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Low Complexity Equalization for Indoor Wireless Communications

by

Chandra Sekhar Bontu, B. Tech., M. Tech.

A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of the requirements
for the degree of

Doctor of Philosophy

**Ottawa - Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
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Abstract

This thesis is an investigation of low complexity and high performance equalization methods for indoor wireless communications. Adaptive equalization is recommended to improve the system performance in two distinctly different scenarios: (1) transmission at low data rates in the digital cellular frequency bands and (2) transmission at very high data rates above 20 Mbps in the millimeter wave frequency band.

For low data rates, typically below 1 Mbps, indoor channels are essentially frequency non-selective. Diversity transmission with equalization is proposed as a low complexity solution to improve the system performance. Various equalization strategies such as decision feedback equalization (DFE), maximum likelihood sequence estimation (MLSE) and their improved forms are compared based on their performance and computational complexity. The adaptation algorithm is limited to the simple LMS or its modified forms such as the normalized LMS. Due to this restriction, the proposed system performance is also evaluated in rapidly varying outdoor channels before arriving at the best equalization strategy. The effect of FEC coding and Viterbi decoding with finite bit-interleaving, as proposed in the IS-54 standard, is also investigated when an equalizer precedes the decoder.

For high data rates, typically above 20 Mbps, the indoor channel is characterized by severe multipath propagation. Due to technology and power limitations, nonlinearities cannot be avoided during the process of transmission and reception at millimeter wave frequencies. Therefore, constant envelope modulation with noncoherent detection is attractive. A combination of adaptive equalization and noncoherent demodulation is a leading candidate for data transmission for these systems. Continuous phase FSK (CPFSK), which preserves its constant envelope property even after hard-limiting and band-limiting, is selected as the modulation technique. Optimum and sub-optimum equalization techniques are derived for FSK signals received over dispersive radio channels. Equalization structures are derived for two different cases by: (1) directly processing the received signal in the absence of carrier phase information and (2) processing the noncoherently demodulated received signal. Performance evaluation is carried out for the low complexity receivers considering: (1) two path quasi-static channel and (2) measured channel responses in large open-concept office building, which represent a worst case scenario. A performance comparison with commutation signalling based receiver with codes of length 5 is also presented.

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List of Abbreviations and Acronyms

ACI	Adjacent channel interference
ACS	Add compare and select
AWGN	Additive white Gaussian noise
BDFE	Bi-directional DFE
BER	Bit error rate
BayDFE	Bayesian DFE
CCI	Co-channel interference
CPFSK	Continuous phase FSK
CPM	Continuous phase modulation
CS	Commutation signalling
DD	Differential detection
DDME	Differential detector followed by a memory based equalizer
DDT	Delayed diversity transmission
DFE	Decision feedback equalizer
DFSE	Decision feedback sequence estimation
DPSK	Differential phase shift keying
DQPSK	Differential QPSK
DS-SS	Direct sequence spread spectrum system
DSK	Double phase shift keying
FBME	Filter bank detector followed by a memory based equalizer
FDD	Frequency division duplex
FEC	Forward error control
FER	Frame error rate
FSK	Frequency shift keying
GHz	Giga cycle per second (10^9 cycles per second)

GMSK	Gaussian MSK
GSM	Group Spécial Mobile/ Global System for Mobile Communication
GaAs	Gallium Arsenide
HD	Hard decision
HEMT	High electron mobility field effect transistors
Hz	Cycles per second
IDF	Integrate and dump filter
IF	Intermediate frequency
IS-54	Interim Standard 54 (U. S.)
ISI	Intersymbol interference
K band	18 - 26.5 GHz
KMPH	Kilo-meters per hour
Ka band	26.5 - 40 GHz
LD	Limiter-discriminator
LDI	Limiter-discriminator-integrator
LMS	Least mean squares
LOS	Line of sight
MAP	Maximum a posteriori
MHz	Mega cycles per second (10^6 cycles per second)
ML	Maximum likelihood
MLP	Multi-layer perceptron
MLSE	Maximum likelihood sequence estimation
MSE	Mean squared error
MSK	Minimum shift keying
Mbps	Mega bits per second (10^6 bits per second)
NLMS	Normalized LMS
PDC	Personal Digital Cellular (Japan)
PSK	Phase shift keying

PSP	Per-survivor processing
QPSK	Quadrature phase shift keying
R+	Real addition
RAM	Random access memory
RF	Radio frequency
RLS	Recursive least squares
RMS	Root mean square
RSSE	Reduced state sequence estimation
RX	Real multiplication
SC	Sign change
SD	Soft decision
SE	Squared error
SER	Symbol error rate
SINR	Signal-to-interference + noise ratio
SIR	Signal-to-interference ratio
SNR	Signal-to-noise ratio
SRRC	Square root raised cosine
SYNC	Synchronization
TDD	Time division duplex
TDMA	Time division multiple access
dB	Decibels
kbps	Kilo bits per second (10^3 bits per second)
m	Meter
millimeter wave band	40 GHz - 300 GHz
ns	Nano second (1×10^{-9} s)
s	Second

List of Mathematical Symbols and Notation

All the bold letters represent a vector

ξ_s	Symbol energy
$\xi_k^{(l \rightarrow m)}$	Error associated with the transition from $s_{k-1}^{(l)}$, the l^{th} state at time instant $k-1$, to $s_k^{(m)}$, the m^{th} state at time instant k (refer to page 22)
$\delta(.)$	Dirac delta/ Unit impulse function
$\gamma_k^{(l \rightarrow m)}$	Branch metric associated with the transition from $s_{k-1}^{(l)}$, the l^{th} state at time instant $k-1$, to $s_k^{(m)}$, the m^{th} state at time instant k (refer to page 22)
μ	Step size parameter for adaptation
ω_c	Carrier frequency in radians/ s.
Ω	Initial phase of the carrier
$\alpha_n = \sum_{k=0}^{n-1} I_n$	Initial phase of the CPFSK signal for the n^{th} signalling interval
ϕ	Sum of the initial phase, Ω , and the uncertainty in the carrier phase
$\sigma_n^2 = E\{n_k^2\}$	Noise variance
$\Gamma(\phi)$	Likelihood/ log likelihood function when the parameter ϕ is known
Γ_k	Likelihood function at time kT_s
τ	Detection delay for the differential detector
τ_c	Time delay between the channel paths
τ_D	Transmission delay between the antennas

β	Filter roll-off factor for SRRC filter
β_i	Phase of received signals at time iT_s
ρ	Number of states of MLSE
$\tilde{s}_{k-1}^{(l)}$	Survivor sequence corresponding to $s_{k-1}^{(l)}$.
v	Number of channel estimators involved using PSP
x^*	Complex conjugate of x
$A \bullet B$	Inner product of vectors A and B
$A \cup B$	Union operation on sets A and B
A^H	Hermitian transpose of the vector A
$\text{Arg}[x]$	Phase of x
$ x $	Magnitude of x
$E[x]$	Expected value of the random variable x
$\ln(x)$	Natural logarithm of x
$\langle x \rangle$	Average value of x
$p(A B)$	Probability of occurrence of A given that B has already occurred
$p(A)$	Probability of A
$\text{Re}\{x\}$	Real part of complex variable x
$\text{Im}\{x\}$	Imaginary part of complex variable x
$()^{-1}$	Inverse operator

$ x $	Magnitude of x
$\lceil \epsilon \rceil$	Smallest integer greater than ϵ
$\otimes$	Convolution operator
$\hat{x}$	Estimated value of x
B_h	Parameter to control the spectral characteristics of GMSK signal
$c(t)$	Channel impulse response in continuous time
c_k	Channel in discrete time representation at time kT_s
$\hat{c}_k$	Estimated channel in discrete time representation at time kT_s
$d(t)$	Pulse amplitude modulated (PAM) signal
d	Detection delay in symbol durations for an MLSE
e_k	Instantaneous error between the estimated and actual sample value
$f(I, c_{k-1})$	Output of the channel, whose impulse response is c_{k-1} , to CPFSK signal transmission for an input sequence of I
f_d	Frequency deviation
$\{f_k(t)\}$	A complete orthonormal set of complex functions in $L_2[0, BT_s]$
$g(\cdot)$	Frequency response function of CPFSK signal
h	Modulation index
$I_0(\cdot)$	Zero th order modified Bessel function of the first kind
I_k	Sample of transmitted signal at time kT_s
$\hat{I}_k$	Sample of the estimated signal at time kT_s

$J_0(\cdot)$	Zero th order Bessel function of the first kind
K	Number of survivors per trellis state
L	Number of symbol periods over which the channel spans
$L_2[0, BT_s]$	A function space of finite energy signals in the time interval $[0, BT_s]$
M	Size of the symbol alphabet
N_0	Power spectral density of complex valued AWGN
N_a	Number of antennas
N_B	Number of symbols in a sequence/ block to be detected
N_b	Number of feedback taps of a DFE
N_{C-1}	Number of symbol periods over which the combined channel, transmit and receive filters spans.
N_D	Number of information symbols
N_f	Number of feed forward taps of a DFE
N_T	Number of training symbols
N_s	Number of sinusoids used in modelling the Jake's model
n_k	Sample of noise at time kT_s
$q(\cdot)$	Phase response function of CPM signal
$r(t)$	Received signal in continuous time representation
r_k	Sample of received signal at time kT_s
$\hat{r}_k$	Estimated sample of the received signal at time kT_s

$s(t)$	Transmitted RF signal
$s_{k-1}^{(l)}$	The l^{th} state at time $(k-1)T_s$
T_b	Bit duration
T_s	Symbol duration
v	Number summations of exponential terms involved in Bayesian DFE
W	Transmission bandwidth
$\{w_{b,i}\}$	Coefficients of the feedback filter for a single tap filter
w_f	Forward tap coefficient for a single tap DFE
$X_n(t)$	Low pass representation of the noise-free received signal
$z(t)$	Low pass representation of $s(t)$

Chapter 1

Introduction

Increasing demand for wireless personal communications has necessitated a search for reliable low complexity transceivers for indoor and mobile environments. Depending on the channel coherence bandwidth and the transmission rate, two communications system configurations can be defined: (1) narrow band systems; where the coherence bandwidth of the channel is much larger than the transmission bandwidth and (2) wide band systems; where the coherence bandwidth of the channel is smaller than the transmission bandwidth. The existing digital cellular radio standards, viz. the North American IS - 54 [45], [105], the European GSM [50], [105], the Japanese PDC [105], are TDMA systems defined for low rate digital information transmission. These existing system standards may also be extended to in-building communications. These systems can be classified as narrow band TDMA systems in a typical indoor environment. However, the existing portable handsets need to be simplified to reduce the size and power consumption, which are the challenging issues for present day research. Furthermore, the difference in the channel behavior is a factor to be considered in this process. The aforementioned standards were designed for a typical outdoor environment, where multipath propagation and severe Doppler frequency shifts are the main detrimental factors. Some add-on techniques are helpful to the present cellular transceivers for successful and cost-effective operation in an indoor environment. For higher data rates above 20 Mbps, i.e. wide band systems, higher transmission frequency bands, typically above K band are required to simplify the design of image rejection filters, etc. For example, for data rates above 20 Mbps, an ideal selection would be K , Ka and millimeter wave frequency bands. With the recent advances in the solid-state technology (GaAs and HEMT devices) efficient design of both RF and base band subsystems is possible at these frequencies and data rates [32], [114], [115]. Severe multipath and nonlinear distortion are problems in the case of wide band systems.

In this thesis, low complexity solutions to these problems involved with both narrow band and wide band systems in a typical indoor environment are addressed. For narrow band systems, the applicability of an existing digital cellular standard (IS-54), at the recommended data rate, to a typical indoor environment is investigated. For wide band systems, digital data transmission in the range of 20 - 160 Mbps in millimeter wave frequency band is considered. Novel low complexity adaptive equalization methods are considered as the main tool to improve the system performance in both cases.

Before stating the particular objectives of this thesis, the channel behavior followed by a comprehensive report on low complexity signal processing strategies for narrow band

and wide band systems is given in the following sections.

1.1 Communication in the indoor environment

The channel behavior at the transmission frequencies of interest is important in selection of optimum system parameters, such as coding, modulation, etc. The experimental results [4], [72], [114], [129] - [131], [141] show that the channel characteristics in an indoor environment are a function of various factors, such as the material of the walls, dimensions of the building, the population density of the building, etc. Depending on the size of the building, the channel dispersion varies from 10 ns to 200 ns. Therefore, for data rates of less than approximately 1 Mbps the channel is essentially frequency nonselective.

For wide band systems, where the data rates are typically above 20 Mbps, the channel exhibits severe multipath propagation. Furthermore, at transmit frequencies above 20 GHz the transmission experiences severe shadowing effects. Due to the above mentioned factors, anti-multipath signal processing methods, such as adaptive equalization [109] or anti-multipath modulation methods, such as double phase shift keying (DSK) [9] or commutation signalling [137] are required to achieve acceptable system performance. An assessment of the feasibility of implementation of these methods with the existing technology is most important and is needed for optimum selection of system parameters.

The experimental results reported in the literature indicate that the statistical modelling of the channel above 20 GHz is difficult. The channel can be simulated assuming approximate channel models. However, it is more appropriate to use experimental measurements of channel responses in system performance simulations.

1.2 Brief background on equalization, diversity and modulation

In this section, applicability of various signal processing methods, such as modulation, equalization etc., is discussed for system performance improvements and low complexity. Since this thesis deals with equalization for two distinctly different applications, narrow band systems, which are characterized by essentially frequency non-selective or flat fading and wide band systems, which are characterized by severe multipath propagation, the following discussion is organized into two sub-sections, corresponding to each of these systems. Note that these two issues, narrow band and wide band systems, respectively look into (1) the applicability of existing digital cellular standards to a typical indoor environment without much modifications and (2) the multi-media communications capability in a typical indoor environment.

Narrow band systems:

Recent advances in digital cellular radio communications show that the combination of adaptive equalization and antenna space diversity plays an important role in improving the system performance in a typical fading and multipath environment [16], [17], [53], [66], [87]. Adaptive equalization improves the system performance by exploiting the time diversity in the received signal, whereas, space diversity combining improves the system performance by exploiting the spatial correlation between the multiple copies of the received signals. The requirement of multiple antennas and the duplication of RF front ends makes spatial diversity combining difficult and costly at the portable terminal. With present day technology, adaptive equalizers can be implemented at reasonable data rates with reasonable power consumption [55], [91], [96]. However, adaptive equalization without the help of antenna diversity combining does not give system performance improvements in the absence of multipath propagation. This situation occurs when the channel dispersion due to reflections from the surroundings is typically less than 20% of the symbol period. This is the situation for data rates less than approximately 1 Mbps in a typical indoor wireless transmission.

Wide band systems

On the other side of the coin, adaptive equalization does offer excellent gains in system performance in wide band systems, where severe intersymbol interference (ISI) is the main detrimental factor. But present day technology limitations in the high frequency bands restrict the selection of carrier modulation/ demodulation methods. Circuit nonlinearities can hardly be avoided at these high transmit frequencies without a large increase in RF circuit complexity and power consumption. Therefore, spectrally efficient phase/ amplitude modulation methods may not be a good selection because of the following difficulties: (1) linear amplification at these frequencies is a challenge to the present day solid-state technology. Even if it can be achieved, it will be more power hungry because of the presence of add-on circuitry [31], [90] and transmitted power back-off, (2) any nonlinearities in amplification or in the process of transmission increase the transmission signal bandwidth, and (3) the carrier phase recovery is also a difficult task at data rates above 20 Mbps making coherent demodulation difficult. The last difficulty suggests that non-coherent demodulation methods are preferable. The first two difficulties suggest that constant envelope modulation methods are preferable. But the application of constant envelope modulation with noncoherent demodulation complicates the succeeding adaptive equalization process.

As an alternative to equalization, anti-multipath modulation methods, such as DSK

or commutation signalling, can be used. Double phase shift keying [9], in which the phase is changed twice within a symbol period, is shown to be very effective in suppressing ISI. However, since the information is carried by the phase of the signal, this cannot be recommended for high frequency transmissions. Another alternative is commutation signalling [137], in which different orthogonal codes are assigned to each symbol in the alphabet and therefore the symbols can be detected using correlators at the receiving end. But the fundamental aspect of selection of modulation still remains. For example, if a constant envelope modulation is selected, in view of power efficiency and nonlinearity issues, then the transmission bandwidth required is processing gain times the original bandwidth requirement for a constant envelope modulation. Now the selection between constant envelope modulation with equalization and constant envelope modulation with orthogonal coding depends on three factors: (1) transmission bandwidth requirement, (2) complexity of the receivers, and (3) performance in typical indoor channels with impairments, such as, co-channel and adjacent channel interferences.

In the previous paragraphs, the main problems and constraints associated with both the narrow band and wide band systems are clearly stated. In the following section, we state the specific thesis contributions, after a brief description of the suggested solutions to the above problems.

1.3 Thesis contributions

From the brief system aspects discussed in the previous section the suggested techniques to improve the system performance followed by specific thesis contributions are listed below.

(1) Suggested solutions for narrow band systems:

In this thesis, a low complexity alternative to spatial diversity combining for narrow band systems is investigated. This work facilitates a study into the applicability of the existing digital cellular standards for inbuilding communications. Since the available portable receivers are already equipped with equalizers the main idea is to use equalization to improve the forward link performance with the help of some add-on techniques at the base station. Diversity transmission from the base station is projected as a way to introduce time-correlation in the transmitted information signal, which in turn is exploited by an equalizer at the portable. Various improved equalization methods suggested for fading channels perform differently depending on the operational environment and the constraints set forth by the communication link. In this thesis, performance of symbol-by-symbol equalization methods, such as conventional DFE, bi-directional DFE and Bayesian DFE

etc., and sequence detection methods, such as MLSE etc., are investigated by defining a TDMA frame following the IS-54 standards and considering link performance, computational complexity and information throughput as the main criteria. Note, however, that the proposed system configurations are equally applicable to other TDMA digital cellular standards, such as GSM. The major constraint employed throughout this research work is restricting the channel tracking algorithm to the simple least mean squares (LMS) algorithm. A detailed evaluation and computational requirements including architectural issues are given for various equalization methods.

The following are the specific contributions:

- (i) Diversity transmission from the base station is suggested as a low complexity alternative to diversity combining at the portable. Of all the diversity transmission methods, as explained in Chapter 2, delayed diversity transmission from the base station and adaptive equalization at the portable is recommended for reducing overall complexity. (Section 3.1)
- (ii) Various factors, such as transmission delay between the antennas, channel acquisition and tracking algorithms etc., that influence the system performance are listed. Optimum transmission delay between the antennas is derived from computer simulations assuming an arbitrary TDMA frame structure. (Sections 3.2 - 3.4)
- (iii) System performance considering the practical constraints, such as limited training, channel variations etc., is also evaluated. A TDMA frame structure following the existing digital cellular standard in North America, IS-54, is considered in this work. A major constraint employed in system evaluation is that the adaptation algorithm used is only limited to the simple LMS algorithm. For faster acquisition, the RLS algorithm is used during the training period. The equalization methods considered in this work are: (a) conventional decision feedback equalizer (b) bi-directional equalizer (c) Bayesian decision feedback equalizer (d) maximum likelihood sequence estimation (MLSE) with conventional channel tracking using the tentative decisions and (e) MLSE with combined channel tracking and sequence estimation algorithm, also known as the per-survivor processing (PSP) algorithm. (Sections 3.5 and 3.6)
- (iv) The portable with the above mentioned equalizers should also work successfully in a typical outdoor environment, where the channel variations are severe. Since we are only limiting the channel tracking algorithm to LMS, which is not recommended at fast channel variations, this evaluation is essential. Therefore the performance of those equalizers, which perform satisfactorily in indoor environment, is also evaluated in an outdoor environment at various vehicle speeds. (Section 3.7)

(v) Of all the equalization methods considered, we conclude that the best equalization strategy is the MLSE-PSP based on the system performance and computational complexity in both indoor and outdoor environments. (Sections 3.6 and 3.7)

(vi) Finally, an attempt is made to use the LMS algorithm during the acquisition phase also. This is demonstrated by considering an arbitrary frame structure and evaluating blind acquisition properties of the MLSE-PSP. The results show that using MLSE-PSP, the channel can be acquired within 30 symbol durations even at a Doppler frequency of 90 Hz, which corresponds to 108 KMPH. In addition, no training sequence is needed during this acquisition mode. (Sections 3.6 and 3.7)

(vii) Implementation of MLSE-PSP algorithm using a parallel structure is also addressed. The computational complexity and memory requirements of MLSE-PSP for two antenna transmission are shown to be comparable to a conventional DFE operating with the LMS algorithm. Furthermore, the overall computational complexity during the acquisition phase for MLSE-PSP with the LMS algorithm is shown to be significantly less compared to a DFE with the RLS algorithm. Note that the DFE acquires the channel based on a known training sequence, whereas, MLSE-PSP does not require any training sequence. (Section 3.6)

(viii) Another aspect which is investigated in this thesis is the relevance of error control coding for indoor radio channel with finite interleaving as recommended in the IS-54 standard. All the above mentioned performance evaluations are carried out for the described TDMA system with the recommended interleaving and coding. The results demonstrate that (a) coding offers marginal gains with respect to system frame error rates even at Doppler frequencies as low as 2 Hz and (b) even though ideally speaking coding should offer additional advantages at higher Doppler frequencies, this is not necessarily true when a signal processor, such as an equalizer, precedes the decoder. (Sections 3.5 - 3.8)

(2) Suggested solutions for wide band systems:

At millimeter wave frequencies, the information transmission using a constant envelope modulation scheme minimizes power consumption, spectral side-lobes regrowth due to nonlinearities and signal distortion. A combination of non-coherent detection and adaptive equalization is a challenging task for non-linear modulation schemes. Selection of modulation scheme, noncoherent demodulation and equalization is the main issues here. Optimal sequence estimation strategies for different combinations of demodulation and equalization are derived. Depending on the signal statistics at the output of different noncoherent demodulators, such as differential detector, filter bank detector etc., optimal/

sub-optimal sequence detectors are suggested. Performance of these receivers is evaluated by using a simple two path channel model and compared with a reference system. Furthermore, the system performance, defined as outage probability as a function of location, is evaluated based on the experimentally obtained channels in a large indoor area (85 m length X 37 m width X 4 m height). The feasibility of high rate data transmission using the proposed receiver structures is evaluated.

