not affect the data and is not affected by the data itself.

40. Synchronisation in a communication system may be required at different levels to establish symbol, bit and word periodicity. The ability to extract information to decipher which periods of the input signal is being received and the rate at which the transmission is occurring is a mandatory requirement of many communication systems. In the work carried out for this thesis there is a need to extract only bit and/or symbol synchronisation information from the incoming signal, as raw transmission is used.

41. The use of raw data transmissions (no bit coding) is undertaken to allow the assessment of the modulation schemes without the effects of additional processing. Higher level synchronisation (code-word level) is not implemented as raw transmission is used. Coding techniques can provide increased resilience to channel interference, usually with a small reduction in data throughput. A number of interference combating techniques (spatial diversity, directional antennas, coding of a number bits into code-words, etc.) do not fall within the scope of this thesis, as they do not fit the project aims.

4.3.1. Synchronisation Transfer

42. It is necessary to establish the instant that a new symbol period begins, whereas the need for bit synchronisation (where the bit period is larger than the symbol period) and word synchronisation is dependent on the workings of the particular system. Synchronisation can not be achieved in this environment by setting the receiver and transmitter off initially synchronised, or by utilising a locally generated signal without some way of locking the receiver frequency to the transmitter frequency.

43. Synchronisation can be achieved in a number of ways if the environment allows; the process can operate continuously, periodically or with a one shot synchronisation approach. Continuous synchronisation uses the technique of conveying the synchronising information simultaneously with the data signal. The use of continuous methods, in which a sub-modulation (like amplitude modulation) is used to convey the synchronisation data, is sometimes carried out. The shallow underwater channel is not an ideal medium for such a scheme, with multipath having a randomising effect on the amplitude of the signal.

44. Periodic synchronisation can be inserted in between data signals at regular intervals. Stability of such a scheme is dependent on the transmission and reception rates being constant between updates. There is a reduction in the overall data rate achievable because of the need for multiple repetitions of the synchronising element. Therefore there is an effective wastage of part of the transmitted signal power.

45. One-off synchronisation can be used by transmitting the synchronisation information at the start of transmission before the data, placing a high constraint on the stability of the clock throughout transmission. The schemes that require additional frequencies for the synchronisation process take up additional bandwidth. Synchronisation may also be established intrinsically, by examining the data itself with the help of coded data that provides the clocking information.

46. A synchroniser consists of two distinct components: a linear circuit that generates a carrier (or clock waveform) and a narrow-band device to separate the signal from any background disturbance. A phase-locked-loop PLL or a tuned-filter can be used as the narrow-band device.

47. Throughout this work the synchronisation information is transferred between the transmitter and receiver by one of two methods. Both of the methods incorporate the synchronisation information within the structure of a modulation technique. The synchronisation information is passed across the communications link in unison with the data signal, allowing continuous transmission of the data to take place. Each scheme's structure is defined by its symbol alphabet and by the way the symbols are assigned to each bit value. With this type of synchronisation, long strings of single-bit value transmissions, do not prohibit synchronisation from being maintained.

48. The modulation techniques have been developed utilising modulation coding. This character-to-frequency coding is the assignment of multiple symbol frequencies from the alphabet to each bit value. This opens up the possibility of providing synchronisation information in each bit period as well as the ability to distinguish bit values through the assignment of the symbol frequencies. The transfer of synchronisation information has been designed to provide minimum interference with the data transmission.

4.3.1.1. First Synchronisation Approach and its Intricacies

49. The first synchronisation approach involves direct transmission of the synchronisation information, in parallel with the data signal. This is achieved by having an additional orthogonal synchronisation symbol in the modulating frequency alphabet, that is not dependent on the bit value, and therefore can be transmitted continuously. This additional symbol frequency intrinsically calls for a wider transmission bandwidth or a tighter spacing of the symbol frequencies within the pre-defined minimum spaced orthogonal bandwidth placement. This from of synchronisation is limited where frequency selectivity is a problem; therefore a second synchronisation scheme was developed (section 4.3.1.2).

50. The overall bandwidth requirements for the communication systems drawn up for this work with BG plc inevitably mean that a tighter spacing of the symbol alphabet frequencies is required; when an additional symbol is introduced. The limiting factors on the spacing of the symbol frequencies within the available bandwidth include:

1. There is a need to maintain orthogonal spacing between each symbol frequency in the alphabet (including any synchronisation symbol) to minimise interference and thus allow optimum matched filter behaviour (Appendix A).

2. The limitations of the DSP, in-terms of its ability to handle (in real-time) the increased number of matched filter taps associated with decreased symbol spacing.

51. There is a minimum frequency spacing that can be implemented for any specific symbol rate. The spacing is preferably maximised, not minimised within the systems limited bandwidth allocation. This allows the matched filter implementations to utilise the full bandwidth available to them limiting the number of taps required for their implementation.

52. The basis of extraction of the bit rate at the receiver is the PLL (Appendix B) which is used to 'lock' a reference frequency to the incoming synchronisation signal. This is carried out by first band-pass filtering the input signal centred at the synchronisation symbol frequency. This limits the frequencies that can be fed to the PLL and allows the frequency to adjust to the incoming signal-to-noise ratio (SNR). The signal is then passed to one of the second order PLL's detector inputs (the other stems from the VCO via the feedback loop). The detector gives an error difference between the input and feedback signal. The detector output is passed through a loop filter, the output of which controls the voltage-controlled-oscillator (VCO) at the incoming bit rate. The VCO output is also passed to a pulse generator that produces a pulse at the bit rate. Pulses are also produced for use in the envelope detector to define appropriate periods for each decision.

4.3.1.2. Second Synchronisation Approach and its Intricacies

53. The second synchronisation transfer approach involves the transmission of synchronisation information within the structure of the modulation scheme but without the need for parallel transmission of a synchronisation symbol. There is no additional bandwidth requirement above those required for the symbols used in the data transmission. This is possible because the modulation schemes developed exhibit redundancy in the assignment of modulating frequency symbols to bit values. Instead of transmitting the 'redundant symbols' in parallel (as in the systems that use the 'first synchronisation approach'), sequential transmission is implemented. In the sequential transmissions the redundant symbols are used to provide a symbol change within each bit period (Fig.4.9 (a)). This change of symbol frequency can be detected at the receiver. Note that the so-called redundant symbols are not actually redundant as they are also used to produce a distinctive sequence of symbols that is 'correct' for either of the bit value transmissions.

54. The extraction of the synchronisation information is based on the ability to detect the sequence of arrival of 'chips' within each bit period. By having one change of chip frequency in the same place in every bit period (no matter what bit value is sent), the rate of change and thus the bit rate can be extracted at the receiver. The symbol rate can then be calculated from this. To maintain the bit evaluation properties of the 'chip' sequence, all chip symbols associated with different bit values are chosen to be unique.

55. The sequence of chips is detected by the receiver, then a reference frequency is locked to the rate of change of chips (which is equivalent to the bit rate) using a Phase Locked Loop (PLL) (Appendix B). This is carried out by first establishing the sequence of incoming chips by looking at the outputs of the matched filters. The signals associated with chip one/two output (same output), are combined and then compared with the signal generated from the combination of the chip three matched filter outputs. This allows the receiver to detect the change from the chip one/two portion of the bit period to the chip three portion; this is achieved independently of the transmitted bit value. Chip one/two are assigned a value of one (+1) and 'chip' three is assigned a value of minus one (-1), (Fig.4.9 (a)). The signal is then passed through a filter matched to the appropriate bit rate and passed to one of a second order PLL's detector inputs. The other input stems from a VCO via the feedback loop. The detector output produced is a measure of the error between the input and the feedback signal. The detector output is passed through a loop filter, the output of which controls the VCO at the incoming bit rate (Fig.B1), (Appendix B). The VCO provides the feedback to the second detector input. The VCO output is also passed to a pulse generator that produces a pulse at the bit rate that is used to define the appropriate instants to make a bit decision. Pulses are also produced for use in the envelope detector to define appropriate periods for each decision.

4.3.2. Implementation of synchronisation scheme

56. The NE567 phase locked loop (PLL) integrated circuit forms the infrastructure for the synchronisation signal extraction at the receiver in both techniques used. It includes an adjustable filter, a detector and a current controlled oscillator (CCO). It has a frequency range of 0.01Hz to 500kHz. The CCO works in a similar way to a VCO oscillator.

57. The principle of operation is as with any such PLL. When in the 'locked state' average D.C. of the detector output is directly proportional to the input frequency. The filter is used to extract the D.C. level from the detector output and apply it to the CCO. Changes in the input frequency cause the detector output to adjust accordingly and shift the controlled oscillator frequency to match the input frequency.

58. The NE567 is set up for 'high input level' mode which stabilises any possible bandwidth variations that may be attributed to in-band signal amplitude variations. In this mode the IC becomes sensitive to harmonics of the centre frequency. Pre-filtering is used in the synchronisation implementations to attenuate the unwanted signals before applying the input to the PLL. When operating in high input mode the loop is in a highly damped state, thus negating the effects of lower damping, which would cause the loops lock-up times worst scenario to degrade.

Concluding Remarks

This chapter presents an overview of the communications systems and principles developed for each of the modulation techniques presented in this thesis. This follows onto an in depth presentation of the individual modulation schemes in the following chapters.

Introduction to the Three-Chip Coded Four-Frequency Shift Keying (3C4FSK) Modulation Technique chapter

The Three-Chip Coded Four-Frequency Shift Keying (3C4FSK)-modulation techniques chapter follows, introducing a modulation technique that uses a sequence of frequency tones (character-to-frequency coding) to represent a single bit. The chapter presents an in-depth description of the 3C4FSK-modulation technique that was chosen from this research as the 'most appropriate' for further investigation. The technique utilises a combination of frequency diversity and time diversity to reduce the effects of the channel characteristics on bit identification. Whilst providing a way of intrinsically transmitting synchronisation information without explicitly transmitting a dedicated symbol frequency (explicit frequency transmission can be problematic with the frequency fading characteristics that are sometimes evident). This technique does not over complicate the receiver implementation and lends itself to ML matched filter detection and DSP realisation.

5. The Three-Chip Coded Four-Frequency Shift Keying (3C4FSK) Modulation Technique

1. The Three-Chip Coded Four-Frequency Shift Keying (3C4FSK) modulation technique uses a sequence of frequency tones (character-to-frequency coding) to represent a single bit. Each bit period is divided into three equal time periods; called 'chips'. Two frequencies are required for the transmission of each bit. When a particular bit value is transmitted the same frequency tone occupies the first two consecutive 'chips' in the bit period (chip three and chip two). This initial pair of 'chips' is followed by a second frequency tone in the final chip period (chip one). As there are two possible bit values, four frequencies are required in the system's modulating frequency alphabet (symbol alphabet). Two frequencies are used for the transmission of a bit one and the other two frequencies are used for the transmission of a bit zero (Fig.5.1).

2. Modulation coding has been incorporated into the modulation scheme by assigning a sequence of frequencies to each bit value (character-to-frequency coding). A reduction of the bit transfer rate is an inevitable by-product of character-to-frequency coding; this is offset by having reduced length individual symbol periods ('chips') allowing a number of them to be placed within each bit period. The relationship between chip periods (T_c), symbol periods (T_s) and bit periods (T_b) in a scheme is described by:

$$T_c = T_s = T_b/3 \qquad \textbf{(5.1)}$$

Modulation coding aims to provide more robust communications, in return for a reduction in the system data rate, when compared to single frequency assignment based schemes like simple FSK.

Symbol Frequency	Bit Value	Value Assigned
F_1	0	(+1)
F_2	0	(-1)
F_3	1	(+1)
F_4	1	(-1)

Table 5.1 Bit Timing Frequency Assignments

3. In a scheme the modulation coding is provided without data rate reduction by utilising 'chips' as described above. The symbol period is reduced and the bandwidth requirement is increased but the data rate is maintained. The modulation coding is used to provide some form of resistance to the noisy channel. The protection provided against noise is achieved by providing a unique sequence of frequencies to representing each bit, thus provides an element of redundancy which can be used to 'check' each bit value decision. In other words the whole frequency sequence does not have to be received correctly for an informed decision to be made. The error has to be able to simulate or at least be more similar to the sequence of the bit value that was not transmitted to result in an incorrect decision. As the sequence of frequencies (assigned to each bit) involves transmission of more than one chip, the receiver decision can be made in terms of the number of 'chips' that are 'correct'; and the likelihood that a certain bit was transmitted can be decided with a higher level of certainty. This technique is more robust in terms of its susceptibility to frequency-specific fading, because of its use of sequences of 'chips'. This is due to the fact that more than one frequency is assigned to each bit value. Thus not all the frequencies have to be detected correctly for a correct decision to be made.

4. At the beginning of the transmission a 'unique' sequence of bit values is sent (which does not appear in the data transmissions); the receiver would look for this sequence to establish the beginning of data transmissions. Any reoccurrence of the sequence redefines the start of the data transmissions. The receiver can 'look out' for the initial sequence to occur for a limited time (about one second); longer than the expected delay for the most significant multipath signal to occur. In the most extremely delayed cases, where a minimum phase channel is being used, the last arrival is the 'most significant', assuming the significant signal arrives within the 'look out' period then the start of transmissions is defined by the last arrival. The data transmissions follows the last 'unique' initial sequence detected in the 'look out' period.

5. The synchronisation scheme used within the Three-Chip Four-Frequency Shift Keying (3C4FSK) modulation scheme is a form of intrinsic synchronisation. A three chip coded scheme incorporates a mechanism for the transfer of bit synchronisation information (for bit timing at the receiver) without having to include additional signalling. The synchronisation information is present in the transmitted signal no matter what sequence of bits is sent. This is achieved by the regular change of frequencies that the chip sequence provides; once within each bit period. Each of the bit periods is made up of three sub-periods ('chips'). The regular transition (change in frequency) in every bit period is provided by assigning different symbols to 'chips' within a bit period; this is used at the receiver for clock recovery. If each pair of symbol frequencies F_1, F_3 (for bit zero) and F_2, F_4 for (for bit one) are assigned a +1 and a -1 respectively (Table 5.1); then the clock waveform (Fig.4.9 (a)) can be used to extract the bit transmission frequency at the receiver.

6. It is advisable that the coded modulation sequence (chip sequence), for each bit does not include certain combinations of the symbol frequencies. Any implementation of closely related signals is avoided by including assignments that utilise tone repetitions, reverse tone sequences, cross-bit symbol repetitions and cross-bit symbol assignment:

1. Tone repetition in a system is when the same frequency is repeated in all 'chips' associated with a particular bit period. The same frequency repeated would be easily simulated by the underwater channel; delayed versions of the same signal (multipath) could result in a single symbol transmission being mistaken for the sequence of three 'same' symbol transmissions that a tone repetition bit assignment requires for a particular bit detection. The intrinsic synchronisation that is provided by a scheme needs is not achieved using tone repetition.

2. Reverse tone sequences occur when the sequence of frequencies that represent a bit zero is the same as the frequency sequence that represents the one bit, when the first sequence is reversed. Systems that use a pair of 'chips' for each bit are particularly susceptible to problems associated with reverse tone pairing. An error in the detection of the chip sequence could result in the incorrect bits being detected at the receiver thereafter. A three-chip implementation is potentially a more robust option, in terms of dealing with reverse tone sequences; but their use is still discounted.

3. Cross-bit symbol repetition is defined here as when two equally positioned 'chips' in both the bit value sequence assignments are given the same symbol. This can limit the benefits of the modulation coding, as the redundancy that is provided is reduced, as the detection of one of the chip periods provides no 'extra' bit distinction information. Cross-bit symbol repetition could provide the synchronisation information but the modulation coding is made less effective in terms of bit distinction. This assignment could especially affect any detection decisions that compare the chips in the different bit values to make a decision, as is done with energy detection.

4. Cross-bit symbol assignment is similar to cross-bit symbol repetition. It is the assignment of the same symbol to different bit values within any of the chip periods. This can compromise the modulation coding benefits incorporated in the scheme, as the redundancy is essentially reduced. ISI can result in the receiver detecting that a transmitted symbol is present in a later chip period. This assignment could especially affect any detection decisions that compare the 'chips' in the different bit values to make a decision, as is done with energy detection.

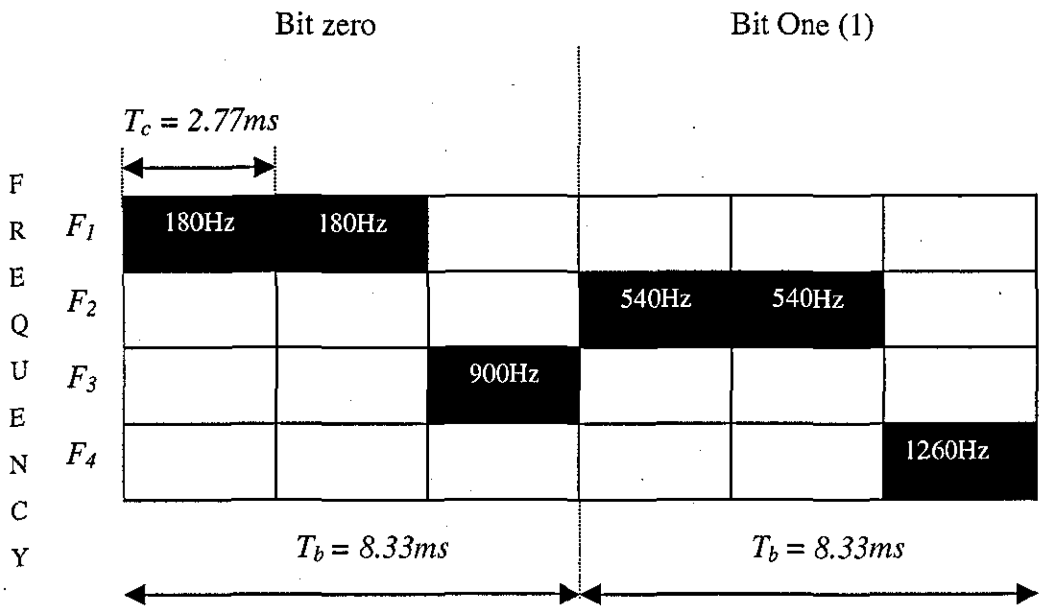


Fig.5.1 (a) The 120bps Schemes Transmission Symbol Assignment

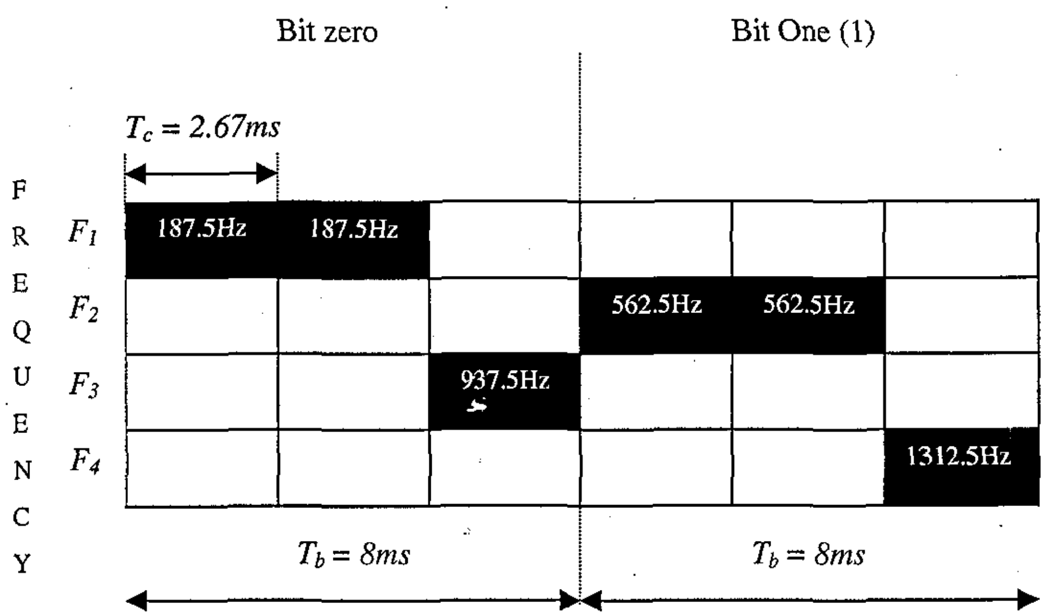


Fig.5.1 (b) The 125bps Schemes Transmission Symbol Assignment

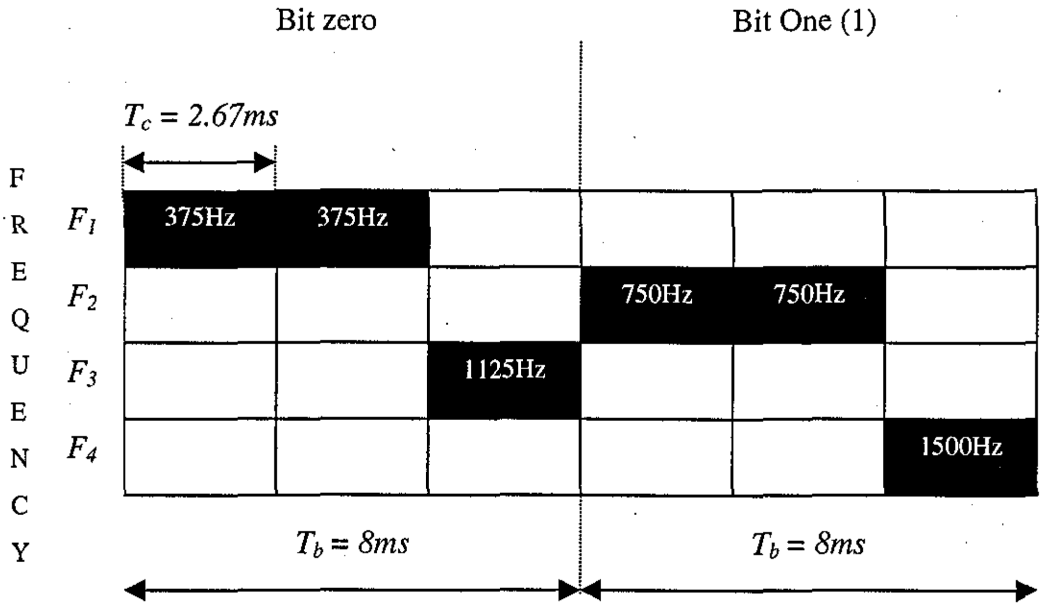


Fig.5.1 (c) The 125bps 'Shifted' Schemes Transmission Symbol Assignment

5.1. Implementation

7. The choice of the implementation was dictated primarily by pre-defined system restrictions and the platform that was used (DSP). The restrictions are described in the form of system requirements, drawn up together with BG plc. The system requirements are highly restrictive to the implementation of a system. There are a number of parameters that are not explicitly defined by the system requirements, but are restricted by them. These 'semi-flexible' parameters include, the choice of the exact values for the four modulating frequency tones in the modulating frequency alphabet. The frequency separation between the frequencies in the alphabet is limited by the use of an orthogonal set of frequencies that all lie within the allocated bandwidth. Orthogonal separation is defined by the symbol rate (Equation 4.5); therefore the frequencies are indirectly defined by the system requirements in the form of the required bit rate.

8. The design of a scheme is dictated by the platform chosen for its implementation. The DSP platform chosen has sampling limitations, in the form of maximum attainable sampling rates. The sampling frequencies used in the DSP implementations were chosen while taking into account the Nyquist's theorem and the DSP sampling limitations.

9. Initial transmission and reception implementations were carried out together on a single DSP with the receiver processing carried out on the locally generated signals to verify the 'correct' receiver operations. This was carried out without real-time input and output processing, which was incorporated later on for the final working system implementations. The high sampling frequencies used in the initial implementations enable 'coincident point decision' processing to be used. The coincident point decision process was modified for implementation in the real-time input-output systems; which require a lower sampling rate intensive decision to minimise the use of the limited DSP processing power. The real-time input-output systems were also evaluated with locally generated signals (as with the early implementations); the transmission output was passed through the output ports to a DAC, ADC and back through the input port. The coincident point schemes (non real-time input and output) have additional restrictions on the sampling rate that can be used. A relatively high sampling frequency is required and the period of each symbol frequency must have a whole number of samples within it. This number is divisible by the number of coincident points required.

10. For the 120bps coincident point system that was initially developed, a sampling frequency of 37800Hz was appropriate as it satisfies the Nyquist criterion and is within the AT&T DSP32C sampling frequency range. This 37800Hz sampling frequency was chosen because it could be divided by the half-cycle frequencies (twice the actual frequency) of each of the modulating frequencies (symbols) in the 120bps alphabet (Table 5.2 (a)). The 125bps system initially developed was assigned a sampling frequency of 39375Hz, which is above Nyquist's requirement and within DSP sampling limitations. Division of the 39375Hz sampling frequency by frequencies that are twice the symbol frequencies (with periods of half a cycle the symbols) in the 125bps coincident point systems alphabet (Table 5.2 (a)) results in an integer. The initial 125bps 'shifted' frequency system developed carried out symbol decisions over a whole cycle of each modulating frequency (symbol). Each frequency from the 125bps 'shifted' symbol alphabet can divide into the sampling frequency of 58500Hz (Table 5.2 (b)) to produce an integer.

11. The modulating frequency alphabet is chosen to fit the relationship for orthogonal operation specified for non-coherent FSK (Equation 4.5), restricting the alphabet to a limited sets of frequencies with usable frequencies of separation (symbol spacing). In the 120bps and 125bps schemes, the symbol period (T_s) is equal to the chip period (T_c) and is equal to half a cycle of the lowest frequency in the relevant modulating frequency alphabet. The lowest frequency (F_1) in each schemes alphabet defines the minimum separation (Table 5.2 (a)), as the chip period (equivalent to the symbol period) is equal to half of one its cycles; the envelope detector decisions are made over half a cycle of the lowest frequency. The chip period of the system is dependent on the bit rate. The chip rate is exactly three times the bit rate (Equation 5.1). There is a specific chip period for a required bit rate, thus a 'pre-defined' lowest symbol frequency and minimum frequency separation (Equation 4.4). The four modulating frequency tones in a system are fully defined by the system bit rate and the available bandwidth. The available bandwidth (1.5kHz per side-band) restricts the choice of frequencies that can be chosen as symbol frequencies, to the set of frequencies with the minimum frequency separation. Wider spaced orthogonal frequency sets are too bandwidth inefficient to be implemented.

12. The 125bps 'shifted' scheme was designed to maximise the bandwidth available for the matched filter implementations (Fig 5.2 (c)). Maximum bandwidth utilisation was achieved by 'shifting' the 125bps systems modulating frequency alphabet 187.5Hz up the frequency band, placing the highest frequency (F_4) right on the side-band limit (1500Hz), for a 3kHz full side-band transmission scheme (Table 5.2 (b), Fig 5.2 (c)). The lowest frequency in the 125bps 'shifted' modulating alphabet (375Hz) equates to a bit-rate of 250bps (for symbol decisions over half a cycle of each frequency). By leaving the bit-rate at 125bps (with symbol decisions over a whole cycle of each frequency) the matched filter implementations can to be carried out with more taps, thus achieving better filter characteristics.

13. As mentioned above the lowest frequency (F_1) in the modulating frequency alphabet is defined by the bit rate. The bit rate in the 120bps and the 125bps schemes are equivalent to one and a half frequency cycles of the lowest modulating frequency (F_1) in the respective alphabets. In a 120bps system, the chip period (with a period 2.7778ms long) equates to a lowest symbol frequency of 180Hz (F_1) (Fig.5.1 (a), Table 5.1(a)). In a 125bps scheme the chip period (which is 2.6667ms long) equates to a lowest symbol frequency of 187.5Hz (F_1) (Fig 5.1 (b), Table 5.2 (a)). In the 'shifted' frequency 125bps scheme, the chip period of 2.6667ms was not assigned a lowest symbol frequency of 187.5Hz; instead a higher frequency was used. A 'shift' of 187.5Hz which was introduced, assigning a highest symbol frequency of 1500Hz and placing the lowest symbol frequency at 375Hz (Fig.5.1 (c), Table 5.2 (b)). A lowest modulation symbol (frequency (F_1)) of 375Hz means that the 125bps 'shifted' scheme's chip period is equal to one full cycle of F_1 in the relevant symbol alphabet; decisions are thus made over a full cycle of each symbol frequency. The choice of the three remaining frequencies in each of the symbol alphabets was carried out using the formula for minimum frequency separation (Equation 4.4); the resulting alphabets are shown (Table 5.2 (a), (b)).