The following are the specific thesis contributions:

- (i) Continuous phase FSK (CPFSK), a constant envelope modulation, is considered in this work. Optimal/ sub-optimal sequence estimation strategies are derived considering the demodulation and equalization as a single operation. When the carrier phase is known at the receiver, i.e. coherent reception, the optimal sequence estimation is shown to be the method suggested in [64]. For noncoherent reception, i.e. when the receiver does not try to retrieve the carrier phase, an optimal receiver is derived. Both these receivers are concluded to be implementable, but are very complex. (Section 4.4)
- (ii) Alternatively, optimal/ sub-optimal receivers are derived for equalization of noncoherently demodulated CPFSK signals. The noncoherent demodulation methods considered here are: (1) differential detection with fractional symbol detection delays and (2) filter bank detection for noncoherently orthogonal signalling. The investigations are limited to four level CPFSK signalling. Depending on the signal statistics at the output of the demodulator, an optimum sequence estimation strategy is suggested. The sub-optimal receivers consist of a noncoherent demodulator followed by a memory based equalizer. (Section 4.5)
- (iii) The performance of the proposed receivers is evaluated in a simple two path quasi-static channel. These results are compared with a known system with QPSK transmission and reception using conventional DFE. The bandwidth requirement and achievable performance are compared. (Section 4.6)
- (iv) Performance of the proposed systems based on equalization of noncoherently demodulated signals is compared with a system employing commutation signalling and DPSK modulation. The channel is assumed to be a three path quasi-static one. (Section 4.7)
- (v) The memory requirement for the memory based equalizer depends on the number of symbol periods the channel spans. For the data rates and the delay spreads considered in this work the channel spans from 4 to 32 bit periods. In this situation, the suggested methods are very complicated. However, FSK signaling and demodulation is well-known

for its capture effects. Thus, the residual ISI may be suppressed somewhat by the FSK demodulator before equalization. Therefore, we suggest a sub-optimal equalizer, spanning only 2 symbol periods. The performance criterion is the outage probability. Here, we use the channel frequency responses reported in [33] to obtain the channel impulse responses. (Sections 4.8.1 and 4.8.2)

(vi) Capture properties of the FSK signals is investigated by defining a parameter, α , a measure of the power outside the equalizer-span. Assuming the same E_b/N_0 of 27 dB at all the locations, if the ratio of signal power within the equalizer window to the signal power outside the equalizer window is 20 dB, the filter band detector as suggested in Chapter 4 can suppress the residual ISI, irrespective of how the power outside the equalizer span is distributed. Based on these results, a feasibility evaluation is carried out as a function of location and data rate. (Section 4.8.3)

(vii) The proposed MLSE receiver is modified to improve the system performance further. A decision feedback sequence estimator (DFSE) consisting of separate RAMs to learn the ISI from the previously detected symbols, which lie outside the equalizer's window, is shown to improve the system performance. A simple memory based DFE is also configured and the results are shown in Chapter 4. (Section 4.8.4)

1.4 Thesis organization

This thesis is organized into 5 chapters. Of the remaining chapters, the second chapter reviews recent advances in equalization methods presented in the literature. Chapters 3 and 4 present the proposed narrow band and wide band systems respectively. Chapter 5 summarizes the work, outlining some directions for future research.

In Chapter 2, the key elements in channel equalization, which affect the system performance are introduced. In the recent literature, novel equalization methods were suggested to improve the system performance and/or decrease the computational complexity in typical fading channels. Some of these equalization methods are outlined. Note that the presented equalization methods are equally applicable to both narrow band systems with diversity transmission and wide band systems. Depending on the implementation feasibility and performance and various envisaged problems, such as nonlinearity issues as mentioned in Section 1.2 the reviewed equalization methods are appropriately recommended for either narrow band or wide band systems. In the initial part of the chapter, the presentation of equalization methods is independent of the type of modulation used. Towards the end of the chapter combined demodulation and equalization methods are discussed in relation to PSK and FSK signals.

In Chapter 3, equalization for narrow band wireless systems is addressed. Here, the applicability of diversity transmission from the base station, as an alternative low complexity method to diversity reception at the portable, is explored. Optimum transmission delay between the transmitting antennas is derived through computer simulations. Effects of limited training and LMS adaptation are investigated considering a practical TDMA system following IS-54 standard. Comparison among various equalization methods is carried out at a channel data rate of 48.6 kbps at 900 MHz and a maximum Doppler frequency of 6 Hz, which corresponds to a walking speed of 2 m/s. Among the equalization methods evaluated, some of the methods are selected as the best based on their performance in both indoor and outdoor environments (using LMS algorithm) and their computational complexity. Finally, the relevance of forward error control coding, with finite interleaving in a typical indoor channel is addressed. During this evaluation, soft decision decoding is used whenever the process in obtaining soft decisions from the equalizer does not involve a considerable increase in the computational complexity.

In Chapter 4, equalization for wide band wireless systems is addressed. The areas explored here include the selection of demodulation and the equalization methods. These two operations, demodulation and equalization can be considered as a single operation, or alternatively as separate operations. Continuous phase FSK (CPFSK), a constant envelope modulation, is selected as the modulation technique. After showing the enormous computational complexity involved in coherent demodulation and equalization of CPFSK signals, in particular, for the data rates considered in this work, optimal detectors for both the above mentioned approaches are derived. When demodulation and equalization are considered as a single operation, an optimum receiver depending on the statistics of the received signal, without the knowledge of carrier phase, is derived. In the second case, when the demodulation and equalization are considered as separate operations, two types of receivers are possible: (1) equalization followed by demodulation and (2) demodulation followed by equalization. However, the first scheme is a coherent case, where the equalizer recovers the phase of the carrier. In the second scheme, an optimum equalization strategy is devised by deriving the signal statistics of different noncoherent demodulation methods. In this work we considered two types of noncoherent demodulation methods: (1) differential detection and (2) filter bank detection for noncoherently orthogonal signals. The performance of the receivers is compared with a well known system considering QPSK transmission and a complex tap DFE in a two path quasi-static channel. Notice that the known system with QPSK transmission and DFE is a challenge to implement at the data rates and transmission frequencies considered in this work. A performance comparison of the proposed system with DPSK commutation signalling is also carried out. For this purpose, we considered the system performance results reported in [80] for commutation

signalling. Finally, at 20, 40, 80 and 160 Mbps data rates, we evaluated the performance of the proposed systems in a large indoor environment, using the measured channel responses at a carrier frequency of 40 GHz. These channel measurements were provided by Bultitude et. al. [33].

In Chapter 5, the thesis work is summarized. Some directions for future work are given.

Chapter 2

Review of Advances in Adaptive Equalization for Fading Channels

As mentioned in Chapter 1, the proposed solutions for both narrow band and wide band systems consist of an equalizer as the pivotal subsystem at the receiver to improve the system performance. Therefore, it is appropriate at this point to review some important equalization methods proposed in the earlier literature. Note that, the equalization methods presented in this chapter are applicable to a typical digital cellular environment, where intersymbol interference (ISI) and fading are the main concerns. Selection among these equalization methods depends on problems associated with the process of transmission and reception, such as the nature of ISI, modulation/ demodulation techniques used, feasibility of implementation, etc. This chapter starts with a more general discussion on channel equalization. In the context of the problems associated with narrow band and wide band systems, as listed in Chapter 1, the applicability of these equalization methods to the respective systems is discussed at different stages.

Equalization is a well known method for combating ISI introduced by band limiting and multipath propagation [109]. For portable communications, where the characteristics of the transmission medium change significantly with time, the equalizer parameters are continuously adapted. Equalization offers additional advantages by suppressing unwanted cyclo-stationary interferences, such as cochannel interference (CCI) and adjacent channel interference (ACI), which are common in a typical cellular scenario [53], [88], [89], [107], [108]. In this chapter, various issues involving adaptive equalization at a portable in a digital cellular environment are addressed. A thorough over-view of the evolution of various advanced techniques from the classical approaches is presented. Initially, various considerations which influence the performance of the equalizer and the associated efficiency measures which judge the performance of the equalizer are listed. The equalization strategies presented in this chapter are classified mainly into two main categories: (1) symbol-by-symbol detectors and (2) sequence detectors. The applicability of these equalization strategies to narrow band and wide band systems considered in this thesis are discussed based on the expected nature of ISI. Subsequently, the effect of modulation and demodulation on equalization is addressed with respect to receiver complexity. Based on the suggested modulation and demodulation methods, the applicability of the reviewed equalization methods to narrow band and wide band systems is further discussed at the end of this chapter.

2.1 Various considerations and efficiency measures for channel equalization

In this section, various considerations which affect the performance of equalization are discussed. The selection of these functions depends on some efficiency measures which are defined later in this section.

2.1.1 Various considerations that influence the performance of an equalizer

The performance of an equalizer in a specific environment is strongly dependent on the following:

(1) Cost function:

An equalizer tries to optimize the system parameters by minimizing a predefined cost function. The optimization process may involve (i) minimization of the mean squared error (MSE) between the estimated and the expected information signal or the estimated and the observed received signals, (ii) maximization of the a posteriori (MAP) or the maximum likelihood (ML) function, etc.

MAP estimation of a symbol reduces to minimizing the probability of symbol error when the information symbols are equally likely and independent (which can also be considered as ML estimation). Since MSE is not completely representative of system symbol error rate, which is generally considered to be the most important performance measure of a communications system, MSE equalizers are always expected to perform worse compared to MAP/ ML equalizers. MAP and ML estimation are classified as nonlinear estimation methods. MSE estimation is a linear estimation method, since it involves only second order statistics in the estimation process. For a particular case, when the probability distribution function of the thermal noise is Gaussian, ML estimate and MSE estimate are identical.

(2) Detection scheme:

The information symbols can be detected symbol by symbol or as a sequence of symbols. In symbol-by-symbol detection scheme, each symbol is detected independently, whereas, in sequence detection a block of information symbols is detected. Therefore, the cost function is calculated over a symbol period in the case of symbol-by-symbol detectors and over a period lasting a block of symbols in the case of sequence estimation. Sequence detectors use probabilistic algorithms to decode, thus, incur large delays in the detection process. These detection delays might in turn affect the channel estimation process. Therefore, selection between the two detection schemes depends mainly on whether the

system is affected further by the large decision delays.

(3) Adaptation algorithms:

The nature of the adaptive algorithm used has a direct impact on the system performance, in particular, in a time varying transmission medium, such as a fading radio channel. Classical algorithms such as the least mean squares (LMS) algorithm [67], [109], the recursive least square (RLS) algorithm [67], [109] etc can be quoted as typical examples. The selection among these algorithms depends on the system performance and computational complexity.

(4) Adaptation procedure:

The deleterious effects of the transmission medium can be suppressed by adaptively learning the characteristics of the medium. This learning process can be performed either by observing the corrupted received signal or by identifying an equivalent system whose output is the same as the received signal. The former case is referred to as the direct adaptation process, whereas the latter is referred to as indirect adaptation. The following equations interpret these two schemes for the LMS algorithm.

Direct adaptation:

$$\begin{aligned}\hat{l}_k &= \hat{c}_k \bullet r_k \\ e_k &= l_k - \hat{l}_k \\ \hat{c}_{k+1} &= \hat{c}_k + \mu e_k r_k^*\end{aligned}\tag{2.1}$$

where l_k , $\hat{l}_k$, r_k , e_k and $\hat{c}_k$ are the transmitted signal, estimated signal, observed received signal vector, estimated error and equalizer parameter vector at time kT_s , respectively. μ is the adaptation step size and the superscript $(*)$ indicates the complex conjugate operation.

Indirect adaptation:

$$\begin{aligned}\hat{r}_k &= \hat{c}_k \bullet l_k \\ e_k &= r_k - \hat{r}_k \\ \hat{c}_{k+1} &= \hat{c}_k + \mu e_k l_k^*\end{aligned}\tag{2.2}$$

where I_k , r_k , $\hat{r}_k$, e_k and $\hat{c}_k$ are the transmitted signal vector, observed received signal, estimated received signal, estimated error and channel impulse response vector at time kT_s respectively.

In the case of fading channels, indirect adaptation is expected to perform significantly better than direct adaptation [89]. For direct adaptation, a small change in channel coefficients may lead to a huge change in the coefficients of the forward filter, since the function of the forward filter is to estimate the inverse of the channel frequency response. This is especially apparent when the channel coefficients are small (i.e., in a deep fade). Since the steady state tracking error is proportional to the time variation in the filter coefficients, the channel tracking error is significantly less than the equalizer tracking error. Some of the results presented in Chapter 3 strengthen this statement. In addition, the multiplications denoted by (2.2) can be performed by sign changes and few real multiplications, since the transmitted symbols are usually amplitude quantized signals. Depending on the required performance and affordable complexity, the factors outlined above are selected, accordingly. For example, if low complexity is the main criterion, then the LMS algorithm is preferred among the adaptation algorithms.

2.1.2 Measures of efficiency

There are some operations, mandatory for successful operation of an equalizer, which have an impact on the overall system operation. These operations emanate from two important phases in the equalizer operation: (1) training phase, during which the equalizer acquires the channel characteristics using a known transmitted sequence and (2) detection phase, during which the equalizer tracks the channel characteristics assuming that the detected symbols are correct. The requirements during these two phases are quite different. During the training phase, the equalizer is required to acquire the channel quickly, since a lengthy training sequence reduces the effective information rate. In the detection phase, the equalizer is required to track the channel variations using the detected symbols. In the case of fading channels, periodic training avoids the divergence of the channel tracking algorithm due to previous erroneous decisions. Typically, the information is sent in packets, which is appended by a reasonable length known data, intended for equalizer training. Keeping the redundant bits within the transmitted packets to a minimum is also a criterion to be considered in selecting a particular equalizer to maintain high throughput efficiency.

Therefore, from the above discussion, the following are the main concerns in the selection of an equalizer for a typical portable communication system:

(1) Error rate performance (symbol / packet error rate)

(2) Computational complexity

In the following sections, some important equalization methods are outlined. These mathematical descriptions are needed to relate these equalization methods to the present research problems. The following description considers the classical and well-known equalization methods, followed by some modified methods, suggested in the recent literature, which promise to improve the system performance or reduce the computational complexity in a typical fading environment. Subsequently, in the context of problems foreseen with narrow band and wide band systems, such as computational complexity and nonlinearity, the applicability of these methods to the respective systems is discussed. Note that the type of modulation and demodulation affects the equalizer complexity and performance of the equalizer. This problem is discussed later in this chapter. In the following section, the presented equalization methods are organized based on detection scheme as symbol detection and sequence detection (though this can be organized in numerous ways).

2.2 Symbol-by-symbol detectors:

One of the well known symbol-by-symbol detectors is the decision feedback equalizer (DFE) [109]. The output of the equalizer (the input to the succeeding decision device) is a linear combination of channel output samples and of previously detected symbols. Decision feedback equalizer is non-parametric, in the sense that the adaptation process is independent of the statistics of the signal and the additive noise. Other types include the maximum a posteriori probability (MAP) symbol-by-symbol detector or Bayesian detector [35], [36], which is parametric in the sense that the adaptation process is dependent on the probability density functions of the signal and the additive noise. A DFE tries to minimize the mean squared error (MSE) between the transmitted and the recovered symbols, whereas, the Bayesian detector maximizes the a posteriori probability (equivalent to minimizing the probability of symbol error when the transmitted symbols are equally probable). There are various modifications to these classical methods, such as bi-directional equalizers, Bayesian DFEs etc., suggested in the recent literature, to improve their performance and/or decrease the computational complexity in a typical portable wireless environment. A brief description of these equalization methods is included in the following paragraphs.

Bi-directional decision feedback equalization:

Consider a simple DFE structure as shown in Fig. 2.1. It is known that a DFE requires forward taps to optimize the system performance in a non-minimum phase channel. To improve the system performance further and decrease the complexity by decreasing the number of forward taps, bi-directional equalizers [83], [84], [133] or DFE with time reversal structure [12] were proposed in the literature. In the bi-directional equalization, the received user packet is scanned twice, in the forward and reverse directions in time (Fig. 2.2). The received signal samples at time kT_s can be interpreted as follows. (Here, to simplify the nomenclature, we assume a symbol spaced forward filter in the following review.)

$$r_k = \sum_{i=0}^L c_i I_{k-i} + n_k \quad \text{for } k = 0, 1, \dots, N_B - 1 \quad (2.3)$$

where L represents the channel dispersion in symbol periods and N_B is the received packet length. r_k , I_k and n_k are the samples of the received signal, transmitted signal and the noise at time kT_s , respectively. $\{c_i\}$ represents the channel impulse response.

$$\bar{r}_k = r_{N_B-k-1} = \sum_{i=0}^L c_{L-i} I_{N_B+i-k-L-1} + n_{N_B-k-1} \quad \text{for } k = 0, 1, \dots, N_B - 1 \quad (2.4)$$

where $\bar{r}_k$ represents the received signal samples read in the reverse order. Equation (2.4) can be represented as follows.

$$\bar{r}_k = \sum_{i=0}^L \bar{c}_i \bar{I}_{k-i} + \bar{n}_k \quad \text{for } k = 0, 1, \dots, N_B - 1 \quad (2.5)$$

where $\{\bar{c}_i\} = \{c_{L-i}\}$, $\{\bar{I}_i\} = \{I_{N_B-i-L-1}\}$ and $\{\bar{n}_i\} = \{n_{N_B-i-1}\}$.

Therefore, the DFEs operating on the forward and reverse scans track the channel impulse responses given by $c(t)$ and $c^*(-t)$, respectively. If there is a deep fade in the middle of the data frame, the data before the fade is correctly recovered by the forward equalizer and the data after the fade by the reverse equalizer. However, for initial channel acquisition, two training sequences are required, one at the start and the other at the end of the user

Fig. 2. 1. Structural representation of a DFE

Fig. 2. 2. Structural representation of Bi-directional DFE

frame. The training sequence of the next frame can be used to train the reverse equalizer. The reverse equalizer can improve the system performance in a maximum phase channel with fewer forward taps. The bi-directional equalizer may yield acceptable performance with few tap coefficients for channels which are neither minimum nor maximum phase. Therefore, the complexity of the equalizer is reduced along with performance improvements.

Another problem with the classical DFE, shown in Fig. 2.1 is the error propagation during the data detection mode. The channel tracking performed using these detected symbols creates further errors for rapidly varying channels, where the adaptation step size is high. In [88], [89], an alternate approach to channel estimation was suggested. The channel impulse response between two training sequences is interpolated and used in subsequent detection. However, since we are concerned with indoor channels, where the fading rate is very low, this type of DFE may not offer any additional advantage.

MAP symbol-by-symbol detection (Bayesian DFE):

The other important type of symbol-by-symbol detector is the MAP detector. As outlined in [109], this is a probabilistic algorithm assuming the statistics of the signal and additive thermal noise are known to the receiver. Thus, unlike the DFE, the MAP detector is a parametric symbol-by-symbol estimator. In the following paragraphs this type of equalizer is briefly described.

For estimating I_k , the transmitted signal at time kT_s , the received signal samples $r_k = \{r_k, \dots, r_{k+L}\}$ are required, assuming that the channel spans LT_s in time. According to the MAP criterion, I_k is detected by maximizing the a posteriori probability function defined below.

$$p(I_k | r_k) = \frac{p(r_k | I_k) p(I_k)}{p(r_k)} \quad (2.6)$$

The denominator in the above equation is independent of I_k , and therefore can be neglected. When the transmitted symbols are equally probable, the term $p(I_k)$ can also be neglected. Therefore, I_k is estimated by maximizing the function $p(r_k | I_k)$. However, $\{r_k, \dots, r_{k+L}\}$ depend on the transmitted symbols $\{I_{k-L}, \dots, I_{k+L}\}$. Hence, I_k is estimated by maximizing the following function.

$$p(r_k|I_k) = \sum_{I_{k-L}} \sum_{I_{k-L+1}} \dots \sum_{I_{k-1}} \sum_{I_{k+1}} \dots \sum_{I_{k+L}} p(r_k|I, I_k) p(I) \quad (2.7)$$

where $I = \{I_{k-L}, \dots, I_{k-1}, I_{k+1}, I_{k+L}\}$.

The computations in the algorithm increase with increase in the channel dispersion (the span of the channel's impulse response). The computational load can be decreased significantly by assuming the past detected transmitted signals, i.e., $\{I_{k-L}, \dots, I_{k-1}\}$ are known [35], [36]. Furthermore, the term $p(I)$ in (2.7) is constant assuming that the transmitted symbols are independent and equally probable.

$$p(r_k|I_k) = G \sum_{I_{k-1}} \dots \sum_{I_{k-L}} p(r_k|I, I_k) \quad (2.8)$$

where G is a constant. Assuming the thermal noise is a zero mean Gaussian distributed random variable with variance σ_n^2 , the probability density function, $p(r_k|I, I_k)$ is given as ...

$$p(r_k|I, I_k) = \frac{1}{(2\pi\sigma_n^2)^{(L+1)/2}} \exp \left\{ - \left[\sum_{i=k}^{k+L} \frac{(r_i - \hat{r}_i)^2}{2\sigma_n^2} \right] \right\} \quad (2.9)$$

where $\hat{r}_k = f(c_{k-1}, I_k)$ is the output of the estimated channel. c_k represents the channel impulse response at time kT_s and $I_k = \{I_k, I_{k-1}, \dots, I_{k-L}\}$.

Therefore, for MAP symbol-by-symbol detection, the metric, $\Gamma(I_k)$, shown below is calculated for all the possible M symbols and the one with maximum metric is chosen as the most probable transmitted symbol. Notice that for evaluating the metric corresponding to each I_k , there are M^L summations.

$$\Gamma(I_k) = \sum_{I_{k-1}} \dots \sum_{I_{k-L}} p(r_k|I, I_k) \quad (2.10)$$

For a soft output, assuming MSE cost function the MAP estimate can be shown to be the following, when the transmitted symbols are equally probable.

$$\hat{I}_k = \frac{\sum_{I_k} I_k p(I_k | r_k)}{\sum_{I_k} p(I_k | r_k)} = \frac{\sum_{I_k} I_k p(r_k | I_k)}{\sum_{I_k} p(r_k | I_k)} \quad (2.11)$$

This equalization strategy is referred to as Bayesian DFE (BayDFE) as in [35], [36], since the estimate in (2.11) is a soft Bayesian estimate. In the case of fading channels, where the channel changes, the channel at time kT_s is updated. Assuming linear ISI, using, for example, the LMS algorithm, the channel impulse response is updated as follows.

$$c_k = c_{k-1} + \mu e_k I_k \quad (2.12)$$

where μ is the step size parameter and $e_k = r_k - \hat{r}_k$ is the error between the observed and estimated channel outputs.

The structural representation of Bayesian DFE is given in Fig. 2.3. The exponentiation operation can be performed using table look-up method. As shown, the channel status is updated indirectly. The previously detected symbols are used for this purpose. Notice that, the summation of exponential functions as indicated by (2.10) are obtained iteratively. Thus, the exponential terms indicated in the figure should be obtained at M^L times faster than the symbol rate. Therefore, with the suggested architecture, the speed of likelihood metric computation increases with an increase in the channel dispersion.

2.3 Sequence estimation:

The other class of equalization method is the maximum likelihood sequence estimation (MLSE) [57], [109], [138]. However, a problem with MLSE is the channel tracking. As mentioned in Section 2.1, there is an inherent delay in data detection. The classical method for channel tracking uses tentative decisions from the MLSE, which are not always reliable. Furthermore, there will be a delay in the estimated channel. In the following, a brief mathematical description of MLSE with conventional channel tracking method is given. Later, a combined channel estimation and sequence detection method introduced by Seshadri is given [119], [122]. Here, the analysis is limited to a symbol spaced MLSE. A fractionally spaced MLSE is proposed in the literature [86] for further improvements in system performance. However, these techniques involve significant increase in computational load.

Fig. 2. 3. Structural representation of Bayesian DFE

2.3.1 MLSE with channel tracking using the tentative decisions

Given a sequence of channel output samples, of length N_B , the maximum likelihood estimate of the transmitted sequence is found by maximizing the metric over all the possible transmitted sequences [109]. The number of possible sequences is M^{N_B} when the size of symbol alphabet is M . If the channel status is also not known, the channel is estimated first. This is an exhaustive search process which turns out to be impractical in situations where the sequence is long. A simple and suboptimal method to perform maximum likelihood detection is to use the Viterbi algorithm.