14. The minimum frequency of separation ($f_{s\,\min}$) is 360Hz for the 120bps system and 375Hz for the 125bps and 125bps 'shifted' schemes (Fig.5.2 (a), (b), (c)). Larger values for the integer m in Equation 4.5 can not be implemented because of the limited bandwidth allocation. The second orthogonal set of frequencies (with m=2 in Equation 4.5) gives modulating frequency values outside the 1.5kHz side-band bandwidth. Frequencies outside the allocated bandwidth would be assigned to symbols F_3 and F_4 (F_3=1620Hz, F_4=2240Hz) in 120 bits/second system. A 125bps system would be affected with frequencies displaced even further outside the allocated bandwidth; as the frequency spacing is larger than in the 120bps scheme. The 125bps 'shifted' scheme is designed to utilise the available bandwidth fully.

Modulating Frequency alphabet	**120bps system** Sampling frequency: coincident point scheme 37800Hz	**125bps system.** Sampling frequency: coincident point scheme 39375Hz.	**Number of samples in a half-cycle. (at coincident point schemes sampling frequency)**
F_1	180Hz	187.5Hz	105
F_2	540Hz	562.5Hz	35
F_3	900Hz	937.5Hz	21
F_4	1260Hz	1312.5Hz	15

Table 5.2 (a) Modulating Frequency Alphabet (120bps/125bps)

Modulating frequency alphabet	**125bps 'shifted' system** Sampling frequency: Coincident point scheme 58500Hz. Real-time schemes: 4800Hz and 4500Hz	**Number of samples in a whole cycle. (at coincident point schemes sampling frequency)**
F_1	375Hz	156
F_2	750Hz	78
F_3	1125Hz	52
F_4	1500Hz	39

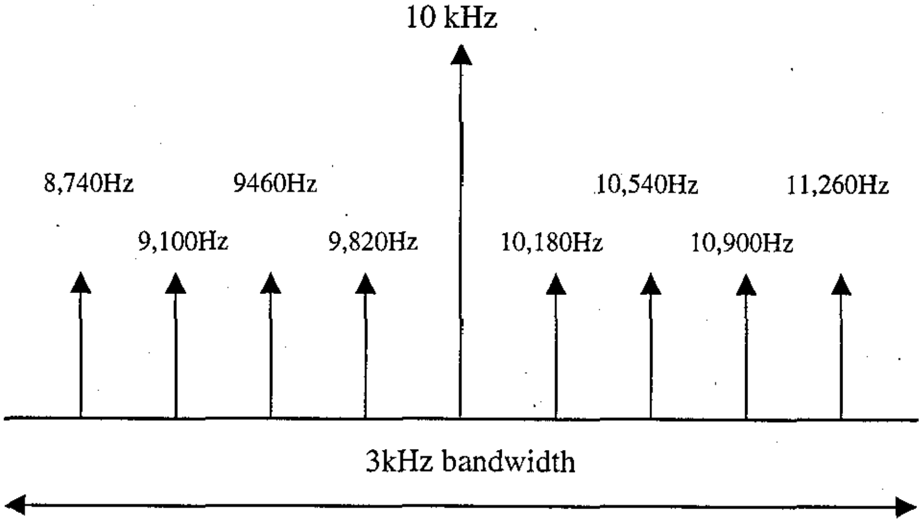
Table 5.2 (b) Modulating Frequency Alphabet (125bps 'shifted')

15. The real-time implementations developed do not utilise coincident point decision processing techniques. A 'full' decision period decision is used because of the lower sampling frequencies required with real-time input-output. For a 125bps system a sampling frequency of 4500Hz is appropriate, as it is above the Nyquist requirement (in-fact three times the highest frequency) and can be implemented by the AT&T

DSP32C. A 4800Hz implementation was also developed as this sampling frequency is in line with PC serial interface requirements (which could be useful at some point). This sampling frequency also satisfies the Nyquist requirement and is within the DSP limitations.

16. The real-time scheme's modulating frequency alphabet facilitates orthogonal operation. Maximum bandwidth utilisation was achieved by placing the highest frequency (F_4) right at the side-band bandwidth limit, which is 1500Hz for a 3kHz full side-band transmission scheme (Table 5.2 (b)). The schemes are essentially 'shifted' schemes; with a symbol frequency on the limit of the bandwidth. The definition 'shifted' is not explicitly used for the real-time input-output systems as they all could be described as 'shifted'. The largest orthogonal frequency separation within the bandwidth allocation is used that can provide the four symbol frequencies required. A lowest frequency (F_1) is assigned of 375Hz, the frequency has symbol decisions carried out over a period equal to one of its one cycles.

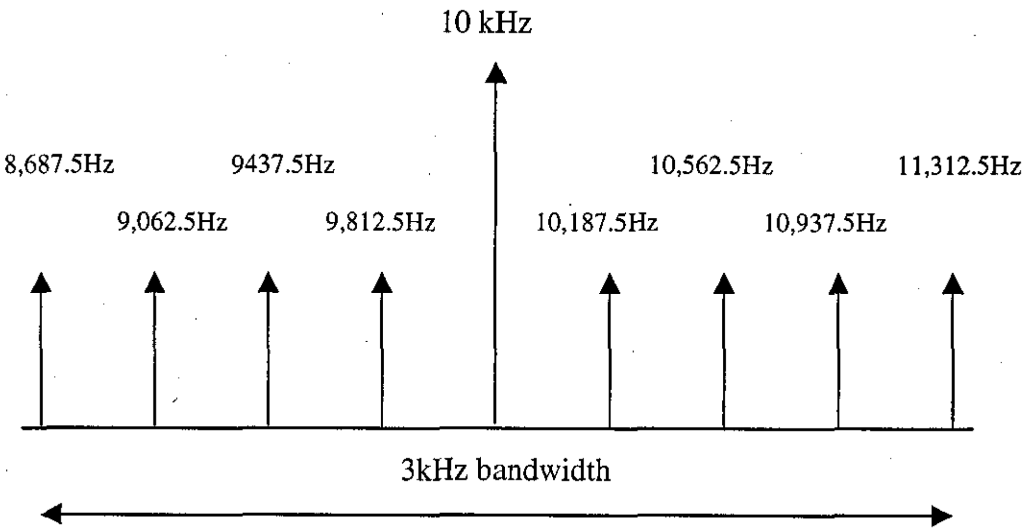
17. The real-time input-output 125bps schemes have a chip period (T_c equal to one third of a bit period T_b) of 2.6667ms which equates to one cycle of the lowest modulating frequency 375Hz (F_1); two cycles of 750Hz (F_2); three cycles of 1125Hz (F_3); and four cycles of 1500Hz (F_4). The modulating frequency alphabet is defined in Table 5.2. The separation of the modulating frequencies is fully defined by the system requirements and is not flexible; the frequencies themselves could have been placed at a number points along the allocated bandwidth. The choice of modulation frequency was carried out to uses the bandwidth most effectively in terms of the bandwidth that is made available for receiver matched filter implementations. The real-time input-output 125bps scheme uses the available bandwidth fully when the minimum orthogonal frequency separation is implemented, with larger orthogonal spacing requiring a larger bandwidth allocation.



Symbol separation: 360Hz

See table 5.2 (a), for symbol alphabet.

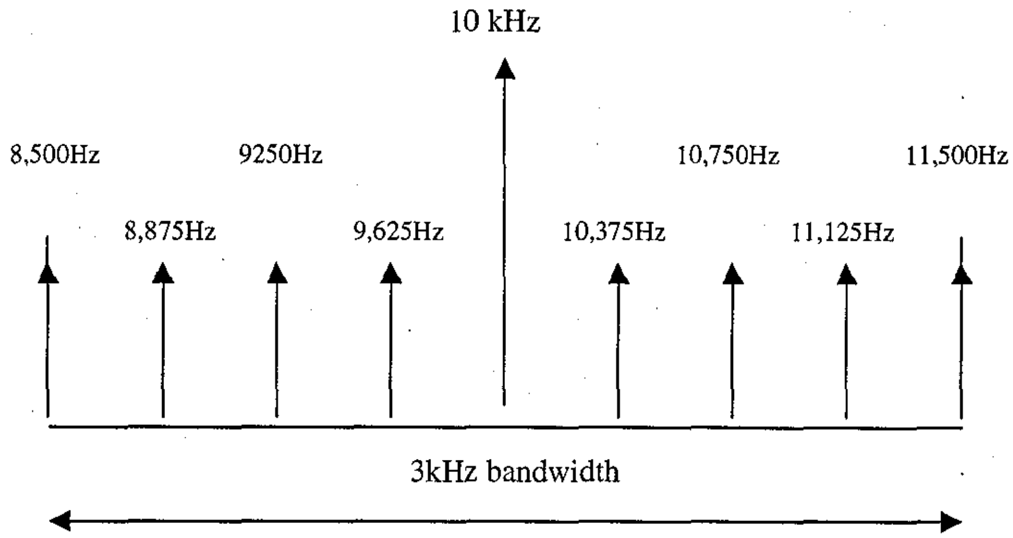
Fig.5.2 (a) Bandwidth Utilisation of the alphabet (120bps)



Symbol separation: 375Hz

See table 5.2 (a), for symbol alphabet.

Fig.5.2 (b) Bandwidth Utilisation of the alphabet (125bps)



Symbol separation: 375Hz

See table 5.2 (b), for symbol alphabet.

Fig.5.2 (c) Bandwidth utilisation of the 'shifted' alphabet (125bps)

5.1.1. Practical Design and Implementation of a System

18. There are two main components of the communications system that are required to work in synergy to facilitate transfer of data through the sub-sea environment; these are transmission and reception. The transmitter system provides the ability to generate the modulating frequency alphabet; generate the carrier waveform; utilise the 'correct' sequence of symbols dependent on the data; modulate the modulating signal on to the carrier waveform; and transmit the output signal into the medium. The receiver system's function is to; extract the acoustic signal, which will have been modified from interaction with a long-range shallow, sub-sea channel; distinguish between individual symbol transmissions; and evaluate the incoming bit sequence from the detected symbol (chip) sequence.

5.1.1.1. The Transmitter (Tx) Implementation

19. The transmission sub-system can be divided into six main processes that can be looked at as separate sub-systems (Fig.5.3). Four of the six processes are generation blocks.

A. Modulating frequency generator

B. Carrier frequency generator

C. Cyclic code generator (chip sequence generator)

D. Data generator

The other two transmission system blocks provide:-

E. Chip selection

F. Carrier modulation

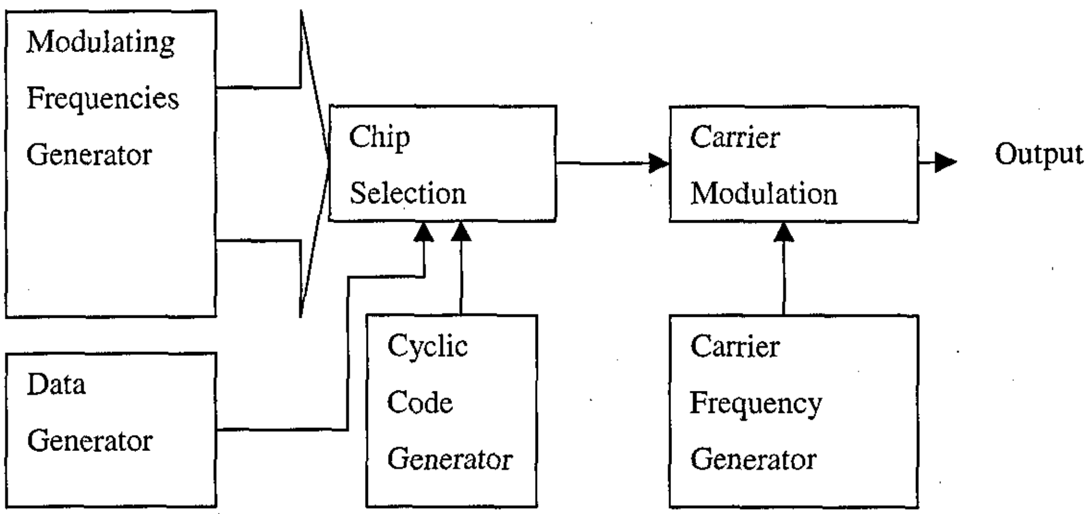


Fig.5.3 Transmitter System

5.1.1.1.1. The Modulating frequency Generator

20. The modulating frequency generator is used to generate the frequencies required for the alphabet (Table 5.2). Full band transmission limits the four frequency tones to a 1.5kHz side-band bandwidth (Fig.5.4). The conceptual design involves the generation of the modulating frequency alphabet from a source oscillator. The use of a single source oscillator allows the transmission of a continuous phase signal if required. This is achieved with the use of the chip selection sub-system, which using the fact that the frequencies that are generated all have the same initial phase when they are produced from a single source oscillator.

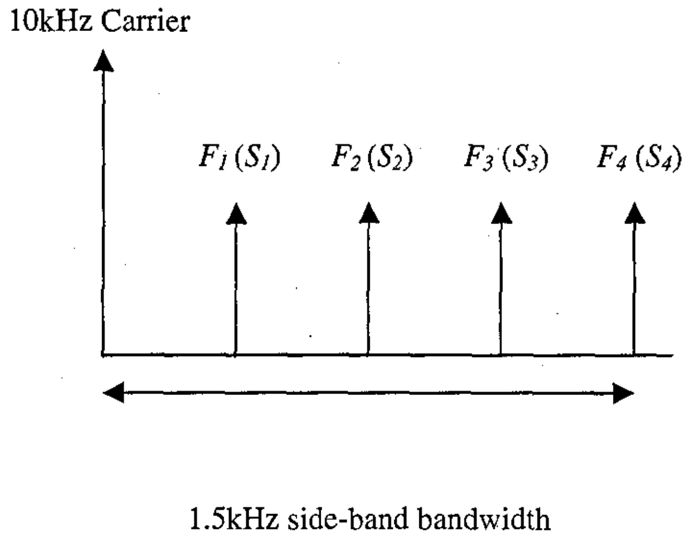


Fig.5.4 Four frequency tones side-band assignment

21. The ADSP2115 'specific' implementation of the modulating frequencies (symbols) generator is produced using analogue wave generation (AWG) software, which assigns values to each sample directly. The modulating frequencies are generated separately. Individual sample points are calculated, dependent on the frequency being generated, and the appropriate output produced. This technique produces symbol frequencies with specified initial phases. A continuous phase signal can be produced if required.

5.1.1.1.2. Cyclic Code Generator, chip Selection and Data Generation

22. The cyclic code sequence generator (chip sequence generator) and the chip selection block are described together as they work in synergy to produce the output, with the former being used to manipulate the latter. The chip selection block uses a signal (cyclic code) to differentiate (at the transmitter) between the different chip periods. Each bit period is made up of three 'chips'; two of which are assigned the same frequency tone within a particular bit period. There is a need to differentiate between 'chips' that hold unique frequency tones. The cyclic code used is effectively a pulse signal, i.e. 1 for chip three, 0 for chip two and 0 for chip one. The first chip to be sent for any bit is chip one, then chip two followed by chip three.

23. The 'chip selection' is effectively a bank of 'switches'; that allow the appropriate frequency tone to go through; depending on the value of the cyclic code. The chip selection has to be able to change the frequency tone assigned to each chip period dependent on the data's bit value. This is done by the inclusion of more 'switches' that only pass the appropriate frequency tones required for a specific bit transmission.

24. Initial DSP-based implementation of the transmission processes was carried out on the same AT&T DSP32C that is used for the receiver implementations. This was carried out to allow the receiver processes to be tested with signals generated locally. The implementation of the cyclic code was straightforward, simply involving the generation of a pulsed signal. The 'chip selection' was carried out by multiplication of the cyclic code 'pulse' with each of the frequencies in the modulating frequency alphabet (Fig.5.5). The pulse signal that is to be multiplied with the chip two and chip one frequency tones is first 'inverted' by subtraction of +1 and squaring the output. The correct frequency for chip one and chip two, or chip three is generated for both a bit 1 and bit 0. The chip three output is then added to the chip one/two output separately for bit 1 and bit 0. One of the two outputs (chip one/two or chip three) will always be zero. The outputs are added together to produce the correct chip frequency. This is done for both bit 1 frequencies and bit 0 frequencies, the bit 1 and bit 0 chip signal is then multiplied by the data bit or the inverted (*modulo*-2) data bit respectfully; the resultants are then summed together to output the appropriate chip frequency.

25. The final DSP 'specific' transmitter implementations are carried on both the Analogue Devices ADSP2115 and the AT&T DSP32C. The implementation of the cyclic code and chip selection is not required on the Analogue Devices platform as the placement of symbols into the appropriate chip periods can be carried out directly using AWG. Each sequence of symbols appropriate to each bit transmission is generated (samples are directly assigned values) by the AWG software.

26. The data is used to select the correct signals. The data value to be multiplied with the zero bit frequencies output is first 'inverted' as when using the cyclic code (Fig.5.5). The two multiplier outputs are then added together (one output is always zero), thus producing the correct chip sequence for either bit one or bit zero. The data generation in the final DSP 'specific' implementations is done independently of the transmitter using the DSP32C; the data is generated and saved on a DAT tape recorder. In real-time the DAT output signal is used as an input to the ADSP2115 platform where the signal is processed to produce a 'pulsed' sequence in which bit one transmissions have a value of one, and bit zero transmissions have a value of zero. This 'pulsed' signal is multiplied with the bit one symbols AWG generator output, and inverted before multiplication with the bit zero symbol's AWG generator output. The two outputs are added together and then the signal is modulated onto the carrier.

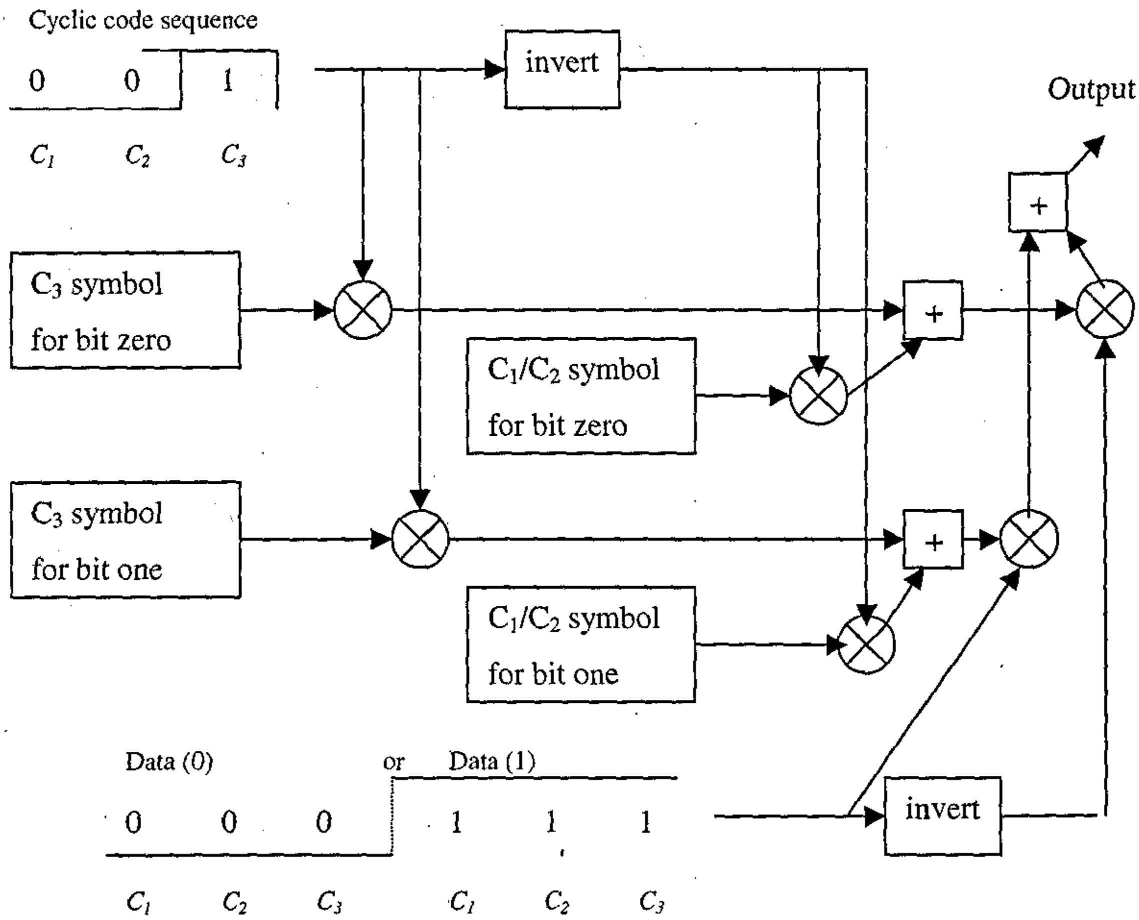


Fig.5.5 Chip selection process

5.1.1.1.3. The Carrier Frequency Generator and Carrier Modulation

27. Once the chip sequence has been produced, then modulation onto a carrier takes place. This involves two operations, carrier frequency generation and the actual modulation. The implementation is straightforward with the output of an oscillator (producing the carrier frequency) being multiplied directly with the signal (sequence of chip frequencies) to produce the output. The modulation performs the inverse operation of the receiver's carrier de-modulation, and is carried out within the transmitter transducer module. The chip sequence is passed from the Analogue Device's ADSP2115 platform (through the real-time serial interface) to the transducer module. The modulation shifts the modulating frequencies up to the 3kHz band dissected by the 10kHz carrier (Fig.5.2).

5.1.1.2. Receiver (Rx) Implementation

28. The receiver sub-system (Fig.5.6) involves more complex operations than are required in the transmission sub-system. The relatively complex structure of the receiver is required to enable evaluation of the transmitted sequence. Seven distinct processes are needed in the non-coherent receiver structure to enable a decision for each bit to be made.

A. Demodulation from the carrier waveform
B. Matched filtering of the demodulated signal
C. Envelope detection
D. 'Zero' bit value decision and 'One' bit value decision
E. Energy decision (Equality decision)
F. Bit evaluation (Choice)
G. Bit synchronisation (PLL)

5.1.1.2.1 Demodulation from the carrier and matched filtering

29. A non-coherent receiver format was developed due to the acoustic signal phase instability found in shallow-water long-range channels. Use of a non-coherent receiver negates the need for the locally generated carrier frequency to have a phase that is in-synch with the transmitter's carrier, as is required with coherent reception. The multiplication of the input signal with a locally generated carrier whose phase is not synchronised is appropriate for non-coherent detection.

30. The actual implementation of the demodulation process is carried out by the direct multiplication of the received signal with a 10kHz carrier, that is generated at the receiver. The demodulation performs the inverse operation of the transmitter's carrier modulation i.e. shifting the signal back to the bandwidth that is occupied before modulation. The demodulation process is carried out in the receiver transducer module before the signal is passed to the AT&T DSP32C (through the real-time serial interface), where matched filtering takes place (Fig.5.6).

31. After demodulation the instantaneous decision at the receiver is an evaluation to decide which of the modulating signals (symbols) are present. The decision is more accurately viewed as the non-instantaneous evaluation of the symbol that is present within a chip period. A set of orthogonal matched filters is used to produce a measure of the frequency content of the incoming signal. The filter outputs are then passed onto a set of envelope detectors, that utilise the instantaneous matched filter outputs to evaluate the symbol that is present in a particular chip period.

32. The matched filters correspond directly to the modulating frequency alphabet frequency tones (produced at the transmitter); the centre frequency of each of the band-pass filters corresponds to one of the symbol frequencies. The real-time 125bps scheme allows the 3kHz bandwidth (Fig.5.2) to be used comprehensively; the individual matched filter implementations are allotted a 375Hz bandwidth. The bandwidth allocation given to each matched filter is in line with the minimum orthogonal frequency separation (Equation 4.4).

5.1.1.2.2 DSP data processing requirements (3C4FSK)

33. The receiver implementation is relatively complex because at high sampling rates the receiver operation becomes DSP processing-power-intensive. Reduction of the number of operations carried out was achieved by limiting the use of high sampling frequencies for the scheme's receiver implementations.

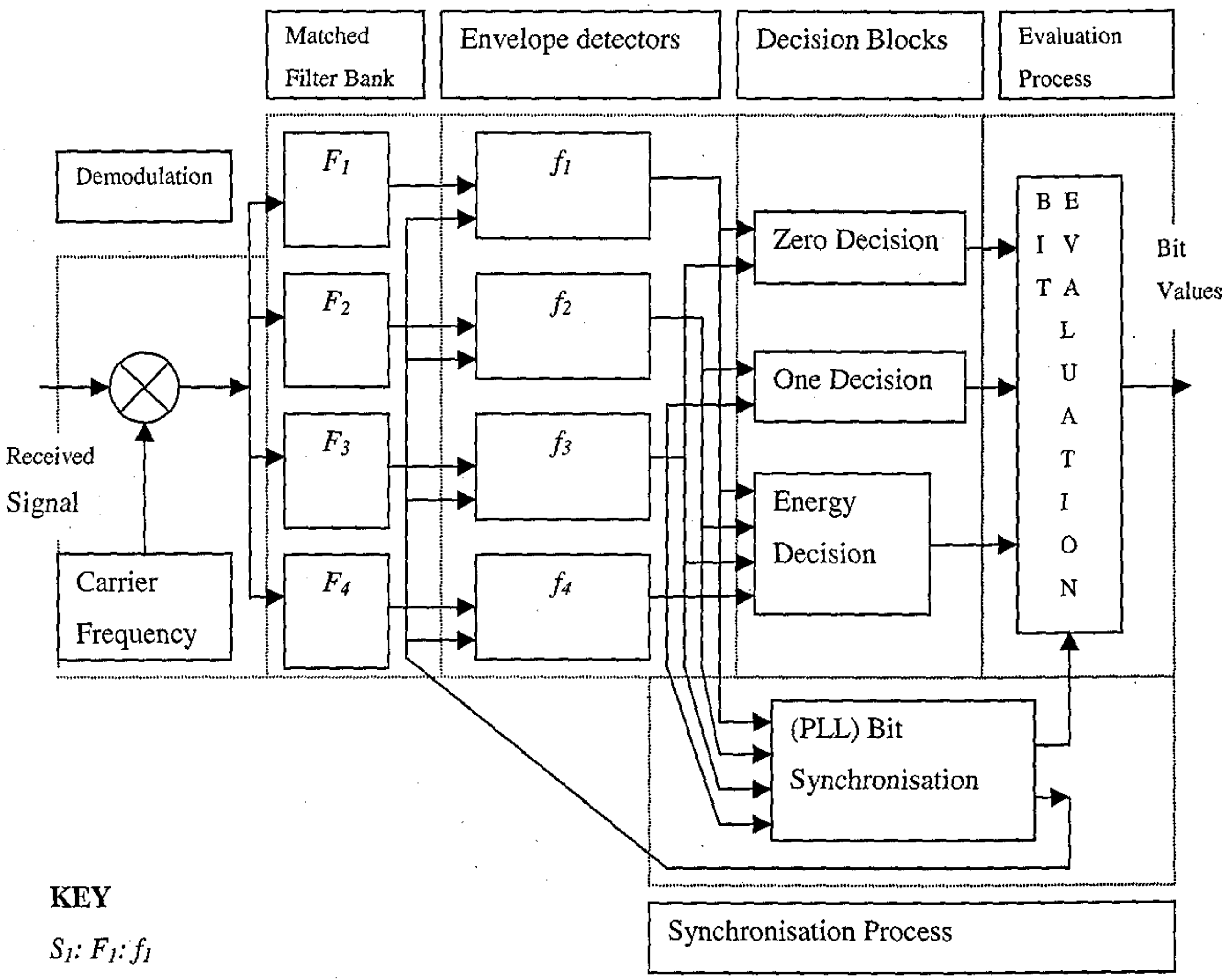
34. In the initial designs, higher sampling frequencies were used and reduced data processing was achieved by reduction of the sampling rate in the envelope detector sub-systems. The 'zero' decision, 'One' decision and Energy decision were effectively reduced to one sample for every chip period. The synchronisation processes require a 'fully sampled' signal (at the system sampling rate) for the PLL input. The input to the synchronisation block stems from the envelope detector outputs in a system (Fig.5.6), so the signals (output from the envelope detectors) were reconstructed for synchronisation. The rest of the receiver (decision sub-systems) only requires the appropriate signal to be reconstructed in the final sub-system (bit evaluation process) before multiplication with the synchronisation output. The final implementations (real-time input-output systems) utilise lower sampling frequencies; the processing is reduced without the addition of the extra complexity of sampling frequency reduction and reconstruction of the sampling frequency. This leads to implementations that do not exceed DSP processing and memory capabilities.