First let us assume that the combined response of the transmit filter, the channel and the receive filter, which spans L symbol periods, c , is known to the receiver. Then the number of states of the MLSE trellis is $\rho = M^L$. Each state of the MLSE is represented by a symbol vector of length L as shown below.

$$s_{k-1}^{(l)} = \{I_{k-1}^{(l)}, I_{k-2}^{(l)}, \dots, I_{k-L}^{(l)}\} \quad (2.13)$$

where the subscript indicates the time index and superscript indicates the number of the state. Unless otherwise stated the same notation is followed throughout this section. Here l varies from 1 to ρ . The transmitted signals are represented by I .

If r_k is the sample of the received signal at time kT_s , then for all the possible state transitions from $s_{k-1}^{(l)}$, the l^{th} state at time $k-1$, to $s_k^{(m)}$, the m^{th} state at time k , the errors and metrics are calculated as follows.

$$\xi_k^{(l \rightarrow m)} = r_k - f(S^{(l \rightarrow m)}, c_{k-1}) \dots \text{for } l = l_1, l_2, \dots, l_M \quad (2.14)$$

$$\gamma_k^{(l \rightarrow m)} = |\xi_k^{(l \rightarrow m)}|^2 \dots \text{for } l = l_1, l_2, \dots, l_M \quad (2.15)$$

where $\xi_k^{(l \rightarrow m)}$ and $\gamma_k^{(l \rightarrow m)}$ respectively represent the error and the metric associated with the transition from $s_{k-1}^{(l)}$, the l^{th} state at time instant $k-1$, to $s_k^{(m)}$, the m^{th} state at time instant k . $s_{k-1}^{(l_1)}, s_{k-1}^{(l_2)}, \dots, s_{k-1}^{(l_M)}$ represent the only states that reach $s_k^{(m)}$ upon the arrival of the k^{th} symbol¹. $c_{k-1} = \{c_0, c_1, \dots, c_L\}_{k-1}$ represents the combined channel

impulse response at time $k-1$ and $S^{(l \rightarrow m)} = \{s_{k-1}^{(l)} \cup s_k^{(m)}\}$ is the symbol vector and $f(S^{(l \rightarrow m)}, c_{k-1})$ represents the estimated channel output. $A \cup B$ represents the union operation on sets A and B . When the ISI is expected to be linear, then $f(S^{(l \rightarrow m)}, c_{k-1})$ can be expressed as follows.

$$f(S^{(l \rightarrow m)}, c_{k-1}) = c_{k-1} \bullet S^{(l \rightarrow m)} \quad (2.16)$$

where $A \bullet B$ represents the dot product (or inner product) of vectors A and B .

For all these M possible transitions, the cumulative metrics are calculated and minimized over all the transitions.

$$\Gamma_k^{(l \rightarrow m)} = \Gamma_{k-1}^{(l)} + \gamma_k^{(l \rightarrow m)} \dots \text{for } l = l_1, l_2, \dots, l_M \quad (2.17)$$

$$\Gamma_k^{(m)} = \min_{s_{k-1}^{(l)}} \Gamma_k^{(l \rightarrow m)} = \Gamma_{k-1}^{(l_o)} + \gamma_k^{(l_o \rightarrow m)} \quad (2.18)$$

The surviving path corresponding to $s_{k-1}^{(l_o)}$, which gives the least cumulative metric $\Gamma_k^{(l_o)}$, of the previous M states, is updated as follows.

$$\zeta_k^{(m)} = \{\zeta_{k-1}^{(l_o)} \cup s_k^{(m)}\} \quad (2.19)$$

where $\zeta_{k-1}^{(l_o)}$ represents the survivor sequence corresponding to $s_{k-1}^{(l_o)}$. The procedure outlined above is repeated for $m=1, \dots, p$ to get all the p survivors at time k . This completes one step of the Viterbi algorithm. The sequence detection process is depicted in Fig. 2.4. Notice that it is assumed that the channel is known at the receiver.

Channel update

After one step of the algorithm, the channel vector is updated using tentative decisions on the sequence. The tentative decisions are extracted by performing trellis

1. The number of possible transitions for each state in an MLSE is limited to M , since there are only M possible symbols.

decoding. Assuming a decoding depth of d symbols, the decisions obtained are...

$$\hat{\mathbf{l}}_{k-d} = \{\hat{l}_{k-d}, \hat{l}_{k-d-1}, \dots, \hat{l}_{k-d-L}\} \quad (2.20)$$

where $\hat{\mathbf{l}}$ represents the vector consisting of the recovered transmitted signals. Assuming that the ISI is linear in nature, using the classical LMS algorithm, the channel vector can be updated as follows.

$$e_{k-d} = r_{k-d} - \hat{\mathbf{c}}_{k-1} \bullet \hat{\mathbf{l}}_{k-d} \quad (2.21)$$

$$\hat{\mathbf{c}}_k = \hat{\mathbf{c}}_{k-1} + \mu e_{k-d} (\hat{\mathbf{l}}_{k-d})^* \quad (2.22)$$

where e_{k-d} is the error between the actual and the estimated channel outputs. The superscript $(*)$ indicates complex conjugate operation.

The above mentioned procedure is depicted in Fig. 2.5. As it can be observed, there is a delay of d in the channel estimation. As the decision depth is increased the decisions are increasingly reliable, but at the same time the channel estimate is delayed. A compromise between the two factors, i.e. reliability and delay, depending on how fast the channel is changing, is crucial for the success of this method.

2.3.2 Combined channel estimation and sequence detection algorithm (Per-survivor Processing) [122]

In the combined channel tracking and sequence estimation algorithm, the likelihood metric is calculated over all the possible sequences, assuming one estimated channel vector for each state. Channel tracking is accomplished by updating each channel vector along the surviving path corresponding to the associated state. Equation (2.14) is modified as follows.

$$\xi_k^{(l \rightarrow m)} = r_k - f^{(l)}(S^{(l \rightarrow m)}, \hat{\mathbf{c}}_{k-1}^{(l)}) \dots \text{for } l = l_1, l_2, \dots, l_M \quad (2.23)$$

and as defined previously $S^{(l \rightarrow m)} = \{s_{k-1}^{(l)} \cup s_k^{(m)}\}$ is the symbol vector. Notice the superscript l for $f(\dots)$ indicating the state dependency of the function. $\hat{\mathbf{c}}_{k-1}^{(l)} = \{c_0^{(l)}, c_1^{(l)}, \dots, c_L^{(l)}\}_{k-1}$ represents the channel impulse response corresponding to $s_{k-1}^{(l)}$, which is time dependent as well as state dependent.

Fig. 2. 4. Channel tracking using known data

Fig. 2. 5. Channel tracking using tentative decisions

The cumulative metric calculation and the optimum state selection are identical to (2.14) - (2.19), given previously. The channel update, using the LMS algorithm, is as follows.

$$\hat{c}_k^{(m)} = \hat{c}_{k-1}^{(l_0)} + \mu \xi_k^{(l_0 \rightarrow m)} (S^{(l_0 \rightarrow m)})^* \quad (2.24)$$

For one step of the algorithm, the above steps are to be repeated for all the p survivors ($m=1, \dots, p$). As can be seen from the above equations the error between the desired output and the obtained output need not be calculated separately, as in the conventional channel tracking method. The error obtained in the metric calculations is used directly as shown in Fig. 2.6.

2.3.3 Generalized PSP

For a better channel tracking K survivors (instead of one survivor) are retained per state. Now the number of channel vectors become $v = K M^L$ (more than p , the number of states. This generalized PSP is represented as PSP - K , meaning K survivors per each state [122].

2.4 Applicability to narrow band and wide band systems

In this section, the feasibility of applying these equalization methods to narrow band and wide band systems is discussed considering the associated problems mentioned in Chapter 1. For narrow band systems, all the above presented equalization methods can be implemented using existing technology. Some of the previous references [91], [96] strengthen this statement. However, the complexity and performance of these equalizers differ.

In the case of wide band systems, as mentioned in Chapter 1, the receiver may be required to combat nonlinear distortion, which introduces nonlinear ISI. Therefore, the output of the channel may not be expressed as a linear combination of the transmitted symbols as in (2.3) or as in (2.16). A nonlinear DFE can be devised as outlined in [52]. However, the number of coefficients to adapt is exponentially proportional to the order of the nonlinearity. In the case of Bayesian DFE and MLSE, since the channel tracking is performed indirectly, the nonlinear relation between the channel output and the input can be learned using memory based methods. In the following section, a brief description of equalizers for combatting nonlinear ISI, including memory based equalizers, is given.

Fig. 2. 6. Channel tracking using PSP

2.5 Equalization for combatting nonlinear ISI

In (2.3) and (2.16), the estimated channel output is expressed as a linear combination of data symbols, which may not be true in a nonlinear ISI channel. There were some reports on equalization for nonlinear ISI channels in the earlier literature [19], [52], [97], [101], [117], [147]. An optimal ML receiver is derived for nonlinear channels in [97], [147]. The number of matched filters required at the receiver increases exponentially with the order of nonlinearity. This receiver structure is very complex for the work under consideration.

One well known, conceptually simple and efficient nonlinear equalization scheme is called random access memory (RAM) or memory based equalizer (Fig. 2.7) [148], [149]. In this scheme for each known transmitted signal vector at time k , I_k , the output of the channel is stored in a RAM using the input vector as the address. These stored values are updated using the LMS algorithm over time. If the channel dispersion is over L symbol intervals, then the RAM will have M^{L+1} addresses, where M is the size of the symbol alphabet. The following equations represent the detailed procedure.

$$e_k = r_k - f_k(I_k) \quad (2.25)$$

where r_k represents the output of the unknown nonlinear channel for input vector I_k and $f_k(\dots)$ is the expected channel nonlinearity involving the information symbols at time kT_s . e_k is the error between the observed and the estimated channel outputs. The estimated value is adapted using the LMS algorithm as follows.

$$f_{k+1}(I_k) = f_k(I_k) + \mu e_k \quad (2.26)$$

where μ is the step size.

As expected this type of equalizer can learn any order of nonlinearity. A DFE or MLSE can be configured using this *table look-up* method. In the case of a MLSE, assuming the thermal noise at the input of the receiver is Gaussian distributed, the cumulative branch metrics at time k , Γ_k , are calculated as shown in Fig. 2.8.

$$\Gamma_k^{(l \rightarrow m)} = \Gamma_{k-1}^{(l)} + |r_k - f_k(S^{(l \rightarrow m)})|^2 \quad \text{for } l = 1, \dots, l_M \quad (2.27)$$

Fig. 2. 7. Memory based equalization technique

Fig. 2. 8. Block schematic of RAM-MLSE

where, $S^{(l \rightarrow m)} = \{s_{k-1}^{(l)} \cup s_k^{(m)}\}$ and l_M represents the number of transitions from each previous state. Notice that in the detection mode, $S^{(l \rightarrow m)}$ replaces the address. Since there are l_M transitions from each state, the vector distance for all l_M specified addresses are calculated and minimum distance among those values is used in updating the cumulative metric. The rest of the approach follows the well known Viterbi algorithm. The contents of the RAM can also be updated in the detection mode by using the minimum distance and a smaller value of the step size.

Fig. 2.9 shows the memory based DFE configuration. The output of the DFE, y , can be expressed as follows.

$$y_k = c_k^* \bullet r_k + f_k(\hat{l}_{k-1}) \quad (2.28)$$

Where c_k and $r_k = \{r_k, r_{k-1}, \dots, r_{k-N_f+1}\}$ are the weight vector of the forward filter and the received signal sample vector of dimension N_f , respectively at time kT_s . $\hat{l}_{k-1} = \{\hat{l}_{k-1}, \hat{l}_{k-2}, \dots, \hat{l}_{k-N_b+1}\}$ is the vector consisting of the past N_b recovered transmitted signals, where N_f and N_b are the forward and backward taps of the DFE, respectively.

It is evident that in the DFE the nonlinearity compensation is limited to previous N_b symbols. If the nonlinearity also includes the present symbol, then a modified form of DFE called multiplicative RAM-DFE [117] can be used. The difficulty with all these structures is the learning time involved. There are some methods suggested to improve the speed of convergence of this type of equalizer [148]. This method consists of a linear transversal filter in parallel with a memory based equalizer. As expected, the linear part of the ISI will be learned by the transversal equalizer very fast and the nonlinear part of the ISI by the RAM relatively slowly. However, this method will be successful when the ISI is predominantly linear. All the memory based equalization methods are suitable for channels which vary very slowly. Similar structures were reported in [19], [101]. The performance of these equalizers was evaluated through computer simulations in [19], [101] considering a linear modulation and approximating the channel response by discrete time Volterra series.

Another alternative to tackle nonlinear ISI is use of multilayer perceptron (MLP) equalizers [61], [128]. Depending on the order of nonlinearity of the channel, the number of hidden layers in the MLP equalizer are increased. However, these equalizers have

Fig. 2. 9. Block schematic of RAM-DFE

disadvantages, such as longer acquisition time and slow tracking and also very high complexity. There are some methods to improve the acquisition time of these equalizers[113].

Notice that to this point the type of modulation and demodulation methods and their effect on equalization are not discussed. For example, if a nonlinear modulation method such as FSK is used, in all the equalization methods presented above the MSE or metric calculations using the symbol will be replaced by the corresponding MSE or metric calculations using a low pass representation of the modulated waveform. Furthermore, when the receiver uses noncoherent demodulation methods, which are usually nonlinear methods, the implications on equalization are to be investigated further. In the following section, we present combined demodulation and equalization methods. The methods presented are limited to PSK and FSK signals.

2.6 Combined demodulation and equalization

In the previous sections, various equalization strategies are presented without imposing any constraints on the type of demodulation. In fact, the complexity involved in the equalization process is related to the type of modulation and demodulation method used. When the modulation is nonlinear in nature, such as in FSK, the amount of processing required is significantly increased as can be seen later in this section.

Phase modulated signals:

For PSK modulated signals the transmitted signal phase is linearly related to the information symbol. Hence, the low pass equivalent of the PSK modulated waveform is a constant over the symbol duration. Differentially coded PSK is generally attractive because these signals can be recovered coherently or noncoherently depending on the requirement. For differential PSK modulation schemes, such as DQPSK- $\pi/4$ [63], [38], a combination of demodulation and equalization can be performed either coherently or noncoherently. In these methods, the demodulation forms an integral part of the equalization process. The equalization strategy with coherent demodulation is shown schematically in Fig. 2.10. As shown, the equalization is performed before demodulating the signal. Some computer simulation results for this strategy were presented in [37], [100], [123]. The other strategy; noncoherent demodulation - equalization, as suggested in [62], is shown in Fig. 2.11².

2. In the strict sense, these two methods are called differentially coherent detection. The reason for this is that the equalizer, which is acting on the sampled complex envelope of the received signal, performs carrier phase recovery also. Though differential detection is a noncoherent demodulation method, since carrier phase information has already been used in the process of equalization, this method can be interpreted as differentially coherent detection.

Fig. 2. 10. Coherent detection and equalization of DQPSK signals

Fig. 2. 11. Noncoherent demodulation and equalization of DQPSK signals

Computer simulation results on this issue can be found in [73]. In this scheme, care has been taken so that the nonlinearities of the differential detector do not affect the linear estimation process.

In the above architectures, assuming that the equalization is performed at base band, the equalization involves complex arithmetic, constituting a significant part of the receiver complexity. When the demodulator precedes the equalization, then the equalization involves only real arithmetic. When demodulation is performed coherently (prior to equalization), then demodulation requires additional carrier phase recovery circuitry. This function is done by equalizer in the earlier structures. Therefore, noncoherent demodulation is preferred for low complexity. However, if noncoherent demodulation precedes the equalization, then there will be significant nonlinearity at the input of the equalizer. For differential QPSK signals, a nonlinear equalizer structure, based on the technique proposed in [52], is proposed in [93], [94], [95]. Another innovative, low complexity nonlinear equalization strategy is suggested by Ariyavisitakul in [11]. He suggested a simple suboptimal nonlinear phase equalizer, which acts on the phase information of the received signal. This equalizer is more suitable for a typical indoor environment, where the channel variations are very slow. The following equations describe the theory of phase equalization. The symbol rate sampled received signal, r_n , corrupted by ISI and thermal noise, after passing through the matched filter at the receiver can be represented as follows.

$$r_i = \sum_{k=0}^L I_{i-k} c_{i,k} + n_i \quad (2.29)$$

where $c_i = \{c_{i,k}, k = 0, 1, \dots, L\}$ represents the symbol rate sampled combined impulse response of the channel and the transmit and receive filters and n_i represents the noise samples at time iT_s . I_k is the information modulated signal at time k , which can be represented as $e^{j\varphi_k}$ where $\varphi_k - \varphi_{k-1} = m(\pi/2) + \vartheta$, $\vartheta = \pi/4$ for $\pi/4$ -shift DQPSK, $m = 0, 1, 2, \dots, M-1$, (M is the size of the symbol alphabet; $M = 4$ for DQPSK). Here it is assumed that $c_{i,k}$ is time limited i.e. $c_{i,k} = 0$ for $k < 0$ and $k > L$. The proposed detector of [11] operates on signal phase only. The phase of the signal represented by (2.29), β_i , can be represented as follows.

$$\beta_i = \text{Arg} \left[\frac{r_i}{|r_i|} \right] \quad (2.30)$$

β_i , the phase of received signals at time iT_s , which can be obtained using a limiter-discriminator type of receiver, can be expressed in signal phase at time iT_s as follows.

$$\beta_i = \varphi_i + \theta_i \quad (2.31)$$

where θ_i is defined as the excess phase term.

Notice that in (2.30), when $|r_i|$ becomes small in the process of channel deep fade, the limiter output becomes very noisy, and uncertainty prevails during a deep fade. β_i can be expressed as follows.

$$\beta_i = \varphi_i + \text{Arg} \left[\sum_{k=0}^L I_{i-k} c_{i,k} e^{-j\varphi_i} + n_i e^{-j\varphi_i} \right] \quad (2.32)$$

The excess phase term in (2.32) can further be divided into accumulated phase, $\hat{\theta}_i(I_i, c_i)$, and hypothesized phase error, $\Delta\theta_i$, which are defined as follows.

$$\beta_i = \varphi_i + \hat{\theta}_i(I_i, c_i) + \Delta\theta_i \quad (2.33)$$

where ...

$$\hat{\theta}_i(I_i, c_i) = \text{Arg} \left[\sum_{k=0}^L I_{i-k} c_{i,k} e^{-j\varphi_i} \right] \quad (2.34)$$

$$\Delta\theta_i = \text{Arg} \left[1 + \frac{n_i e^{-j\varphi_i}}{\sum_{k=0}^L I_{i-k} c_{i,k} e^{-j\varphi_i}} \right] \quad (2.35)$$

$\hat{\theta}_i(I_i, c_i)$ depends on the present and the past L symbols represented by $I_i = \{I_i, \dots, I_{i-L}\}$ and the channel, c_i , at time iT_s . Notice that in the expression for β_i , $\Delta\theta_i$ is the only random variable, when the channel is assumed fixed. The probability density function of the phase error, $\Delta\theta_i$, can be approximated as follows [11], [109].

$$p(\Delta\theta_i) = \sqrt{\frac{S_i}{\pi}} \exp(-2\gamma_i [1 - \cos(\Delta\theta_i)]) \quad (2.36)$$

where S_i is the instantaneous signal to noise ratio (SNR), defined as ...

$$S_i = \frac{\left| \sum_{k=0}^L c_k I_{i-k} e^{-j\phi_i} \right|^2}{2\sigma_n^2} \quad (2.37)$$

and σ_n^2 is the noise power spectral density.

Hence, the power spectral density of β_i is given as follows.

$$p(\beta_i) = \sqrt{\frac{S_i}{\pi}} \exp(-2\gamma_i [1 - \cos(\beta_i - \hat{\theta}(I_i, c_i))]) \quad (2.38)$$

Assuming that the instantaneous signal-to-noise ratio (SNR), S_i , is assumed be unknown at the receiver, then a sub-optimal log-likelihood function over a sequence of N_B symbols, Γ_L , can be defined as follows:

$$\Gamma_L = -\sum_{i=0}^{N_B} 2(1 - \cos(\Delta\theta_i)) = -\sum_{i=0}^{N_B} (2 - e^{j(\beta_i - \phi_i - \hat{\theta}(I_i, c_i))} - e^{-j(\beta_i - \phi_i - \hat{\theta}(I_i, c_i))}) \quad (2.39)$$

which can be simplified further as follows...

$$\Gamma_L = \sum_{i=0}^{N_B} \left| e^{-j\frac{(\beta_i - \hat{\theta}(I_i, c_i))}{2}} e^{-j\frac{\phi_i}{2}} [e^{j(\beta_i - \hat{\theta}(I_i, c_i))} - I_i] \right|^2 \quad (2.40)$$

or...

$$\Gamma_L = \sum_{i=0}^{N_B} \left| e^{j(\beta_i - \hat{\theta}(I_i, c_i))} - I_i \right|^2 \quad (2.41)$$

Using this likelihood function, a sequence estimator can be suggested using the Viterbi algorithm. The branch metrics are calculated as shown in Fig. 2.12. A DFE structure can also be suggested as in Fig. 2.13. These receivers are expected to be robust against device and channel nonlinearities also.

In both the receivers in Fig. 2.12 and 2.13, the accumulated phase term is iteratively learned during the training sequence using the table look-up method as outlined in Section 2.5. Since the number of possible sequences are M^{L+1} in number, all these sequences should occur at least once during the training sequence. Therefore, the length of the training should be at least $L+1+M^{L+1}$. When the channel dispersion is more than a couple of symbol durations, the size of training sequence may become impractical. When the channel is varying, the accumulated phase term is adapted during the detection mode using a smaller step size.

Frequency modulated signals

FSK signals can be equalized similar to the methods proposed for DQPSK as in Fig. 2.10 - Fig. 2.11. However, due to the nonlinear relationship between the information symbols and the modulated waveform, the low pass representation of the modulated signal is not a constant over the symbol duration as in the case of PSK modulation. Therefore, the signal has to be sampled at higher than the symbol rate. The sampling rate, R , depends on the modulation index of the modulation scheme. The modulation index tends to increase the spacing between the frequency of the symbol waveforms thus by increasing the maximum frequency of the signal, therefore the sampling frequency. As shown in Fig. 2.14, the reconstructed signal at the output of the equalizer, which should also operate at rate R , is mapped to the corresponding symbol alphabet. Note that a slicer or a decision circuit as shown in Fig. 2.10 cannot be used as demodulator. Instead, the mapping can be performed by measuring Euclidean distance between the signal waveforms. This process is further complicated for continuous phase FSK (CPFSK) signals because of the phase memory involved from past symbols [64].

Similar to the receiver depicted in Fig. 2.12 and 2.13, if the FSK demodulator precedes the equalization operation, then the sampling rate can be reduced to the symbol rate. However, due to the nonlinear relationship between the information symbols and the modulation waveforms, the ISI which is linear prior to the demodulation, becomes nonlinear. The memory based equalization methods, similar to the ones outlined in Section 2.5, may be devised to combat nonlinear ISI. However, the probability density function of the noncoherently demodulated signals is required to derive this type of receiver.

Fig. 2. 12. Block schematic of the phase equalizer (MLSE)

Fig. 2. 13. Block schematic of the phase equalizer (DFE)

Fig. 2. 14. Coherent demodulation and equalization of FSK signals

2.7 Equalization for the narrow band and wide band systems

For narrow band systems, as discussed in Chapter 1, all the outlined equalization methods are applicable. However, since PSK signals do not require high sampling speeds and linear amplifiers can be implemented at the transmitted frequencies up to ~ 2 GHz, PSK modulation may be preferred. The complexity and performance of the narrow band systems using some of the above mentioned equalization methods is the main topic of Chapter 3. An additional requirement of all these equalizers is that the same system should work in a typical outdoor environment, where severe Doppler frequency is also a concern.

For wide band systems, constant envelope modulation methods, such as CPFSK, whose detectability is not lost by hard limiting, are preferred. As mentioned at the end of Section 2. 6, optimum detectors for demodulated ISI corrupted signals is considered in Chapter 4. The emphasis is given on those methods which are simple and less complex due to extreme data rates which are considered in this work.

Chapter 3

Equalization for Narrow Band Indoor Wireless Systems

Recent advances in digital cellular radio show that the combination of adaptive equalization (temporal filtering) and space diversity combining (spatial filtering) at the receiver offers improvements in system performance [17], [53], [54] by combatting both frequency selective and frequency non-selective fading. Various methods of combined adaptive equalization and diversity reception, presented in the literature, show that the equalizers are to be implemented for each of the antenna branches fully or partially to achieve the best system performance. This is true for minimum mean squared error (MSE) equalizers, such as decision feed back equalizers (DFE) [17], as well as for maximum likelihood sequence estimation (MLSE) methods [86]. However, the combination of equalization and diversity combining may not be cost effective to implement at the portable for the following reasons: (1) the requirement of multiple antennas and the duplication of the RF front end and IF signal processing equipment makes the receiver more complex and increases power consumption and (2) adaptive equalization does not give any additional advantage when the channel exhibits frequency non-selective fading, which is true in the indoor environment.

Diversity reception at the portable is necessary to improve the system performance for frequency non-selective fading channels [125], [134], [140], [144]. Alternatively diversity transmission methods can be used, [22], [23], [25], [27], [69], [99], [104], [120], [121], [145], [146]. Typically, diversity transmission would be implemented for forward link channels (base station to mobiles), since the cost and the size of extra antennas can more readily be absorbed by the base station. In both methods, diversity combining and diversity transmission, uncorrelated signals at the antennas are assumed. In an indoor environment, where the base station is mounted on a ceiling or wall, obtaining independent fading at the base station is easy relative to the outdoor environment. Therefore, the main issue is how well diversity transmission methods perform when compared to diversity reception. This question will be answered in this chapter.

In Section 3.1, various diversity transmission methods are reviewed. The advantages and disadvantages of these diversity transmission methods are discussed. One of the diversity transmission techniques, known as delayed diversity transmission, is discussed in Section 3.2 and found to be the best choice based on hardware requirements. In Section 3.3, various parameters that influence the performance of a system with delayed diversity transmission from the base station and an equalizer at the portable are listed. In Section 3.4, the effect of these parameters is investigated through computer simulations.

Performance in a practical scenario is studied by considering the North American digital cellular standard architecture in the presence of Doppler frequency shifts. Here, we assume a portable mobility of 2 m/s which corresponds to a maximum Doppler frequency shift of 6 Hz at a carrier frequency of 900 MHz. An in-depth comparison of various equalizers that are discussed in Chapter 2 is presented in Section 3.5. In Section 3.6, different equalizers are compared based on their performance and computational complexity. The best equalization methods resulting from this comparison are further evaluated in an outdoor environment in Section 3.7, since one of the objectives is to employ existing digital cellular radio portable handsets in an indoor environment. The evaluation is in terms of the required SNR per information bit to attain an acceptable error rate. Based on performance in an outdoor environment, a system with the best performance and the lowest complexity is selected for the subsequent system evaluations. In Sections 3.5 to 3.7 we consider two systems, one without forward error control (FEC) coding and the other with FEC coding, in order to investigate the coding gain on the system performance. In Section 3.8, the effect of finite depth interleaving and coding are investigated at low Doppler frequencies below 6 Hz, i.e. at portable speeds of less than 2 m/s. In Section 3.9, the performance of equalization and diversity transmission is compared with diversity reception employing optimal combining. In Section 3.10, we summarize the results.

3.1 Diversity transmission as an alternative to diversity reception

The basic concept in diversity transmission is to place multiple antennas at the base station instead of at the portable. The information transmission from the base station and the subsequent recovery at the portable are configured to obtain diversity advantage at the portable. The basic configuration for antenna diversity combining and antenna diversity transmission are depicted in Fig. 3.1 and 3.2 respectively. For diversity transmission, the combination of *transmitter signal processing* and *receiver signal processing* is appropriately designed to ensure the advantage of diversity gain at the receiver. Here, as in the case of diversity reception, the antennas are assumed to be placed sufficiently far apart so that the transmitted signals from the antennas follow independent paths. The efficiency of the diversity transmission depends on the type of signal processing at the transmitter and the corresponding type of signal processing at the receiver. Since the purpose of this research is to find a low complexity solution, the two basic criteria, as stated in Chapter 2, to select an optimum combination of transmitter signal processing and receiver signal processing are: (1) system performance (2) hardware complexity and power consumption and (3) the effective information rate.

Diversity transmission methods, reported in the literature, can be classified broadly into three categories [69]–[71], [120], [121], [146]. In the first category, feedback

Fig. 3. 1. Portable with antenna diversity combining

Fig. 3. 2. Antenna diversity transmission

information from the receiver is used to configure the transmitter. A scheme proposed in [69], as depicted in Fig. 3. 3, is applicable to time division duplex (TDD) systems. In the reverse transmission (from the portable to the base station), the weights of the antennas are adapted according to a known information sequence and, in the forward transmission, the information is transmitted using the complex conjugate of the same weights. In this method, no special signal processor is required at the portable. The main assumption made here is that the interval between bursts is so short that there is no significant change in the channel characteristics from the time of reverse transmission to the forward transmission. On the other hand, the burst on the reverse transmission should be long enough for the base station to learn the channel properties. A similar method proposed in [143] can be used for frequency division duplex (FDD) systems.

In the second category, feed-forward information is used to train the receiver. A method proposed in [146] is to transmit the same information through multiple antennas simultaneously from the base station but with different transmission delays. Fig. 3. 4 shows the system block diagram. Here, the portable receives multiple replicas of the same signal with the information dispersed in time. Therefore, the situation is similar to transmission in a frequency selective fading environment. As is well known, an equalizer improves the system performance in this situation. Using a feed forward training sequence, the equalizer parameters are initially adapted to acquire the channel information. Referring to Fig. 3. 4, the transmitter signal processing consists of different delay elements, incorporated in the signal path to the antennas, and the receiver signal processing consists of an equalizer. This method is applicable to both TDD and FDD systems.

In the third category, channel coding and multiple antennas are used to obtain a diversity advantage. In a method proposed in [71], [120], the information bits are block encoded and each of the encoded symbols are successively transmitted from the base station through a different antenna. So each of the coded symbols undergoes independent fading and the advantage of infinite interleaving is obtained. The scheme is depicted in Fig. 3. 5. At the receiver these independently faded coded symbols are processed using a maximum likelihood decoder. Here the transmitter signal processing consists of a block encoder at the base band and a RF switch, which connects the signal to one of the antennas during different symbol periods. The receiver signal processing consists of a maximum likelihood detector. However, antenna switching produces periodic abrupt phase changes in the received signal, thereby causing periodic errors. In order to avoid this problem, a phase sweeping transmitter diversity method is proposed in [71]. This method can be explained as follows assuming two transmitting antennas at the base station. In phase sweeping diversity, fast fading effects can be created by transmitting the information modulated carrier simultaneously from both the antennas. The carrier transmitted from one

Fig. 3. 3. Category - I: Receiver feedback diversity

Fig. 3. 4. Category - II: Delayed diversity transmission

Fig. 3. 5. Category - III: Coded diversity transmission

of the antennas is phase modulated independently of the original signal modulation as shown below.

$$r_1(t) = R_1(t) \cos [\omega_c t + \Theta_s(t) + \Theta_1(t) + \kappa(t)] \quad (3.1)$$

$$r_2(t) = R_2(t) \cos [\omega_c t + \Theta_s(t) + \Theta_2(t)] \quad (3.2)$$

Where R_i and Θ_i are the random envelope and the random phase due to fading. $\Theta_s(t)$ is the information bearing phase and $\kappa(t)$ is the swept phase. Notice that we are utilizing the decorrelation property in the frequency domain also by varying the signal phase of one antenna. Further details can be found in [71].

Therefore, by using the phase sweeping transmitter diversity, the advantage of coded diversity can be obtained at the receiver with limited interleaving [71] even at low Doppler frequencies. For this scheme, the channel information is not necessary.

Diversity transmission methods are mainly intended to simplify the portable receiver by eliminating the multiple antennas and the duplication of signal processing equipment. The advantages and disadvantages of the diversity transmission schemes, presented previously, with respect to spatial diversity combining are described below. Complexity is defined in terms of the hardware requirements at the base station as well as at the portable receiver.

In category I, the base station transmitter should be equipped with a RF processor in each antenna branch. Therefore, compared to diversity reception at the receiver, this category has the same hardware complexity. In category II, the base station requires additional components in the form of delay elements and the receiver requires an equalizer. Though the delay elements are not difficult to realize, the effect of these delays on the system performance is a parameter to be investigated. In category III, the base station is equipped with a block encoder and a RF switch or modulators on each branch and the portable requires a ML decoder.

Category I & II require less hardware compared to category III. Between category I & II, the former is very sensitive to the presence of Doppler frequency shift and timing control.

Therefore, even though all these schemes seem to be promising in rendering diversity gains comparable to diversity reception on the basis of hardware complexity, the second category seems to be preferable. Furthermore, since the existing digital cellular

standards already make provision for an equalizer at the receiver - no additional hardware is required at the portable.

A thorough system performance evaluation of this scheme, named delayed diversity transmission (DDT), is presented in the sections to follow.

3.2 Delayed diversity transmission

As shown in Fig. 3.4, the base station transmitter employs diversity transmission and the portable receiver uses an equalizer. The transmitting antennas at the base station are assumed to be sufficiently far apart so that the transmitted signals from the antennas undergo independent fading. The information modulated carrier is divided into N_a parts in amplitude and fed to N_a antennas. A transmission delay of $(i-1)\tau_D$ is incorporated for the i^{th} antenna¹ (an equal incremental time delay between the antennas is assumed). Assuming that each of these signals follow independent paths, the complex envelope of received signal at the portable can be expressed as...

$$r(t) = \sum_{i=1}^{N_a} C_i(t) z(t - [i-1]\tau_D) + n(t) \quad (3.3)$$

where $z(t)$ and $C_i(t)$ represents the low-pass equivalents of the information signal and the time varying channel gain for the i^{th} antenna, respectively. $n(t)$ represents the complex envelope of the zero mean additive white Gaussian noise (AWGN) with double-sided power spectral density $N_0/2$. $n(t)$ satisfies the following properties:

$$E\{n(t)\} = 0 \quad (3.4)$$

$$E\{n(t)n(u)\} = 0 \quad (3.5)$$

$$E\{n(t)n^*(u)\} = N_0\delta(t-u) \quad (3.6)$$

where the superscript $(*)$ indicates complex conjugate operation. $\delta(\cdot)$ is the Dirac delta or the unit impulse function. The complex path gains for the antennas $C_i(t)$ $i = 1, \dots, N_a$ are independent zero mean complex Gaussian process whose real and imaginary parts are independent and identically distributed.

1. The delays could be implemented digitally if the transmitter units are implemented with digital signal processing.

$C_i(t)$ satisfies the following properties:

$$E[C_i(t)] = 0 \quad \text{for} \quad i = 1, 2, \dots, N_a \quad (3.7)$$

$$E[C_i(t) C_j^*(t - \tau)] = K_C(\tau) \delta(i - j) \quad \text{for} \quad i, j = 1, 2, \dots, N_a \quad (3.8)$$

$$E[C_i(t) C_j(t - \tau)] = 0 \quad \text{for} \quad i, j = 1, 2, \dots, N_a \quad (3.9)$$

where $K_C(\tau)$ is the auto-correlation function of $C_i(t)$. Several types of auto-correlation functions (equivalently power spectra) have been reported in the literature [14], [76]. A widely accepted model, given in [76], is considered in this work, in which the auto-correlation function is given as:

$$K_C(\tau) = \sigma_{C_i}^2 J_0(2\pi f_{dm} \tau) \quad (3.10)$$

where $J_0(\cdot)$ is the zeroth order Bessel function of the first kind, $\sigma_{C_i}^2$ is the average power of $C_i(t)$ and f_{dm} is the maximum Doppler frequency, given by $f_{dm} = v_H/\lambda$, v_H and λ are the speed of the portable in meters per second and the wavelength in meters, respectively.

From (3.3), It is equivalent to say that the signal $z(t)$ is passed through a dispersive channel whose delays and power delay profile are known, and which can be controlled by the transmitter. Novel optimum and suboptimum receiver structures, reported in [2], [3] for two-path Rayleigh channel, can be employed in the present scenario. However, whenever there is an occasional occurrence of multipath fading due to the movements in the surrounding media, the assumption of known delay between the paths may not be valid. It is well-known that an equalizer is an appropriate receiver in these situations. However, since the channel from the base station to the mobile is unknown, a finite training time is required for the receiver to acquire the channel information. In addition, the training should be repeated periodically in situations where the channel varies with time.

3.3 Constraints on delayed diversity transmission:

As mentioned earlier, the performance of the forward link, i.e. from the base station to the portable is dependent on the transmitter signal processing, which consists of time delay elements, and the receiver signal processing, which consists of an equalizer. The success of DDT depends mainly on the following factors:

- (1) Relative transmission time delay between the antennas.

- (2) Equalization strategy used at the receiving end.
- (3) Adaptive algorithm used to acquire and track the channel variations.

As mentioned previously, the effect of the relative transmission delay between the antennas on the system performance and computational complexity is to be evaluated. Details of the derivation of an optimum delay are given in the next section.

The effect of the second factor, the type of equalization strategy, can be examined by considering various techniques viz., decision feedback equalization (DFE), maximum likelihood sequence estimation (MLSE) or different techniques as presented in Chapter 2.

The third factor is very important in examining the feasibility of DDT to a practical TDMA indoor wireless system. The acquisition of the channel information should be done within a short time so that the overhead is not increased. The tracking of the channel should be done as efficiently as possible so that the system performance is not degraded. These two factors reflect on the system efficiency and system performance respectively. As complexity is also an issue here, only the simple LMS algorithm or its modified forms like the normalized least mean squares (NLMS) algorithm are considered for channel tracking.

The effect of these three factors on the system performance is evaluated in the following sections.

3.4 Optimum relative transmission time delay between the antennas

To arrive at an optimum delay, the system performance and complexity are evaluated as a function of the relative transmission delay. Note that the optimum delay may also be a function of the type of equalization employed at the receiver. In this section, we evaluate the optimum delay for some of the equalization methods, discussed in Chapter 2.

A: DFE at the portable:

In Chapter 2, DFE and its modified forms are discussed. Since the bi-directional DFE methods are based on the fundamental structure of DFE, it is sufficient to consider a conventional DFE for arriving at an optimum delay. Analytical performance evaluation of this type of system has previously been reported in the literature. In [104], [120] analytical performance results for simulcasting system were presented for DFE using the Chernoff bound equations given in [16], [17], [116]. However, it is known that the Chernoff bound is not a tight bound in fading channels [42], [68], [109]. An alternative is to use the semi-analytical method proposed in [42], [68]. However, this method does not consider the

acquisition and tracking capabilities of equalization algorithms and also error propagation effects in a DFE. Therefore, to obtain a more accurate system evaluation computer simulations are used.

In this section an optimum value for the relative time delay between the transmitting antennas is derived for DDT assuming that (i) the receiver has sufficient time to acquire the channel characteristics (ii) the channel variations are assumed to be very slow compared to the symbol rate (i.e. the channel is quasi-static) (iii) a DFE is operating at the receiver (iv) the normalized LMS (NLMS) algorithm is used to adapt the equalizer.

System under consideration:

In the following computer simulations an arbitrary TDMA frame format is considered with 1000 training symbols and 2000 data symbols. The 1000 training symbols ensure the convergence of equalizer parameters in a mean squared error (MSE) sense².

In the computer simulations, linear systems and perfect timing and carrier recovery are assumed. The DFE is denoted as DFE(N_f, N_b) where N_f and N_b are the number of forward and backward taps of the DFE respectively. The simulations are initially run to extract the optimum step size parameter for the normalized LMS (NLMS) adaptive algorithm. This procedure is repeated whenever the order of the DFE is changed. The average signal-to-noise ratio (SNR), defined as in the following equation, is measured at the output of the receive filter.

$$\langle \text{SNR} \rangle = \frac{\frac{1}{T_s} \int_0^{T_s} E_{ch} [|X(t) \otimes h_r(t)|^2] dt}{N_0 \int_{-\infty}^{\infty} |h_r(t)|^2 dt} \quad (3.11)$$

The transmit and receive filters, whose impulse responses are $h_t(t)$ and $h_r(t)$ respectively, are square root raised cosine (SRRC) filters with roll-off factor, β , 35%. $E_{ch}[\cdot]$ represents the ensemble average over the channel statistics. These filters are modelled as a 64-tap finite impulse response (FIR) filters. The data is QPSK modulated. The simulation results are compared with the analytical error bounds given in [16], [17], [104], [116].

-
2. The training period in a practical system would be much less than 1000 symbols. This lengthy training sequence used here ensures that the results are not affected by the training period at this stage of investigation. Later, a much shorter training sequence is used.

Since the computational complexity is an important issue, the system performance is also evaluated as a function of the number of feed forward taps of the DFE. The system performance improves as the time delay between the antennas increases, but with a corresponding increase in the required number of forward taps to achieve this performance. An optimum time delay between the antennas can be deduced from these results. The system performance is evaluated in the presence of additive white Gaussian noise (AWGN) and co-channel interference (CCI). The co-channel interference considered here consists of a single strong interferer, transmitted by a different base station identically configured as the base station under evaluation. We assume a frequency non-selective fading channel which is stationary over a duration considerably larger than the symbol period. Therefore, the complex path gain $C_i(t)$ as given in (3.10) can be assumed to be a Gaussian random variable. Therefore, the auto-covariance function of $C_i(t)$ is given as:

$$K_C(\tau) = \sigma_{C_i}^2 \delta(\tau) \quad (3.12)$$

Results [22]:

Fig. 3. 6 shows the system bit error rate (BER) performance in a complex Gaussian fading, AWGN environment for $N_a = 1, 2$ and 4 transmitting antennas with inter-antenna time delay of T_s , the symbol duration. For comparison, the theoretical error bounds, obtained by Monte-Carlo averaging the iterative equations given in [16], [17], [104], [116] are also presented. From these results, it is clearly evident that the analytical bounds are pessimistic, in particular, when N_a is small. Figs. 3. 7 and 3. 8 show the system performance as a function of the relative time delay between the antennas, τ_D , and the number of forward taps, N_f , respectively. Here the system performance results are presented for both AWGN and co-channel interference environments. Fig. 3. 7 shows that for a fixed receiver complexity, (referring only to the signal processing portion of the receiver) the system performance is optimum at $\tau_D = T_s$. Fig. 3. 8 shows that with an increase in complexity, i.e. more forward taps, better performance can be achieved for $\tau_D = 2T_s$ compared to $\tau_D = T_s$. The improvement in system performance, when compared to the increase in number of forward taps, is marginal. Therefore, it can be concluded that the optimum time delay for a reasonable complexity is, T_s , a symbol duration. For delays larger than T_s , the improvement in system performance due to increased computational complexity, i.e. number of feed forward taps, N_f , is not significant. Detailed results can be found in [22].

Fig. 3. 6. System BER performance for $\tau_D = T_s$, $\beta = 35\%$.

Fig. 3. 7. System BER performance as a function of τ_D

- (1) $N_a = 2$
- (2) $\beta = 35\%$
- (3) DFE(5.2)

Fig. 3. 8. System BER performance as a function of N_f for $N_b = 2$

B: MLSE or Bayesian DFE at the portable:

For MLSE and Bayesian DFE, where the channel variations are tracked indirectly, as mentioned in Section 2. 1, we can arrive at an optimum transmission delay intuitively. For MLSE, a delay of more than a symbol period between the transmitter antennas increases the number of states of the trellis, which in turn increases the receiver complexity. A delay of less than a symbol period gives inferior diversity performance because the equalizer may not resolve the paths which are closely spaced. Therefore, a relative transmission delay of T_s , a symbol period, is a good selection. This is also true in the case of Bayesian DFE. As described in Chapter 2, the number of summations involved for each of the possible symbols is exponentially proportional to L , L is equal to $\lceil (N_d - 1)\tau_D / T_s \rceil$ for diversity transmission, where $\lceil x \rceil$ represents the smallest integer greater than x . Therefore as τ_D increases, the complexity of the receiver increases exponentially. Increasing τ_D does not improve the system performance since the minimum distance does not change for a delay of more than a symbol period. Simulation results presented for MLSE detector in [40] show that the system performance improves as τ_D is increased from 0 to T_s . Therefore, we conclude that the optimum delay is a symbol period.

The effects of the second and third factors discussed in Section 3.3, the type of equalization and the channel acquisition and tracking method respectively, can be evaluated by considering a standard TDMA frame structure. For this purpose, the IS-54 format structure is considered in this work.

3. 5 Feasibility of DDT in a practical TDMA system

In this section, the system performance is evaluated for different equalization strategies considering all the practical constraints discussed in Section 3. 3. The results obtained in Section 3. 4 are used in the following performance evaluation, i.e. the relative time delay between the antennas is assumed to be T_s . The performance of various important equalization strategies that are discussed in Chapter 2 is evaluated. In the following sections, the system performance is evaluated assuming a maximum Doppler frequency of 6 Hz, which corresponds approximately to 2 m/s portable speed at 900 MHz. Performance at different portable speeds is reported later in this chapter.

3. 5. 1 System configuration and architecture

To test the feasibility of DDT in a practical TDMA frame format the North American IS-54 standard system, where the modulation technique is $\pi/4$ shifted DQPSK, is considered. The performance of this system is evaluated in a slow fading environment,

typical of indoor wireless systems, in two configurations.

Based on the IS - 54 standard, two system configurations are defined as follows:

1. System-1 does not include the rate-1/2 forward error control (FEC) coder as recommended in IS-54 [45]. The same frame format of 40 ms is still used. Therefore, in this case the user slot consists of 260 information bits with 64 overhead bits, necessary for the synchronization (SYNC) etc. The number of training symbols per block is 14, as in the IS-54 system.
2. System-2 includes the rate-1/2 FEC coder with the specified 26X10 interleaving and deinterleaving [45]. Here, the specifications of the IS-54 are followed except for one discrepancy: all the information bits from the source are encoded using the rate - 1/2 FEC coder (equal protection), whereas IS-54 provides for unequal coding protection for some speech bits. The user slot consists of 130 information bits which are coded into 130 symbols (the 64 overhead symbols are not coded).

The reason for considering these two systems is to investigate the effect of the error statistics at the output of the equalizer on the decoder performance. Though ideally, FEC coding with a finite interleaving depth gives a diversity advantage in the presence of Doppler frequency shifts, this may not necessarily be true in situations where a decoder is preceded by a signal processor, such as an equalizer. The random error and block error correcting capabilities of the FEC coding, used with Viterbi decoding, depend on the free distance of the code, the interleaving depth, and also on the error statistics at the output of the signal processor, in the present case an equalizer.