5.1.1.2.3 Envelope detection

35. The output of each matched filter is passed to an envelope detector, which carries out an integration over the chip period. The envelope detectors are used to provide a comparable magnitude measurement of the content of each symbol frequency in the received signal.

36. Initial designs used coincident point envelope detector processing. The coincident point integration required the same number of sampled points to be used for each frequency tone (symbol); each sample is in a coincident position in the cycle of each frequency tone. The integration was done over a set number of cycles of each symbol frequency tone. Each of the symbol's decision period holds the same number of coincident samples spread across different half-cycle periods (or whole cycle periods in some cases). There were a limited number of coincident points that could be used for envelope detection purposes as the DSP sampling rate is limited. The modulating frequency alphabet (Table 5.2 (a)) for the 120bps system and 125bps system were implemented using both three and five coincident point envelope detectors. Two different numbers of coincident points were used as not all the modulating frequencies had the same number (three or five) of evenly spread points over a half-cycle at the sampling frequency used

37. Two envelope detectors were implemented at the output of the appropriate matched filters. One of each pair of is a three coincident point detector (*fla, f4b*) (Fig.5.7), and one is a five coincident point detector (*flb, f4a*) (Fig.5.7). The single envelope detectors placed on the outputs of the filters matched to matched filter F_2 and matched filter F_3 are a five coincident point envelope detector and a three coincident point envelope detector respectively (Fig.5.7). All comparisons that are required between envelope detectors can be made with the equivalent coincident point envelope detectors. The outputs of the matched filters matched to symbol S_2 and symbol S_3 do not have to be compared directly in the receiver, so it does not matter that they do not have the same number of coincident points. The 'shifted' 125bps modulating frequency alphabet (Table 5.2) can all be divided by thirteen (using 54000Hz as the system sampling frequency) so only four envelope detectors are required (Fig.5.6). The low sampling rate systems that were developed for the real-time implementation do not provide adequate numbers of coincident points, so another method of envelope detection was used. The implementation of a magnitude-integration, over the full chip period was adopted.



KEY

S_1: F_1: f_1

S_2: F_2: f_2

S_3: F_3: f_3

S_4: F_4: f_4

Symbol: Associated Matched Filter: Associated Envelope Detector

The 125bps real-time input-output scheme structure.

Fig.5.6 Receiver System

38. The initial receiver implementations used sampling frequency reduction techniques to obtain the coincident points required for each envelope detection process. These negated the unwanted samples as well as reducing the DSP processing time. The reduced sampling frequency envelope detector incorporates the actual sampling rate

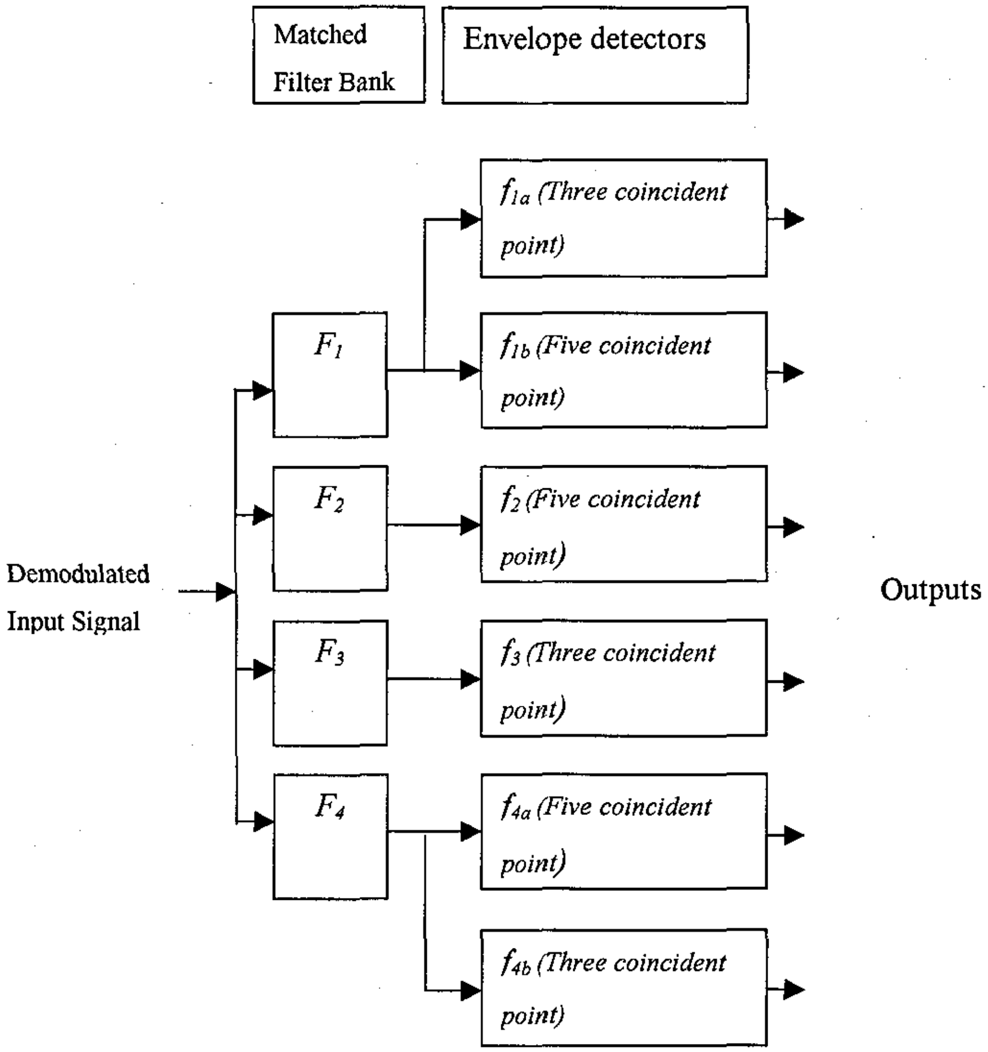


Fig.5.7 Dual Envelope Detector structure

reduction of the signal as well as the processes required to perform envelope detection. The reduced sampling frequency envelope detector's operation involved first discarding the unwanted cycles (or half-cycles) of the matched filter signal, retaining the required decision half-cycle (or whole cycle in the 'shifted' scheme). The retained signal was then rectified and its sampling frequency was reduced to leave only the coincident points that are required. The points from one decision period are summed together by a magnitude integration to produce an output for comparison with the output of other envelope detectors to distinguish which symbol frequency is present (Fig.5.8).

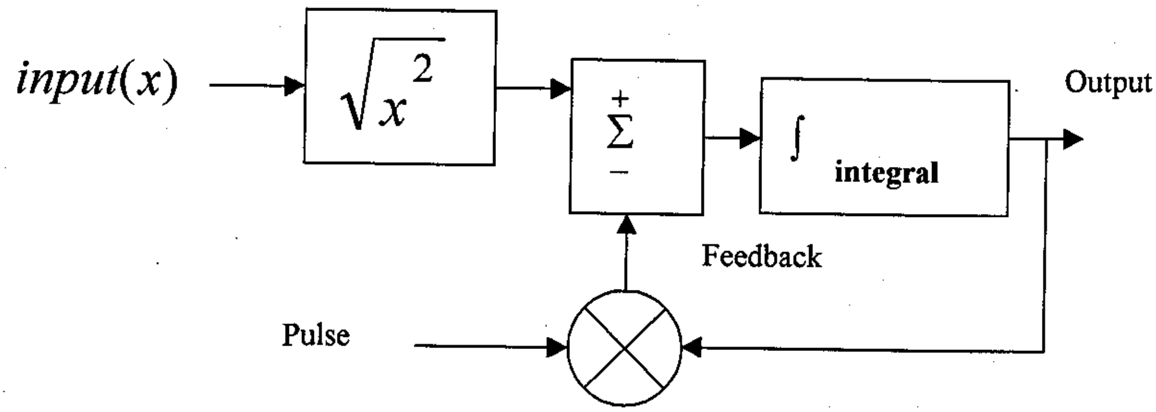


Fig.5.8 **Envelope Detection Process**

39. The DSP reduced sampling frequency envelope detector implementations achieved input signal rectification by squaring and square rooting the samples. The signal was then reduced to the required number of coincident points by performing the reduction of the sampling frequency over half a cycle (or a whole cycle in the case of the 'shifted' scheme) of the appropriate symbol (dependent on which matched filter output is being processed). The unwanted samples within the chip period were discarded. The coincident points were summed with a feedback signal, which sets the integrator to zero at the beginning of each chip period and at the beginning of the envelope detectors decision period (which dependent on the number of cycles of the frequency being used) (Fig.4.7 (a)). The signal was then fed into a magnitude integrator whose output provided the feedback signal. The feedback worked by feeding back the last output (peak value) of the integrator in each chip period, so it can be subtracted at the start of the following chip period. The last value produced by the samples outside the decision period but within the chip period was also fedback, setting the integrator to zero at the beginning of the decision period. A pulsed signal derived from the synchronisation process (the signal being one only when a feedback signal is required), is multiplied with the integrator output before the delayed resultant is subtracted from the input to the integrator (Fig.5.8). The resultant is delayed by one sample position so the subtraction is made at the appropriate instants. The peak value (which is last value in a chip period) of the integrator from a chip period is passed as an output of the envelope detector.

40. The reduced sampling frequency envelope detectors all have basically the same structure, with differing sampling frequency reduction factors and pulse values dependent on which matched filter frequency they are operating on and on the number of coincident points being used. The reduced sampling frequency envelope detector's operation for the 120bps and 125bps schemes uses half-cycle decision periods. The 125bps 'shifted' scheme has the same operations, but the detection is carried out over a whole cycle instead of a half-cycle and the same number of coincident points is used for all of the envelope detectors.

41. The envelope detectors used in the real-time input-output systems involve an integration over the full chip period ('full period' envelope detection); as opposed to a section of the chip period being used as in the coincident envelope detection process (Fig.4.7 (a), (b)). Implementation of 'full period' envelope detection is relatively less complex, as reduction of the sampling rate is not required and the integration uses all samples within a 'chip period'. The DSP 'specific' envelope detector implementation performs input rectification on the matched filter output by squaring and then square rooting the samples. The signal is then summed with a feedback signal, which is equal to zero minus the value of the integral of the samples from the previous chip period, initialising the integration (getting rid of any previous values). The signal is then feed into a magnitude integrator whose output provides the feedback signal. The feedback works by delaying the largest output (peak value) of the integrator in the chip period, so it can be subtracted at the start of the next chip period. A pulse signal derived from the synchronisation process is multiplied with the integrator output (which is one only for the last sample in the chip period) before the delayed resultant is subtracted from the input. The resultant signal is delayed by one sample position, so the subtraction is made from the first sample of the following chip period. The integrator output is passed as the output of the envelope detector.

5.1.1.2.4 Zero and one bit decision

42. The zero decision and one decision are treated as totally separate evaluations in most of the schemes that have been developed (Fig.5.9); they are described together in this section because they perform identical operations, but on different input signals. The decision is a measure of 'how close' the sequence of 'chips' (symbol frequencies) is to that expected for either a zero or a one bit transmission. The evaluation is carried out for both bit decisions (zero and one bits) in parallel.

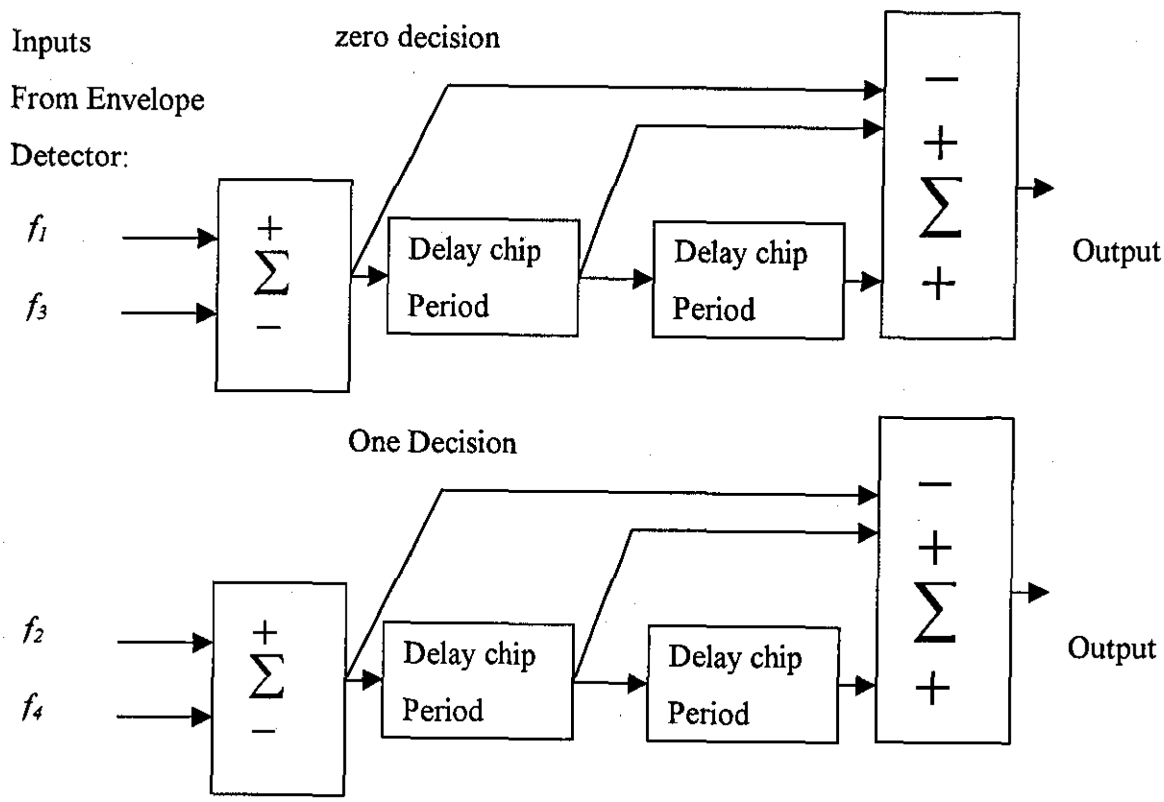


Fig.5.9 Zero and one decisions (bit decision)

43. The zero and one decisions are made independently, by focusing only on the chip frequencies appropriate to either of the possible bit values. The zero decision compares only the zero bit chip frequencies (outputs of the envelope detectors $f1$ and f_3 shown in Fig.5.6) and assigns -1 or +1 for a particular chip dependent on their comparative magnitudes. The one decision does exactly the same operation but uses the outputs of the envelope detectors f_2 and f_4 (Fig.5.6).

44. Both the decision operations (zero decision and one decision) perform an addition of the first two 'chips' (chip one, chip two) in a bit period and subtracts the third chip (chip three) (Fig.5.9). The calculation generates a value +3 for a correct chip sequence (chip three = -1, chip two = +1, chip one = +1). A value between +3 and -3 can be produced, with +3 representing the 'most likely' correct and –3 the 'least likely' correct. The 'chips' are moved on one place in the decision calculation when the next chip arrives. The new incoming chip (chip four) becomes chip three and the old chip three becomes new chip two, the old chip two becomes new chip one and finally the old chip one is discarded (Fig.5.10). This is carried on as new 'chips' arrive.

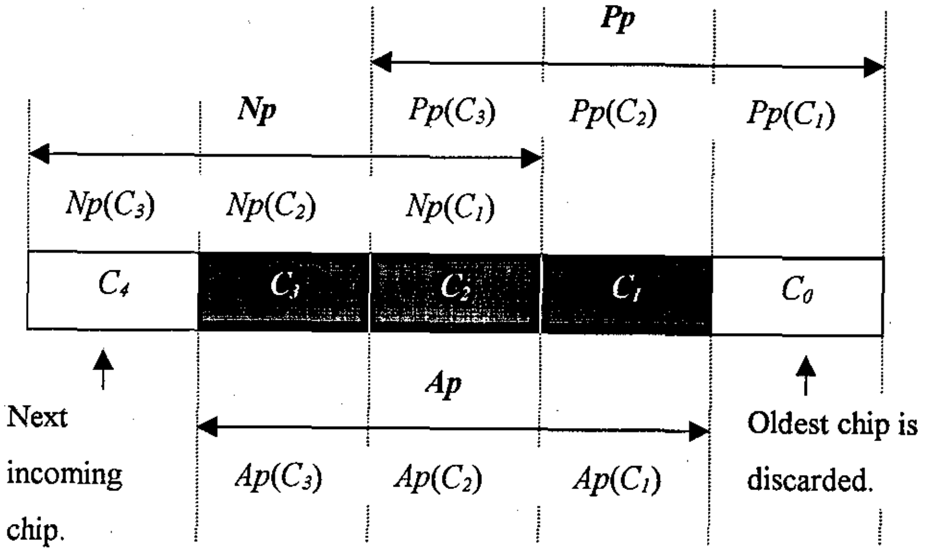
45. There are multiple outputs produced per-bit-period for both the zero and one decisions, but the bit evaluation block is only used when three 'chips' from the same bit period are involved in the calculation. The synchronisation module is used to decide when the output is correct; when all 'chips' in the bit evaluation stem from the same bit period.

46. The initial receiver 'DSP-specific' implementations used zero and one decision implementations that operate on a reduced sampling frequency output, produced by the coincident point envelope detectors (one sample per-chip-period). A subtraction of the chip three value from the summed value of chip two and chip one values can be carried out in a straight forward manner as there is only one sample per-chip-period . The output signal was then replaced with -1 or +1 dependent on if it is negative or positive respectively. The signal was either delayed by a single chip period or two chip periods to allow the summation of chip one and chip two to take place. The two outputs where added together and the third value (chip three) was subtracted from the result. Two output values are produced separately, one by the zero decision and one by the one decision.

47. The real-time input-output implementations ('DSP-specific' implementations) use zero and one decisions that operate on the 'fully sampled' signal (sampled at the schemes sampling rate Table 5.2), that is produced by the envelope detectors (Fig.5.6). A summation of the chip one and the chip two values is carried out together with a subtraction of the chip three value. The values for each of the 'chips' are required once every chip period (one sample) for the bit decisions to operate correctly. A 'fully sampled' signal means that there are additional samples within each chip period that are not required for the bit decisions; the samples are all nullified except for one sample in ever chip period. The synchronisation signal is used to produce a pulse once every chip period. This is multiplied with the input signal. The pulses are generated as a value of +1 for chip one and chip two, and -1 for the chip three. A magnitude integral is made over the bit period, after the previous bit period's integrals last value is subtracted. Multiplication of the integral output with a pulse which is only one (+1) for last pulse generated in a bit period (chip three's pulse) takes place, enabling only the last integral value to be output (a value between 3 and minus -3). The decision outputs are compared by the subtraction of the bit one decision output from the bit zero decision output; a positive result indicates that the zero decision is dominant and a negative result indicates that a one decision is dominant. The signal is limited to +1, -1, and 0 values; indicating a zero bit, a one bit and a tie respectively at the decision sample instant. The signal is then passed to the bit evaluation process; in the case of a tie the energy process is used to decided between a zero bit and a one bit.

5.1.1.2.5 Energy decision

48. There are times when the implementation of the parallel bit decision processes can leave the bit evaluation unresolved. It is possible to achieve a situation where both decisions are equally likely. An energy decision is then implemented (Fig.5.11). The energy decision is a way of distinguishing between the two possible bit values when the bit decision is unresolved. A measure of the energy content stemming from a zero and a one bit is made across all three chip periods in a bit period. The energies are compared to establish the 'winner' of any contention that may occur in a particular bit decision.



Ap Active 'perceived' bit period, *Pp* Previous 'perceived' bit period,
Np Next 'perceived' bit period.

Fig.5.10 'Chip' allocation throughout bit decision's

49. The DSP-specific implementation of the energy decision involves subtraction of the one bit decision chip one/two output (of the envelope detector f_2 associated with modulating frequency F_2 shown in Fig.5.6), from the bit zero, chip one/two output (of envelope detector $f1$ associated with modulating frequency F_1). There is also a subtraction of the bit one, chip three output (of envelope detector f_4 associated with modulating frequency F_4) from bit zero, chip three output (of the envelope detector f_3 associated with modulating frequency F_3). The outputs are replaced with +1 or 0 dependent on whether the subtraction yields a positive or negative result. The first subtraction result (for chip one) is delayed for two chip periods. The same subtraction (using the same envelope outputs as the first subtraction) is carried out for chip two and the result is delayed one chip period. A further subtraction is carried out, for chip three and then the result is summed with the delayed results from the chip one and chip two subtractions (Fig.5.11). An output value between 0 (for a bit zero) and 3 (for bit one) is produced.

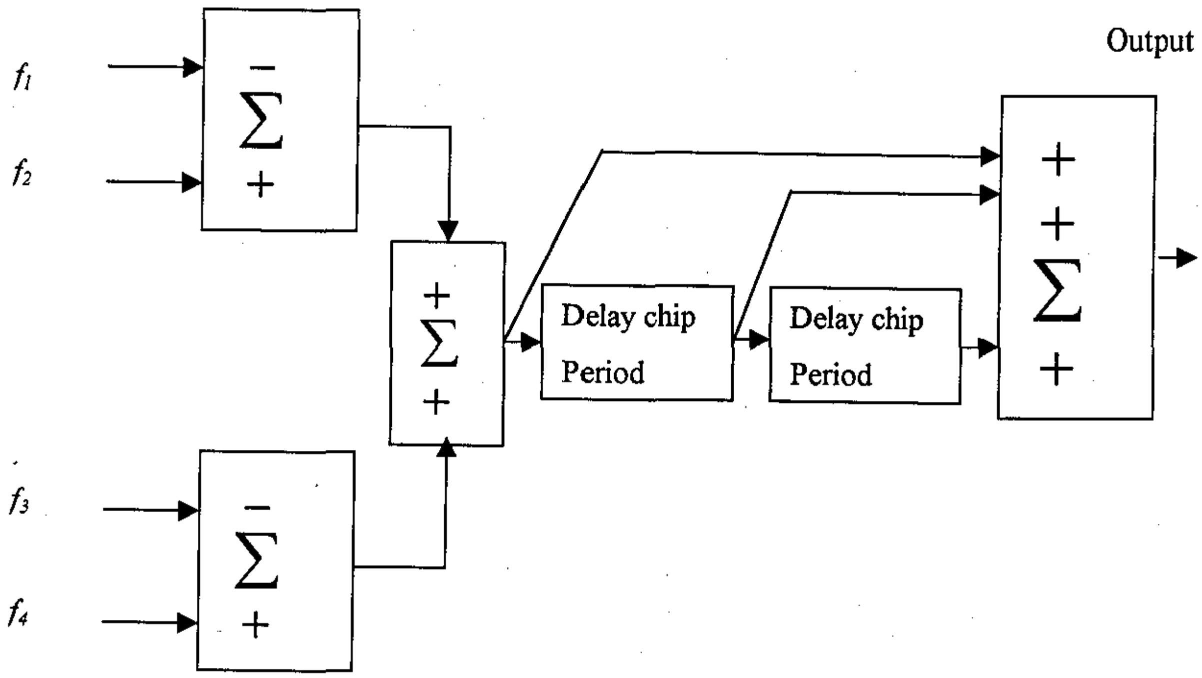


Fig.5.11 Energy Decision

5.1.1.2.6 Synchronisation MFSK (3C4FSK)

50. The output data is "correctly" evaluated within the bit evaluation block when the three-chip periods involved in the zero and one bit decisions all originate from the same bit period transmission. The output therefore is not required at instants when the three 'chips' do not originate from the same chip period. The modulation intrinsically transfers the synchronisation information with the data. The chip decoding carried out by the bit decisions incorporates the ability to distinguish between the 'correct' and 'incorrect' sequence of 'chips' for each bit value. The one change of symbol frequency included within each bit period provides a mechanism for the transfer of the bit rate to the receiver within the sequences of 'chips' without additional transmissions. The output can thus be obtained once every bit period. The bit rate (frequency) extraction is carried out utilising a second order phase locked loop (PLL); which locks to the frequency of change of the regular sequence of chip (symbols) in each bit period. The PLL input signal is provided by a comparison of the chip one/two and chip three envelope detectors outputs. The output is filtered and made into a square-wave format to be passed to the first input of a PLL's detector. The synchronisation information extraction has been implemented utilising a NE567 Integrated Circuit. The synchronisation methodology follows the second synchronisation methodology (Section 4.3.1.2).

51. To establish the rate of change of the chip frequencies the Phase Locked Loop compares the input signal with a feedback signal using a detector (Appendix B). The feedback originates from the oscillator output. The detector output is passed through a loop filter (LPF), the output of which controls the oscillator at the incoming bit rate. The oscillator output is feed back to the detector. Pre-filtering of the PLL's input signal with a band pass filter (BPF) is performed to allow the PLL to adjust to the Signal to Noise Ratio (SNR), negating noise effects by restricting the range of input frequencies that the PLL can encounter.

Concluding Remarks

This chapter follows on from the previous overview chapter, and presents an in-depth description of the 3C4FSK-modulation technique. Providing an insight into the transmission and receiver implementations adopted. The following two chapters present the other modulation techniques that have been developed.

Introduction to the Five-Frequency Four-Space-Frequency Shift Keying (5F4SFSK) Modulation Technique

The Five Frequency Four-Space Frequency Shift Keying (5F4SFSK) modulation technique chapter follows, introducing a modulation technique that uses transmission of parallel frequency tones to represent a single bit. The chapter presents an in-depth description of the 5F4SFSK-modulation technique. The technique utilises frequency diversity to reduce the effects of the channel characteristics on bit identification. Whilst providing a way of synchronisation information by explicitly transmitting a dedicated symbol frequency. The technique does not over complicate the receiver implementation and lends itself to ML matched filter detection and DSP realisation. This technique was not chosen as the 'most appropriate' for further investigation as the use of explicit frequency transmission for transmission of synchronisation information can be problematic with the frequency fading characteristics that are sometimes evident in this environment.

6. The Five-Frequency Four-Space-Frequency Shift Keying (5F4SFSK) Modulation Technique

1. The Five Frequency Four-Space Frequency Shift Keying (5F4SFSK) modulation technique uses two unique frequency tones for each bit-value. Each pair of bit frequencies is transmitted together, along with an additional frequency tone that is used for bit synchronisation. The synchronisation tone is transmitted continuously throughout transmission (Fig 6.1).