The system performance measures considered here are: (1) average bit error rate (BER) and (2) average frame error rate (FER). In the IS-54 TDMA format, the first fourteen symbols of the user slot are used for acquiring the channel characteristics. The recursive least squares algorithm (RLS) is used for faster acquisition. For the subsequent channel tracking and detection, the least mean square (LMS) algorithm is used. When we compare the FERs of uncoded and coded systems, there are two important points to be noted, assuming that the SNR per information bit is same for both the systems: (1) the effective SNR at the input of the equalizer for system 2 is 3 dB less compared to system 1 and (2) the TDMA frame consists of 130 information bits for system 2 compared to 260 for system 1 (32 symbol overhead is the same for both the systems).

In all the computer simulations, a maximum Doppler frequency of 6 Hz is assumed at the transmit frequency of 900 MHz for a channel data rate of 48.6 kbps. The BER and

FER are measured over 25,000 frames, which are configured as in [45]. At low FERs the number of transmitted frames are increased to 100,000. The simulated results presented here are for a *full rate* user [45].

The transmit and receive filters are square root raised cosine (SRRC) filters, modelled as 64 tap FIR filters. Linear systems and perfect timing are assumed. The system error rates are measured against the signal to noise power ratio (SNR), which is calculated at the output of the receive filter. A delay of forty symbol durations is assumed at the output of the Viterbi decoder (which is more than six times the code's constraint length). The simulation block diagram is shown in Fig. 3.9. The channel modelling is explained below.

Channel model

Each channel response is modelled as a sum of sinusoidal waveforms as illustrated in [76]. Jakes has suggested a method to generate independent channels using a set of oscillators with recommended phase shifts. It has been shown that the number of independent channels that can be obtained using this method is limited to N_s , the number of sinusoids [76]. But it was observed in the simulations that significant correlation still exists between the channel waveforms, thus generated. This fact is also mentioned in [106]. In this work, the phases of each of the oscillators are selected such that the cross correlation between any two channel waveforms is approximately zero. Furthermore, the number of sinusoids, N_s , used in the generation of each channel waveform, is much greater than the number of independent channels required for the system evaluation. In the present case, the number of independent channels required is equal to the number of transmitting antennas at the base station, N_a , multiplied by one more than the number co-channels, i.e. $N_s > (N_I + 1) N_a$, where N_I represents the number of interfering signals.

Co-channel interference model

The interference comprises of one strong interferer transmitted by an identically configured base station in an adjacent cell. The relative delay between the interferer and the desired signal is assumed to be random and always an integral number of symbol periods. The system performance evaluation in the presence of co-channel interference is measured against signal to interference plus noise ratio (SINR) which is measured at the output of the receive filter.

Fig. 3. 9. Simulation block diagram for DDT using the IS-54 frame format

$$SINR = \frac{S}{N_0 + I} \quad (3.13)$$

where N_0 is the noise power and I is equivalent noise power contributed by the interferer.

3.5.2 DFE at the portable receiver

The system block diagrams of the base station and the portable are shown in Figs. 3. 10 and 3. 11 respectively. As discussed in Chapter 2, two methods were suggested for combined equalization and demodulation for $\pi/4$ shifted DQPSK modulation. The non-coherent demodulation [73] performs slightly worse than the coherent demodulation which is employed here. The decision boundaries of the decision device, shown in Fig. 3. 11, are shown in Fig. 3. 12. These two decision boundaries are used alternatively to detect the DQPSK symbols. To extract the soft decisions from the DFE, the method suggested in [84], [85] can be used. In this method, the likelihood ratios are calculated for each symbol and are fed to the deinterleaver along with the hard-decision outputs of the DFE. But the calculation of likelihood ratios involves additional computation. Therefore, unlike [84], [85], the soft decisions from the DFE are extracted as shown in Fig. 3. 11 and subsequently processed by the soft decision Viterbi decoder.

The recursive least mean squares (RLS) algorithm is used for initial acquisition using the training sequence and the normalized LMS (NLMS) algorithm is used for subsequent data recovery.

Figs. 3. 13 - 3. 14 show the system performance in AWGN for both the system configurations. In the case of single antenna, a DFE with 5 forward taps and a single backward tap is used to estimate the phase of the channel³. Fig. 3. 13 shows the BER performance for one, two and three transmitting antennas for system 1. No significant improvement in the system performance is obtained by using three rather than two transmit antennas. This can be attributed to the limited number of training symbols. The number of training symbols should be at least double the number of the taps of the DFE. Therefore with the IS-54 TDMA frame format the system performance does not improve with an increase in the number of feed forward taps greater than five. Fig. 3. 14 shows the FER for system 1. Fig. 3. 15 and 3. 16 show the average BER and FER for the coded system respectively. These results show that the system FER performance is significantly improved by the FEC coding and finite bit interleaving. For the coded system, as can be seen from Figs. 3. 15 and 3. 16, the system gain due to diversity transmission (in particular

3. In fact, a single tap equalizer can suffice this purpose.

Fig. 3. 10. Block diagram of the base station

Fig. 3. 11. Block diagram of the portable

Fig. 3. 12. Decision boundaries for decision device of Fig. 3. 11

The IS-54 frame format without FEC coding (SYSTEM-1)

Fig. 3. 13. Average BER performance of system - 1

The IS-54 frame format without FEC coding (SYSTEM-1)

Fig. 3. 14. Average FER performance of system - 1

The IS-54 frame format with FEC coding (SYSTEM-2)

Fig. 3. 15. Average BER performance of system - 2

The IS-54 frame format with FEC coding (SYSTEM- 2)

Fig. 3. 16. Average FER performance of system - 2

for dual diversity) is lost at high SNRs. This is due to the occurrence of lengthy bursts at the output of the DFE, which cannot be corrected by the FEC decoder. For three transmitting antennas, however, the performance is significantly better even at high SNRs. But at low SNRs, using three transmitting antennas does not seem to give any additional gain compared to dual diversity transmission. From these results, it can be concluded that the performance improvements by increasing the number of transmitting antennas is limited by training allotted in the transmission frame. Therefore, with a DFE at the portable, the number of transmitting antennas is limited to two for the IS-54 standard TDMA frame structure. Computer simulations are also performed in the presence of a single strong interferer. Table 3.1 lists the relative gains in SNR per information bit obtained by DDT with two transmit antennas in the presence of thermal noise or CCI. These results are presented in [23].

The computational complexity of the DFE in terms of the number of real multiplications (RX) and real additions (R+) involved in recovering one symbol (during the tracking phase) are listed in Table 3.2. The multiplications involving the detected symbols are performed by sign changes (SC).

3.5.3 Bi-directional DFE at the portable receiver

In this section, the system performance of the uncoded and coded systems, described in Section 3.5.1, is evaluated by considering a bi-directional DFE at the portable receiver. The allotted fourteen synchronization (SYNC) symbols of the present user slot are used to train the forward DFE and the fourteen SYNC symbols of the slot immediately following the user slot are used to train the reverse DFE.

Using the procedure outlined in Section 2.2 on bi-directional equalizers, two equalization schemes are defined. These two schemes are explained considering a user frame structure, consisting of N_D information symbols preceded and followed by N_T training symbols, as shown Fig. 3.17. In the first method, the forward and backward DFEs are trained using the first and the last N_T symbols of the received frame respectively. At the end of the training period, the squared magnitude of the error (SE) between the estimated and the expected transmitted signal is measured. Based on this squared error, the direction of equalization, either forward DFE or reverse DFE, is decided for the subsequent data detection as depicted in Fig. 3.18. This method is referred to as bi-directional DFE type - 1 (BDFE-1).

In the second configuration, the user slot is fully (both in training and data detection modes) scanned in both directions, irrespective of the SE at the end of training. The SEs

Table 3.1 SNR per information bit for a given BER/ FER for DFE(5, 1) at a maximum Doppler frequency of 6 Hz

Performance in AWGN/ CCI	$N_d=1$ System-1	$N_d=2$ System-1	Gain in dB	$N_d=1$ System-2	$N_d=2$ System-2	Gain in dB
For BER of 1×10^{-1}	27 dB / 27 dB	20.5 dB/ 21.5 dB	6.5 dB/ 5.5 dB	25.5 dB/ 25.5 dB	21.0 dB/ 22.5 dB	4.5 dB/ 3.0 dB
For FER of 2×10^{-2}	27 dB/ 27 dB	22.0 dB 22 dB	5.0 dB/ 5.0 dB	21.0 dB 21 dB	17.5 dB 18.5 dB	3.5 dB/ 2.5 dB

Table 3.2 Computational complexity per symbol for DFE(N_f , N_b) using LMS algorithm

Function	RX	R+	SC
Estimation	$4N_f$	$4(N_f + N_b)$	$4N_b$
Channel update	$4N_f + 2$	$4(N_f + N_b)$	$N_f + 5N_b$

Fig. 3. 17. Frame structure for bi-directional equalizer

Fig. 3. 18. Structural representation of Bi-directional DFE # 1

corresponding to all the recovered symbols, by the forward and the reverse DFEs, are stored in buffers. The final data is detected at the end of the slot by choosing either the symbol recovered by the forward DFE or the symbol recovered by the reverse DFE, depending on whichever has the smaller SE at the corresponding time, as depicted in Fig. 3. 19. This method is referred as bi-directional DFE type - 2 (BDFE-2). Notice that in all these methods, the instantaneous squared error is considered in recovering the information.

The convergence behavior of the bi-directional equalizer is observed at a maximum Doppler frequency of 6 Hz. The mean squared error (MSE) of the BDFE1 and BDFE2 as a function of time is shown in Fig. 3. 20. The MSE is obtained by averaging the SE over 25000 frames. In the case of BDFE2, the lower squared error of the forward and reverse tracking is selected at each instant. We did not show the convergence during the acquisition period of the reverse track. The number of forward taps, N_f , is varied from 1 to 5 and the simulations are repeated. Note that whenever the number of taps are changed, the optimum step size for the NLMS algorithm during the tracking period is evaluated through repeated computer simulations. The criterion in selecting the optimum step size is the average BER or FER (not the MSE). These results indicate that BDFE2 converges better than BDFE1. As can be seen, for BDFE2 the number of forward taps does not affect the resulting MSE. However, this behavior is not consistent with the system BER performance. Figs. 3. 21 and 3. 22 show the average BER for system 1 and 2 respectively as a function of forward taps. These results indicate that there is a significant improvement in the system performance as the number of forward taps are increased until $N_f = 3$. Above three forward taps, the improvement in system performance is not significant. This statement is true for both the systems, BDFE1 and BDFE2. A similar phenomenon is also observed for frame error rates. As can be seen, BDFE2 performs significantly better than BDFE1. This is expected because in BDFE1, only one of the DFEs, forward or reverse, is used in the tracking mode depending on their SEs at the end of the training period, which is not very reliable.

However, from Fig. 3. 21, a single forward tap filter BDFE1 gives a better error rate than BDFE2. This can be explained by observing the operation of a single forward tap DFE as explained below. In the case of single forward tap, i.e. $N_f = 1$, the system is equivalent to a simple receiver structure where the contributions from the past N_b decisions is removed as shown below.

$$I_k = w_f r_k + \sum_{i=1}^{N_b} w_{b,i} \hat{I}_{k-i} \quad (3.14)$$

2

PM-1 3 1/2" x 4" PHOTOGRAPHIC MICROCOPY TARGET
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PRECISIONSM RESOLUTION TARGETS

Fig. 3. 19. Structural representation of Bi-directional DFE # 2

Fig. 3. 20. Convergence behavior of bi-directional equalizers as a function of forward taps

Fig. 3. 21. System - 1: Average BER performance of bi-directional equalizers as a function of the number of forward taps

Fig. 3. 22. System - 2: Average BER performance of bi-directional equalizers as a function of the number of forward taps

where I_k , $\hat{I}_k$, r_k are the transmitted, recovered information symbols and the received signal at time kT_s and w_f and $\{w_{b,i}\}$ are the forward and backward tap coefficients of the equalizer respectively. N_b represents the number of backward taps. Notice from (3.14) that the operation of the DFE is simplified to removing the contribution from the past symbols. Assuming N_a antennas separated by a transmission delay of T_s , the received sample at time kT_s can be expressed as follows.

$$r_k = c_0 I_k + \sum_{i=1}^{N_a-1} c_i I_{k-i} + n_k \quad (3.15)$$

where $\{c_i\}$ are the path gains and n_k is the sample of the thermal noise at time kT_s . Therefore the information symbol can be recovered as...

$$\hat{I}_k = \frac{1}{c_0} r_k - \sum_{i=1}^{N_a-1} \frac{c_i}{c_0} \hat{I}_{k-i} - \frac{1}{c_0} n_k \quad (3.16)$$

Notice that the difference between the transmitted and the recovered symbols constitutes the noise term, assuming that the past detected symbols are correct. In the case of conventional DFE, most of the information symbols can not be recovered correctly whenever c_0 is very small, since noise is amplified. In the case of bi-directional DFE, the forward DFE evaluates I_k using r_k as shown in (3.16), whereas the reverse DFE evaluates I_k using r_{k+N_a-1} as follows.

$$\hat{I}_k = \frac{1}{c_{N_a-1}} r_{k+N_a-1} - \sum_{i=0}^{N_a-2} \frac{c_i}{c_{N_a-1}} \hat{I}_{k+N_a-1-i} - \frac{1}{c_{N_a-1}} n_{k+N_a-1} \quad (3.17)$$

In the case of BDFE1, the SEs of the forward DFE, which is equal to $|n_k/c_0|^2$ or the reverse DFE, which is equal to $|n_{k+N_a-1}/c_{N_a-1}|^2$, are compared at the end of the training sequence and decided in favor of the lowest. In the case of BDFE2, the SEs are compared for all the symbols. Therefore, there is an increasing possibility for BDFE2 to make errors compared to BDFE1. This explains why BDFE2 is inferior to BDFE1 in the case of single tap DFE as shown in Fig. 3. 21. Note that this phenomenon is only true for a single forward tap DFE, since the forward filter of the DFE does not estimate the inverse folded spectrum of the received signal as is the case for a DFE with more than two forward taps [109].

The system BER and FER performance for system - 1 is shown in Fig. 3. 23 and 3. 24 respectively. As shown, BDFE2 (3, 1) performs significantly better than BDFE1(3, 1) and DFE (5, 1). Figs. 3. 25 and 3. 26 show the corresponding system performances for the coded system. BDFE1 gives inferior performance compared to the conventional DFE, however, this is with fewer forward taps. This is because the selection between the forward and backward DFEs at the end of the training period is dependent on the instantaneous MSE, which is not a good indicator of the symbol error rate. At high SNPs, the system gain from using diversity transmission diminishes because of the burst errors at the output of the DFE. The diversity gain for BDFE2 (3, 1) is listed in Table 3.3. Identical figures are obtained in the presence of CCI also.

The computational complexity of the bi-directional DFEs can be calculated following Table 3.2. During the data detection phase, the computational complexity of BDFE1 is less than the conventional DFE, because of fewer forward taps. For BDFE2, the computational complexity is approximately the same as that of the conventional DFE. For dual antenna transmission, the number of computations are tabulated later in this chapter. In the following sections, we consider only BDFE2 for comparison with the other types of equalization methods, since BDFE1's performance is worse than the conventional DFE.

3. 5. 4 Bayesian DFE at the portable receiver

In this section, forward link performance results with dual diversity transmission from the base station and Bayesian DFE (BayDFE) at the portable are presented.

Bayesian DFE for diversity transmission:

Assuming flat faded signals from N_a transmitting antennas simultaneously with different delays and M possible transmitted symbols, the number of summations of exponential terms, as given in Section 2. 2, become $v = M^L$, where L is the number of symbol periods spanned by the combined response of the equivalent channel, the transmit and receive filters. When the combined response of the transmit and receive filters follow any of the Nyquist pulse shapes, there will not be any additional dispersion resulting from the filters when the receiver samples the signal at the symbol rate. Therefore, when the base station transmitter is equipped with N_a antennas with transmission delays as described in

Section 3. 2, then $L = \left\lceil \frac{(N_a - 1) \tau_D}{T_s} \right\rceil$, where $\lceil \xi \rceil$ represents the smallest integer larger

than or equal to ξ . Since, as previously deduced, a transmission delay of T_s is the optimum, the dispersion of the equivalent channel is $(N_a - 1) T_s$. Therefore, the number of

Fig. 3.23. Average BER performance for system - 1 with dual diversity transmission from the base station and bi-directional equalization at the portable.

Fig. 3. 24. Average FER performance for system - 1 with dual diversity transmission from the base station and bi-directional equalization at the portable.

Fig. 3. 25. Average BER performance for system - 2 with dual diversity transmission from the base station and bi-directional equalization at the portable.

Fig. 3. 26. Average FER performance for system - 2 with dual diversity transmission from the base station and bi-directional equalization at the portable.

Table 3.3 SNR per information bit for a given BER/ FER for RDFE2(3,1) at a maximum Doppler frequency of 6 Hz

Performance in AWGN/CCI	$N_d=1$ System-1	$N_d=2$ System-1	Gain in dB	$N_d=1$ System-2	$N_d=2$ System-2	Gain in dB
For BER of 1×10^{-3}	27 dB	19.4 dB	7.6 dB	25.5 dB	20.3 dB	5.2 dB
For FER of 2×10^{-2}	27 dB	20.2 dB	6.8 dB	21.0 dB	16.8 dB	4.2 dB

summations of exponential terms are $v = M^{N_s-1}$.

Thus, for dual diversity transmission the number of summations of the exponential terms are four. The performance results for system BER and FER are presented in Fig. 3. 27 and Fig. 3. 28 respectively for system - 1. Figs. 3. 29 and 3. 30 show the system BER and FER performance respectively for system - 2. Here, we considered both hard and soft decision decoding. As can be seen, there is a significant diversity gain by dual antenna transmission. Compared to the conventional DFE and the bi-directional DFE, the Bayesian DFE performs significantly better. Also observe that for single antenna transmission, Bayesian DFE performs significantly better than the conventional DFE. The reason, as already noted in Section 2.1 and in [89], is that the channel can be tracked efficiently through indirect adaptation. Furthermore, the soft decision decoding improves the system performance substantially. Table 3.3 lists the diversity gains for system - 1 and 2 for both soft and hard decision decoding schemes. Notice that the diversity gains for systems with soft decision decoding (SD) are less than for hard decision decoding (HD) schemes. Table 3.5 gives the computational complexity involved per symbol for extracting both hard and soft decisions. As can be seen obtaining soft decisions does not involve significant additional complexity compared to hard decisions and performance is significantly improved.

For dual antenna transmission with DQPSK modulation, where $N_a=2$ and $M=4$; $v=4$, the number of computations are listed in Table 3.6. As shown, the soft decisions from the Bayesian DFE can be obtained with a few real additions. Based on the performance improvements observed, soft decisions decoding is recommended. Therefore, in the following sections, the Bayesian DFE with soft decision decoding is considered for comparison purpose. From hereon, whenever Bayesian DFE is referred to, it implies that it is with soft decision decoding.

3. 5. 5 MLSE at the portable receiver

The system configuration is similar to the one shown in Figs. 3. 10 and 3. 11, except that the DFE is replaced by a MLSE. To extract soft decisions from the MLSE, the method proposed in [65], can be used. However, there will be a significant increase in the computational complexity involved with this method. As mentioned earlier, we are only interested in low complexity solutions. Therefore, only hard decision decoding is considered for MLSE. As before, the LMS algorithm is used for channel tracking. Two different techniques are used to track the channel: (1) using the delayed decisions from the output of the MLSE and (2) using the principle of per-survivor processing (PSP)[49], [111], [112], [119],[122]. The system performance in the former method is a function of the delay

Fig. 3.27. Average BER performance for system - 1 with dual diversity transmission from the base station and Bayesian DFE at the portable.

Fig. 3. 28. Average BER performance for system - 1 with dual diversity transmission from the base station and Bayesian DFE at the portable.

Fig. 3. 29. Average BER performance for system - 2 with dual diversity transmission from the base station and Bayesian DFE at the portable.

Fig. 3.30. Average FER performance for system - 2 with dual diversity transmission from the base station and Bayesian DFE at the portable.

Table 3.4 SNR per information bit for a given BER/ FER - for BayDFE at maximum Doppler frequency of 6 Hz

Performance in AWGN/ CCI	$N_a=1$ System-1	$N_a=2$ System-1	Gain in dB	$N_a=1$ System-2	$N_a=2$ System-2	Gain in dB
For BER of 1×10^{-3}	26.0 dB	17.4 dB	8.6 dB	30.0 dB (HD)	19.6 dB	10.4 dB
				23.3 dB (SD)	16.5 dB	6.8 dB
For FER of 2×10^{-2}	23.3 dB	17.8 dB	5.5 dB	22.5 dB (HD)	16.3 dB	6.2 dB
				17.1 dB (SD)	14.1 dB	3.0 dB

Table 3.5 Computational complexity per symbol for Bayesian DFE using the LMS algorithm

Function	RX	R+	SC	exp(.)	Com pare	R/ M
Estimation	$(2N_a+1)vM$	$(8N_a-3)vM + (v-1)M$	$4N_avM$	vM	0	0
Addition for SD	0	$(5M-3)$	$4M$	0	0	1
Addition for HD	0	0	0	0	1 out of M	0
Channel tracking	2	$8N_a$	$5N_a$	0	0	0

Table 3.6 Computational complexity per symbol for Bayesian DFE for dual diversity transmission

Function	RX	R+	SC	exp(.)	Compare	R/
Estimation	80	220	128	16	0	0
Addition for SD	0	17	16	0	0	1
Addition for HD	0	0	0	0	1 out of 4	0
Channel tracking	2	16	10	0	0	0

of the decisions that are fed back from the output of the MLSE. If the delay is too small the decisions are not reliable and, on the other hand, if the delay is large then there is a time offset between the estimated channel and the actual channel. This is one of the reasons for degraded performance of decision feedback tracking for MLSE in fast varying channels. The PSP is a special case of a combined channel tracking and the sequence estimation method described in Chapter 2. In the PSP each state is assigned a set of weight parameters, which represent the channel status. These weights are updated every iteration of the algorithm according to the surviving symbol sequence for that particular state. The updating of these weights can be done by any of the classical adaptive algorithms, such as the LMS or the RLS. In [49] it was mentioned that both the LMS and the RLS algorithms have similar performance. This method is also independently proposed by Fukawa and Suzuki in [58]. In [58] and subsequently in [150], this algorithm is referred as RLS - MLSE, in which RLS algorithm was used in channel tracking.

Generalized PSP for diversity transmission:

Assuming flat faded signals with different delays from N_a simultaneously transmitting antennas with M possible transmitted symbols the number of states become $\rho = M^L$, where L is the number of symbol periods spanned by the combined response of the equivalent channel, the transmit and receive filters. There are $v = K\rho$ channel vectors which are continuously updated along the v surviving paths. When the combined response of the transmit and receive filters follow any of the Nyquist pulse shapes, there will not be any additional dispersion resulting from the filters when the receiver samples the signal at the symbol rate. Therefore, when the base station transmitter is equipped with N_a antennas with transmission delays as described in Section 3. 2, then $L = \left\lceil \frac{(N_a - 1) \tau_D}{T_s} \right\rceil$, where $\lceil \xi \rceil$ represents the smallest integer larger than or equal to ξ .

Results:

In evaluating the system performance, as in the previous cases, the investigations are limited to two transmitting antennas. A delay of 20 symbol intervals is assumed at the output of the MLSE. Fig. 3. 31 gives the system BER results for system 1 (no FEC coding) for two transmitting antennas at a maximum Doppler frequency of 6 Hz. As shown, at low bit error rates there is a significant advantage in SNR required to achieve a specified BER using the PSP. Typically, at a BER of 2×10^{-4} the PSP channel tracking gives an advantage of more than 2 dB. As is also shown, the PSP channel tracking is very near to the attainable optimum limit of channel tracking using the correct data. Fig. 3. 32 shows the FER

Fig. 3.31. System BER performance with MLSE for system 1

Fig. 3.32. System FER performance with MLSE for system - 1

performance for system 1 for all three types of channel tracking methods. There is a significant SNR advantage by using the PSP when compared to the conventional delay tracking method. Figs. 3. 33 and 3. 34 show the average BER and FER performance respectively for system - 2. Table 3.7 shows the gain in SNR by using the diversity transmission. These results clearly show that; (1) there is significant improvement in the system performance using diversity transmission from the base station and (2) the FEC coding improves system FER, although there is lower SNR per coded symbol. Notice that for the coded system, the SNR per symbol is 3 dB less at the input of the equalizer compared to the uncoded system.

Acquisition capabilities of the PSP:

At this point, it is instructive to note that PSP type channel tracking was initially introduced as a blind adaptation algorithm [119], [122]. Therefore, it is relevant at this stage to investigate the unsupervised learning capabilities of PSP. As mentioned earlier, in all the previous simulations, SYNC symbols are used for channel acquisition. Similar performance is also expected with a blind acquisition strategy. This can be ascertained by some tests on the capabilities of the PSP for blind acquisition. For this purpose, an arbitrary frame with $\pi/4$ shifted DQPSK modulated information symbols is considered. At the start of each frame the PSP algorithm is initialized with random channel weight vectors and random states. The squared error over all the frames is averaged to get an ensemble average value. The LMS algorithm is used to update the channel vectors. Fig. 3. 35 show the mean squared error (MSE) as a function of symbol index for different values of K , ranging from 1 to 4 for a maximum Doppler frequency of 6 Hz. These results show that if the receiver is turned on slightly before its allotted slot, then the receiver can easily acquire the channel characteristics. Alternatively, the received signal samples can be stored in a buffer and processed iteratively. The attractive feature of this algorithm is that it needs only simple LMS algorithm update. Therefore compared to DFE, where the more complex RLS algorithm is needed for channel acquisition, the overall complexity of the equalization process is reduced for PSP. In addition, PSP has the advantage of unsupervised learning. Therefore, there is a two fold advantage by using the PSP algorithm; (1) computational complexity is significantly reduced during the acquisition mode and (2) the number of overhead symbols can be reduced further.

However, during the unsupervised learning mode, there is a power penalty due to the additional power consumption from turning on the RF front end and the receiver ahead of the allotted time slot. If the iterative processing approach is adopted, there is a delay involved in decision making.

Fig. 3. 33. System BER performance with MLSE for system 2

Fig. 3. 34. System average FER performance with MLSE for system 2

Table 3.7 SNR per information bit for a given BER/ FER for MLSE-PSP at a maximum Doppler frequency of 6 Hz

Performance in AWGN	$N_d=1$ System-1	$N_d=2$ System-1	Gain in dB	$N_d=1$ System-2	$N_d=2$ System-2	Gain in dB
For BER of 1×10^{-3}	25.9 dB	16.2 dB	9.7 dB	30.0 dB	19.0 dB	11.0 dB
For FER of 2×10^{-2}	23.3 dB	16.2 dB	7.1 dB	22.4 dB	15.8 dB	6.6 dB

Fig. 3.35. Blind acquisition performance of PSP - K at 6 Hz maximum Doppler frequency

A detailed analysis of the computational and memory requirements of this algorithm is considered next.

Implementation and computational complexity of PSP - k:

In this section the implementation of the PSP algorithm is discussed briefly. The structural block diagram is given in Fig. 3. 36. The parallel structure, as shown in the figure, facilitates efficient implementation of the Viterbi algorithm at high throughput rates. In the case of the IS-54 frame structure, the channel data rate is 48.6 kbps the user's effective information rate is 13.0 kbps for a *full rate* user and 6.5 kbps for a *half rate* user. At these low bit rates a pipeline structure using available technology suffices. However, a general approach is presented here, which can also be used for higher bit rates [55].

As shown, the *transition metrics* are calculated using the trellis information and the channel vectors. After calculating the ML metrics for all the possible state transitions, these metrics are passed on to the add-compare-select (ACS) units. The ACS units calculate the cumulative maximum likelihood (ML) metrics, compare them and select the best K metrics for each state. Each ACS unit, in fact, consists of K compare-select operations. The path information of the K_p best paths is stored in the survivor path memory. When the path information exceeds $N_B + T$, where T is the delay involved in detecting the symbols⁴, the sequence detector starts scanning the path, stored in the survivor memory unit, backwards using the cumulative metric stored in the feedback unit at time $k + N_B + T$ and the trellis information. At the end of this cycle the survivor memory locations are shifted by N_B units to the left. When the new cycle starts, the memory locations are filled from time $N_B + 1$ onwards. The channel units are updated continuously as shown in the figure. The detected sequence of the present frame is stored in a buffer until the next frame arrives. The complete data is then passed on to the deinterleaver.

Computational requirements:

For the IS-54 standard architecture, where the modulation scheme is $\pi/4$ shifted DQPSK, the details of the computations per symbol are listed in Table 3.8. The corresponding computational complexity for dual diversity transmission is tabulated in Table 3.9.

4. T is a function of the total time dispersion during the transmission, which includes the dispersion of the channel and the artificially created dispersion by employing DDT.

Fig. 3. 36. Structural representation of the PSP algorithm

Table 3.8 Computational complexity of PSP (acquisition/ detection) per symbol

Operation	real multiplications	real additions	sign changes	comparisons
Error calculation	$2Mv$	$4vN_a + 8vM - 6v$	$4v(N_a + M - 1)$	$p(K \text{ out of } 4)$
Channel update	$2v$	$4vN_a$	$5vN_a$	-----

Table 3.9 Computational complexity of PSP-K with dual antenna transmission per symbol

Operation	real multiplications	real additions	sign changes	comparisons
Error calculation	$32K$	$136K$	$80K$	$4(K \text{ out of } 4)$
Channel update	$8K$	$32K$	$40K$	-----

Memory requirements for PSP-K:

Approximate expressions for calculating the memory requirements are listed below.

Number of memory locations for storing floating point variables: $2v(N_a+1) + \rho(3M+K)$

Number of memory locations for storing integer variables: $2vN_a + 4(N_B+T)$

where...

v represents the number of survivors ($= K\rho$)

ρ represents the number states of the trellis

K represents the number of survivors per state

N_B represents the length of the sequence to be detected

T represents the delay required in detection.

N_a represents the number of transmitting antennas

M represents the size of the symbol alphabet

For example, if a 4 byte memory is allowed for a real variable and 1 byte memory for an integer variable, a total of 1048 bytes of memory are required for dual antenna transmission in a flat fading channel and PSP - 1. A detection delay, T , of 20 symbol periods and a sequence length, N_B , of 162 symbols (i.e. a complete IS-54 frame) are assumed in these calculations. In arriving at the above expressions, the buffers required for the arithmetic, logic operations and the frame multiplexing (i.e. for holding the present data until the arrival of the next frame) operations are considered.

3.6 Comparison of various equalization methods for dual diversity transmission

In this section, the performance and computational complexity of various equalization methods, presented in the previous sections, are compared for dual diversity transmission at a maximum Doppler frequency of 6 Hz. The equalization methods considered are:

- ◆ Bi-directional equalizer (BDFE2)
- ◆ Bayesian DFE
- ◆ MLSE with classical channel tracking
- ◆ MLSE with PSP channel tracking ($K = 1$)

The comparison is carried out for both system - 1 (without FEC coding) and system - 2 (with rate - 1/2 FEC coding). As observed in Sections 3. 5. 2 and 3. 5. 3, the conventional DFE and BDFE1, compared to other equalization methods perform poorly. Therefore, we will not consider these equalizers in the following comparison. In the case of MLSE, due to the significant computational complexity involved in extracting soft decisions from MLSE, we only use hard decision decoding. For symbol by symbol detectors, we use soft decision decoding, since the difference in complexity for obtaining the soft decisions compared to obtaining hard decisions is very small. As shown in Section 3. 5. 4, for Bayesian DFE, soft decisions can be obtained at the cost of a few real additions.

A. Performance comparison:

The results presented in Section 3. 5 are superimposed on one graph to show the relative performance. Figs. 3. 37 and 3. 38 show the average BER and FER performance respectively of the different equalizers for system - 1. Here, the results when the base station transmits the signal using a single antenna are also presented. As shown, for single antenna transmission, the MLSE and Bayesian DFE, where the indirect adaptation is employed, track the channel phase more efficiently than a DFE, where direct adaptation is employed. Furthermore, as expected, the MLSE and the Bayesian DFE give identical results for single antenna transmission. For dual antenna transmission, MLSE-PSP-1 shows significant gains when compared to the other equalization methods. Figs. 3. 39 and 3. 40 show the average BER and FER performance respectively of system - 2. Here, the Bayesian DFE with soft decisions offers large gains compared to the other systems. This is mainly attributed to the soft decision decoding. MLSE-PSP performs well even with hard decision decoding. For single antenna transmission, DFE(5, 1) performs better than MLSE-PSP-1 because of soft decision decoding. From Figs. 3. 38 and 3. 40, it is evident that the FEC coding improves the FER performance at these low Doppler frequency shifts of 6 Hz. Table 3.10 shows the required average SNR per information bit (which is equal to the ratio of energy per information bit to the noise power spectral density, E_b/N_0) for all the equalizers considered here. We also simulated the system performance in the presence of a single strong cochannel interferer. System performance in the presence of CCI for all the

Fig. 3.37. Performance comparison of various equalization methods for system -1

Fig. 3.38. Performance comparison of various equalization methods for system -1

Fig. 3.39. Performance comparison of various equalization methods for system -2

Fig. 3. 40. Performance comparison of various equalization methods for system -2

Table 3.10 Required average SNR per information bit to achieve a FER of 2×10^{-2} at a Doppler frequency of 6 Hz

Equalizer type	Average SNR per information bit for System - 1	Average SNR per information bit for System - 2
$N_a = 2$ (Dual antenna transmission); $\tau_D = T_r$		
MLSE-PSP-1	16.2	15.7
MLSE-conv	16.9	15.7
BayDFE	17.8	14.1
BDFE2 (3, 1)	20.2	16.8
$N_a = 1$ (Single antenna transmission)		
MLSE-PSP	23.3	22.5
DFE (5, 1)	27.0	20.6
BayDFE	23.3	17.1

equalizers is observed to be not significantly different compared to AWGN environment. Therefore, we do not present these results here.

B. Computational complexity comparison:

The other performance criterion in this evaluation is the computation complexity of the different equalization methods. Table 3.11 shows the number of different operations required per symbol during the detection phase for all the equalization methods for $\pi/4$ shifted DQPSK and dual diversity transmission. Of all the methods, Bayesian DFE requires the most number of computations, where as, MLSE-PSP requires the least number of computations. MLSE-PSP-1 is approximately 50% less complex than Bayesian DFE.

Furthermore, as shown previously, MLSE-PSP with LMS channel update has the added advantage of fast blind acquisition capabilities. For the other methods, a more complicated RLS algorithm is required for channel acquisition using the allotted training symbols. The computational complexity of PSP using LMS algorithm during the acquisition phase is discussed further later in this chapter.

3.7 System performance in an outdoor environment

At this point, we can conclude that the MLSE-PSP, the MLSE with conventional tracking and the Bayesian DFE perform well in a slowly fading indoor environment with the help of dual diversity transmission. However, we have not yet proven that the IS-54 system employing these equalization methods can be used in both the indoor and outdoor environments. We need to show that the equalizers perform within the required specifications in the outdoor environment also. In the literature, some of the work reported in [35], [36], [49], [82], [111], [112], [119] can be referred to for the performance evaluation of these equalizers in an outdoor environment. References [35] and [36] deal with the Bayesian DFE's performance in two path fading channel at a maximum Doppler frequency of 100 Hz. But the channel data rate considered here is 300 kbps, and furthermore, these results are not for the IS-54 frame structure. Different channel models and frame structures are used in [49] for the performance evaluation of PSP, which is referred as the combined channel tracking and sequence detection algorithm as introduced in [119]. Similar kind of evaluations are reported in [111] and [112]. In [82], Lin et. al. reported the performance of PSP (again referred to as combined channel tracking sequence detection algorithm) for the IS-54 standard frame format without FEC coding. In the following section, a complete evaluation of these equalizers is reported. We have presented the performance of PSP- K in an outdoor environment in [25], [27].

Table 3.11 Number of computations required per symbol ^a

Equalizer type	(X)	(+)	SC	C	exp	div
MLSE-PSP	40	168	120	4	0	0
MLSE-Conv	40	152	80	4	0	0
BayDFE	82	253	154	0	16	1
BDFE2	52	64	24	0	0	0

a. (X) - real multiplication; (+) - real addition; (SC) - sign change; (C) - 1 out of 4 comparison; (exp) - exponentiation; (div) - real division.

In this section, computer simulation results for maximum Doppler frequencies of 18.5 Hz, 55.5 Hz and 90 Hz, which correspond to vehicle speeds of 22.2, 66.6 and 108 KMPH respectively at 900 MHz. As stated before, our investigations are limited to the LMS algorithm in the outdoor environment. Here we show the results in terms of the required SNR per information bit to achieve a specified error rate performance. Figs. 3. 41 to 3. 42 show the required SNR to achieve a specified error rate as a function of the Doppler frequency for the uncoded system. Simulation results for MLSE with the conventional channel tracking with delay $d = 2$, the Bayesian DFE and the MLSE-PSP- K for all the possible values of K are presented. For comparison, we considered the performance of MLSE with channel tracking using the correct symbols as the reference system. The results presented in Fig. 3. 41 and 3. 42, show that the MLSE-PSP outperforms all the other equalization methods. As shown in these figures, there are diminishing returns as K is increased above 2 for system - 1. Another observation is that the average FER for PSK - K for $K > 1$ is better compared to the channel tracking using correct data symbols (see Fig. 3. 42). This paradoxical observation is explained by examining the probability of error bursts as a function of the burst length as shown in Fig. 3. 43. These results indicate that though the bit error rates are the same for both cases, there are fewer error bursts for PSP channel tracking since longer error bursts occur more frequently. Figs. 3. 44 and 3. 45 give the performance results for system - 2. From these results, it is clearly evident that, though ideally the coded system is expected to perform much better as the Doppler frequency increases due to the increased effectiveness of finite interleaving, this is not necessarily true due to tracking limitations. As shown in these figures, there are diminishing returns as K is increased above 2 for coded system also. The average FER performance improves employing FEC coding and interleaving, in particular at high Doppler frequencies. Table 3.12 shows the gain in SNR per information bit due to coding as a function of maximum Doppler frequency at a FER of 2×10^{-2} . For comparison, we showed the gain obtained for the reference system also.

Previously, the blind acquisition capabilities of the PSP have been presented at a maximum Doppler frequency of 6 Hz. It is worthwhile to carry out similar investigations at higher Doppler frequency shifts also. The computer simulations are repeated at a maximum Doppler frequency of 90 Hz. Fig. 3. 46 shows the mean squared error as a function of symbol index. These results show that PSP - 3 or 4 with the LMS update can acquire the channel properties with sufficient accuracy in 30 symbol periods. Thus, an added benefit with MLSE-PSP is that a training sequence is not required and issues of cell-site synchronization are avoided. Therefore, we can suggest a system with MLSE-PSP- K operating during the acquisition phase with $K = 3$ and with $K = 1$ or 2 in the subsequent detection phase. However, as mentioned previously, since the acquisition time for the PSP is longer than the allowed training period in the IS-54 frame structure, the received frame

Fig. 3.41. SNR per information bit required for a BER of 1×10^{-3} for system -1 as a function of Doppler frequency for different equalization strategies

Fig. 3. 42. SNR per information bit required for a FER of 2×10^{-2} for system-1 as a function of Doppler frequency for different equalization strategies

Fig. 3. 43. The probability of error bursts versus the length of the error burst

Fig. 3. 44. SNR per information bit required for a BER of 1×10^{-3} for system-2 as a function of Doppler frequency for different equalization strategies

Fig. 3. 45. SNR per information bit required for a FER of 2×10^{-2} for system-2 as a function of Doppler frequency for different equalization strategies

Table 3.12 Coding gain (dB) for PSP-K at different Doppler frequencies at a FER of 1×10^{-2}

SYSTEM	6 Hz	18.5 Hz	55.5 Hz	90 Hz
PSP - 1	0.5	1.4	3.5	6.0
PSP - 2	0.7	1.0	3.1	4.8
PSP - 3	0.7	1.2	3.3	5.3
PSP - 4	0.8	1.4	3.5	5.5
Reference	1.2	2.5	6.2	8.2

Fig. 3. 46. Blind acquisition performance of the PSP at 90 Hz maximum Doppler frequency.

is stored in a buffer and processed iteratively. Part of the work described in this section is reported in [25], [27], [28].

Comparison of computational complexity of PSP - 3 in blind acquisition mode with DFE - RLS in trained acquisition mode:

Now the question is what is computational complexity of blind acquisition compared to training assisted RLS acquisition. To answer this question, we considered a conventional DFE (5, 1) with RLS algorithm update during the acquisition scheme.

For dual antenna transmission, the number of computations per symbol as a function of K are given in Table 3.9 (page 96). Tables 3.14 and 3.13 give the corresponding computational complexity per symbol in acquisition and detection mode respectively for a DFE with 5 forward taps and one backward tap. Note that in the TDMA frame considered here, the acquisition mode is over 14 symbol periods and data detection is over 148 symbol periods. As an example the computational complexity of: (1) DFE(5, 1) with RLS for trained channel acquisition and LMS for channel tracking with (2) PSP with $K = 3$ for blind channel acquisition and with $K = 1$ for channel tracking, are compared. Blind acquisition using PSP-3, for example, requires approximately 30 symbols to acquire the channel with sufficient accuracy. The ratio of real multiplications required for PSP-3 over 30 symbols and DFE(5,1) with RLS over 14 symbols is $(30 \times 120) / (14 \times 426)$, which is around 0.6. Therefore, assuming that multiplications constitute the major part of the computational load, then PSP-3 has 60% of the complexity of RLS channel acquisition. During the data detection mode, PSP-1 requires nearly the same number of multiplications as the DFE(5, 1) with LMS update. For further performance improvements, PSP- K with $K > 1$ can be used during data detection mode at the expense of additional complexity. Therefore, considering the performance of all the equalizers, we conclude that MLSE with PSP channel tracking is superior for both system -1 and system -2. The presented results show that the average FER performance is significantly improved by using FEC coding, specifically at high Doppler frequencies.

3.8 Coding gain at low Doppler frequencies

Table 3.15 shows the required SNR per information bit for different maximum Doppler frequencies. These results show that the FEC coding improves the system FER even at a low maximum Doppler frequency of 2 Hz. However, the coding gain at low Doppler frequencies is less than 1 dB due to the ineffectiveness of the interleaving depth used. This marginal gain is achieved by reducing information rate to half. Therefore, using FEC coding in indoor environment is disputable. However, if the IS-54 portables, equipped

Table 3.13 Computational complexity of DFE (acquisition using RLS algorithm)

Operation	real multiplications	real additions	sign changes	comparisons
Error calculation & Channel update	426	426	40	0

Table 3.14 Computational complexity of DFE (detection using LMS algorithm)

Operation	real multiplications	real additions	sign changes	comparisons
Error calculation & Channel update	42	48	14	0

Table 3.15 Required SNR per information bit to achieve a specified performance at different maximum Doppler frequencies

System		2 Hz	4 Hz	6 Hz
System - 1	BER of 1×10^{-3}	15.6 dB	15.6 dB	16.2 dB
	FER of 2×10^{-2}	15.8 dB	15.8 dB	16.2 dB
System - 2	BER of 1×10^{-3}	18.5 dB	18.5 dB	19.1 dB
	FER of 2×10^{-2}	15.0 dB	15.0 dB	15.7 dB

with MLSE-PSP are used in the indoor environment without modification, the results presented indicate that the combination of FEC coding with the recommended interleaving provides a marginal gain.

3.9 Equalization with diversity transmission vs. Diversity combining at 6 Hz Doppler frequency

Finally, it is important to show how diversity transmission with equalization performs compared to diversity reception. In [145], it was shown that the diversity transmission with equalization can perform within 0.1 dB of diversity reception. However, these results were shown through analytical approximations and in the absence of channel variations. Here, we show computer simulation results for the IS - 54 standard system at 6 Hz maximum Doppler frequency considering (1) dual diversity transmission from the base station and MLSE-PSP with $K=1$ at the portable and (2) dual diversity reception with optimal combining at the portable. The first fourteen SYNC symbols are used to train the weights of the receive antennas or to train the equalizer as before. Fig. 3.47 and 3.48 compare the average FER performance for uncoded and coded systems respectively. In the case of optimal combining, we considered the total average received SNR per information bit for both the antennas. It is evident from these results that the diversity transmission and equalization achieve system performances very near to optimal combining at the receiver. In particular, at a FER of 2×10^{-2} , the forward link performs within 0.6 dB and 1.0 dB for uncoded and coded systems respectively. However, if the comparison is carried out by assuming equal transmitted signal powers, optimal combining achieves additional advantage of 3 dB.

3.10 Summary and conclusions

The proposed system with diversity transmission from the base station and equalization at the portable is shown to improve system performance significantly compared to single antenna transmission. The relative transmission delay of a symbol period between the antennas is shown to be optimum considering system performance and computational complexity as the selection criteria. The other parameters which affect the performance of diversity transmission are (1) the type of equalization and (2) the acquisition and tracking algorithms. Since most of the well known equalizers belong to a class of MSE type, such as DFE, or MAP detectors, or MLSE, we considered one from each of these categories for the system performance evaluation. A thorough performance evaluation indicates that MLSE-PSP can give significant diversity gains. This is also shown to be true even in the outdoor environment. We restricted the tracking algorithm to the LMS algorithm for its low complexity. Further restriction on the number of training symbols is

Fig. 3. 47. Performance comparison of diversity combining and diversity transmission with MLSE-PSP for system - 1

Fig. 3. 48. Performance comparison of diversity combining and diversity transmission with MLSE-PSP for system - 2

imposed by adopting the North American IS-54 standard architecture. We found, however, that the MLSE-PSP can acquire the channel blind within 30 symbols using the LMS algorithm.

From the results presented in this chapter, we recommend: (1) MLSE with PSP channel tracking at the portable receiver and (2) dual antenna transmission with a transmission delay of one symbol from the base station. This combination achieves system performance improvements equivalent to diversity reception at the portable receiver but with low complexity. The same portable receiver with MLSE-PSP type equalizer and the LMS update can also be used in a typical outdoor environment with single or dual antenna transmission.