2. Modulation coding can be used to establish more robust performance in inhospitable environments, including shallow long-range sub-sea channels. A 5F4SFSK scheme provides parallel redundancy instead of 'serial redundancy' provided by a 3C4FSK scheme. In theory, only one frequency is required per bit value but by assigning an additional ('redundant') frequency to each bit-value the scheme is made more robust in terms of susceptibility to frequency-specific fading. The relationship between chip periods (T_c), symbol periods (T_s) and bit periods (T_b) in a 5F4SFSK scheme is described by:

$$T_c = T_s = T_b \qquad (6.1)$$

3. The protection is provided without any need to reduce symbol rates or reduce the bit rate. The intrinsic synchronisation benefits of the character-to-frequency coding used in the 3C4FSK scheme are not provided in a 5F4SFSK scheme, so an additional frequency is transmitted for bit synchronisation purposes. The complex receiver decision, which involves deciphering a sequence of frequencies for each bit-value (carried out for a scheme) is not required here as there are no sequences of symbols within a bit period (longer symbol periods are implemented). In a 5F4SFSK scheme there is a price paid for the loss of synchronisation information (which could be provided by a sequence of symbols within each bit period) of a tighter bandwidth restriction. Five frequencies within the limited bandwidth makes the equivalent matched filter implementations are more taps-intensive.

4. At the beginning of the transmission a 'unique' sequence of bit values is sent (which does not appear in the data transmissions); the receiver can look for this sequence to establish the beginning of data transmissions. Any reoccurrence of the sequence redefines the start of the data transmissions. The receiver 'looks out' for the initial sequence to reoccur for a limited time (about one second), which is longer than the expected delay for the most significant multipath signal to occur. In the most extremely delayed multipath environments under consideration, where a minimum phase channel is in use (the last multipath arrival is the 'most significant'), it is assumed that the 'most significant' signal arrives within the 'look out' period. The data transmission follows the last occurrence of the 'unique' initialisation sequence detected within the 'look out' period.

5. Bit synchronisation information is transferred explicitly by the continuous transmission of a frequency tone throughout communications, specifically for bit timing. The bit timing data is present in the transmitted signal no matter what sequence of bits is sent. The receiver would be required to perform frequency synthesis if the synchronisation frequency transmitted was not the same as the bit rate but in the 5F4SFSK schemes implemented, the synchronisation frequency was chosen to be equal to the bit-rate and orthogonal to all the other symbol frequencies.

6. The bit synchronisation information (frequency tone) is transmitted with both bit 1 and bit 0 transmissions. If the bit synchronisation information was only available from one of the 'bit-value transmissions', then synchronisation (PLL lock) may be lost in long sequences of transmission of the 'other bit-value'. Thus there is a need for an extra synchronisation frequency, which has to be placed within the system's predefined bandwidth allocation (Fig.6.5). A 5F4SFSK scheme's longer symbol periods, which are three times larger than for 3C4FSK, allow less separation between orthogonal frequencies (Equation 4.5) thus extra frequencies can be accommodated.

7. It is advisable that the assignment of frequencies (symbols) to bit-values involves the placement of adjacent symbols with opposing bit values. Therefore adjacent symbols can be paired off, and used for individual decision processes. This is appropriate as frequencies that are more closely positioned are 'more likely' to have similar fading characteristics. Non-adjacent symbol assignment has been used (Fig.6.1).

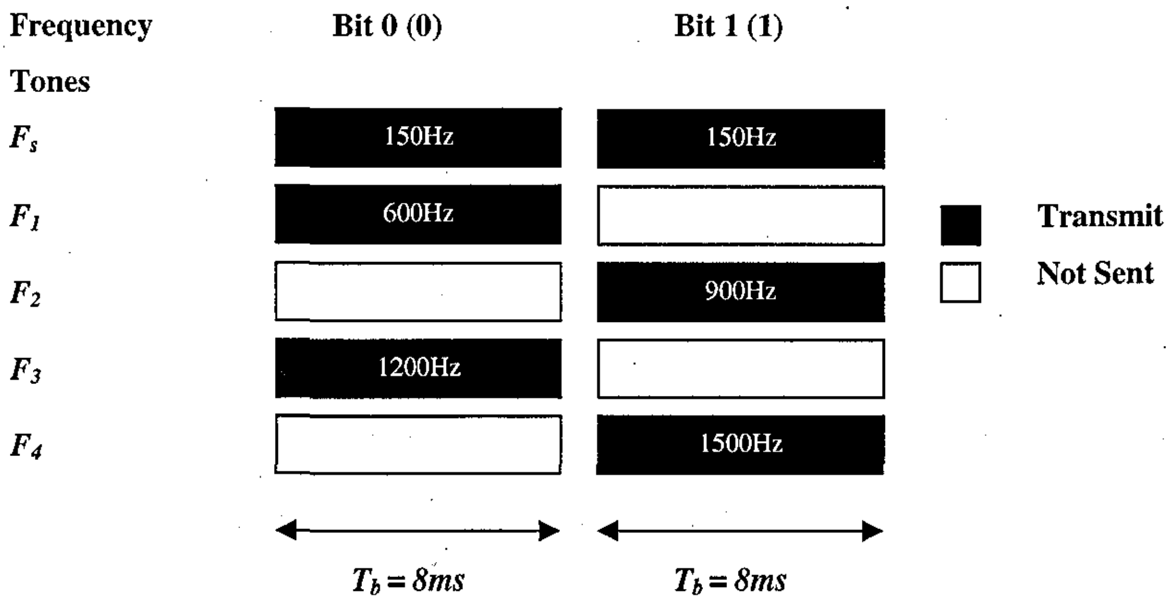


Fig.6.1 5F4SFSK Scheme's Transmission Symbol Assignment

6.1. 5F4SFSK Implementation

8. The 5F4SFSK implementation has to accommodate the same system requirements as defined for the 3C4FSK scheme; the requirements were drawn up together with BG plc. The relatively long symbol periods (compared to the 3C4FSK implementation) associated with this scheme means that the spacing required to obtain orthogonal frequencies is decreased. A greater number of orthogonal frequencies, up to twelve for a 125bps scheme and up to ten for a 150bps scheme, can therefore be placed within the allocated bandwidth.

9. As with the 3C4FSK scheme the transmission and reception implementations of the 54SFSK scheme were first of all carried out on a single DSP, with the receiver processing locally generated signals to allow verification of the 'correct' receiver operation. Real-time input and output processing was incorporated at a later stage of development. The 'coincident point decision' processing used in the 'prototype' systems required high sampling frequencies. Modifications to the decision processes allowed lower sampling rates to be implemented, thus enabling real-time input and output, thus minimising the use of DSP processing power. These real-time input-output systems were evaluated with locally generated signals (as with the systems). The implementation of modulating frequencies that all have at least two cycles (half the frequency) in the system bit period was chosen for the coincident point 5F4SFSK scheme to allow symbol decisions to be made over this period. The two-cycle symbol decisions (compared to one-cycle or half-cycle used in the 3C4FSK coincident point schemes) are used to allow adequate matched filter implementations in the more intensively used bandwidth of a 5F4SFSK scheme.

10. The system sampling rate is dictated by the Nyquist theorem requirements (at the lower end) and the AT&T DSP32C's limitations (at the higher end). The choice of system sampling rate is also determined by the amount of processing that the receiver has to undertake; the more processing required the longer the periods in between sampling instants are required to be. The sampling frequency implemented for the 5F4SFSK coincident point system that was initially developed was 54kHz which is above the Nyquist requirement and within the AT&T DSP32C operating range. This sampling frequency can be divided wholly by the two-cycle frequency (half the frequency) of all the modulating frequencies. The sampling frequency also provides an equal number of coincident points (six) for each of the modulating frequency's decision periods. When the number of samples in two cycles of each symbol frequency (Table 6.1 (a)) is divided by six an integer is the resultant.

11. The 5F4SFSK scheme's modulating frequency alphabet is chosen to fit the relationship for orthogonal operation specified for a non-coherent FSK system (Equation 4.5). Symbol spacing is thus restricted to certain frequencies of separation. In a 5F4SFSK scheme the symbol period (T_s) is equal to the bit period (T_b); the choice of the lowest data symbol's frequency (F_1), where at least two cycles are wanted in a symbol period, can be chosen as any frequency equal to or higher than twice the bit rate. The lowest modulating frequency has to be placed so that four other orthogonal (higher) frequencies can also fit within the bandwidth.

12. The frequencies in the modulating frequency alphabet (Table 6.1), were chosen by placing the highest frequency (F_4) on the edge of the available bandwidth (1500Hz). This ensures that the largest amount of bandwidth is available for any matched filter implementations (the schemes are thus 'shifted' to the edge of the available bandwidth). Frequencies F_s, F_1, F_2, F_3 and F_4 are positioned at an orthogonal distance from each other whilst all frequencies are within the bandwidth allocation (Fig.6.2). The synchronisation frequency (F_s) was chosen to be equal to the bit rate, which is outside the allocated bandwidth for the lowest modulating frequency's matched filter (which extends down to 300Hz as four data symbols have equal allocations). If any other synchronisation frequency was sent, frequency synthesis would have to be performed at the receiver to produce the bit rate.

13. A highest symbol frequency (F_4) of 1500Hz was first assigned, then the choice of the three remaining frequencies in each of the modulating frequency alphabets (symbol alphabets) was carried out using the formula for orthogonal separation (Equation 4.5). The resulting modulating frequency alphabet is shown (Table 6.1 (a)). The minimum frequency of separation ($f_{s\,\min}$) is 150Hz for a 150bps system. With this minimum spacing the bandwidth available for each matched filter is reduced and thus the characteristics of the matched filter implementations worsens. Larger values for the integer m in Equation 4.5 can be used to make more frequencies available. Higher values of m reduces the number of orthogonal frequencies within the scheme's bandwidth allocation. A value of m equal to two was chosen for the data symbols and m equal to three for the synchronisation symbol, allowing a synchronisation symbol of 150Hz to be chosen.

14. The real-time implementations developed do not use coincident point decision processing techniques. A 'full' decision period is used because of the lower sampling frequencies required for real-time input-output operation. Versions utilising 'half' decision periods and an equal number of cycles of each frequency have also been implemented. The 5F4SFSK 150bps system was implemented with a sampling frequency of 4500Hz or 4800Hz as appropriate, as this is above the Nyquist requirement (three times the highest frequency) and can be implemented by the AT&T DSP32C.

15. The 5F4SFSK 150bps scheme is designed to maximise the bandwidth available, allowing implementation of relatively wide bandwidths for each receiver matched filter, thus limiting the number of filter taps required. The modulating frequency alphabet (symbol alphabet) is defined primarily by the need for orthogonal spacing in effective matched filter detection (Appendix A) and secondarily by the system's bandwidth and bit rate requirements. The implementation of the 'maximum' orthogonal separation of frequencies within the allocated system bandwidth is preferred (Equation 4.5). In the real-time input-output 150bps scheme a lowest data symbol of 600Hz (F_1) and a synchronisation symbol (F_s) of 150 Hz are used. The modulating frequency alphabet is fully defined in Table 6.1 (b). The minimum orthogonal frequency separation ($f_{s\,\min}$) is 150Hz; a non-minimum separation is used. The frequencies themselves could have been placed at a number of points along the allocated bandwidth. The choice of modulation frequency was carried out by first placing the highest frequency on the boundary of the allocated bandwidth ('shifted scheme') then placing the synchronisation symbol at 150Hz. The rest of the bandwidth was shared equally between the data symbols F_1, F_2, F_3, F_4, whilst making all the symbols orthogonal. Any higher values of orthogonal separation than that used would require a larger bandwidth allocation.

Modulating Frequency alphabet	**150bps system**	**Number of samples in a two-cycles:** 54000Hz. Sampling frequency.
F_1	600Hz	180
F_2	900Hz	120
F_3	1200Hz	90
F_4	1500Hz	72
F_s	150Hz	

Table 6.1 (a) Modulating Frequency Alphabet (150bps coincident point system)

Modulating frequency alphabet	**150bps system:** Sampling frequency of 4500Hz or 4800Hz.
F_1	600Hz
F_2	900Hz
F_3	1200Hz
F_4	1500Hz
F_s	150Hz

Table 6.1 (b) Modulating Frequency Alphabet (150bps systems)

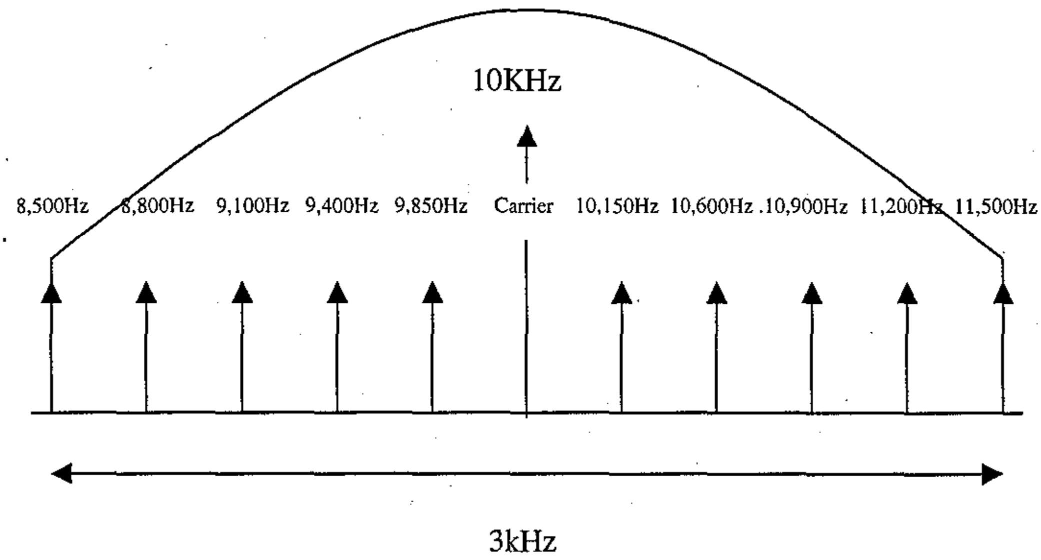


Fig.6.2 5F4SFSK Bandwidth utilisation (150bps)

6.1.1. Practical Design and Implementation of a 5F4SFSK System

16. The transmitter and receiver sub-systems work in synergy to transfer the data through the sub-sea environment. The 5F4SFSK transmitter and receiver implementation descriptions follow. The transmitter is used to generate the modulating frequency alphabet; generate the carrier waveform, utilise the 'correct' pair of symbols dependent on the data, modulate the signals onto the carrier waveform; and transmit the output signal into the medium. The receiver is used to extract the acoustic signal that has been modified from interaction with a long-range shallow, sub-sea channel, distinguish between the possible pairs of symbol transmissions, and evaluate the incoming bit sequence from the detected symbol sequence.

6.1.1.1. The 5F4SFSK Transmitter Implementation

17. The transmission sub-system can be divided into four main processes that can be looked at as separate sub-systems (Fig.6.3). Three of the four processes can be described as generation blocks:

A. Modulating frequencies generator and 'combiner'

B. Carrier frequency generator

C. Data generator

The other transmission system block provides:-

D. Carrier modulation

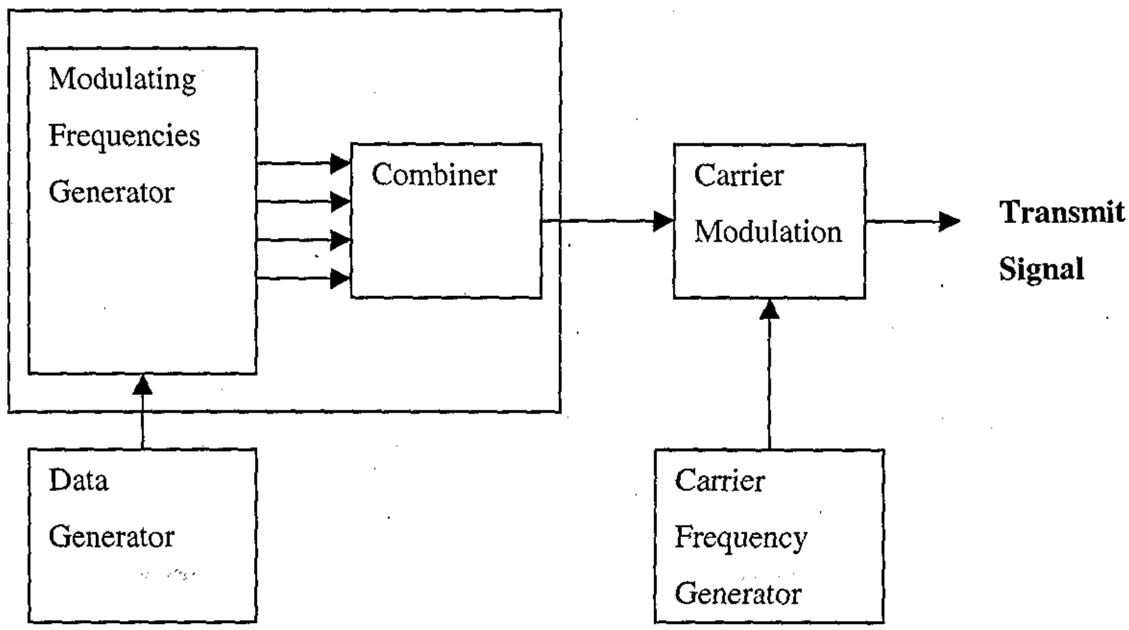


Fig.6.3 5F4SFSK Transmission system

6.1.1.1.1 The Modulating frequency's Generator and Combiner

18. The modulating frequencies generator and the combiner are used to generate all the frequencies in the 5F4SFSK alphabet (Table 6.1). This includes the four data symbol frequencies (F_1, F_2, F_3, F_4) and the synchronisation frequency (F_s) (Fig.6.5). The orthogonal frequencies in the 5F4SFSK modulating frequency alphabet allow optimum exploitation of the matched filter responses at the receiver (Appendix A).

19. The conceptual design involves the generation of the modulating frequency alphabet from a source oscillator. The modulation frequencies are produced, by dividing the source frequency. The two data symbols for each bit value and the synchronisation symbol are combined within the combiner sub-system, then the appropriate symbols are summed before the data is used to pass the correct signal to the modulator (Fig.6.4).

20. The DSP-specific implementation of the modulating frequencies generator is simplified with analogue wave generation (AWG) development software. The symbol frequencies are produced by directly assigning the appropriate values to each sample point. The combiner works as a bank of switches, allowing particular frequencies (symbols) to pass-through, depending on the value of the data sequence. The symbols assigned to each bit-value are summed prior to the application of the 'switching' process (to the resultant signals produced by the summation) the appropriate signal is duly passed onto the modulation sub-process.

6.1.1.1.2 Carrier Modulation, Carrier Frequency and Data Generators

21. Once the symbol sequence has been produced then the signal is modulated onto a carrier. This involves two operations, carrier frequency generation and actual modulation of the signal onto the carrier. The implementation is straightforward, with the output of an oscillator (producing the carrier frequency) being multiplied directly with the modulating signal (sequence of 'summed' symbol frequencies) to produce the output. The multiplication shifts whichever one of the modulating alphabet frequencies (symbols) is present up to the 3kHz band centred on a 10kHz carrier (Fig.6.2).

22. Initial DSP-based implementation of the transmission processes carried out on the AT&T DSP DSP32C platform that is also used for the receiver implementation. This was carried out to allow the receiver processes to be tested with signals generated locally. The final transmitter implementation was carried out on a ADSP2115. AWG software is used to directly generate the carrier frequency (sample value by sample value) instead of the use of an oscillator.

23. The data that is used by the combiner to select the correct output is generated prior to testing. Real-time production of the data is carried out using a *PN* sequence generator on the ADSP2115 platform. The signal is processed to produce 'pulsed' sequence of bit 1 transmissions (value one) and bit 0 transmissions (value zero). This 'pulsed' signal is multiplied by all the bit 1 symbol's AWG generator outputs and inverted before multiplication with all the bit 0 symbol's AWG generators outputs. The two outputs are added together and modulated onto the carrier.

6.1.1.2 The 5F4SFSK Receiver Implementation

24. The receiver sub-system's (Fig.6.6) complex structure is required to allow correct reception to take place in the acoustically inhospitable environment provided by a long-range shallow-water channel. There are six distinct processes that are carried out by the non-coherent 5F4SFSK receiver to enable a decision for each bit to be made:

A. Demodulation from the carrier and pre-processing of the signal
B. Matched filtering of the demodulated signal
C. Envelope detection
D. Bit evaluation
E. Energy decision
F. Synchronisation (PLL)

6.1.1.2.1 Demodulation from the carrier and matched filtering

25. The non-coherent receiver does not require the carrier frequency (that is used for demodulation) to have an exact phase as in coherent reception, thus simple multiplication of the input signal with a locally generated carrier is adequate. The real-time implementation of the demodulation block is achieved by direct multiplication of the input signal with the locally generated carrier. This performs the inverse operation of the carrier modulation block at the transmitter i.e. shifting the signal back to its original bandwidth as occupied before modulation (as carried out for the other modulation schemes). The demodulation process is carried out in the receiver transducer module before the signal is passed to the AT&T DSP32C platform (through the real-time serial interface), where matched filtering takes place (Fig.6.6).

26.. After demodulation, the instantaneous decision at the receiver becomes an evaluation of which of the modulating signals from the alphabet is present in any bit period. A set of orthogonal matched filters is used to produce a measure of the frequency content of the incoming signal. The filter outputs are then passed onto a set of envelope detectors. The matched filters correspond directly to the modulating frequency alphabet tones (symbols), with the centre frequency of each of the band-pass filters corresponding to one of the frequencies. The synchronisation frequency is also assigned a matched filter (BPF), which provides the input to the synchronisation sub-system.

27. The scheme uses the allocated 3kHz bandwidth the 1.5KHz-sideband bandwidth is shown in Fig.6.5. The four data symbol matched filters are allocated 300HZ bandwidth each, which is equivalent to the symbol's orthogonal spacing. The synchronisation symbol could have be placed anywhere within the remaining bandwidth (0Hz-450Hz); the synchronisation matched filter is also allocated 300Hz and centred on 150Hz frequency in this case.

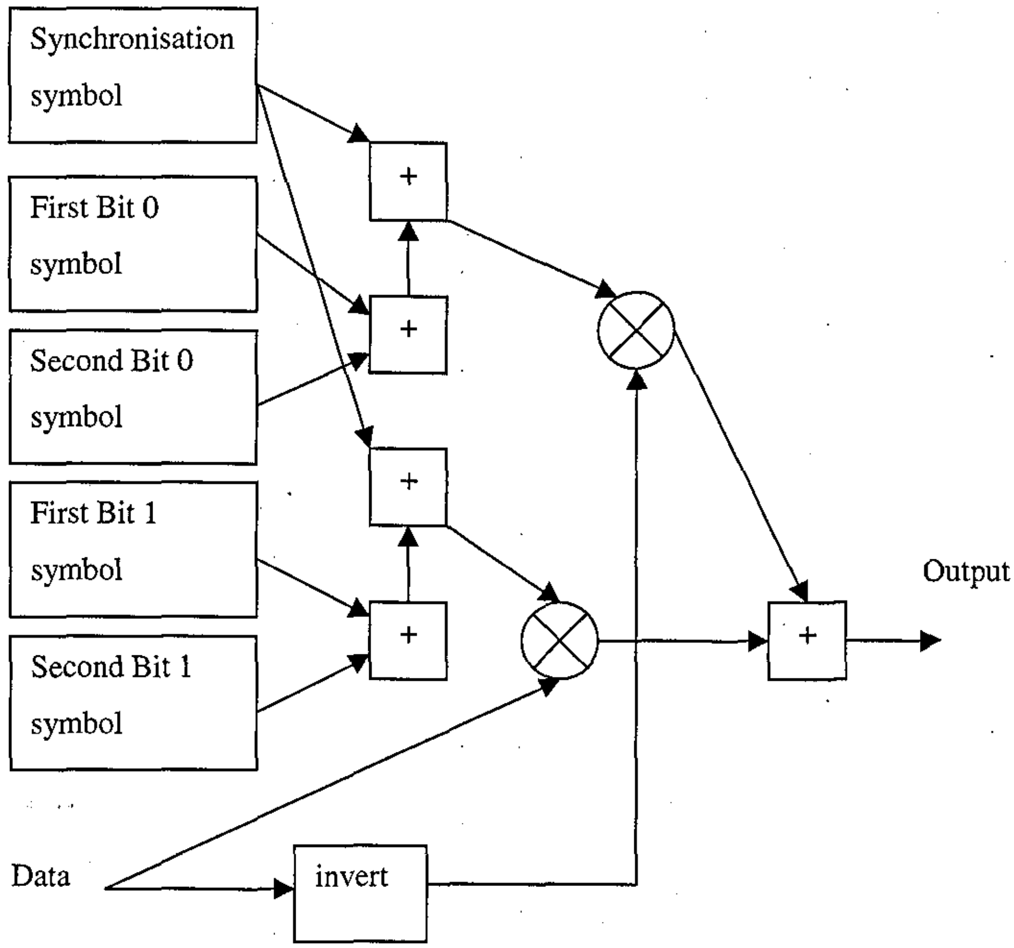


Fig.6.4 **Combiner process**

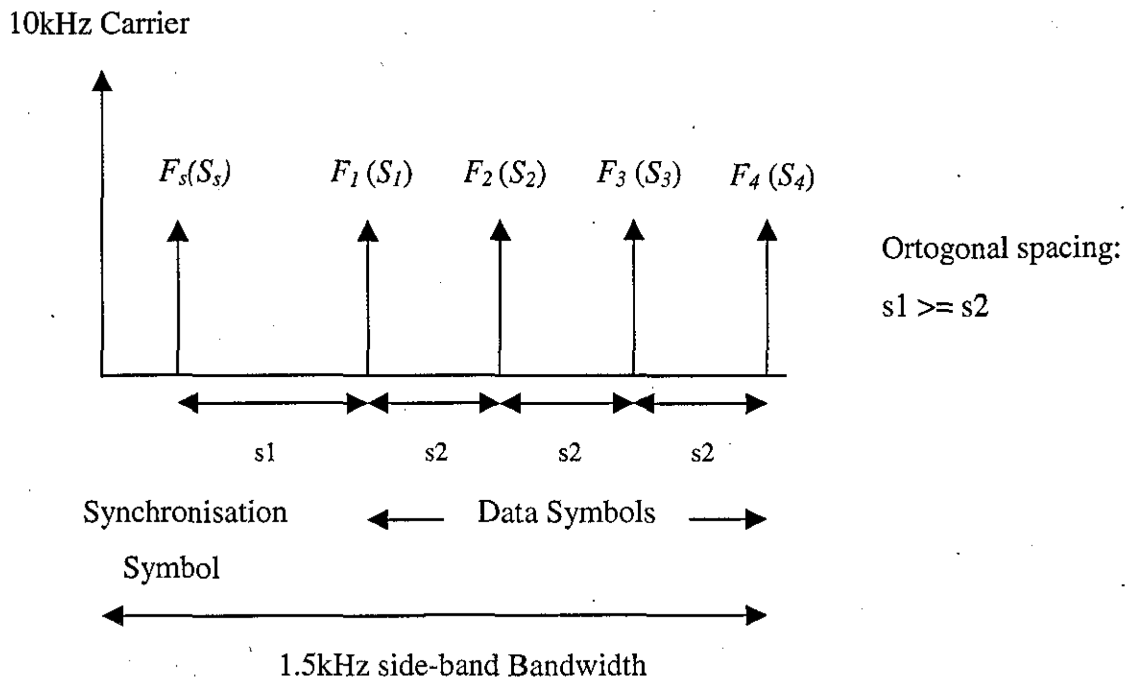


Fig.6.5 **Four frequency tones side-band assignment**

6.1.1.2.2 DSP data processing requirements (5F4SFSK)

28. The receiver implementation is complex and requires a reduction in the number of calculation carried out to reduce the processing burden. Reduced data processing is achieved in a coincident point 5F4SFSK scheme by reduction of the sampling rate within the envelope detector sub-systems. The decision sub-systems are effectively reduced to one sample per 'chip' period. The synchronisation sub-system requires a 'fully sampled' signal (full number of samples for the system sampling rate) for the PLL to work; thus the synchronisation signal's sampling frequency is not reduced at any point. The synchronisation signal is not obtained via the envelope detectors as it is in a scheme, but directly from the demodulated input. The 5F4SFSK receiver's bit evaluation sub-system only requires the sampling frequency of the signals associated with the data to be reconstructed just before the introduction of the synchronisation signal to the bit evaluation sub-system .