We also evaluated the performance of a coded system, following the IS-54 standards. The results show that the coding further enhances the diversity gains by 0.5 dB at a portable speed of 2 m/s. Note that this gain is achieved at a 3 dB lower SNR per coded symbol compared to uncoded system. The effectiveness of the MLSE in removing the ISI is affected by the increased thermal noise relative to the uncoded system. Coding gain is observed to be further increased in the outdoor environment.

Chapter 4

Equalization for Wide Band Indoor Wireless Systems in the Millimeter Wave Frequency Band

As stated in Chapter 1, the channel dispersion varies from 10 ns to 200 ns depending on the physical dimensions of the building, material of the walls etc. Therefore, for high rate data transmission in the range of 20 Mbps and above, the signals experience severe intersymbol interference (ISI). Equalization improves the system performance in these situations [109]. Due to the large bandwidth requirements, we consider the millimeter wave frequency band (40 GHz - 300 GHz) in this work. The channel variation normalized to data rate at these transmit frequencies, at the expected portable mobility and changes in the surrounding environment is not significant. However, present day technology limitations in the millimeter wave frequency band restrict the selection of easily realizable carrier modulation/demodulation methods. In particular, circuit nonlinearities are difficult to avoid at these high transmit frequencies without a large increase in circuit complexity and power consumption. Therefore, spectrally efficient phase/amplitude modulation methods may not be a good selection because of the following difficulties: (1) linear amplification at these frequencies is a challenge for present day solid state technology. Even though it can be achieved, linear amplification will be power hungry because of the add-on linearization circuitry and/or the requirement to back-off the power amplifier; (2) any nonlinearities in amplification or in the process of transmission increase the transmission signal bandwidth and (3) carrier phase recovery is a difficult task at these high data rates and transmit frequencies making coherent demodulation difficult. The last difficulty suggests that non-coherent demodulation methods are preferable. The first two difficulties suggest that constant envelope modulation schemes are preferable. Therefore, in view of power conservation and nonlinearity problems, modulation methods which are constant envelope and which can be demodulated noncoherently are preferred. This leads to a consideration of differentially coded PSK or FSK signals. As is well known, compared to FSK signals, PSK signals lose the constant envelope property after band limiting, which in turn generates undesirable harmonics and intermodulation products after signal limiting, (mainly due to amplification at the transmitter). The effects of amplifier nonlinearities on PSK system performance have been extensively studied in the literature. Some nonlinearity studies relating to DQPSK- $\pi/4$ modulation scheme have been reported in [10], [20]. FSK schemes, which can be demodulated using simple noncoherent methods [5], [92], can be better choices. One such FSK modulation scheme is the well-known continuous phase modulation (CPM) [5], which is considered in this work. Therefore, the combination of noncoherent demodulation and equalization is the main objective in this chapter.

In Section 4.1, mathematical representation of some important forms of the CPM signals, such as continuous phase FSK (CPFSK) and Gaussian MSK (GMSK), are presented. In Section 4.2, a brief review of the performance of various CPM signalling schemes in typical fading environment, both in frequency non-selective and frequency selective channels, is presented. A review of various equalization methods for CPM signals, reported in the literature is presented in Section 4.3. All these reported methods require high computational complexity in the context of the present work. Therefore, we examine two novel methods in our search for a low complexity (or as low as possible) solution by considering: (1) noncoherent demodulation and equalization as a single operation (Section 4.4) and (2) noncoherent demodulation and equalization as separate operations (Section 4.5). The resulting optimum sequence estimation strategy when noncoherent demodulation and equalization are considered as a single operation, can be realized for the present scenario, though with high computational complexity. In Section 4.5, we derive suboptimum sequence detection schemes for noncoherently demodulated CPFSK signals, considering two types of noncoherent demodulation methods: (1) differential detection and (2) filter bank detection for noncoherently orthogonal signals. In Section 4.6, we present the computer simulation results for the proposed low-complexity receiver structures in a two path quasi-static channel. The system performance is also compared with a known system with QPSK signalling and a complex tap DFE at the receiver. Note that, the reference system is not attractive to implement in the present scenario. In Section 4.7 we present a performance comparison of the proposed systems with conantulation signalling using the performance results presented in [80] for a three path quasi-static channel. In Section 4.8, the performance of the proposed systems using measured channel responses reported in [33] is presented. The performance evaluation considers two cases: (1) without power control and (2) with power control. Section 4.9 concludes the work on wide band systems.

4.1 Continuous phase modulation

In continuous phase modulation schemes, the RF signal envelope is constant and the phase varies in a continuous manner. All CPM signals can be described by the following equation [5], [109].

$$s(t) = \begin{cases} \sqrt{\frac{2E_s}{T_s}} \cos \left[\omega_c t + 2\pi \sum_{i=0}^n I_i h_i q(t - iT_s) \right] & \text{for } nT_s \leq t < (n+1)T_s \\ 0 & \text{elsewhere} \end{cases} \quad (4.1)$$

where ξ_s , T_s are the symbol energy and symbol duration respectively and ω_c is carrier frequency in radians/s.

The data $\{I_i\}$ are M -ary data symbols, M even, taken from the alphabet $\{\pm 1, \pm 3, \dots, \pm(M-1)\}$; h_i is the modulation index, which may vary from interval (pre-defined) to interval; $q(\cdot)$ is the phase response function. CPM scheme is denoted by its phase response function or by its derivative $g(\cdot)$, the frequency response function. Depending on these response functions, various modulation schemes can be defined. An excellent overview of this subject can be found in [5]. Continuous phase FSK (CPFSK) is an important modulation scheme which belongs to this family. CPFSK is a constant h modulation with the frequency response function a rectangular pulse over a symbol duration, i.e.,

$$g(t) = \begin{cases} 1/(2T_s) & \text{for } 0 \leq t \leq T_s \\ 0 & \text{elsewhere} \end{cases} \quad (4.2)$$

Another important binary modulation scheme which belongs to the CPM family is Gaussian minimum shift keying (GMSK). GMSK is a binary CPFSK scheme with modulation index, h , equal to 0.5. The frequency response pulse is of infinite duration as shown below [5].

$$g(t) = \frac{1}{2T_s} \left\{ Q \left[2\pi B_b \frac{\left(t - \frac{T_s}{2}\right)}{\sqrt{\ln(2)}} \right] - Q \left[2\pi B_b \frac{\left(t + \frac{T_s}{2}\right)}{\sqrt{\ln(2)}} \right] \right\} \quad (4.3)$$

where

$$Q(t) = \frac{1}{\sqrt{2\pi}} \int_t^{\infty} e^{-x^2/2} dx \quad (4.4)$$

In practice, this pulse is truncated to span L_p symbol intervals. This pulse shape is used to achieve compact spectral characteristics which are controlled by the parameter B_b . GMSK is the recommended modulation scheme for the GSM standards.

It is relevant at this point to review the literature for the performance of the CPM signals in the channel types under consideration. Here the review is restricted mainly to noncoherent detection schemes.

4.2 Continuous phase modulation in a fading environment

The CPM signals are attractive for their well established and simple noncoherent demodulation methods. Frequency modulation is also less sensitive against ISI because of its *capture* properties [92]. The well known noncoherent demodulation methods are: (1) limiter discriminator (LD) or limiter discriminator integrator (LDI) and (2) differential detection [92]. Some innovative methods, like coherent differential detection, were reported in [56]. The effects of Rayleigh fading on CPM signals for LD, LDI and DD were evaluated analytically in various papers [13], [46]-[48], [70], [78], [79], [102], [124], [126], [127]. In [13], [46]-[48], [70], [102], [126], [127], the evaluation is limited to flat fading channels. These analytical methods, for evaluating the system error rates, consider the effect of adjacent symbols on the decision of the present data [77]. These effects are more significant when base band pulse shaping is considered. The pulse shaping is needed for various reasons such as (1) bandwidth conservation and (2) robustness against timing errors [56]. Asano and Pasupathy suggested a modified detection method called the fading magnitude-integrate, sample and dump processor in [13], where the LDI receiver is modified by including the effects of magnitude fading. This type of receiver is shown to perform significantly better compared to a LDI receiver in fast fading Rayleigh channels. Korn [78], [79] has evaluated the system performance in a multipath fading, cochannel interference and adjacent channel interference environment. Some computer simulation results for differential detection in the presence of flat fading and CCI were reported in [124]. In all the above reports the investigations are mostly limited to GMSK signals in a frequency non-selective fading channel. The baseband pulse shape introduces ISI in the base band signal. Since this ISI is introduced before the modulation process, this can be regarded as linear ISI at the output of the demodulator. In some of the above mentioned references DFE and MLSE were suggested to improve the system performance further by eliminating the linear ISI at the output of the demodulator. Note that, this ISI is not the result of multipath in the transmission medium.

Performance of noncoherently orthogonal FSK signals in a fading environment has been reported in [15], [34], [60], [98]. For noncoherently orthogonal FSK signals, the demodulator consists of a bank of filters each tuned to the discrete frequency followed by a square law/ envelope detector and an integrate and dump filter (IDF). The bank of filters can be replaced by correlators matched to the symbol waveforms. The filters are preferred to safeguard against Doppler frequency shifts and local oscillator frequency instabilities [34]. Filters are also preferred for receiver implementation cost reduction. These receiver structures are particularly attractive for the present research because the subsystems are simple to implement.

4.3 Equalization of CPM signals

As can be seen from (4.1), CPM signals are nonlinearly related to the information symbols. When these modulated signals are transmitted through time-dispersive channels, ISI may result in two different ways: (1) ISI due to pulse shaping and (2) ISI due to the transmission medium. In a linear modulation method, such as PSK, ISI resulting from pulse shaping remains linear after modulation. Therefore, ISI due to the pulse shaping filter and the transmission channel can be combined. In contrast, in CPM systems, the ISI resulting from pulse shaping becomes nonlinear after modulation. For example, refer to Fig. 4.1, where the nature of the ISI present at various levels of modulation/ demodulation is depicted. The modulator converts the linear ISI introduced by the pulse shaping filter, $f_L(I_k)$, where $I_k = \{I_k, I_{k-1}, \dots, I_{k-L_p+1}\}$ and $L_p T_s$ represents the span of the pulse shaping filter, into some nonlinear function $z_k = f_{NL}(f_L(I_k))$. When this signal is passed through a channel, which spans $L T_s$ symbols, the resulting symbols can be represented as $g_L(z_k)$, where $g_L(\dots)$ is a linear function representing the channel ISI and $z_k = \{z_k, z_{k-1}, \dots, z_{k-L+1}\}$. In the absence of ISI from the transmission medium, the demodulator compensates (not in the strict sense, since not all the nonlinear functions can be inverted in practice) the nonlinearity introduced by the modulator. Therefore, a linear transversal filter can be used to improve the system performance in this situation. In the presence of frequency selective fading the linear ISI introduced by the channel is converted into nonlinear ISI by the demodulator, which cannot be removed by a transversal filter. On the other hand, linear equalization can be employed before demodulating the signal. But this type of equalization for FSK signals requires high sampling rates, thus increasing the computational complexity as discussed in Chapter 2.

Various reports on equalization for noncoherently demodulated FSK signals (in particular for GMSK) were reported in the earlier literature. Some of the related work can be found in [1], [6], [21], [41], [64], [75]. Most of this reported work addresses equalization of the ISI introduced by the pulse shaping filter. The combined effect of the pulse shaping filter and the multipath channel are considered in [1], [64]. In [64], a complex tap reduced-state maximum likelihood sequence estimator (MLSE) is suggested for the GSM standard system, where the modulation scheme is Gaussian minimum shift keying (GMSK). The proposed Viterbi processor operates on the complex low pass waveform, which is approximated by four equally spaced samples. The strategy reported in [1] consists of a complex tap decision feedback equalizer which operates on the input signal of the limiter discriminator and the reconstructed signal of the final decision at the output of the limiter discriminator. Both these methods require very high processing speeds.

Fig. 4. 1. Characteristics of ISI in FSK signals with noncoherent demodulation

The sampling frequency required increases with an increase in the size of the symbol alphabet (i.e., for M -ary CPFSK as M increases). A different strategy was adopted in [21], [118], where the decision criterion is changed adaptively at the output of the limiter discriminator. The decision boundary is, in general, a nonlinear function. But this method is only applicable for known ISI characteristics, such as a band limiting filter. In [18], an equalization strategy for differentially detected CPM signals is proposed. In this method, differential detectors with T_s and $2T_s$ detection delays were considered. However, a major assumption made in [18] is that the channel is always represented by a two path model where the paths are separated by a symbol period with varying power levels. This equalization method uses both the amplitude and phase information at the output of the differential detector (normally in a differential detector, the decision is made depending on the amplitude of the output of the differential detector). In the present work, we can not use this equalization method because the assumption made about the channel in [18] is not valid for the present work. The recommended equalization method may not work in real environments where the inter-path delay is not equal to a symbol period.

In this work, we want to investigate low complexity equalization methods, where the processing speed is comparable to the symbol rate and the computational complexity of the receiver, for the large dispersion delays, is as low as possible. Here, two types of detectors are suggested.

- (i) One approach is to derive an optimum detector considering demodulation and equalization as a single operation. This is investigated by deriving optimal/ suboptimal noncoherent detector for an ISI corrupted CPFSK signalling scheme.
- (ii) A second approach is a receiver with a noncoherent demodulator followed by an equalizer. Here, the demodulation and the equalization are separate operations. Therefore, the equalizer should be able to detect the data symbols acting on the output of the demodulator.

In the following sub-section, we investigate the first type, where the demodulation and equalization are considered as a single operation. Here, we derive optimum coherent and noncoherent ML sequence estimators for CPFSK signals. The coherent ML sequence estimator is derived for comparison purposes. These derivations closely follow the methods suggested in [5], [103], [109], [118]. The optimum receivers, though not necessarily practical, provide a base line for evaluating the effectiveness of low-complexity equalizer structures.

4.4 Optimum ML sequence estimation of M-ary CPFSK signals corrupted by ISI in an AWGN channel

In this section, an optimum coherent/ non-coherent ML sequence estimator for M-ary CPFSK signals corrupted by ISI will be derived (M is the size of the symbol alphabet).

The complex low-pass equivalent representation of the M-ary CPM signal of (4.1) can be written as [109] ...

$$z(t) = \sqrt{\frac{\xi_s}{T_s}} \exp \left\{ j \left[2\pi f_d \int_0^t d(\tau) d\tau + \Omega \right] \right\} \quad (4.5)$$

where T_s is the symbol period, f_d is the frequency deviation, Ω is the initial phase of the carrier (i.e., at $t = 0$) and $d(t)$ is the pulse amplitude modulated (PAM) signal. Considering a finite isolated block of (N_B+1) symbols¹, $d(t)$ can be represented as follows.

$$d(t) = \sum_{n=0}^{N_B} I_n g(t - nT_s) \quad (4.6)$$

(I_n) are the amplitudes mapped from the binary information bit stream to $\{\pm 1, \pm 3, \pm 5, \dots, \pm(M-1)\}$ and $g(t)$ represents the frequency shaping pulse. When $g(t)$ is a rectangular pulse of unit amplitude (as is the case in CPFSK signals), then $z(t)$ can be expressed as follows.

$$z(t) = \sum_{n=0}^{N_B} z_n(t) \quad (4.7)$$

where ...

$$z_n(t) = \begin{cases} \sqrt{\frac{\xi_s}{T_s}} \exp \left\{ j \left[\frac{\pi h}{T_s} I_n (t - nT_s) + \pi h \alpha_n + \Omega \right] \right\} & \text{for } nT_s \leq t \leq (n+1)T_s \\ 0 & \text{elsewhere} \end{cases} \quad (4.8)$$

1. The block is assumed to be preceded and followed by a guard time exceeding the channel delay spread.

$$h = 2f_d T_s \text{ and } \alpha_n = \sum_{k=0}^{n-1} I_k.$$

Notice that, $z_n(t)$ is also a function of the present symbol I_n and the accumulated phase α_n , which is a function of $\{I_{n-1}, I_{n-2}, \dots, I_0\}$. Here, we assume that the initial accumulated phase, α_0 as zero. When this CPFSK signal is passed through a channel, whose impulse response is given as $c(t)$, then the received signal, $r(t)$, can be represented as follows.

$$r(t) = e^{j\phi} z(t) \otimes c(t) + n(t) \quad (4.9)$$

The parameter ϕ represents the uncertainty in the carrier phase. Without loss of generality, the initial phase, Ω , can be absorbed into ϕ . $n(t)$ represents the additive white Gaussian noise

$$r(t) = \sum_{n=0}^{N_B} e^{j\phi} X_n(t) + n(t) \quad (4.10)$$

where ...

$$X_n(t) = \begin{cases} z(t) \otimes c(t) & \text{for } nT_s \leq t \leq (n+1)T_s \\ 0 & \text{elsewhere} \end{cases} \quad (4.11)$$

From (4.8) and (4.11), $X_n(t)$ is dependent on the information sequence $I_n = \{I_n, I_{n-1}, \dots, I_{n-L}\}$ and α_{n-L} , where the channel is assumed to span L symbol durations. Incorporating α_{n-L} , which is a function of the past symbols, $X_n(t)$ can be represented as $X_n(t, I_n, \alpha_{n-L})$. To simplify the representation, the dependency on I_n is removed in the following analysis. However, at an appropriate time, the dependency is reinstated.

4.4.1 Coherent detection

For coherent detection, ϕ is assumed to be known at the receiver. The parameter α_n is dependent on the previous data bits, therefore, this is a part of the estimation process.

Using the complex valued Karhunen-Loève series expansion for $r(t)$, based on a complete orthonormal set of complex functions $\{f_k(t)\}$ in $L_2[0, N_B T_s]$, we obtain the

following expressions (where $L_2[0, N_B T_s]$ represents a function space of finite energy signals in the time interval $[0, N_B T_s]$).

$$r(t) = \lim_{K \rightarrow \infty} \sum_{k=0}^K r_k f_k(t) \quad (4.12)$$

where...

$$r_k = \int_0^{N_B T_s} r(t) f_k(t) dt \quad (4.13)$$

$$r_k = \sum_{n=0}^{N_B} e^{j\phi} X_{nk} + n_k \quad (4.14)$$

where

$$X_{nk} = \int_0^{N_B T_s} X_n(t) f_k(t) dt = \int_{nT_s}^{(n+1)T_s} X_n(t) f_k(t) dt \quad (4.15)$$

and

$$n_k = \int_0^{N_B T_s} n(t) f_k(t) dt \quad (4.16)$$

Assuming that the thermal noise is a white Gaussian random process, the joint conditional probability density function of the observations $\{r_k\}$ can be expressed as follows [109].

$$p(r/l, \phi) = \frac{1}{(\sqrt{2\pi N_0})^K} \exp \left[- \sum_{k=0}^K \frac{\left| r_k - \sum_{n=0}^{N_B} X_{nk} e^{j\phi} \right|^2}{2N_0} \right] \quad (4.17)$$

where $N_0 = E[n_k^2]$, which represents the noise power spectral density.

Then the log likelihood function, $\Gamma(\phi)$, as K approaches infinity, can be expressed as follows:

$$\Gamma(\phi) = \int_0^{N_b T_b} \left| r(t) - e^{j\phi} \sum_{n=0}^{N_b} X_n(t) \right|^2 dt \quad (4.18)$$

which can be simplified as follows [109].

$$\begin{aligned} \Gamma(\phi) = & \int_0^{N_b T_b} |r(t)|^2 dt - 2 \operatorname{Re} \left[\int_0^{N_b T_b} r(t) \sum_{n=0}^{N_b} X_n^*(t) e^{-j\phi} dt \right] + \\ & \int_0^{N_b T_b} \sum_{n=0}^{N_b} \sum_{m=0}^{N_b} X_n(t) X_m^*(t) dt \end{aligned} \quad (4.19)$$

According to the definition (4.11), $X_n(t)$ and $X_m(t)$ are disjoint for $n \neq m$. Therefore, (4.19) can be expressed as ...

$$\Gamma(\phi) = \int_0^{N_b T_b} |r(t)|^2 dt - 2 \operatorname{Re} \left[\int_0^{N_b T_b} r(t) \sum_{n=0}^{N_b} X_n^*(t) e^{-j\phi} dt \right] + \int_0^{N_b T_b} \sum_{n=0}^{N_b} |X_n(t)|^2 dt \quad (4.20)$$

When all the transmitted symbols are equally probable and assuming that the block consists of a long sequence of symbols, the first and the third terms in the above equation can be considered to be constants. Therefore, (4.19) is simplified as ...

$$\Gamma(\phi) = \operatorname{Re} \left[\int_0^{N_b T_b} r(t) \sum_{n=0}^{N_b} X_n^*(t) e^{-j\phi} dt \right] \quad (4.21)$$

The above likelihood function has to be evaluated for all M^{N_b+1} possible sequences. Here, we assume that the composite waveforms, $\{X_n(t)\}$, are available at the receiver. $X_n(t)$ is defined by the state vector $\{I_n, I_{n-1}, \dots, I_{n-L}\}$ and the phase state

$\alpha_{n-L} = \pi h \left(\sum_{i=0}^{n-L-1} l_i \right) \bmod 2\pi$. Therefore, if the channel, represented by $c(t)$, spans L

symbol periods, then we can expect $\eta = pM^{L+1}$ different waveforms, where p represents the number of possible phase states. As given in [5], for $h = 2k/p$, where k and p are integers and prime to each other, there are p phase states with values $0, 2\pi/p, 4\pi/p, \dots, (p-1)2\pi/p$. The complexity of the receiver grows exponentially with the transmitted symbol block length. The likelihood function, independent of block length, can be efficiently calculated by using the Viterbi algorithm.

As N_B tends to ∞ , (4.18) can be expressed as an iterative equation as follows:

$$\Gamma_k^{(l \rightarrow m)}(\phi) = \Gamma_{k-1}^{(l)}(\phi) + \gamma_k^{(l \rightarrow m)} \quad (4.22)$$

where $\gamma_k^{(l \rightarrow m)} = \int_{kT_s}^{(k+1)T_s} |r_k(t) - X_k^*(t, S^{(l \rightarrow m)}, \alpha_{k-L}^{(l \rightarrow m)}) e^{j\phi}|^2 dt$ and $\Gamma_k^{(l \rightarrow m)}$ are the

branch metric and the corresponding cumulative branch metric for a transition from the l^{th} state at time $k-1$ to the m^{th} state at time k . The accumulated metric at the l^{th} state at time $k-1$ is represented by $\Gamma_{k-1}^{(l)}$. $r_k(t)$ represents the received signal in the n^{th} symbol interval (i.e., $nT_s \leq t \leq (n+1)T_s$). Notice that, the dependency of $X_n(t)$ on the information sequence and the accumulated phase is reinstated. $S^{(l \rightarrow m)}$ and $\alpha_{k-L}^{(l \rightarrow m)}$ represent the information sequence and the accumulated phase corresponding to the state transition from the l^{th} state at time $k-1$ to m^{th} state at time k (refer to the representation given in Chapter 2), respectively.

The above iterative equation can be efficiently implemented using the Viterbi algorithm. The structure represented by the above equation is identical to the one considered in [64]. During the training period, $X_k(t)$ is obtained from the training sequence.

During the detection mode, $X_k(t, S^{(l \rightarrow m)}) = c(t) \otimes f(S^{(l \rightarrow m)})$, where $f(\cdot)$ denotes the CPFSK modulation signal mapping operation. The integration in (4.22) can be approximated as a sum of a finite number of samples.