6.1.1.2.3 Envelope detection

29. The output of each matched filter is passed to an envelope detector, which performs an integration. A technique using two-cycles of each symbol frequency is used for the coincident point implementation (Fig.4.7 (a)). In the case of the real-time systems three different integration techniques were experimented with:

1. Integration across the whole bit period
2. Integration across half the symbol period
3. Integration over a number of cycles of each frequency tone.

The envelope detectors are used in all cases to provide a comparable magnitude measurement from each of the matched filter outputs.

30. The coincident point scheme's integration requires the same number of 'coincident' sampled points to be used for each frequency's two-cycle integration. With each of the points being in a coincident position in the cycles for each of the different tones, the two cycles of each tone are thus split into the same number of sections, with every section of a single tone having the same duration (Fig.4.7 (a)). The modulating frequency alphabet (Table 6.1) is implemented with six coincident point envelope detectors. The real-time implementation does not include the complications of coincident point processing because there is a need to limit the sampling frequency used to conserve the processing power. There are a limited number of coincident points that can be used for envelope detection purposes when the sampling rate is lower.

31. The real-time implementations were developed to work without the need for coincident points. Full-bit period, half-bit period and cycle-envelope detection are used (Fig.6.7). The real-time envelope detector implementation performs input rectification on the matched filter output by squaring and then square rooting the samples. The signal is then summed with a feedback signal which is equal to one minus the integral of the samples from the previous period, initialising the integration (i.e. getting rid of any previous values). The signal is then fed into a magnitude integrator whose output provides the feedback signal. The feedback works by delaying the largest output (peak value) of the integrator in the decision period (whole-bit period, half-bit period or two-cycle period) so it can be subtracted at the start of the following bit period. The period used is dependent on which of the three real-time envelope detector techniques mentioned above is being used. A pulse signal derived from the synchronisation process is multiplied by the integrator output (which is one only for the largest sample) before the delayed resultant is subtracted from the input to the integrator. The integrator output is passed as the output of the envelope detector.

6.1.1.2.4 Bit evaluation and energy decision

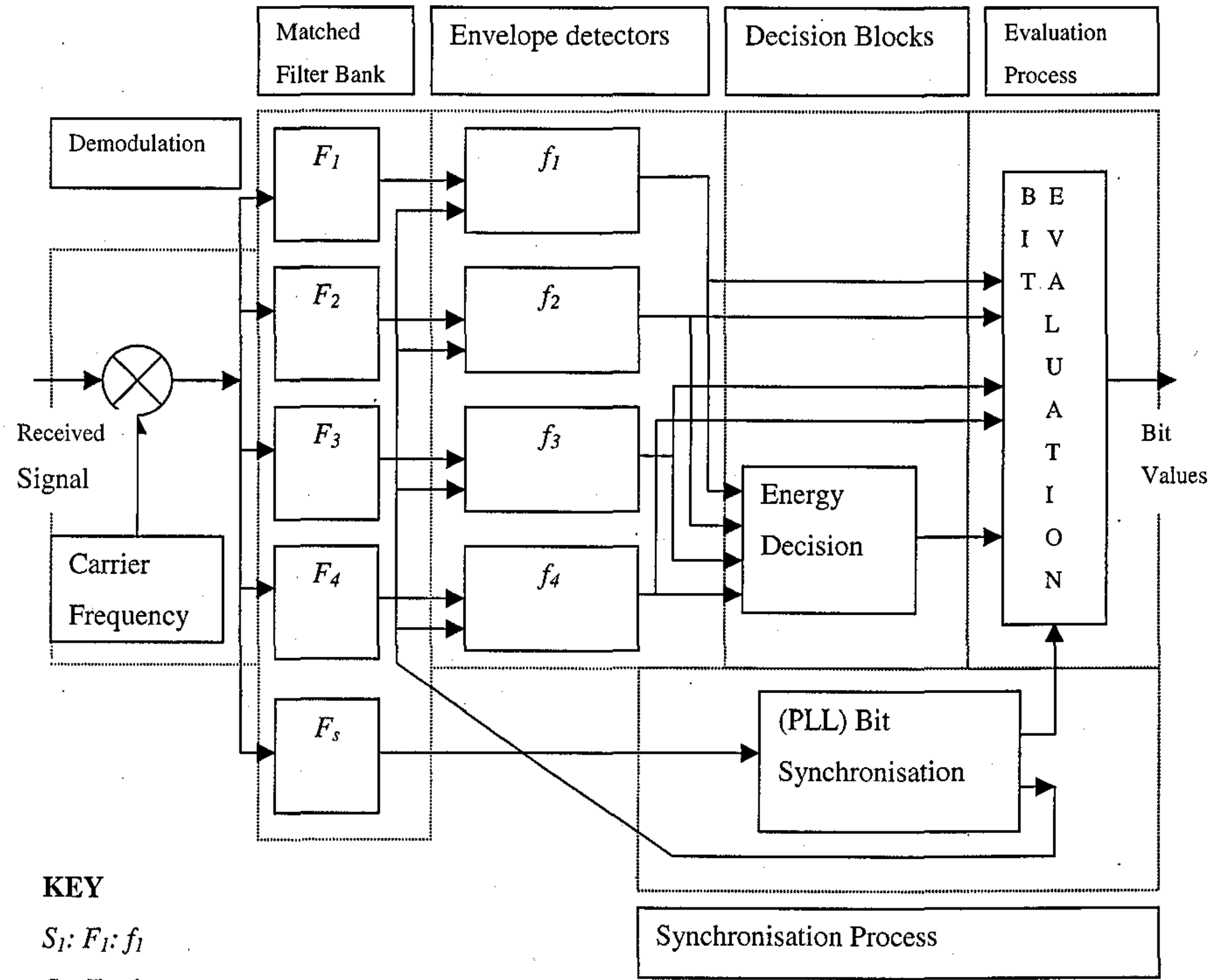
32. The 5F4SFSK bit evaluation process is made up of two decision processes (two symbol decisions) and the energy decision, which work in synergy to evaluate which bit was transmitted. The evaluation is made once every bit period with a synchronisation signal being used to establish the correct instant for the received bit value to be determined. The inputs to the decision sub-systems are the outputs of the envelope detectors. The envelope detector outputs provide a measure of the energy content of each of the modulating frequencies in any bit period. The envelope detectors are paired off (f_1, f_2 and f_3, f_4), one in each pair connected with a symbol frequency representing a different bit-value (Fig.6.6). The outputs of the envelope detectors in each pair are compared to establish the symbols 'most likely' to be part of the received signal; the two resultants are then compared to establish which bit-value was sent. Three decisions are possible:

1. Zero bit received
2. One bit received
3. A contention between the two decisions.

In the case of a contention (where each decision pair points to a different bit-value) an energy decision is implemented.

33. There are times when the implementation of the decision processes can leave the bit evaluation unresolved. It is possible to achieve a situation where both decisions are 'equally likely' after the symbol decisions have been evaluated. The system does not choose either the one or zero bit arbitrarily, though this is acceptable under maximum likelihood detection [Appendix A]. An overall energy decision is implemented instead. The energy decision is a way of distinguishing between the two outputs by using the extra information that is sent when sending more than one symbol-per-bit. A measure of the energy content appropriate to either a zero or a one bit is made by adding the outputs of the envelope detectors associated with a particular value; $f_1 + f_3$ associated with bit 0 and $f_2 + f_4$ associated with bit 1. The resultants are compared to establish the 'winner' of any contention that may occur in a particular bit evaluation.

34. The real-time DSP 'specific' implementation of the bit evaluation process firstly involves the subtraction of envelope detector outputs, f_1 from f_3, and f_2 from f_4 decision. Any negative results are replaced with -1 and any positive results with +1. The values are then summed producing, either +2, -2 or 0. In the unlikely case where one or more of the symbol decisions was inconclusive and a zero was the output at that stage, the summed resultant could be +1, -1 or 0. Therefore positive resultants are output as one bits, negative resultants as zero bits and a zero resultant means the energy decision is required. The resultant is limited, squared and square rooted (making the result always one or zero), and then used to decide if the energy decision is required (zero value means the energy decision is required). The square rooted signal is also used to set the input to zero if the energy decision is to be used for multiplication with the output stemming from the symbol decisions. When the energy decision is not required the square rooted signal has a value of one and multiplication with the output stemming from the bit decisions means the signal is passed on.



KEY

S_1: F_1: f_1

S_2: F_2: f_2

S_3: F_3: f_3

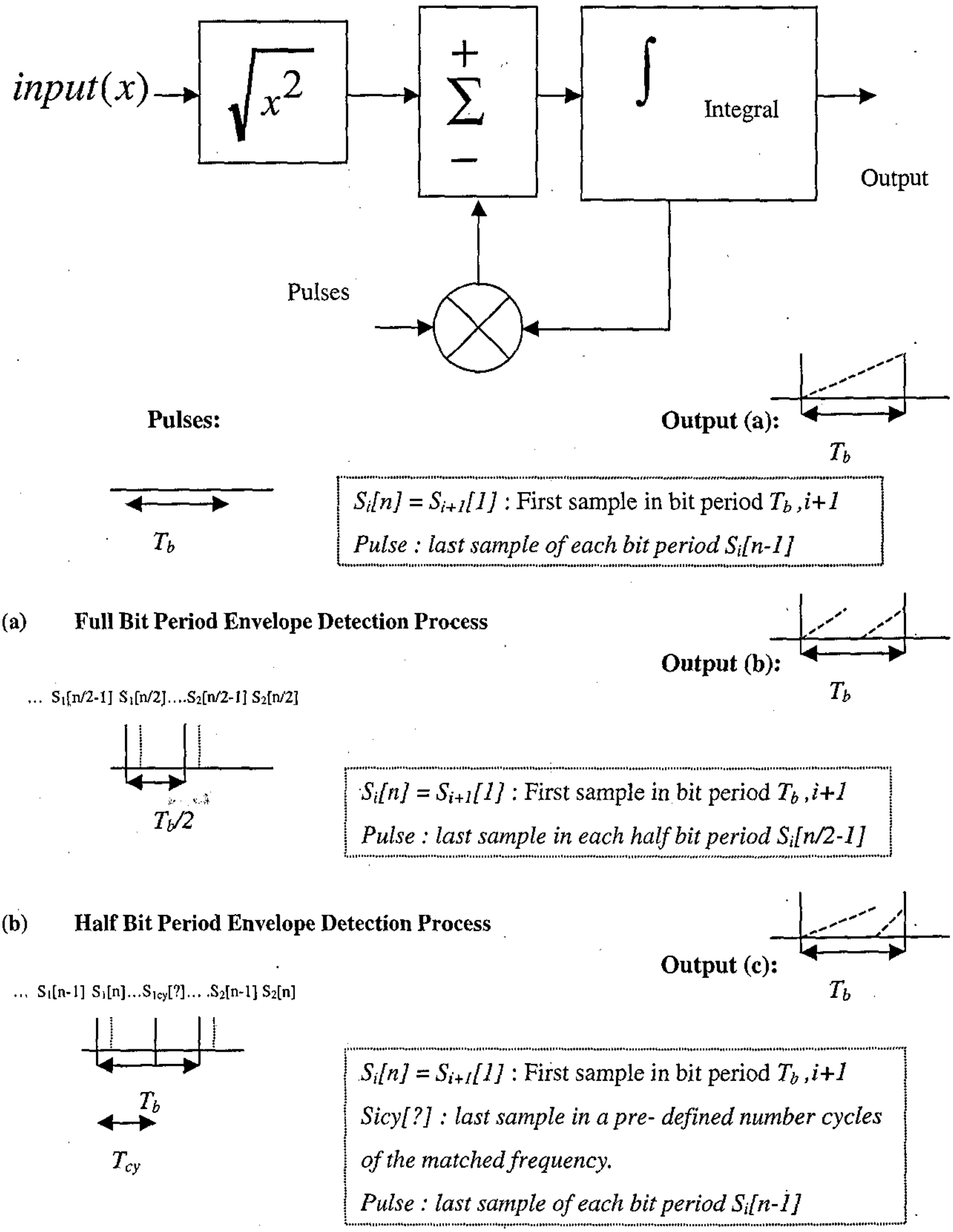
S_4: F_4: f_4

S_s: F_s: -

Symbol: Associated Matched Filter: Associated Envelope Detector

The 150bps real-time input-output scheme structure is shown.

Fig.6.6 5F4SFSK Receiver System



(a) **Full Bit Period Envelope Detection Process**

(b) **Half Bit Period Envelope Detection Process**

(c) **Cycle Envelope Detection Process**

Fig.6.7 Envelope Detection Process

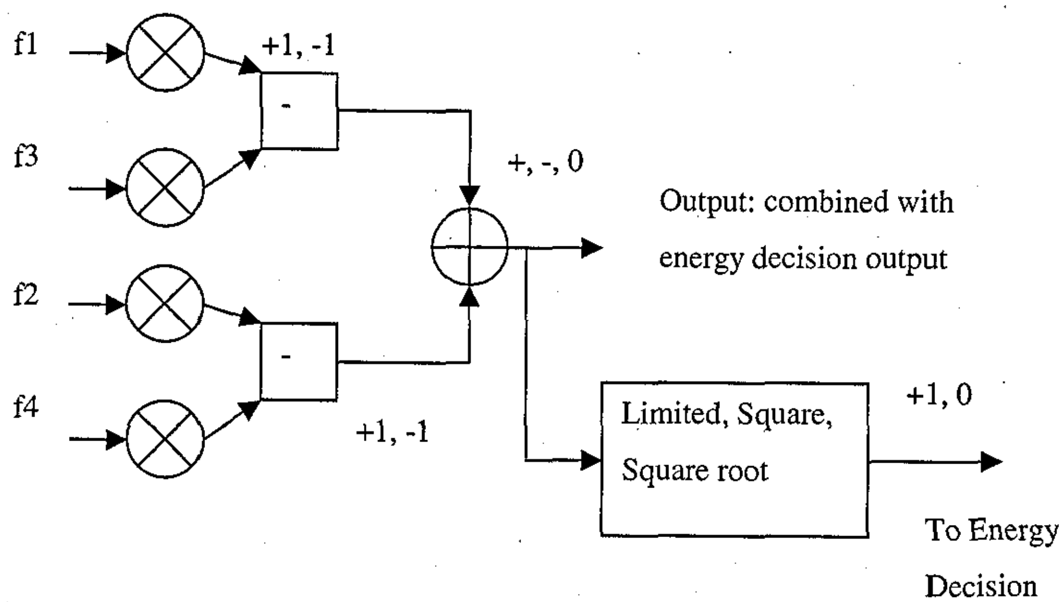


Fig.6.8 **Bit Evaluation Process**

35. An inverted version of the square rooted signal mentioned above (achieved by subtracting a value of one and squaring the result) is multiplied directly with the incoming energy decision. It passes the signal when an energy decision is required. The incoming energy decision uses outputs of the envelope detectors that have been summed together. The $f_1 + f_3$ resultant is subtracted from the $f_2 + f_4$ resultant, negative values are then replaced with -1 and positive values with +1 before multiplication with the inverted version of the square rooted signal takes place. The two outputs of the symbol decision and the energy decision are summed together (one of the two is always zero). The resultant signal's last sample (in a bit period) is then held for a full bit period. This is done by multiplying the signal with a pulse signal that is only one for the last sample in the bit period, then subtracting the value of the previous bit period from the sample. The result is then passed to an integrator to produce an output that holds the last sample in each bit period for the full bit period (so only one bit value can be output in a bit period). The pulse is produced using a bit synchronising signal extracted from the input signal using PLL techniques.

6.1.1.2.5 Bit synchronisation SFSK (5F4SFSK)

36. The output has to be obtained once every bit period, so there is a need to extract the bit rate (frequency) from the incoming data. The 5F4SFSK-modulation technique explicitly transfers the synchronisation information with the data. The bit rate (frequency) extraction is carried out utilising a second order phase locked loop (PLL); which locks to the frequency of the transmitted synchronisation symbol. The PLL input signal is provided by the output of a matched filter matched to the synchronisation symbol. The output is made into a square-wave format to be passed to the first input of a PLL's detector. The synchronisation information extraction has been implemented utilising a NE567 Integrated Circuit. The synchronisation methodology follows the first synchronisation methodology (Section 4.3.2.1).

37. The Phase Locked Loop compares the input signal with a feedback signal using a detector, with the feedback originating from the oscillator output (Appendix B). The detector output is passed through a loop filter (LPF), the output of which controls the oscillator at the incoming bit rate. The oscillator output is feed back to the detector. Pre-filtering of the PLL's input signal with a band pass filter (BPF) is performed to allow the PLL to adjust to the Signal-to-Noise Ratio (SNR), negating noise effects by restricting the range of input frequencies.

Concluding Remarks

This chapter presents the 5F4SFSK-modulation technique; a description of the transmission and receiver implementations that have been adopted is given. The following chapter presents a third technique developed for this research, which represents a hybrid scheme; a number of the techniques used in the 3C4FSK and 5F4SFSK techniques were developed from the hybrid scheme.

Introduction to the Dual Space-Frequency Multiple-Frequency Shift-Keying Modulation Technique

The Dual Space-Frequency Multiple-Frequency Shift Keying (DSFMFSK) modulation technique chapter follows, introducing a modulation technique that uses transmission of parallel frequency tones to represent a single bit. The chapter presents an in-depth description of the DSFMFSK-modulation technique. The technique utilises time and frequency diversity to reduce the effects of the channel characteristics on bit identification. Whilst providing a way of transferring synchronisation information by explicitly transmitting a dedicated symbol frequency. The technique does not over complicate the receiver implementation but is relatively more complex than the other schemes presented in this research. The implementation lends itself to ML matched filter detection and DSP realisation. This technique was not chosen as the 'most appropriate' for further investigation as the use of additional more closely spaced frequency tones lead to more restrictive matched filter implementation and DSP processing requirement which are not justified. It was also noted that the use of explicit frequency transmission for transmission of synchronisation information can be problematic; with the frequency fading characteristics that are sometimes evident in this environment.

7. Dual Space-Frequency Multiple-Frequency-Shift-Keying (DSFMFSK) Modulation Technique

1. The Dual Space-Frequency Multiple-Frequency-Shift-Keying (DSFMFSK) scheme represents each bit with a unique sequence of frequency tones (character-to-frequency coding) and a separate parallel unique symbol transmission throughout the bit period. Six frequency tones are used to transfer data and provide synchronisation information without the need for explicit transfer of a synchronisation symbol. The six frequencies can be looked on as two separate sub-sets of frequencies; one set of four frequencies and one set of two frequencies. The set with four frequencies is used to provide a 2C4MFSK-transmission sequence (variation on a 3C4MFSK scheme) and the other set is used to provide BFSK signalling. The two schemes are combined in a DSFMFSK scheme, thus incorporating SFSK-like signalling, i.e. parallel transmissions.

2. Two of the MFSK (2C4MFSK) frequency tones are assigned to each of the bit values (bit 0 and bit 1). A sequence of frequency tones (character-to-frequency coding) is used to transmit data and provide the synchronisation information. One of the other set of symbols (BFSK symbols) is assigned to a bit 1 and the other to a bit 0. The DSFMFSK bit period is split into two chip periods (which is similar to the application of three chip periods in the 3C4FSK scheme). For transmission of a bit 0, frequencies F_1 and F_3 occupy the first chip period (C_1) and frequencies F_1 and F_5 occupy the second chip period (C_2). For transmission of a bit 1, frequencies F_2 and frequency F_4 occupy the first chip period and frequencies F_2 and F_6 occupy the second chip period (Fig.7.1). The modulation coding aims to achieve a more robust communications.

3. The error protection that is required in an inhospitable environment such as a long-range shallow sub-sea channel; is provided by including parallel redundancy as well as serial redundancy. Error protection is provided by both character-to-frequency coding and parallel redundancy in this scheme. In theory, only one frequency is required per bit, as with a BFSK scheme. By assigning extra ('redundant') frequencies in sequence (MFSK) and having a BFSK transmission (with different symbol frequencies) in parallel, the scheme's resilience to noise is enhanced. The symbol periods for the serially transmitted symbols are equal to a chip period. The parallel BFSK scheme does not use serial transmissions of different symbols (all chip periods have the same symbol for a particular bit value) thus the symbol period can be looked at as being equal in length to three chip periods or a bit period. The relationship between chip periods (T_c), parallel symbol periods (PT_s), serial symbol periods (ST_s) and bit periods (T_b) in a DSFMFSK scheme is described by:

$$T_c = ST_s = PT_s/3 = T_b/3 \qquad \textbf{(7.1)}$$

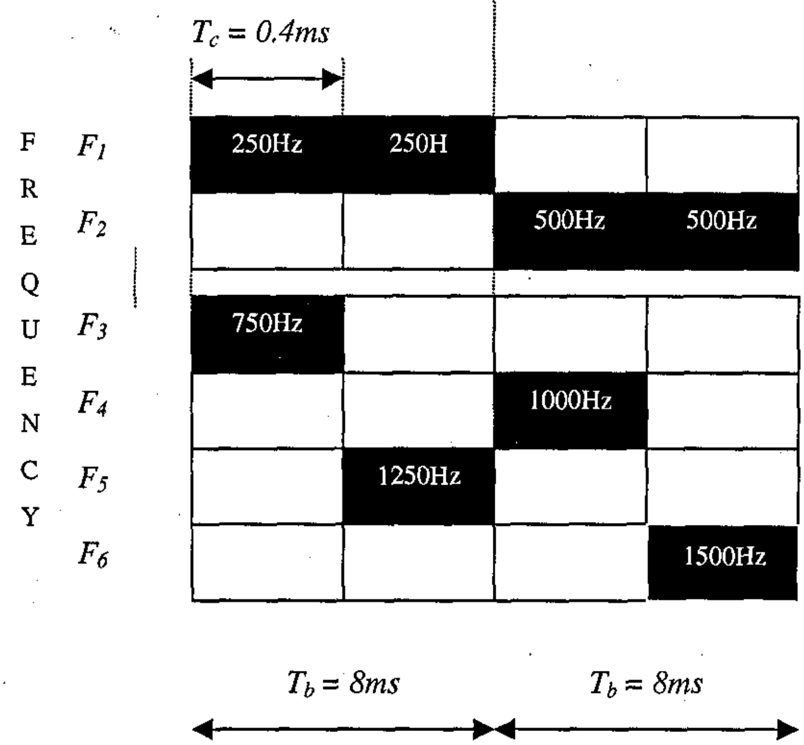


Fig.7.1 DSFMFSK-Frequency Transmission

4. At the beginning of the transmission a 'unique' sequence of bit values is sent (which does not appear in the data transmissions); the receiver looks for this sequence to establish the beginning of data transmissions. The sequence used is determined by the data generator a unique sequence it does not transmit. Any reoccurrence of the sequence redefines the start of the data transmissions. The receiver would 'look out' for the initial sequence to occur for a limited time, which is longer than the expected delay for the most significant multipath signal to occur. In the most extremely delayed cases, where a minimum phase channel is being used, the last arrival is the 'most significant' (in terms of received signal level). Assuming the significant signal arrives within the 'look out' period, the start of transmissions is defined by the last arrival. The data follows the last 'unique' initial sequence detected in the 'look out' period.

5. The synchronisation scheme used for the DSFMFSK system involves the use of intrinsic synchronisation. The bit timing information is present in the transmitted signal no matter what sequence of bits is sent, without explicit transmission of a synchronisation symbol. This is achieved in a similar way to how it is done in a 3C4FSK scheme. A regular transition (change in frequency) in every bit period is used for clock recovery. The symbols in each chip period are assigned either a (+1) or (-1) at the receiver (Table 7.1), which provides the input signal to the synchronisation pre-filter and PLL. The bit synchronisation information is transmitted with both bit 1 and bit 0 transmissions. If the bit synchronisation information was only available from one of the 'bit value' transmissions, then synchronisation (PLL lock) could be lost in long sequences of the 'other' bit.

6. The symbols used for the two bit values in this coded modulation scheme are chosen to establish two unique (non-interfering) sets. The symbols assignment used for the 2C4MFSK sub-scheme are chosen to avoid tone repetitions, reverse tone sequences, cross-bit symbol repetitions and cross-bit symbol assignment. The symbols used for the BFSK implementation exhibit tone repetition across both chips periods, this assignment is used because the BFSK symbol decisions are carried out over the whole bit period (the symbol periods are effectively two chip periods long).

Symbol Frequency	Bit Value	Value Assigned
F_1	0	Na
F_2	1	Na
F_3	0	(+1)
F_4	1	(-1)
F_5	0	(+1)
F_6	1	(-1)

na = not applicable

Table 7.1 Bit Timing Frequency Assignments

7.1. DSFMFSK Implementation

7. The system restrictions drawn up together with BG plc, together with the DSP platform used, restrict the implementation of the DSFMFSK scheme in a number of ways. The restricted parameters include the choice of modulating frequency tones, which are restricted to the allocated bandwidth and their spacing. The choice is limited to one of a finite number of orthogonal frequency sets that are dependent on the symbol rate used (Equation 4.5), and is therefore indirectly defined by the bit rate. The symbol rate stipulation for a DSFMFSK scheme is slightly more complex than for the other schemes. There are effectively two different symbol rates being used; the FSK sub-system symbols are treated as being equal in length to a bit period (at the receiver), whilst the 2C4MFSK sub-system has two symbol periods in a bit period. The orthogonal frequency set has to be applicable to both symbol rates being transmitted.

8. Implementations are restricted by the sampling limitations of the DSP platform used. Abiding by the Nyquist sampling theorem negates the use of the bottom end of the DSP sampling rates, i.e. anything below twice the highest symbol frequency. The system sampling frequency chosen for this DSFMFSK (125bps) scheme is 4500Hz, which is above the Nyquist requirement and within the AT&T DSP32C operating range. The system's evaluation was initially carried out using locally generated signals; the transmission output was passed through the output ports to a DAC, ADC and back through the input port.

9. The 2C4MFSK sub-scheme's modulating frequency alphabet fits the relationship for orthogonal operation specified for non-coherent FSK (Equation 4.5). The 2C4MFSK sub-system's serial symbol period (ST_s) is equal in length (4ms) to a chip period (T_c) and is half the length of a bit period (T_b), whilst the FSK sub-system's parallel symbol period (PT_s) is equal in length (8ms) to a bit period (T_b). The lowest frequency (F_1) is equal to the frequency of the orthogonal separation between symbols in the scheme's alphabet (Table 7.2). The available bandwidth (1.5kHz per-side-band) and the number of symbols required (six), restricts the choice of available frequency sets that can be chosen as symbol frequencies. The choice comes down to one of two frequency separations for the FSK sub-scheme. A set with the minimum frequency separation (which provides up to 12 frequencies to choice from) or the next set-up with $m=2$ (Equation 4.5), which is a sub-set of the minimum orthogonal spaced set with only six frequencies. By setting $m=2$ instead of $m=1$, wider matched filter characteristics can be implemented more effectively as they require fewer filter taps.

10. The minimum frequency of orthogonal separation for the FSK sub-scheme ($f_{s\min}FSK$) is 125Hz, and with m=2, it becomes 250Hz. The 2C4MFSK sub-system symbol's minimum orthogonal separation ($f_{s\min}2C4FSK$) is 250Hz, and there is only one set of six frequencies available within the allocated bandwidth. The second smallest orthogonal spacing (250Hz) for the FSK sub-scheme provides one set of frequencies, that are equivalent to the set with minimum spacing for the 2C4MFSK sub-scheme. Wider spaced orthogonal frequency sets are too bandwidth-inefficient to be implemented. The highest symbol frequency (F_6) is placed right at the edge of the allocated side-band bandwidth, which is 1500Hz for a 3kHz full side-band transmission scheme (Table 7.2, Fig.7.2).

7.1.1. Practical Design and Implementation of a DSFMFSK System

11. There are two main components of the DSFMFSK system that work in synergy to facilitate transfer of data through the sub-sea environment; these are the transmission and reception systems.

The transmitter system provides the ability to:

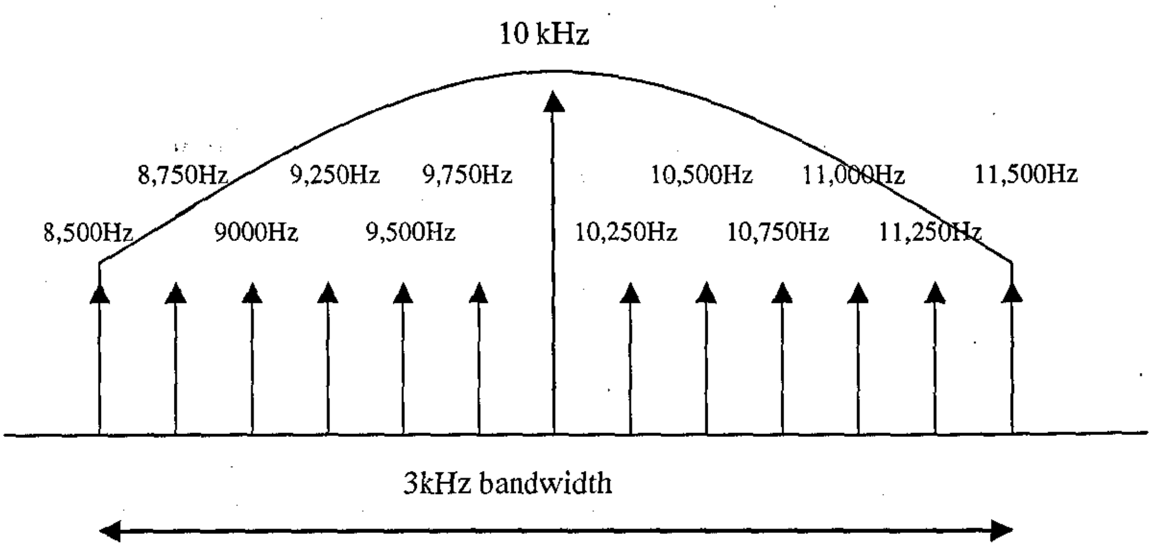
1. Generate the modulating frequency alphabet
2. Generate the carrier
3. Utilise the 'correct' sequence of symbols dependent on the data
4. Combine the correct parallel symbols in the bit period
5. Modulate the modulating signal onto the carrier
6. Transmit the output signal into the medium

The receiver system's function is to:

1. Extract the acoustic signal, which will have been modified from interactions within a long-range shallow, sub-sea channel
2. Distinguish between individual symbol transmissions in both sub-schemes
3. Evaluate the incoming symbol sequence in each bit period as well as evaluating the parallel symbol
4. Combine the two sub-scheme results to distinguish the transmitted bit sequence

Modulating Frequency alphabet	**Associated Sub scheme**	**DSFMFSK alphabet**
F_1	FSK	250Hz
F_2	FSK	500Hz
F_3		750Hz
F_4		1000Hz
F_5		1250Hz
F_6		1500Hz

Table 7.2 Modulating Frequency Alphabet



Symbol separation: 250Hz

See table 7.2, for symbol alphabet.

Fig.7.2 Bandwidth Utilisation of the DSFMFSK alphabet (125bps)

7.1.1.1. The DSFMFSK Transmitter (Tx) Implementation

12. The transmission sub-system can be divided into seven main processes that can be looked on as separate sub-systems (Fig.7.3). Four of the seven processes can are generation blocks:

A. Modulating frequency generator
B. Carrier frequency generator
C. Cyclic code generator (chip sequence generator)
D. Data generator

The other three transmission system blocks provide:
E. Chip selection
F. Bit period assignment and symbol combining
G. Carrier modulation

7.1.1.1.1. The modulating frequency generator

13. The modulating frequency generator is used to generate the frequencies required for the DSFMFSK alphabet (Table 7.2). Full-band transmission limits the six frequency tones to a 1.5kHz side-band bandwidth (Fig.7.4). The conceptual design involves the generation of the modulating frequency alphabet from a source oscillator.

14. The ADSP2115 'specific' implementation of the modulating frequency (symbol) generator, is produced using analogue wave generation (AWG) software as is done with the other modulation schemes in this research. The AWG assigns values to each sample directly. The modulating frequencies are generated separately. Individual sample points are calculated, dependent on the frequency being generated, and the appropriate output produced.

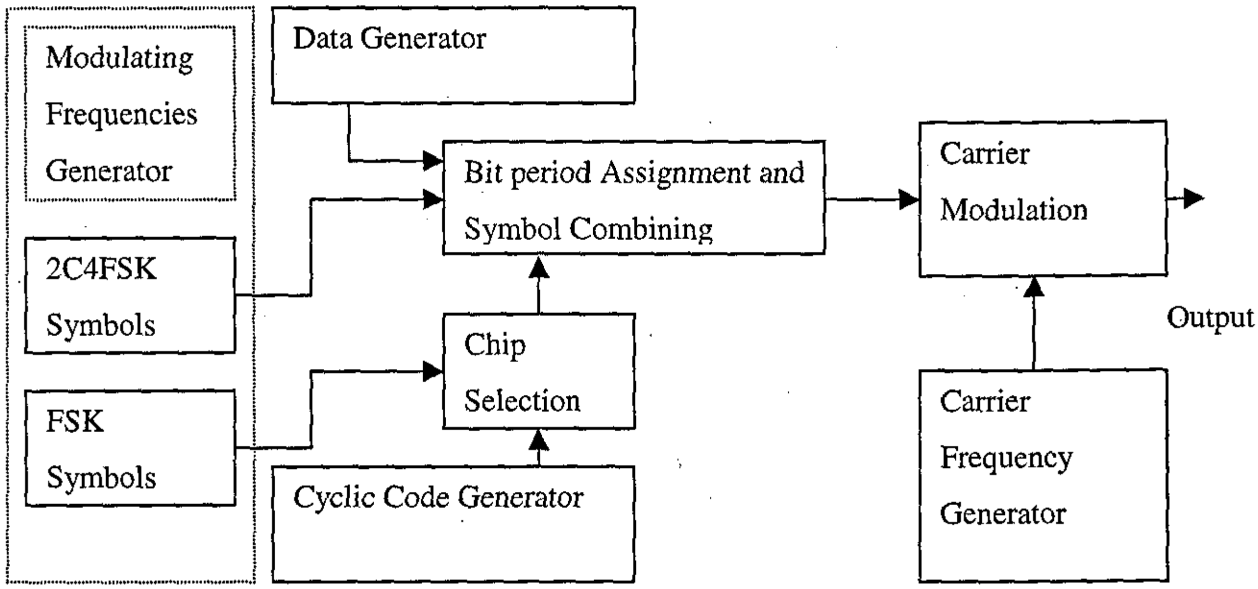


Fig.7.3 The DSFMFSK Transmission System

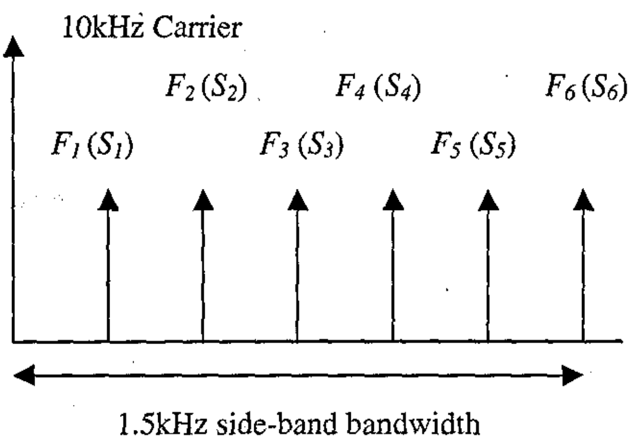


Fig.7.4 Six frequency tones side-band assignment

7.1.1.1.2. The cyclic code generator, chip selection, data generation, bit period allocation and symbol combining

15. The cyclic code sequence generator (chip sequence generator) and the chip selection block are described together, as they work in synergy to produce the output, with the former being used to manipulate the latter. The chip selection block uses a signal (cyclic code) to differentiate between the two different chip periods. The cyclic code used is effectively a pulse signal, i.e. 1 for chip two and 0 for chip one. The first chip to be sent for any bit is chip one followed by chip two.

16. The 'chip selection' is effectively a bank of 'switches' that allow the appropriate frequency tone to go through, depending on the value of the cyclic code. The chip selection has to be able to change the frequency tone assigned to each chip period dependent on the data's bit value. This is done by the inclusion of more 'switches' that only pass the appropriate frequency tones required for a specific bit transmission.

17. Initial DSP-based implementation of the transmission processes was carried out on the AT&T DSP32C that is also used for the receiver implementations. This was carried out to allow the receiver processes to be tested with signals generated locally. The implementation of the cyclic code was straightforward, simply involving the generation of a pulsed signal. The 'chip selection' was carried out by multiplication of the cyclic code 'pulse' with each of the 2C4MFSK sub-scheme's frequencies in the modulating frequency alphabet (Fig.7.5). The pulse signal that is to be multiplied with the chip one frequency tones is first 'inverted' by subtraction of +1 and squaring the output. The correct frequency for chip one or chip two is generated for both a bit 1 and bit 0. The chip-two output is then added to the chip one output separately for bit one and bit zero. One of the two outputs (chip one or chip two) will always be zero. The outputs are added together to produce the correct chip frequency, i.e. for both bit 1 and bit 0 frequencies. The bit 1 and bit 0 chip signal is then multiplied by the data bit or the inverted (*modulo*-2) data bit respectively; the resultants are then summed to output the appropriate chip frequency.

18. The sequence of chip frequencies is passed to the bit period assignment and symbol combining sub-system together with the data. The two symbols used in the FSK sub-scheme are passed as inputs; the bit value (data signal) is used to choose between the symbols. Carried out by multiplying the bit 1 with the data and the bit 0 symbol with an inverted (*modulo*-2) version of the data, and then summing the two resultants. This FSK symbol which has been passed is then summed with the chip sequence for that particular bit period and the result is passed on for modulation.

19. The final DSP 'specific' transmitter implementations are carried out on the Analogue Devices ADSP2115 or the AT&T DSP32C. The implementation of the cyclic code and chip selection is not required on the Analogue Devices platform, as the placement of symbols into the appropriate chip periods can be carried out directly using AWG. A sequence of symbols appropriate to a particular bit transmission is generated (sample values are assigned directly) by the AWG software; the signal is then passed onto the bit period assignment and symbol combining sub-system.

20. The data is used to select the correct signals. The data value to be multiplied with the zero bit frequencies output is first 'inverted' as when using the cyclic code (Fig.7.5). The two multiplier outputs are then added together (one output is always zero), thus producing the correct chip sequence for either bit one or bit zero. The data generation in the final DSP 'specific' implementations is done independently of the transmitter using the DSP32C; the data is generated and saved on a DAT tape recorder. In real-time the DAT output signal is used as an input to the DSP-platform; as is done for the other modulation schemes studied in this research.

7.1.1.1.3. The carrier frequency generator and carrier modulation

21. Once the modulating signal has been produced, then its modulation onto a carrier takes place. This involves two operations, carrier frequency generation and the actual modulation. The implementation is straightforward, with the output of an oscillator (producing the carrier frequency) being multiplied directly with the modulating signal to produce the output. The modulation is the opposite operation of the receiver's carrier de-modulation, and is carried out within the transmitter transducer module. The modulating signal is passed from the Analogue Device's ADSP2115 platform (through the real-time serial interface) to the ADC and on to the transducer module. The modulation shifts whichever combination of the symbol frequencies that are present up to the 3kHz band dissected by the 10kHz carrier (Fig.7.2). This is carried out within the transducer module provided by BG plc.

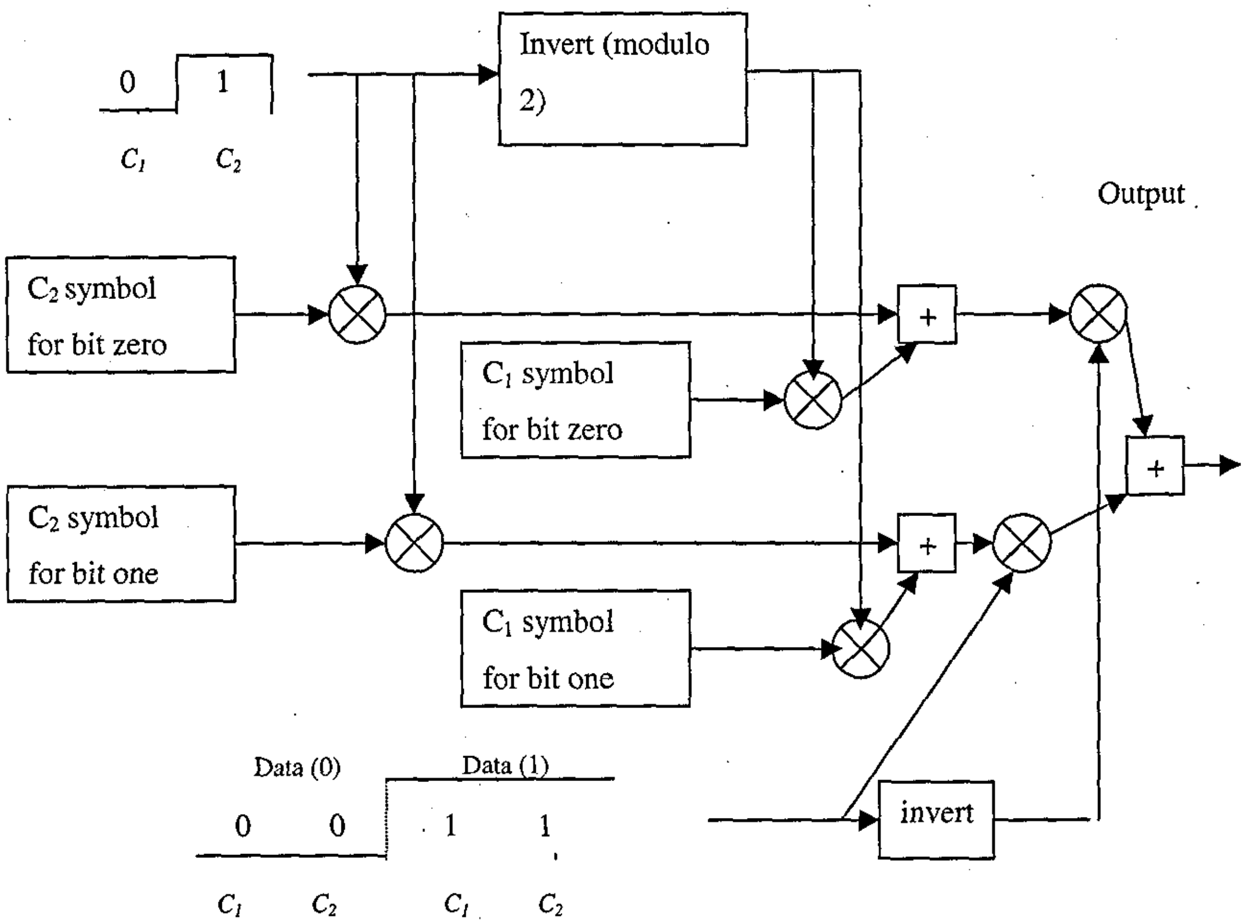


Fig.7.5 **Chip selection process**

7.1.1.2. Receiver (Rx) Implementation

22. The receiver sub-system (Fig.7.6) involves more complex operations than are required in the transmission sub-system. The relatively complex structure of the receiver is required to enable evaluation of the transmitted sequence. Seven distinct processes are needed in the non-coherent DSFMFSK receiver structure to enable a decision for each bit to be made.

A. Demodulation from the carrier waveform
B. Matched filtering of the demodulated signal
C. Envelope detection
D. 2C4MFSK 'Zero' bit value decision and 'One' bit value decision
E. FSK bit decision
F. Bit evaluation (Choice)
G. Bit synchronisation (PLL)

7.1.1.2.1 Demodulation from the carrier and matched filtering

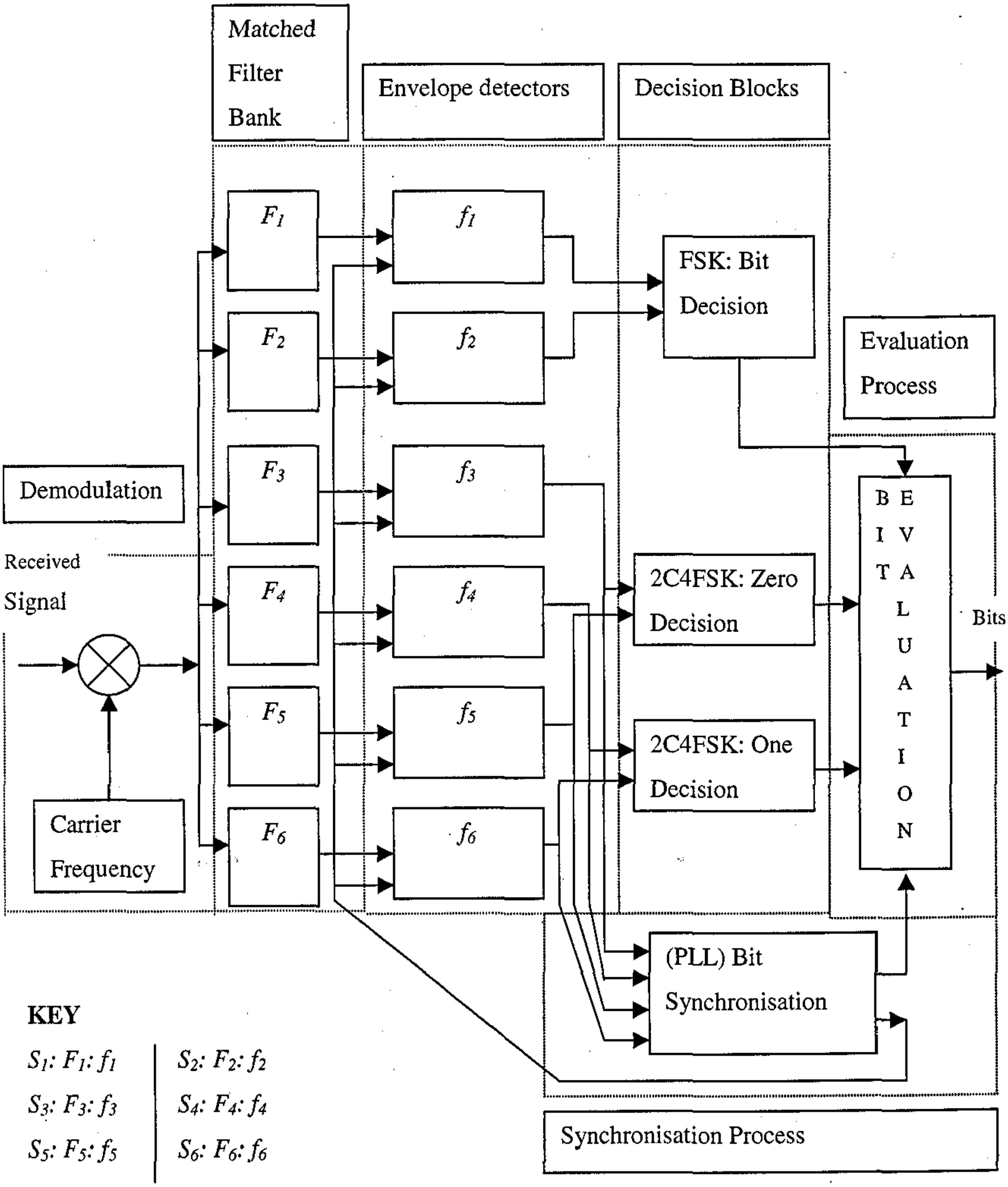
23. A non-coherent receiver format was developed, due to the acoustic signal phase instability found in shallow, long-range sub-sea channels. The multiplication of the input signal with a locally generated carrier whose phase is not synchronised to the transmitter's carrier is appropriate for non-coherent detection. Thus direct multiplication of the received signal with a locally generated 10kHz carrier is carried out for this DSFMFSK scheme. The demodulation process is carried out in the receiver transducer module before the resultant signal is passed to the AT&T DSP32C (through the real-time serial interface), where matched filtering takes place (Fig.7.6).

24. After demodulation, the immediate requirement at the receiver is to evaluate which of the modulating signals (symbols) is present. The decision is more accurately viewed as the non-instantaneous evaluation of the symbol that is present within a symbol period. A set of orthogonal matched filters is used to produce a measure of the frequency content of the incoming signal. The filter outputs are then passed to a set of envelope detectors that utilise the instantaneous matched filter outputs to evaluate the symbol that is present in a particular symbol period.

25. The matched filters correspond directly to the modulating frequency alphabet frequency tones produced at the transmitter; the centre frequency of each of the band-pass filters corresponds to one of the symbol frequencies. The real-time 125bps DSFMFSK scheme uses the 3kHz bandwidth (Fig.5.2) comprehensively; the individual matched filter implementations are each allotted a 250Hz bandwidth.

7.1.1.2.2 Envelope detection

26. The envelope detectors are used to provide a comparable magnitude measurement of the content of each symbol frequency in the received signal. The implementation of magnitude integration over the full symbol period was adopted. The sampling points from one symbol period are summed by magnitude integration to produce an output for comparison with the output of the other envelope detectors; this enables the receiver to distinguish which symbol frequency is present (Fig.7.7).



KEY

S_1: F_1: f_1	S_2: F_2: f_2
S_3: F_3: f_3	S_4: F_4: f_4
S_5: F_5: f_5	S_6: F_6: f_6

Symbol: Associated Matched Filter F?: Associated Envelope Detector f?

The 125bps real-time input-output implementation

Fig.7.6 The DSFMFSK Receiver System

27. The envelope detectors used in the real-time input-output system involve an integration over the full symbol period ('full period' envelope detection). The DSP 'specific' envelope detector implementation performs input rectification on the matched filter output by squaring and then square rooting the samples. The signal is then summed with a feedback signal that is equal to zero minus the value of the integral of the samples from the previous symbol period, initialising the integration (getting rid of any previous values). Next the signal is fed into a magnitude integrator whose output provides the feedback signal. The feedback works by delaying the largest output (peak value) of the integrator in the symbol period, so that it can be subtracted at the start of the next symbol period. A pulse signal derived from the synchronisation process is multiplied with the integrator output (which is 1 only for the last sample in the symbol period) before the delayed resultant is subtracted from the input. The resultant signal is delayed by one sample position, so the subtraction is made from the first sample of the following symbol period. The integrator output is passed as the output of the envelope detector (Fig.7.8).

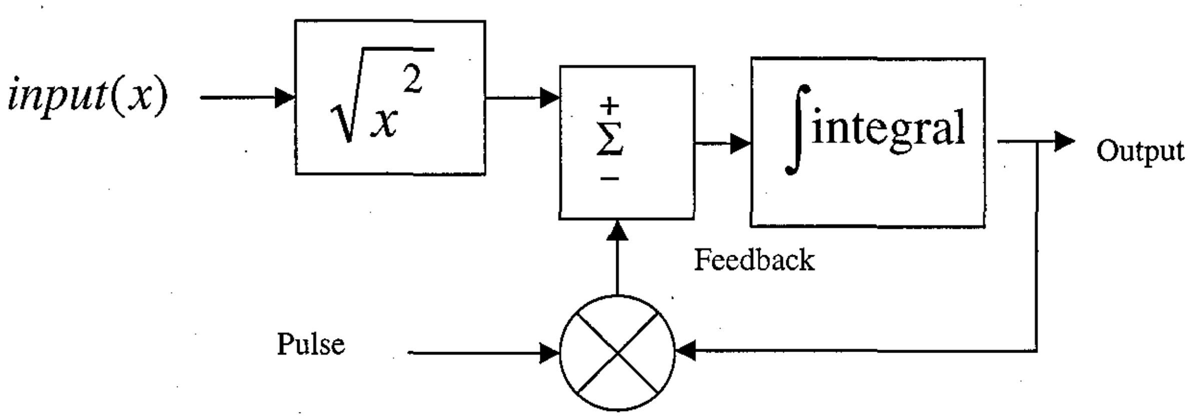


Fig.7.7 **Envelope Detection Process**

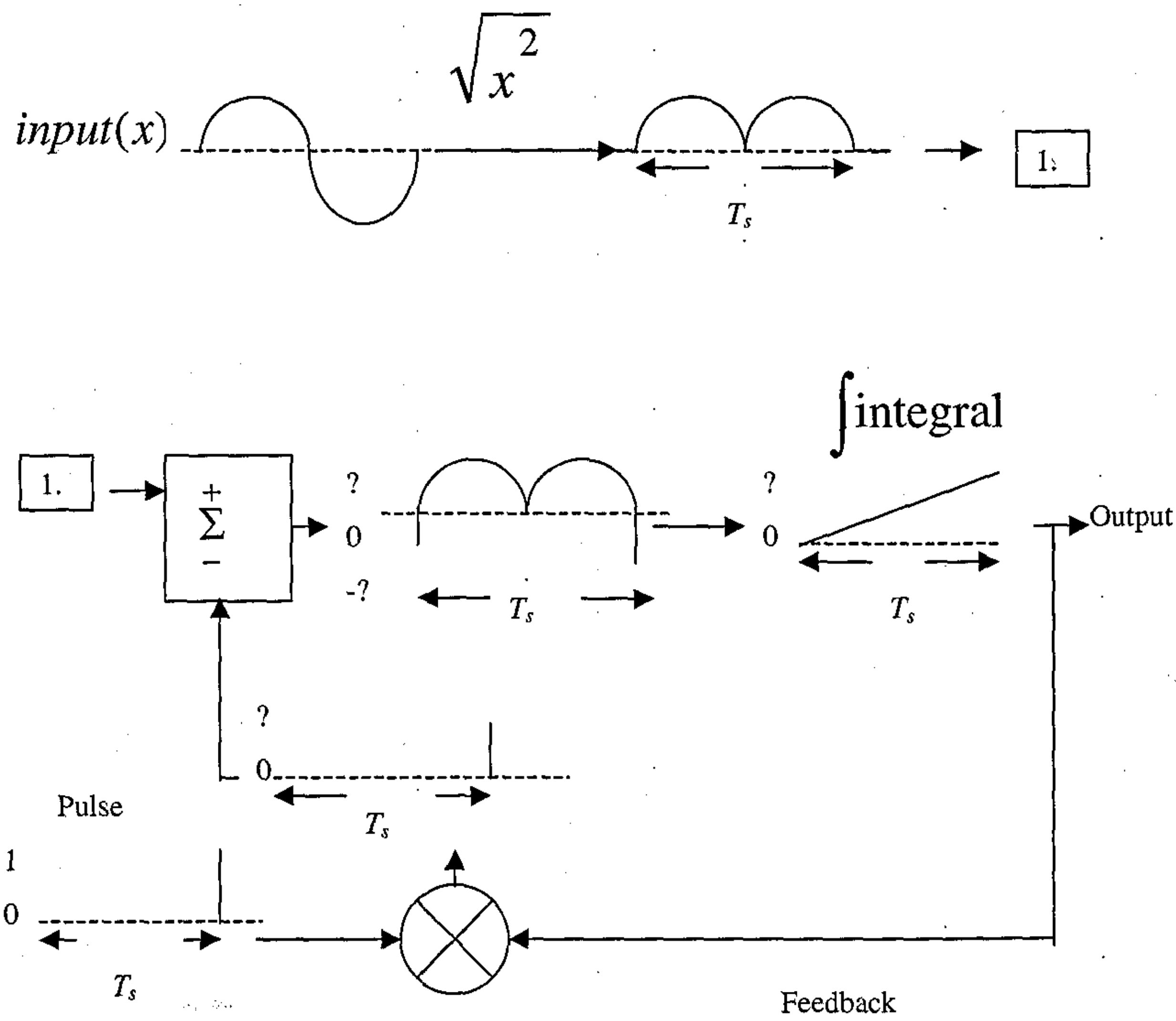


Fig.7.8 **Envelope Detection Signals**

7.1.1.2.3 Zero and one bit decision

28. The bit decision process is made up of two 2C4MFSK processes (zero-bit decision and one-bit decision) and a FSK decision process. The 2C4MFSK processes treat the evaluation of a zero bit or a one bit totally separately, whereas the FSK decision combines the bit decision into one process (Fig.7.9). The 2C4MFSK decision processes are described together in this section because they perform identical operations, but on different input signals, then the FSK decision process is described.

29. Each of the 2C4MFSK decisions is a measure of 'how close' the sequence of chips (symbol frequencies) is to that expected for either a zero-bit or a one-bit transmission. The evaluation is carried out for both bit decisions in parallel. The zero and one decisions are thus made independently by focusing only on the chip frequencies appropriate to either of the possible bit values (only the 2C4MFSK sub-schemes symbols). The zero decision compares only the zero-bit chip frequencies (outputs of the envelope detectors f_3 and f_5 shown in Fig.7.6) and assigns -1 or +1 for a particular chip, dependent on their comparative magnitudes. The one-decision does exactly the same operation but uses the outputs of the envelope detectors f_4 and f_6 (Fig.7.6).

30. Both the bit decision operations perform a subtraction of the second chip (C_2) from the first chip (C_1) (Fig.7.9). The calculation generates a value +2 for a correct chip sequence (chip two = -1, chip one = +1). A value between +2 and -2 can be produced, with +2 representing the 'most likely' correct and –2 the 'least likely' correct. The chips are moved on one place in the decision calculation when the next chip arrives. The new incoming chip (chip three) becomes chip two and the old chip two becomes new chip one and finally the old chip one is discarded (Fig.7.10). This is carried on as new chips arrive.

31. There is more than one output produced per bit period for both the zero and one decision, but the bit evaluation block is only used when two chips from the same bit period are involved in the calculation. The synchronisation module is used to decide when the output is correct when all chips in the bit evaluation stem from the same bit period.

32. As a sampled signal is used there are additional samples in each chip period that are not required for the bit decisions; the samples are all nullified except for one sample in every chip period. The synchronisation signal is used to produce a pulse once every chip period. This is multiplied with the input signal. The pulses are generated as a value of +1 for chip one, and -1 for the chip two (Fig.7.11(a)). A magnitude integral is made over the bit period, after the integral of the previous bit period's last value is subtracted. Multiplication of the integral output with a pulse which is only +1 for last pulse generated in a bit period (chip two's pulse) takes place (Fig.7.11 (b)), enabling only the last integral value to be output (a value between +2 and minus -2). The decision outputs are compared by the subtraction of the bit one decision output from the bit-zero decision output; a positive result indicates that the zero decision is dominant and a negative result indicates that a one decision is dominant. The signal is limited to +1, -1, and 0 values, indicating a zero bit, a one bit and a tie respectively at the decision sample instant. The signal is then passed to the bit evaluation process; in the case of a tie the FSK decision is used to decide between a zero and a one bit.

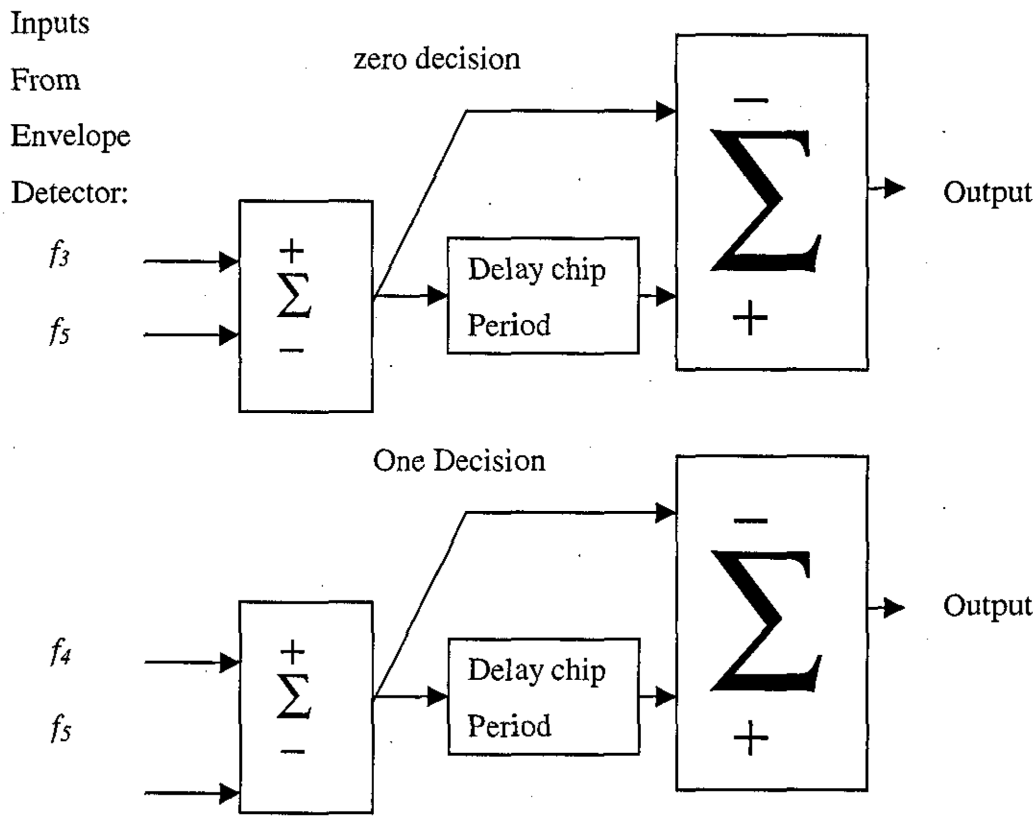
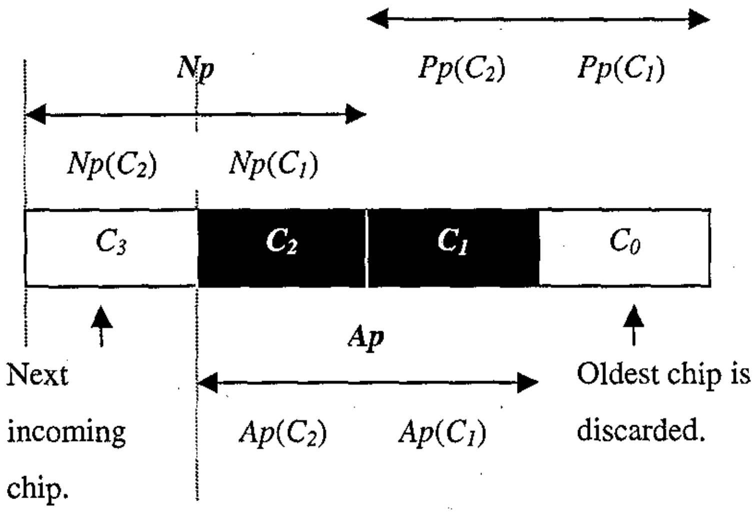


Fig.7.9 Zero and one decisions (bit decision



Ap Active 'perceived' bit period

Pp Previous 'perceived' bit period

Np Next 'perceived' bit period

Fig.7.10 'Chip' allocation throughout bit decision's

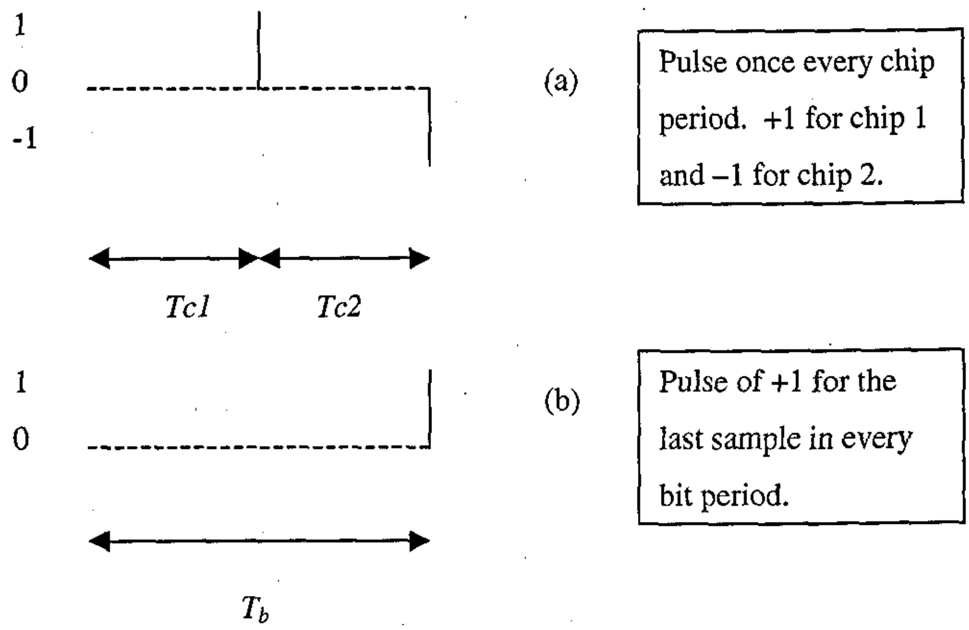


Fig.7.11 Bit decision process signals

33. There are times when the implementation of the parallel bit decision processes can leave the bit evaluation unresolved. It is possible to achieve a situation where both decisions are equally likely; an FSK decision is then implemented (this is replaced in the other schemes in this research by a more robust energy decision). The FSK scheme represents a parallel transmission of the data that is 'independent' of the 2C4MFSK scheme and therefore can be used when it is inconclusive. The FSK decision compares the symbol content of the two symbols assigned to it by looking at the outputs of the appropriate envelope detectors (f_1, f_2). The symbols are compared to establish the 'winner'. The FSK decision process firstly involves the subtraction of the envelope detector output, f_1 from f_2. Any negative results are replaced with -1 and any positive results with +1. In the unlikely case where the 2C4MFSK decisions are inconclusive and a 0 is produced by the FSK decision, a random bit choice can be made, as is appropriate with maximum likelihood detection (Appendix A).

7.1.1.2.4 Synchronisation DSFMFSK

34. The output data is "correctly" evaluated within the bit evaluation block when the two-chip periods involved in the zero-bit and one-bit decisions (2C4MFSK) all originate from the same bit period transmission and the FSK symbols originate from the same bit transmission. The output therefore is not required at instants when the two chips do not originate from the same chip period. The DSFMFSK modulation technique intrinsically transfers the synchronisation information with the data. The chip decoding carried out by the bit decisions incorporates the ability to distinguish between the 'correct' and 'incorrect' sequence of 'chips' for each bit value. The one change of symbol frequency included within each bit period provides a mechanism for the transfer of the bit rate to the receiver within the sequences of 'chips' without additional transmissions. The output can thus be obtained once every bit period. The bit rate (frequency) extraction is carried out using a second order phase locked loop (PLL), which locks to the change in frequency of the regular sequence of chips (symbols) in each bit period. The PLL input signal is provided by a comparison of the chip one and chip two envelope detectors outputs. The output is filtered and made into a square-wave format to be passed to the first input of a PLL's detector. The synchronisation information extraction has been implemented utilising a NE567 Integrated Circuit, using the second synchronisation methodology (Section 4.3.1:2).

35. To establish the rate of change of the chip frequencies the PLL compares the input signal with a feedback signal using a detector (Appendix B). The feedback originates from the oscillator output. The detector output is passed through a loop filter (LPF), the output of which controls the oscillator at the incoming bit rate. The oscillator output is feedback to the detector. Pre-filtering of the PLL's input signal with a band pass filter (BPF) is performed to allow the PLL to adjust to the Signal-to-Noise Ratio (SNR), negating noise effects by restricting the range of input frequencies that the PLL can encounter.

Concluding Remarks

This chapter follows on from the previous chapters that present the 3C4FSK and 5F4SFSK techniques. The hybrid DSFMFSK scheme incorporates both serial and parallel diversity; implemented separately in each of the previous chapter's modulation techniques. This chapter provides an insight into the transmission and receiver implementations adopted for the DSFMFSK scheme. The following chapter describes the theory and tests carried out for this research.

Theoretical Performance, Performance and Evaluation

8. Theoretical Performance and Performance

1. The performance of a modulation technique within a channel can be defined in terms of the achievable *bit-error-rate* (BER). Theoretical BER values can be calculated and depicted graphically as a function of carrier-to-noise ratio (CNR), as is done for White Gaussian Noise (WGN) and Raleigh fading environments (Fig. 8.2, 8.3). Raleigh fading more closely approximates the sub-sea environment than WGN; there are significant differences between the Raleigh case and a practical environment. Significant phenomena are neglected including inter-symbol-interference (ISI) and frequency specific fading (FSF), therefore the theoretical results can only be viewed as guidelines.

2. White Gaussian Noise and Rayleigh Fading performance are relevant to this work as they approximate some of the effects due to the acoustically prohibitive phenomena found in long-range shallow subsea channels. The calculation of performance in WGN is included here as it approximates the combined effects of different sources of additive noise on the acoustic signal. WGN is generally used though out communication engineering to show the relative performance of modulation techniques. Raleigh fading takes into account the signal fading effect on signal level that is generally attributed to multipath propagation.

8.1. Theoretical Performance Calculations

3. Each of the modulation schemes is made up essentially of a number of 'FSK decisions', which are made either in parallel and/or in serial with each other. The theoretical performance of each modulation technique is calculated by first braking down the techniques into underlying 'FSK decisions', then evaluating the overall performance.

4. The probability of symbol error when using an FSK maximum likelihood detector in WGN, is given by [116]:

$$P_{eWGN} = \frac{1}{2}\exp(-\frac{A^2}{2\sigma_n^{\ 2}}) \qquad \textbf{(Eq. 8.1)}$$

The pulse shape (A) is assumed constant over the symbol interval. The variance of the noise is equal to the noise power per-Hertz multiplied by the bandwidth of the matched filters (B):

$$\sigma_n^{\ 2} = N_o B$$

For a filter bandwidth $B = 2/T_s$ where T_s is the symbol period, the probability of symbol error in *WGN* is defined as shown in Equation 8.2. This assumes that the bandwidth of the filters extends to the first zero of the spectrum of the pulsed sinusoidal signal [116].

$$P_{eWGN} = \frac{1}{2}\exp(-\frac{A^2}{2\sigma_n^{\ 2}}) = \frac{1}{2}\exp(-\frac{A^2}{2N_0 B}) = \frac{1}{2}\exp(-\frac{A^2}{4N_0}) = \frac{1}{2}\exp(-\frac{A^2}{4N_0})$$

$$= \frac{1}{2}\exp(-\frac{E}{2N_0})$$

(Eq. 8.2)

5. The average energy (E) is equal to A squared multiplied by the symbol period and divided by two.

$$E = \frac{TA^2}{2}$$

Setting the bandwidth equal to:

$$B = \frac{2}{T}$$

The carrier to noise ratio (CNR) is defined as:

$$CNR = \frac{E}{No}(dB)$$

Where *No* is the Noise power per Hertz.

6. A bandwidth adjustment factor is introduced to allow for larger bandwidth assignments:

$$\Delta = B/(2/T)$$

The probability of error for a FSK scheme in *WGN* environment s thus given by:

$$P_e = \frac{1}{2}\exp(-\frac{E}{2N_0\Delta}) \tag{8.3}$$

Allow calculation of probability of error for different matched filter bandwidths. This is equal to:

$$P_e = \frac{1}{2}\exp(-\frac{E}{2N_0\Delta}) = \frac{1}{2}\exp(-\frac{\lambda}{2}) \tag{8.4}$$

Where:

$$\lambda = \frac{E}{N_0\Delta} = CNR$$

7. A more realistic mathematical representation of a sub-sea channel requires fading characteristics to be incorporated into the performance calculation. For a channel that exhibits flat fading (fading is assumed constant) the probability of error (*Pe*) given for a *WGN* can be modified to take into account the reduction in signal amplitude at the receiver:

$$P_e = \frac{1}{2}\exp(-\frac{xA^2}{2{\sigma_n}^2}) \qquad \textbf{(8.5)}$$

Where x is a constant proportional to the reduction in signal amplitude of the received signal that is attributed to the fading characteristics of a particular channel.

8. The fading found in a sub-sea channel is primarily due to the rapidly varying multipath characteristics of the signal propagation; therefore fading is not flat. The probability density functions of the amplitude found at a receiver in such a multipath environment can be more closely modelled using either a Rayleigh or Rician distribution. The use of Raleigh assumes that all the paths received exhibit a random attenuation. Whereas the use of Rician assumes one of the paths exhibits 'constant' attenuation (within a symbol period), this is normally assumed the 'direct' path [145].

9. Here we look at the pdf of Raleigh fading given by [146]:

$$p(\lambda) = \frac{1}{\Gamma}\exp(-\frac{\lambda}{\Gamma})$$

(Eq. 8.6)

Where:

$$\lambda = CNR,$$

$$\Gamma = CNR(average) = \frac{\frac{E}{N_0\Delta}}{N} = \frac{E}{N_0\Delta}N$$

N is equal to the number of rays (paths).

The error performance can be found using the integral of the resultant, of the multiplication of the Rayleigh probability density function together with the probability of error in *WGN* [146]:

$$P_e = \int_0^\infty p\ (\lambda) P(e) d\lambda = \int_0^\infty \frac{1}{\Gamma} \exp(-\frac{\lambda}{\Gamma}) \frac{1}{2} \exp(-\frac{\lambda}{2}) d\lambda$$

$$= \frac{1}{2\Gamma} \left(\frac{1}{\frac{1}{\Gamma} + \frac{1}{2}} \right) = \frac{1}{2+\Gamma} \tag{8.7}$$

The probability of error for FSK in Raleigh is inversely proportional to the CNR and experientially proportional in WGN:

In WGN.

$$P_e = \frac{1}{2} \exp(-\frac{E}{2N_0\Delta})$$

In Raleigh.

$$P_e = \frac{1}{2 + \frac{E}{No\Delta}}$$

8.1.1. 3C4FSK Theoretical Performance

10. In the case of the 3C4FSK modulation technique each bit decision is defined by two separate sequences of three FSK symbol decisions, one accessing the 'likelihood' of a bit zero and the other the 'likelihood' of a bit one. The resultants are compared to find the detected bit value, in the case of a tie a energy decision is used. The energy decision is a third sequence of FSK decision's which use the combined inputs of the other decisions. The individual symbol decisions (FSK decisions) made as part of the bit zero and bit one decisions are assumed independent of each other, even though a choice between the same two symbols is made for each. This is acceptable as each symbol decision is made in a different portion of the bit period, and is therefore operating on a different input. The two bit decisions (zero and one) are also independent from each other as they operate on separate symbols; the decisions are mutually non-exclusive.

11. The energy decision makes a decision using the combined inputs to the other decisions. The three FSK energy decisions are also independent from each other as with the bit decisions. The energy decision is only used where the comparison of the bit zero and bit one decisions is inconclusive (a tie situation), the decision is assumed independent of the other decisions. The evaluation of the probability of error can thus be calculated (Appendix C):

$P_{e3C2FSK} = 1$ - (((P_c(D0)) and (P_c(D1))) or ((P_c(D0)) and(1-P_c(D1)) and (P_c(DE))) or ((1-P_c(D0)) and (P_c(D1)) and (P_c(DE)))) Where $P_c = 1 - P_e$

8.1.2. 5F4SFSK Theoretical Performance

12. In the case of the 5F4SFSK technique, each bit decision is defined by three parallel FSK symbol decisions. Decisions one (*D1*) and two (D2) look at a different pairs of symbol frequencies one of each pair representing a one or a zero transmission. The decision resultants are compared to find the detected bit value, in the case of a disagreement a third decision is used (*D3*). The third decision is also a FSK decision, all the symbols associated with each bit are combined before a decision is made. The individual symbol decisions are independent of each other. The 5F4SFSK scheme can thus be calculated (Appendix C):

$P_{e5F4SFSK} = 1$ - (((1- P_e(D1)) and (1- P_e(D2))) or ((1- P_e(D1)) and (P_e(D2)) and (1- P_e(D3))) or ((P_e(D1)) and (1- P_e(D2)) and (1-P_e(D3))))

8.1.3. DFSFSK Theoretical Performance

13. In the case of the DFSFSK modulation technique each bit decision is defined by two separate sequences of two FSK symbol decisions, one accessing the 'likelihood' of a bit zero and the other the 'likelihood' of a bit one. The resultants are compared to find the detected bit value, and in the case of a tie a third parallel decision is used. The third decision is also a FSK decision but is made over the full bit period. The individual symbol decisions (FSK decisions) made as part of the bit zero and bit one decisions can be seen as independent of each other, even though a choice between the same two symbols is made for each. This is acceptable as each symbol decision is made in a different portion of the bit period and therefore operates on a different input. The two bit decisions (zero and one) are also independent from each other as they operate on separate symbols; the decisions are mutually non-exclusive. The third decision is only used where the comparison of the bit zero and bit one decisions is inconclusive (a tie situation), the decision is assumed independent of the other decisions. The evaluation of probability of error can thus be calculated (Appendix C):

$P_{eDFSFSK}$ = *1 - ((($P_c(D1)$) and ($P_c(D2)$)) or (($P_c(D1)$) and (1- $P_c(D2)$) and ($P_c(D3)$)) or ((1-$P_c(D1)$) and ($P_c(D2)$) and ($P_c(D3)$)))*

8.2. Performance

14. The performance of modulation techniques that use matched filter detection can differ depending on the filter bandwidth used. The underlying FSK decisions implemented in each of the schemes developed have different theoretical performance this is due to the different filter implementations having different matched filter bandwidth. As the bandwidth of a filter is reduced in size, the theoretical performance improves (Fig. 8.1); this does not account for signal distortion that may be found within the smaller bandwidth implementations. A bandwidth of $T_s/2$ only extends out to the first nulls of the matched frequency and can thus yield distortion[144]. The performance of the larger bandwidth FSK schemes is the most realistic (Fig. 8.1). All the calculations have been made using the theoretical performance, which neglects possible signal distortion. It should also be noted that the data rates of the schemes in Fig 8.1 vary, therefore there is not a direct comparison in there performance.

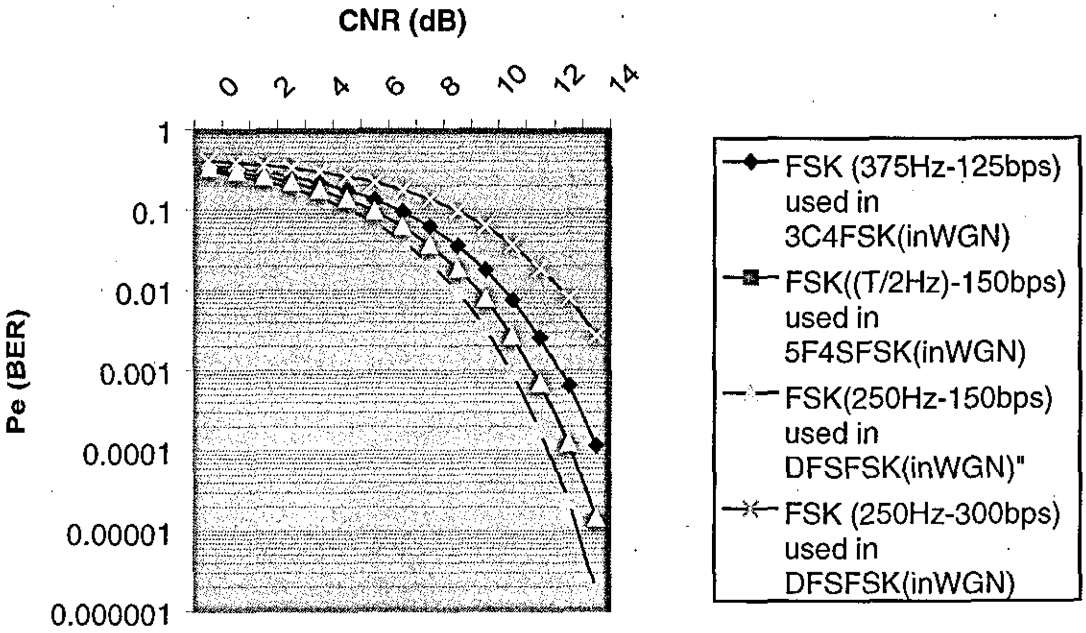


Fig. 8.1 **FSK theoretical performance in WGN**

15. The performance of the modulation techniques within WGN is shown in Fig. 8.2; it is clear that theoretically the 3C4FSK scheme provides the best performance. The 5F4SFSK scheme has the second best performance, with approximately 1dB difference at a BER of 10^{-5} that increases moderately with CNR. The DFSFSK scheme has the faster decrease in performance with an additional 3dB offset (4dB difference compered to a 3C4FSK scheme) at this BER. All three schemes show a rapid exponential tendency to low BER's as the CNR is increased; they also all show a significant increase in performance over the equivalent matched filter bandwidth FSK schemes. The 3C4FSK scheme has a 6db improvement over the 375Hz, matched filter bandwidth FSK scheme at a BER of 10^{-5}, which increases with improving CNR. The 5F4SFSK scheme has a 3dB improvement over the *T/2* matched filter bandwidth FSK scheme at this BER and the DFSFSK scheme also has a small improvement over its associated FSK schemes.

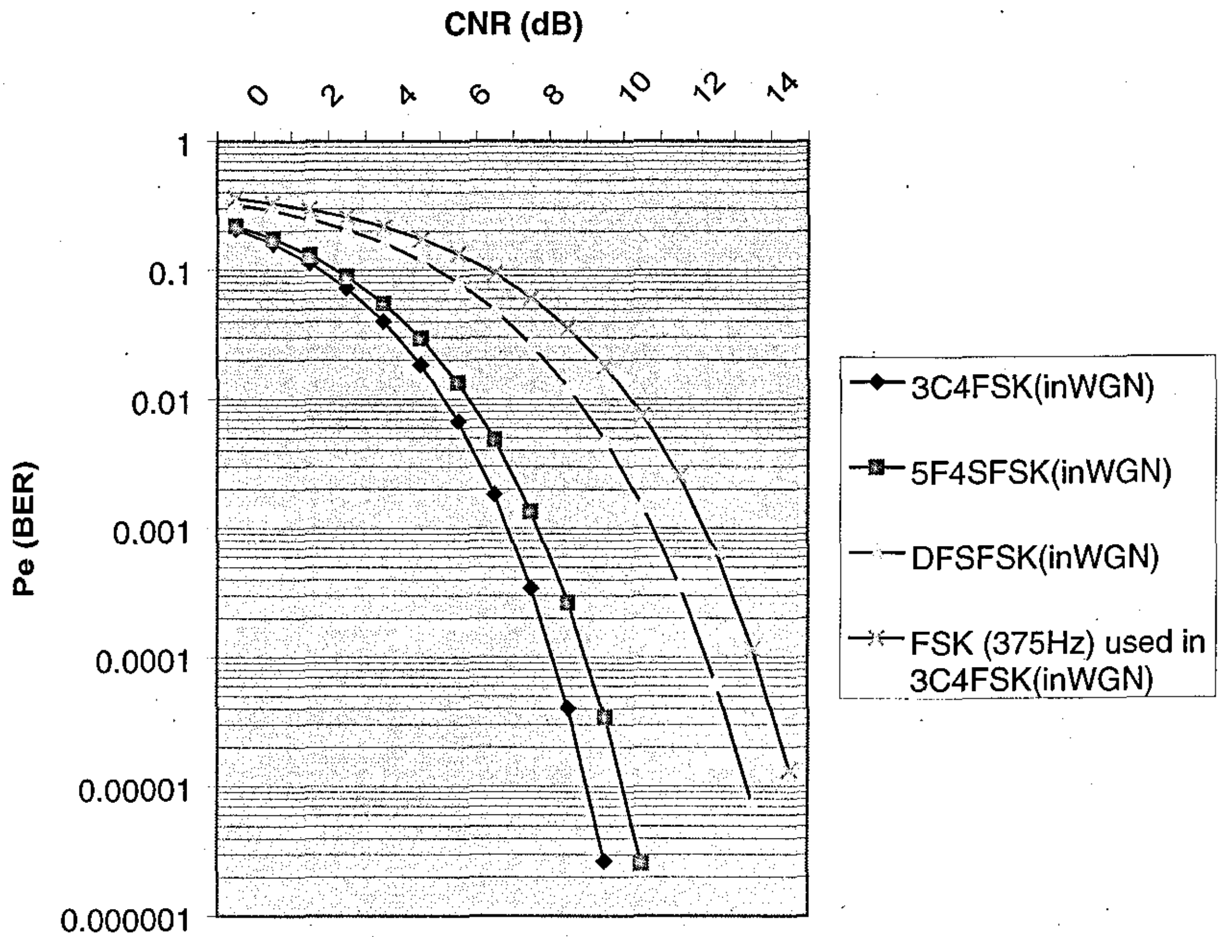


Fig. 8.2 **Modulation schemes theoretical performance in WGN**

16. The performance of the modulation schemes in Raleigh Fading is shown in Fig. 8.3. In this case the BERs do not drop off exponentially with increasing CNR but tend to an inverse relationship. In a multipath environment noise levels are influenced by signal level, therefore an increase in signal level does not improve the BER as rapidly as in a WGN environment. BERs need larger carrier-to-noise ratios then required in the WGN case. Theoretically, the 3C4FSK scheme provides the best performance out of the three schemes with the 5F4SFSK and DFSFSK schemes showing a closer performance to each other than with WGN. The 3C4FSK performance at a BER of 10^{-5} is approximately 10dB better than 5FS4FSK and 13dB better than the DFSFSK scheme. All three schemes show an increase in performance over the equivalent matched filter bandwidth FSK schemes. The 3C4FSK scheme has a 35dB improvement over the 375Hz, matched filter bandwidth FSK scheme at a BER of 10^{-5}.

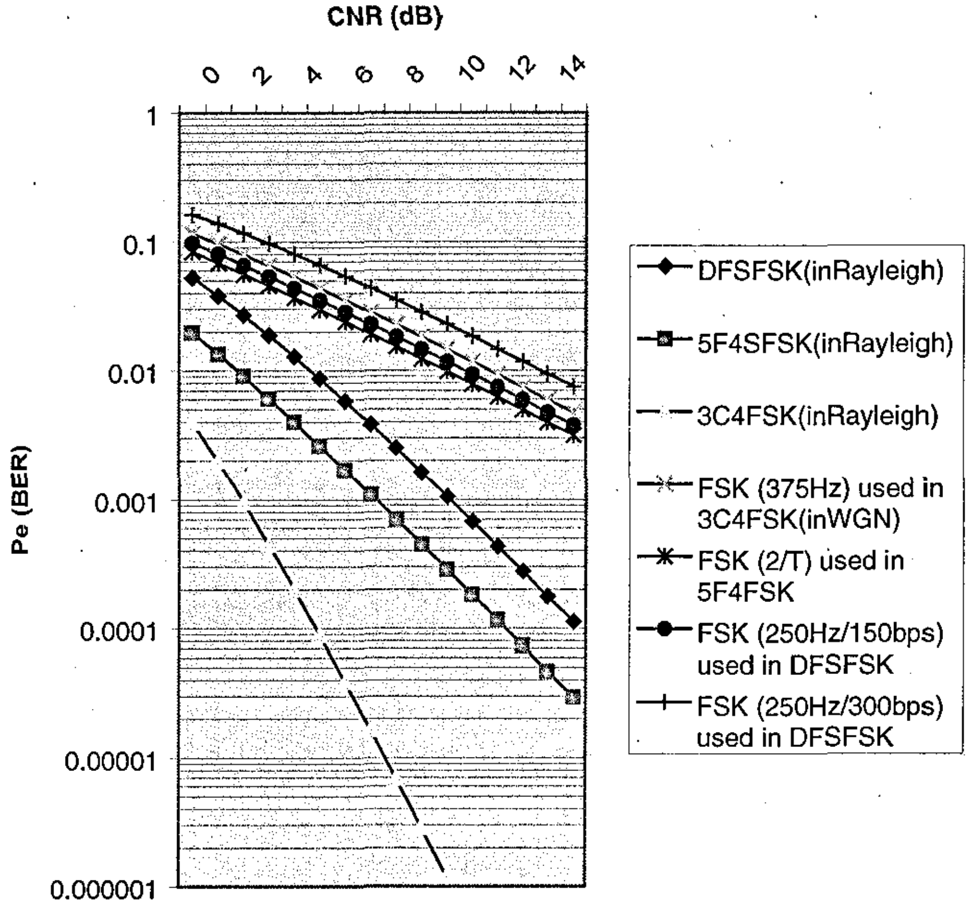


Fig. 8.3 Theoretical performance in Raleigh Fading (ten multipath)

17. As all the theoretical results do not take into account inter-symbol-interference and frequency specific fading, the effectiveness of the schemes may vary from the performance expected in the Raleigh fading model. The signal level used needs to be relatively large compared to the ambient noise, thus allowing low BER to be achievable.

18. The performance of the modulation techniques were investigated in a number of trials, aimed at establishing the practical performance within WGN and within a

Raleigh fading environment (multipath). The trials took the form of practical trials (Gaussian and Raleigh channels Fig.8.4, Fig.8.5). BER were established by comparing the received data with the transmitted data.

19. The performance of the modulation technique has been defined here in terms of the achieved *bit-error-rate* (BER). The BER's have been recorded and graphically displayed as a function of Carrier-to-noise ratio (CNR). The results are superimposed onto the theoretical Raleigh and White Gaussian Noise (WGN) theoretical performance graphs.

20. The practical performance of the modulation techniques within WGN is slightly worse than theoretical calculations. This is nearly a uniform displacement for the carrier to noise ratios tested. It can be seen that the theoretical performance is realistic (Fig. 8.4). The differences between the practical and theoretical performance are put down to a number of things including, 1. The fact that the theoretical analysis is not exact, with a number of phenomena being neglected, 2. Signal distortion that may be produced as narrow matched filters are used for signal discrimination. 3. System self noise that is produced by the receiver itself may also contribute. The theoretical performance can be looked at as what is possible under ideal conditions and therefore there is bound to be some discrepancy from what is achieved in a practical situation. The 3C4FSK scheme achieved the best BER performance as expected from theory, better than 1 in 10^{-5}, at a CNR of 10dB's. At the CNR of 8dB's a BER performance better than 1 in10^{-3} is achieved.

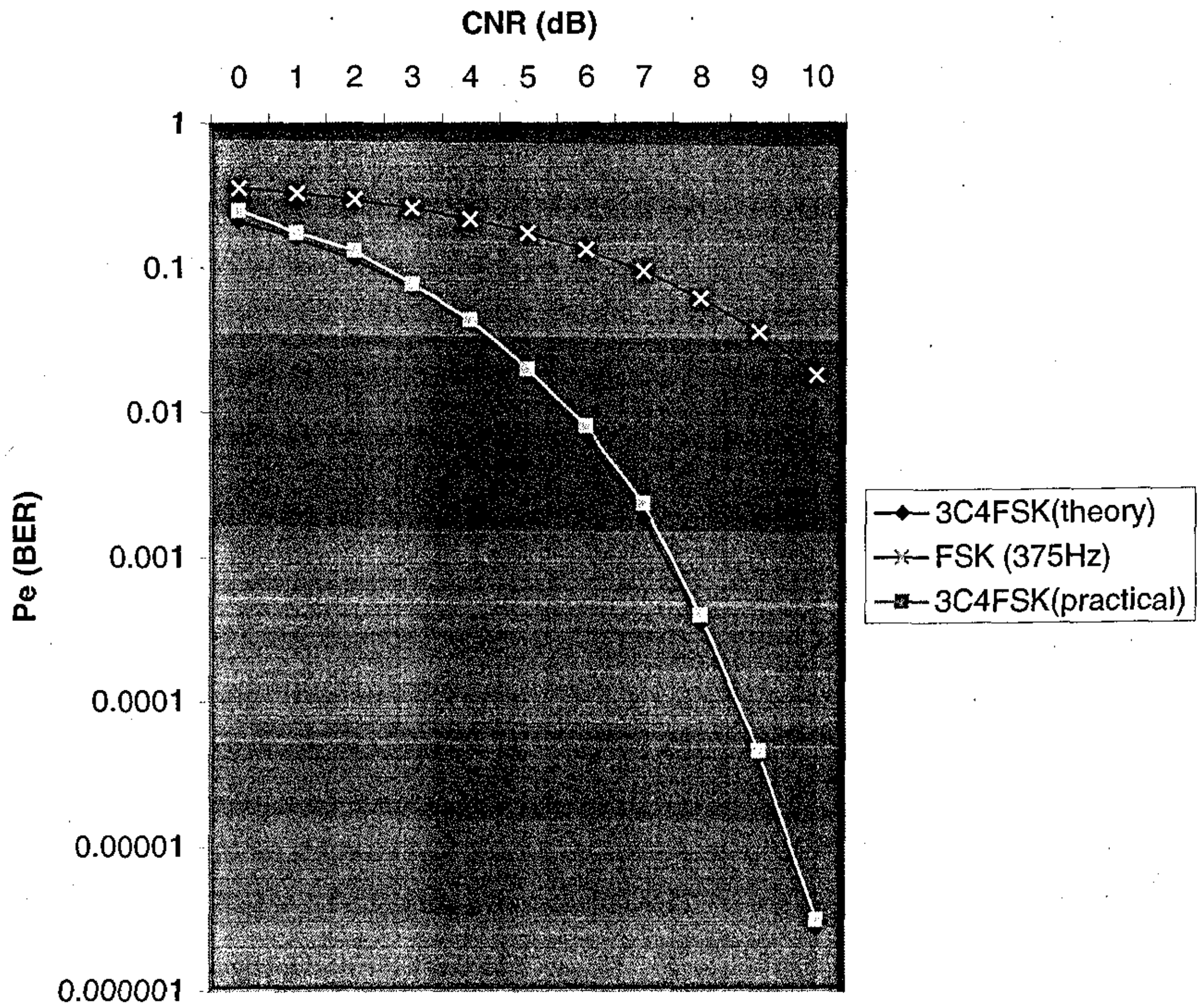


Fig. 8.4 3CFSK schemes practical and theoretical performance in WGN

21. The practical performance of the modulation techniques within a Raleigh channel is an order of 1dB's out from the theoretical calculations. This is also more or less a uniform displacement for the carrier to noise ratios tested. It can be seen that the theoretical performance is realistic (Fig. 8.5), small discrepancy from the theoretical results can be put down to system self noise and matched filter signal distortion as with the WGN result. The 3C4FSK scheme achieved the best BER performance, better than 1 in10^{-5}, at a CNR of 10dB's. At the worse CNR of 8dB's, a BER performance just worse than 1 in10^{-5}.

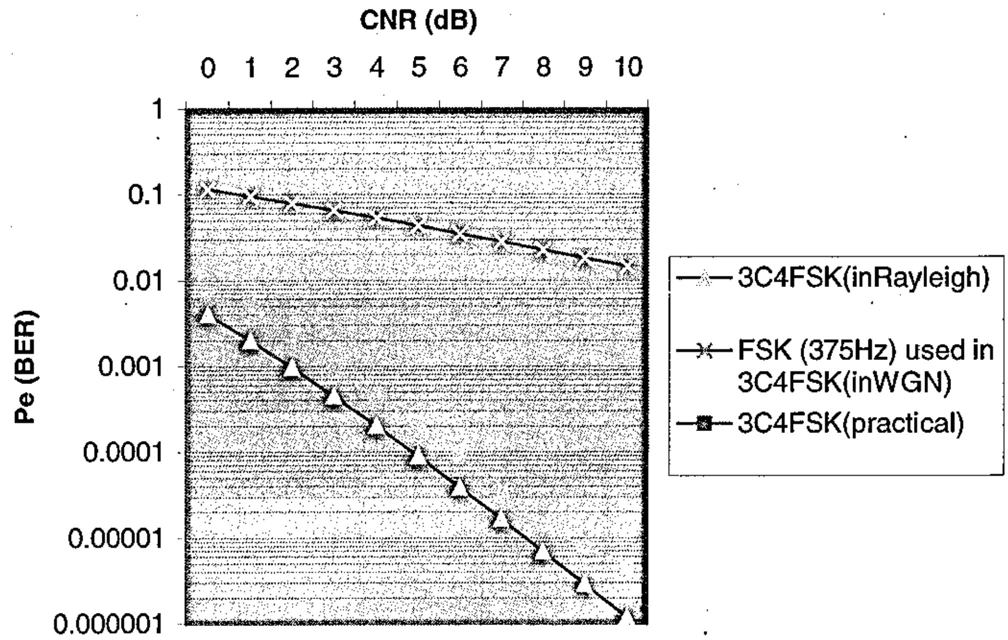


Fig. 8.5 Performance in Raleigh Fading (ten multipath)

Concluding Remarks

This chapter presents the theoretical and practical performance of the modulation techniques. Focusing on the WGN and Raleigh fading performance of each of the schemes. Theoretical performance is shown to be realistic by the practical results. The following chapter presents an evaluation of the work carried out within this body of research.

Evaluation

9. Evaluation

1. This chapter details the conclusions drawn; from the evaluation of the performance trials, background investigation and modulation development process. Comparison of the results with the theoretical WGN and Raleigh fading calculations respectively clearly shows comparable performance (Chapter 8). The results show significant improvements in performance when using each of the developed modulation schemes. The 3C4FSK-modulation technique shows the most improvement over a basic scheme and therefore is the obvious choice for any further investigation. This scheme has the most potential out of the schemes developed to provide a suitable modulation technique for the shallow, long-range sub-sea environment at bit rates of 125-150bps.

2. The serial redundancy in the 3C4FSK decision processes and its use of multiple frequencies helps to make the technique more robust. This combined with the intrinsic synchronisation provided within the scheme produces a modulation scheme that satisfies the criteria set out at the start of this research.

3. The parallel redundancy in the 5F4FSK scheme and the combined use of serial and parallel diversity used in the DFSFSK scheme even though showing improvement over basic modulation schemes are restricted by the bandwidth restrictions placed on this research. The additional frequencies required in the limited bandwidth matched filter implementations, make them more taps intensive. The number of taps that can be implemented is limited and therefore comparable implementations have a lower performance than the 3C4FSK scheme. This may explain the performances not reaching the levels, achieved using the 3C4FSK scheme.

4. Any future investigation should take the form of shallow-water, long-range seawater trials over extended periods of transmission using the 3C4FSK scheme. Such trials are not part of this investigation but are the logical next step for any further investigations that have the required funding and can provide the appropriate trials environments.

5. It is also suggested that any investigation carried out should be widened, from systems that using single omni-directional receiver hydrophones. It should include multiple receiver techniques such as spatial processing of the received signal with multiple widely spaced receiver elements; as these schemes have been shown in the research literature to have potential. This would widen the investigation of modulation techniques to include phased based modulation schemes.

6. The contribution to knowledge contained in this thesis includes the; design; development; implementation and evaluation of three modulation schemes specifically for use within the shallow-water, long-range sub-sea environment. Frequency diversity, serial (time) diversity and parallel diversity are all utilised. The design and application of a number of innovative implementation techniques is also undertaken including a form of intrinsic synchronisation and coincident point envelope detection.

APPENDIX A MAXIMUM LIKELIHOOD DETECTION

APPENDIX A

MAXIMUM LIKELIHOOD DETECTION

1. Maximum likelihood detection involves making an estimate ($\hat{S}_i$) of the transmitted symbol (S_i). The received signal is operated on, producing an observation signal (x) that the estimate ($\hat{S}_i$) is made from. This is carried out whilst attempting to minimise the average probability of symbol error [143]. In other words this procedure is achieved while attempting to maximise the average probability of a correct decision.

2. The decision theory that this procedure is based on involves four hypotheses for a symbol alphabet with four equally likely transmitted symbols, i.e.

1. Symbol one (S_1) is being sent (H_1).
2. Symbol two (S_2) is being sent (H_2).
3. Symbol three (S_3) is being sent (H_3).
4. Symbol four (S_4) is being sent (H_4).

3. The observation signal (x) provides a way of choosing between the hypotheses. The *a priori* probabilities $P(H_1)$, $P(H_2)$, $P(H_3)$ and $P(H_4)$ can be defined as conditional probabilities that are dependent on the observation signal (x) [144].

1. The probability that symbol (S_1) was sent, given the observation signal (x), $P(H_1/x)$.
2. The probability that symbol (S_2) was sent, given the observation signal (x), $P(H_2/x)$.
3. The probability that symbol (S_3) was sent, given the observation signal (x), $P(H_3/x)$.

4. The probability that symbol (S_4) was sent, given the observation signal (x), $P(H_4/x)$.

4. The maximum *a posteriori* (MAP) decision criterion stipulates that a symbol (S_i) was sent, when the probability that this symbol was sent $P(S_i)$ given the observation signal (x) is larger or equal to the probability $P(S_k)$ for all symbols in the symbol alphabet, where $k \neq i$ (Equation A1). The "equal to" stipulation in the equation A1.A below points to the fact that if a decision is not achieved a random selection is appropriate and does not affect the average probability of symbol error [144].

5. Looking at the observation signal as continuous, the likelihood functions $p_1(x)$, $p_2(x)$, $p_3(x)$ and $p_4(x)$ represent the probability density of the observation signal (x) under each of the hypotheses. The maximum likelihood decision can be defined as in Equation A2.

Maximum *a posteriori* (MAP) decision criterion:

For a hypothesis (H_i) to be chosen:

$$P(H_i/x) \geq P(H_K/x) \text{ where } (k = \text{to all except } i,\ k \neq i) \qquad \text{(A1.A)}$$

If $i = 1$ for hypothesis H_1, then $k =$ 2, 3, 4. for a four symbol scheme.

Thus:

$$P(H_i/x) \big/ P(H_k/x) > 1, \ (k = \text{to all except } i,\ k \neq i). \qquad \text{(A1.B)}$$

Maximum likelihood decision:

Using :

$$P(A/B) = {P(A/B)}\big/{P(B)}$$

we get:

$$P(H_i/x) = \frac{p_i(x) \times P(H_i)}{p(x)},\ i = 1 \text{ or } 2 \text{ or } 3 \text{ or } 4.$$

Substituting this into Equation A1.A, we get the maximum likelihood decision:

$$\text{Choose } H_i \text{ when } \frac{p_i(x)}{p_K(x)} \geq \frac{P(H_k)}{P(H_i)},\ (k = \text{to all except } i,\ k \neq i). \tag{A2}$$

With all the all *a priori* probabilities equal, $\frac{p_i(x)}{p_K(x)} \geq 1$ ($k =$ to all except i, $k \neq i$).

The hypothesis whose likelihood is the largest is chosen $p_i(x) \geq p_k(x)$, ($k =$ to all except i, $k \neq i$).

APPENDIX B PHASE LOCKED LOOP

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PHASE LOCKED LOOP

1. The Phase Locked Loop (PLL) forms the basis of both the synchronisation extraction methods utilised in this work (Section 4.3.1.1, Section 4.3.1.2). The second order PLL is a stable synchronising circuit that can be used to track an input signal with a reference signal (oscillator output) and also provides noise suppression. The parameters that are adjusted as the output signal converges towards the input signal are the oscillator's output frequency and phase components. The PLL circuit tends towards a 'locked state' where the error between the output and the input is 'small' or zero. If a difference is detected between the signals a control action is produced that attempts to reduce the error to its minimum. The PLL structure is build up of three fundamental building blocks a phase detector, controlled oscillator (voltage or current) and a loop filter.

2. The basic PLL structure is shown in Fig.B1. The actual performance is dependent on the characteristics of the loop components used. For the practical specifications used, see Section 4.3.2. The locked state is defined here in terms of transfer functions.

3. The phase detector takes two signals as its input, the input signal $i(t)$ and a feedback signal $o(t)$, which is also the output signal. The phase detector's output $d(t)$ includes a 'dc' value that is dependent on the phase error θ_e; and 'ac' signal frequencies at multiples of the input frequency ($2\omega_1$, $4\omega_1$ and so on). The unwanted 'ac' signal components are negated by the low-pass loop filter $F(s)$. The filter output $f(t)$ controls the Voltage Controlled Oscillator's (VCO) output.

4. In the locked state the frequencies of the input signals and output signals are equal, ω_1 and ω_2 respectively. The phase transfer function $H(s)$ is related to the ratio of the input phase θ_1 to the output phase θ_2, and is dependent on the transfer functions of the individual components that make up the PLL. The phase detector has a gain G_d; the output of the phase detector $d(t)$ can be defined as $G_d \sin\theta_e$ if the higher frequencies are neglected. The loop filter negates the higher frequencies. For small phase errors the argument of the sine function is appropriate on its own; thus $d(t)$ can be defined as $G_d\theta_e$. In the locked state the phase detector can be looked upon as a zero order block.

5. The filter transfer function $F(s)$ is dependent on the implementation used. The VCO operation involves a gain G_v, and integration $\frac{1}{s}$ (Laplace transformation representation). The VCO performs an integral over the frequency variation $\Delta\omega_2$, to give the output phase θ_2. The transfer function of the VCO is G_v/s. The phase transfer function of the PLL, $H(s)$, is given by $H(s) = \frac{G_d G_v F(s)}{s + G_d G_v F(s)}$.

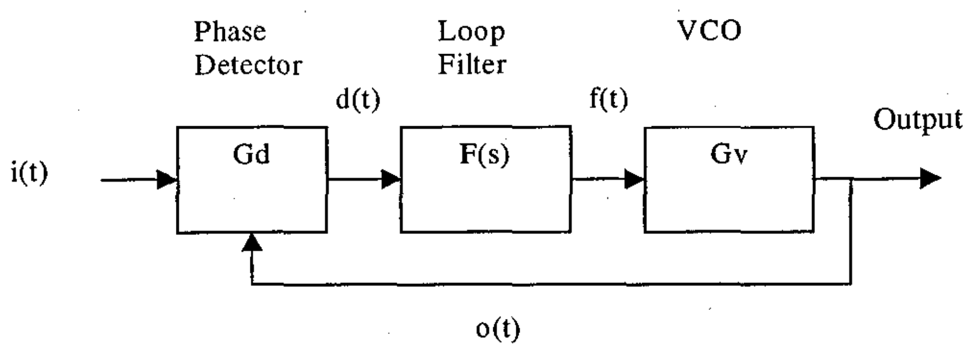

Fig. B1 PhaseLocked Loop

APPENDIX C DECISION EVALUATION

APPENDIX C

EVALUATION DECISION

3C4FSK:

1. Defining decision zero as *D0*, decision one as *D1*, energy decision as *DE* and symbol periods intuitively as A, B, C. Noting that $P_c = 1 - P_e$, 'and' represents the multiplication of two independent probabilities and *'or'* represents the addition of two mutually exclusive events. The probability of a correct decision (P_c) for Decisions D0, D1, DE is given by:

$$P_c(D0)= P_c(D1)= P_c(DE)=$$

$((1-P_e(D0_A))$ *and* $(1-P_e(D0_B))$ *and* $(1-P_e(D0_C)))$ *or* $((1-P_e(D0_A))$ *and* $(1-P_e(D0_B))$ *and* $(P_e(D0_C)))$ *or* $((1-P_e(D0_A))$ *and* $(P_e(D0_B))$ *and* $((1-P_e(D0_C)))$ *or* $((P_e(D0_A))$ *and* $(1-P_e(D0_B))$ *or* $(1-P_e(D0_C))$

$= ((1-P_e(D0_A)) * (1-P_e(D0_B)) * (1-P_e(D0_C))) + ((1-P_e(D0_A)) * (1-P_e(D0_B)) * (P_e(D0_C))) + ((1-P_e(D0_A)) * (P_e(D0_B)) * ((1-P_e(D0_C))) + ((P_e(D0_A)) * (1-P_e(D0_B)) * (1-P_e(D0_C))$

(A3)

The Probability of an error (P_e) for a 3C4FSK scheme can thus be calculated:

$P_{e3C2FSK} = 1 - (((P_c(D0))$ *and* $(P_c(D1)))$ *or* $((P_c(D0))$ *and* $(P_e(D1))$ *and* $(P_c(DE)))$ *or* $((P_e(D0))$ *and* $(P_c(D1))$ *and* $(P_c(DE))))$

$=1 - (((P_c(D0))$ *and* $(P_c(D1)))$ *or* $((P_c(D0))$ *and* $(1-P_c(D1))$ *and* $(P_c(DE)))$ *or* $((1-P_c(D0))$ *and* $(P_c(D1))$ *and* $(P_c(DE))))$

$= 1 - (((P_c(D0)) * (P_c(D1))) + ((P_c(D0)) * (1-P_c(D1)) * (P_c(DE))) + ((1-P_c(D0)) * (P_c(D1)) * (P_c(DE))))$

(A4)

5F4FSK:

2. Noting that the $P_e = 1 - P_c$, $P_c(D0) = P_c(D1) = P_c(D3)$, the probability of an error (P_e) for a 5F4SFSK scheme can thus be calculated:

$P_{e5F4SFSK} = 1 - (((P_c (D1))$ *and* $(P_c (D2)))$ *or* $((P_c (D1))$ *and* $(P_e (D2))$ *and* $(P_c(D3)))$ *or* $((P_e(D1))$ *and* $(P_c(D2))$ *and* $(P_c(D3))))$

$= 1 - (((1- P_e(D1)) \text{ and } (1- P_e(D2))) \text{ or } ((1- P_e(D1)) \text{ and } (P_e(D2)) \text{ and } (1- P_e(D3))) \text{ or } ((P_e(D1)) \text{ and } (1- P_e(D2)) \text{ and } (1-P_e(D3))))$

$= 1 - (((1- P_e(D1)) * (1- P_e(D2))) + ((1- P_e(D1)) * (P_e(D2)) * (1- P_e(D3))) + ((P_e(D1)) * (1- P_e(D2)) * (1-P_e(D3))))$ **(A5)**

DFSFSK:

3. Noting that the $P_e = 1- P_c$, $P_c(D1) = P_c(D2)$, $P_c(D1_A) = P_c(D1_B) = P_c(D2_A) = P_c(D2_B)$ and that $Pc(D3)$ has different size of filter bandwidths thus is different. The probability of a correct decision (P_c) for decisions one and two:

$Pc(D1) = Pc(D2) =((1- Pe(D1_A)) * (1- Pe(D1_B))) + ((Pe(D1_A)) * (1- Pe(D1_B) * (0.5)) + ((1- Pe(D1_A)) * (Pe(D1_B) * (0.5))$

The probability of an error (P_e) for a DFSFSK scheme can thus be calculated:

$P_{eDFSFSK} = 1 - (((Pc(D1)) \text{ and } (Pc(D2))) \text{ or } ((Pc(D1)) \text{ and } (Pe(D2)) \text{ and } (Pc(D3))) \text{ or } ((Pe(D1)) \text{ and } (Pc(D2)) \text{ and } (Pc(D3))))$

$= 1 - (((Pc(D1)) \text{ and } (Pc(D2))) \text{ or } ((Pc(D1)) \text{ and } (1- Pc(D2)) \text{ and } (Pc(D3))) \text{ or } ((1-Pc(D1)) \text{ and } (Pc(D2)) \text{ and } (Pc(D3))))$

$=1 - (((Pc(D1)) * (Pc(D2))) + ((Pc(D1)) * (1- Pc(D2)) * (Pc(D3))) + ((1- Pc(D1)) * (Pc(D2)) * (Pc(D3))))$ **(A6)**

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