4.4.2 Non-coherent detection

For noncoherent detection, the phase parameter, ϕ , which is the sum of the initial phase, Ω , and the uncertainty in the carrier phase, is unknown at the receiver. In practice, ϕ varies very slowly with time due to frequency drift etc., but significantly over a block of (N_B+1) symbol intervals. We assume in the following derivation that the phase, ϕ , varies from symbol to symbol or in other words, ϕ can be represented by $\{\phi_n\}$, where n is the symbol index. The $\{\phi_n\}$ are assumed to be independent and uniformly distributed random variables.

As before, the joint conditional probability density function of the observations $\{r_k\}$ can be written as follows:

$$p(r/I, \{\phi_n\}) = \frac{1}{(\sqrt{2\pi N_0})^K} \exp \left[- \sum_{k=0}^K \frac{\left| r_k - \sum_{n=0}^{N_B} X_{nk} e^{j\phi_n} \right|^2}{2N_0} \right] \quad (4.23)$$

The dependency of $\{\phi_n\}$ can be removed by averaging the above probability density function over the statistics of $\{\phi_n\}$, i.e.,

$$p(r/I) = \left[\frac{1}{2\pi} \right]^{N_B} \int_0^{2\pi} \int_0^{2\pi} \dots \int_0^{2\pi} p(r/I, \{\phi_1, \phi_2, \dots, \phi_{N_B}\}) p(\phi_1) p(\phi_2) \dots p(\phi_{N_B}) d\phi_1 d\phi_2 \dots d\phi_{N_B} \quad (4.24)$$

The above integration can be simplified as follows [139], [109].

$$p(r/I) = \frac{1}{\sqrt{2\pi N_0}^K} \exp \left[\sum_{k=1}^K \frac{|r_k|^2 + \left| \sum_n \sum_m X_{nk} X_{mk}^* \right|}{2N_0} \right] \prod_{n=0}^{N_B} I_0 \left[\left| \sum_{k=1}^K \frac{r_k X_{nk}^*}{N_0} \right| \right] \quad (4.25)$$

Where $I_0(\cdot)$ is the zeroth order modified Bessel function of the first kind.

As $K \rightarrow \infty$, the summation in (4.25) approaches integration. Replacing the summation by integration and neglecting the constant and signal independent factors, the following likelihood function, Γ , is obtained.

$$\Gamma = \exp \left[\frac{1}{2N_0} \sum_n \int_0^{N_s T_s} \sum_m X_n(t) X_m^*(t) dt \right] \prod_{n=0}^{N_s} I_0 \left[\left| \frac{1}{N_0} \int_{nT_s}^{(n+1)T_s} r(t) X_n^*(t) dt \right| \right] \quad (4.26)$$

According to the definition (4.11), $X_n(t)$ and $X_m(t)$ are disjoint for $n \neq m$, i.e.

$$X_n(t) X_m^*(t) \neq 0 \text{ only for } n = m. \text{ So the term } \exp \left[\frac{1}{2N_0} \sum_n \int_0^{N_s T_s} \sum_m X_n(t) X_m^*(t) dt \right] \text{ is a}$$

constant. Therefore, the first term in the above sequence can be neglected since it is a constant. The ML metric simplifies to the following:

$$\Gamma = \prod_{n=0}^{N_s} I_0 \left[\left| \frac{1}{N_0} \int_{nT_s}^{(n+1)T_s} r(t) X_n^*(t) dt \right| \right] \quad (4.27)$$

Due to the monotonic nature of $I_0(x)$, the log likelihood function, Γ , can be expressed as follows [109], [139].

$$\Gamma = \sum_{n=0}^{N_s} \left[\left| \frac{1}{N_0} \int_{nT_s}^{(n+1)T_s} r(t) X_n^*(t) dt \right| \right] \quad (4.28)$$

This equation can be expressed as an iterative equation as follows.

$$\Gamma_n = \Gamma_{n-1} + \left| \int_{nT_s}^{(n+1)T_s} r(t) X_n^*(t) dt \right| \quad (4.29)$$

where Γ_n is the likelihood function at time nT_s .

Therefore, the optimum sequence detection is to correlate the received signal with $X_n(t)$, sum over all possible sequences and select the one with largest magnitude as the most probable transmitted sequence (Fig. 4.2). Note that, as in the case of a coherent detector, there are $\eta = pM^{L+1}$ different functions, $X_n(t)$, resulting from the M symbols and p accumulated phases (as stated on page 135). This receiver can be efficiently implemented using the Viterbi algorithm. When the channel is unknown, the channel is estimated using the delayed decisions at the output of the MLSE (as outlined in Chapter 2).

Fig. 4. 2. Block diagram representation of the optimum noncoherent ML sequence estimator

Special case: Optimum symbol by symbol detection:

For optimum symbol by symbol detection, (4.29) reduces to the following.

$$\Gamma_n = \left| \int_{nT_s}^{(n+1)T_s} r(t) X_n^*(t) dt \right| \quad (4.30)$$

i.e. $\Gamma(n)$ is calculated for all $X_n(t)$ and the symbol corresponding to the maximum metric is selected as the most probable transmitted symbol at time n . For orthogonal FSK, in the absence of channel dispersion, i.e., for non-fading additive white Gaussian noise channel, the number of states or equivalently the number of composite waveforms, $X_n(t)$, reduces to $\eta=M$ and the resulting receiver is the same as the classical filter-bank/ correlator detection receiver for FSK signals [109].

From the outcome of the above derivations, it can be concluded that the noncoherent receiver (Fig. 4.2) is practically realizable, but with high complexity. For channel dispersions of the order of one or two symbol periods, this type of receiver can be implemented with acceptable hardware complexity. The composite waveforms can be built at the receiver by representing each symbol waveform by a finite number of samples. The sampling frequency depends on the modulation index and the size of the symbol alphabet.

4.5 Suboptimum detection of noncoherently demodulated signals

Now, we turn our attention to considering demodulation and equalization as separate operations. It is well known that FSK signalling is associated with simple and efficient noncoherent demodulation techniques. However, these techniques are nonlinear in nature, thus the well known linear signal processing methods, such as linear equalizers are not applicable for post-demodulation processing. The linear filtering operations on the signal imposed by the channel do not remain linear at the output of the demodulator. Therefore, the equalizer succeeding the demodulator must be capable of handling nonlinear ISI, such as the ones discussed in Section 2.5.

To combat post-demodulation ISI, sub-optimal processing methods can be derived by observing the signal statistics at the output of the demodulator. The noncoherent demodulation methods considered are: (1) differential detection and (2) filter bank detection for non-coherently orthogonal FSK signals.

4.5.1 Scheme - I: Differential detector based receiver:

For CPFSK, one of the well known noncoherent demodulation strategies is differential detection. In differential detection, the received signal is multiplied by its delayed version as shown in Fig. 4.3. Due to the involvement of the delayed signal in the process of demodulation, the ISI introduced by the transmission medium is further increased. Thus, the complexity of the succeeding equalizer is also dependent on the delay used in the demodulation process. As shown in Fig. 4.3, if the delay, τ , is comparable to a symbol period, T_s , the ISI resulting from the channel is further increased by T_s . Therefore, for low complexity, smaller detection delays are recommended. In the earlier literature, fractional detection delays were recommended for improving the differential detector performance in the presence of frequency selective fading [7], [8]. Here, we recommend fractional delays to reduce the complexity of the succeeding equalizer. The probability density function of the signal at the output of the differential detector is required to derive an optimum detector. In the following paragraphs, we derive the signal statistics at the output of the demodulator.

The complex low pass equivalent of the received signal at the output of the receive filter, $r(t)$, can be expressed as follows.

$$r(t) = X(t) + n(t) \quad (4.31)$$

where ...

$$X(t) = \sum_{i=0}^{N-1} c_i z(t - \tau_i) \quad (4.32)$$

where $r(t)$, $X(t)$, and $n(t)$ are the low-pass equivalents of the received signal, the received information signal and the thermal noise, respectively. $\{c_i\}$ are the complex path gains for the combined impulse response of the transmit, receive filters and the channel (hereon referred to as the composite channel). The composite channel is assumed to have N paths and spans $\tau_c = \tau_{N-1}$ seconds. Therefore, the received signal at time t depends on the transmitted signal over $(t - \tau_c, t)$ or equivalently on the present and the past L information symbols, where $L = \lceil \tau_c / T_s \rceil$ and $\lceil \xi \rceil$ represents the smallest integer greater than ξ . τ_i is the composite channel path delay for the i^{th} path. $z(t)$ represents the low-pass equivalent of the CPFSK modulated information signal expressed as follows.

Fig. 4. 3. Differential detection of CPFSK signals

$$z(t) = \sum_{n=0}^{\infty} \sqrt{\frac{\xi_s}{T_s}} e^{j\Theta_n(t)} \quad (4.33)$$

ξ_s and T_s are the energy and time duration of a CPFSK symbol. $\Theta_n(t)$ is the accumulated phase of the carrier in the n^{th} symbol interval, defined as follows:

$$\Theta_n(t) = \begin{cases} \frac{\pi h}{T_s} I_n(t - nT_s) + \pi h \sum_{k=0}^{n-1} I_k & \text{for } (n-1)T_s \leq t \leq nT_s \\ 0 & \text{elsewhere} \end{cases} \quad (4.34)$$

where I_n represents the n^{th} information symbol.

The output of the differential detector, $v(t)$, is the product of the signal, $r(t)$, and its delayed version, $r(t-\tau)$, which can be expressed as in [70], [132].

$$v(t) = \frac{1}{2} \text{Re}[-jr(t)r^*(t-\tau)] \quad (4.35)$$

where the superscript $(*)$ represents the complex conjugate operation and $\text{Re}[x]$ and $\text{Im}[x]$ represent the real and imaginary parts of x .

The integrate and dump operation can be approximated by a sum of a finite number of independent samples of $v(t)$ over one symbol period. In particular, if the receive filter has a bandwidth W , then the output of the integrate and dump filter (IDF) can be expressed as a sum of WT_s samples of $v(t)$ over T_s , a symbol period. $v(t_k)$, sample of $v(t)$ at time t_k ($0 < t_k < T_s$) can be expressed as...

$$v_k = v(t_k) = \frac{1}{2} \text{Re}[-jr_{k1}r_{k2}^*] \quad (4.36)$$

where...

$$r_{k1} = r(t_k) = X_{k1} + n(t_k) \quad (4.37)$$

$$r_{k2} = r(t_k - \tau) = X_{k2} + n(t_k - \tau) \quad (4.38)$$

where

$$X_{k1} = \sum_{i=0}^{N-1} c_i z(t_k - \tau_i) \quad (4.39)$$

$$X_{k2} = \sum_{i=0}^{N-1} c_i z(t_k - \tau - \tau_i) \quad (4.40)$$

Note that, X_{k1} and X_{k2} are functions of the transmitted symbol sequence, $I_k = \{I_k, I_{k-1}, \dots, I_{k-L}\}$. For the following analysis, we do not explicitly indicate this dependency in our representation. However, at an appropriate time, we reinstate this dependency in the representation.

The probability density function of v_k can be obtained following Stein's method [132] or Turin's method [136]. Here we follow Turin's method. v_k can be expressed as a linear transformation on b .

$$v_k = b^H Q b \quad (4.41)$$

where the superscript H represents the Hermitian transpose and the vectors b and Q are defined as follows.

$$b = \begin{bmatrix} r_{k1} \\ r_{k2} \end{bmatrix} \quad (4.42)$$

$$Q = \begin{bmatrix} 0 & j/4 \\ -j/4 & 0 \end{bmatrix} \quad (4.43)$$

The auto-covariance matrix of b , K_b , is given as ...

$$K_b = E[(b - X)(b - X)^H] \quad (4.44)$$

where ...

$$X_k = \begin{bmatrix} X_{k1} \\ X_{k2} \end{bmatrix} \quad (4.45)$$

and using the following relations ...

$$E[(r_1 - X_{k1})^* (r_2 - X_{k2})] = \sigma_n^2 \rho_n^*(\tau) \quad (4.46)$$

$$E[(r_2 - X_{k2})^* (r_1 - X_{k1})] = \sigma_n^2 \rho_n(\tau) \quad (4.47)$$

where $\sigma_n^2 = E[|n(t_k)|^2] = E[|n(t_k - \tau)|^2]$ and $\rho_n(\tau) = \frac{E[n(t) n^*(t - \tau)]}{\sigma_n^2}$.

(4.44) can be simplified as follows.

$$K_b = \sigma_n^2 \begin{bmatrix} 1 & \rho_n(\tau) \\ \rho_n^*(\tau) & 1 \end{bmatrix} \quad (4.48)$$

Then, the inverse matrix of K_b is given as ...

$$K_b^{-1} = \frac{\sigma_n^2}{\sigma_n^4 (1 - |\rho_n(\tau)|^2)} \begin{bmatrix} 1 & -\rho_n(\tau) \\ -\rho_n^*(\tau) & 1 \end{bmatrix} \quad (4.49)$$

In [139, Prob. 4. 4. 3., p. 396], [136], it was shown that the characteristic function for this type of Hermitian form of Gaussian random variables (when Q is a Hermitian matrix), is given by the following equation:

$$\Psi_{v_k}(\omega) = E[e^{j\omega v_k}] \quad (4.50)$$

$$\Psi_{v_k}(\omega) = |I_{22} - j\omega K_b Q|^{-1} \exp[-X_k^H K_b^{-1} \{I_{22} - (I_{22} - j\omega K_b Q)^{-1}\} X_k] \quad (4.51)$$

where I_{22} is the identity matrix of order 2×2 . After simplification, the characteristic function of v_k , $\Psi_{v_k}(\omega)$, can be shown to be ...

$$\Psi_{v_k}(\omega) = \frac{1}{1 + jC\omega + D\omega^2} \exp \left[\frac{-(A\omega^2 + jB\omega)}{1 + jC\omega + D\omega^2} \right] \quad (4.52)$$

where ...

$$A = (|X_{k1}|^2 + |X_{k2}|^2 - 2\operatorname{Re}\{X_{k1}^* X_{k2} \rho_n(\tau)\}) \frac{\sigma_n^2}{16} \quad (4.53)$$

$$B = -\frac{1}{2} \operatorname{Re}\{-jX_{k1} X_{k2}^*\} \quad (4.54)$$

$$C = -\frac{\sigma_n^2}{2} \operatorname{Im}[\rho_n(\tau)] \quad \text{and} \quad D = \frac{\sigma_n^4}{16} (1 - |\rho_n(\tau)|^2) \quad (4.55)$$

Therefore, the conditional probability density function of v_k given X_k can be found by evaluating the following integral. Note that, X_k is a function of I_k , the transmitted symbol vector.

$$p_Y(v_k/X_k) = \int_{-\infty}^{\infty} \frac{1}{1 + jC\omega + D\omega^2} \exp \left[-\frac{\{A\omega^2 + jB\omega\}}{1 + jC\omega + D\omega^2} \right] \exp[-j\omega v_k] d\omega \quad (4.56)$$

The first and second conditional moments of v_k are given by ...

$$E[v_k^n/X_k] = -j^n \frac{\partial^n}{\partial \omega^n} \Psi(\omega) |_{\omega=0} \quad (4.57)$$

$$E[v_k/X_k] = -C - B = \frac{\sigma_n^2}{2} \operatorname{Im}[\rho_n(\tau)] + \frac{1}{2} \operatorname{Re}[-jX_{k1} X_{k2}^*] \quad (4.58)$$

$$E[v_k^2/X_k] = j^2 \{-2C^2 - 2D - BC - BC + 2BC - 2A - B^2\} \quad (4.59)$$

$$E[v_k^2/X_k] = 2C^2 + 2D + 2A + B^2 \quad (4.60)$$

Therefore, the conditional mean and variance of v_k are ...

$$m_{v_k} = \frac{\sigma_n^2}{2} \operatorname{Im}[\rho_n(\tau)] + \frac{1}{2} \operatorname{Re}[-jX_{k1} X_{k2}^*] \quad (4.61)$$

and

$$\begin{aligned} \sigma_{v_i}^2 = & \frac{\sigma_n^4}{8} (1 - |\rho_n(\tau)|^2) + \frac{\sigma_n^2}{8} [|X_{k1}|^2 + |X_{k2}|^2 - 2 \operatorname{Re} \{X_{k1}^* X_{k2} \rho_n(\tau)\}] + \\ & \frac{\sigma_n^4}{4} [\operatorname{Im} \{\rho_n(\tau)\}]^2 - \frac{\sigma_n^2}{2} \operatorname{Im} \{\rho_n(\tau)\} \operatorname{Re} \{-jX_{k1} X_{k2}^*\} \end{aligned} \quad (4.62)$$

Here, we limit the analysis for the white noise case (i.e., $\operatorname{Re}[\rho_n(\tau)] = \operatorname{Im}[\rho_n(\tau)] = 0$ for $\tau \neq 0$). Note that, this implies the receive filter in Fig. 4.3 does not introduce significant correlation in the thermal noise generated at the receiver sub-systems prior to filtering. Then the resulting mean and variances are ...

$$m_{v_i} = \frac{1}{2} \operatorname{Re} \{-jX_{k1} X_{k2}^*\} \quad (4.63)$$

$$\sigma_{v_i}^2 = \frac{\sigma_n^4}{8} + \frac{\sigma_n^2}{8} [|X_{k1}|^2 + |X_{k2}|^2] \quad (4.64)$$

Gaussian approximation:

When $W \gg 1/T_s$, the probability density function of y , the output of the integrate and dump filter, which can be expressed as $y_k = \sum_{j=1}^{WT_s} v(t_k + j/W)$, can be approximated by a Gaussian distribution.

$$p_{y_k/I_k}(y_k) = \frac{1}{\sqrt{2\pi\sigma_{y_k}^2(I_k)}} \exp \left[-\frac{(y_k - m_{y_k}(I_k))^2}{2\sigma_{y_k}^2(I_k)} \right] \quad (4.65)$$

where ...

$$m_{y_k}(I_k) = \frac{1}{2T_s} \operatorname{Re} \int_0^{T_s} -jX_{k1}(I_k;t) X_{k2}^*(I_k;t) dt \quad (4.66)$$

and

$$\sigma_{y_k}^2(I_k) = \frac{\sigma_n^4}{8} + \frac{\sigma_n^2}{8T_s} \int_0^{T_s} [|X_{k1}(I_k;t)|^2 + |X_{k2}(I_k;t)|^2] dt \quad (4.67)$$

Note that, in the above equations, the dependency of I_n is incorporated. From (4.65), the log likelihood metric for optimum detection can be written as follows [139].

$$\gamma_k = -\log[\sigma_{y_k}(I_k)] + \frac{(y_k - m_{y_k}(I_k))^2}{2\sigma_{y_k}^2(I_k)} \quad (4.68)$$

However, the branch metric defined in (4.68) is complicated for actual implementation. A sub-optimum sequence detector can be derived by re-writing (4.67) as follows.

$$\sigma_{y_k}^2(I_k) = \frac{\sigma_n^4}{8} \left\{ 1 + \left[\frac{\xi_1(I_k) + \xi_2(I_k)}{\sigma_n^2} \right] \right\} \quad (4.69)$$

where $\xi_1(I_k)$ can be defined as the received signal power averaged over a symbol period conditioned on a transmitted sequence I_k as follows.

$$\xi_1(I_k) = \frac{1}{T_s} \int_0^{T_s} |X_{k1}(I_k;t)|^2 dt \quad (4.70)$$

Similarly, $\xi_2(I_k)$ is the received signal power averaged over $(-\tau, T_s - \tau)$ conditioned on a transmitted sequence I_k . Note that, $\xi_1(I_k) \approx \xi_2(I_k)$ for $\tau \ll T_s$. With these interpretations, we can define a sub-optimum receiver. When $[\xi_1(I_k) + \xi_2(I_k)]/\sigma_n^2 \ll 1$, then the variance of y_k , $\sigma_{y_k}^2$, does not depend on the transmitted symbol vector.

$$\sigma_{y_k}^2(I) \equiv \frac{\sigma_n^4}{8} \quad (4.71)$$

Using this approximation, the likelihood function of (4.68) can be simplified as follows.

$$\gamma_k = (y_k - m_{y_k}(I_k))^2 \quad (4.72)$$

From (4.72), the cumulative branch metric at time kT_s can be expressed as follows.

$$\Gamma_k = \Gamma_{k-1} + \{y_k - m_{y_k}(I_k)\}^2 \quad (4.73)$$

Therefore, the receiver can be implemented as shown in Fig. 4.4. The quantity $m_{y_k}(I_k)$ for all possible transmitted sequences, I_k , can be obtained before hand using the table look-up methods outlined in Chapter 2. A RAM following the differential detector, is trained initially using a known sequence. If the channel dispersion spans L symbols, the RAM has M^{L+1} addresses (assuming $\tau \ll T_s$). During the training period, the output of the demodulator corresponding to each input symbol sequence is adaptively learned and the values are stored in RAM, using the input sequence as the address. Using the least mean squares (LMS) algorithm, the update process can be expressed as follows.

$$m_{y_k}(I_k) = m_{y_{k-1}}(I_k) + \mu e_k \quad (4.74)$$

where $m_k(I_k)$ are the RAM content corresponding to the address I_k at time kT_s , μ is the step size for adaptation and $e_k = y_k - m_{k-1}(I_k)$, where y_k is the output of the demodulator at time kT_s . As mentioned in Chapter 2, this table look-up method requires at least $M^{L+1} + L + 1$ training symbols.

In the tracking mode, the Viterbi algorithm is used to detect the symbols using the values stored in the RAM. Note that the MLSE has M^L states in contrast to M^{L+1} addresses of the RAM. The branch metric, $\gamma_k^{(l \rightarrow m)}$, used in the sequence detection is calculated as follows:

$$\gamma_k^{(l \rightarrow m)} = |y_k - m_k(S^{(l \rightarrow m)})|^2 \quad \text{for } l = 1, 2, \dots, l_M \quad (4.75)$$

where $S^{(l \rightarrow m)} = \{s_{k-1}^{(l)} \cup s_k^{(m)}\}$ and as before l_M represents the number transitions from each previous state.

As the expected channel variations are slow, a very small step size in the tracking mode will suffice. This receiver configuration is represented as DDME($a, [b]$), where a and

Fig. 4. 4. Suboptimal sequence detection of differentially demodulated CPFSK signals (DDME)

b represent the number of states of the MLSE and the number of RAM addresses, respectively.

4.5.2 Scheme II: Filter bank detector based receiver for non-coherently orthogonal FSK

When the symbol waveforms are non-coherently orthogonal, the optimum receiver in additive white Gaussian noise (AWGN) environment in maximum likelihood sense is the well known filter bank detector as shown in Fig. 4.5 [109]. The penalty for using orthogonal waveforms, relative to the differential detector shown in Fig. 4.3, is an increase in the required transmission bandwidth. Similar to the previous analysis, we derive the signal statistics at the output of the filter bank detector.

The signals at the output of the band pass filters, $r_i(t)$ for $i=1, \dots, M$, can be expressed as follows.

$$r_i(t) = X_i(t) + n_i(t) \quad \text{for } i = 1, \dots, M \quad (4.76)$$

where ...

$$X_i(t) = r(t) \otimes r_{xi}(t) = \sum_{j=0}^{N-1} c_j z(t - \tau_j) \otimes r_{xi}(t) \quad (4.77)$$

where $\otimes$ represents the convolution operation, $r(t)$ is the low pass equivalent of the received signal and $n_i(t)$ is the AWGN for each of the branches. $r_{xi}(t)$ is impulse response of the i^{th} receive filter. The output of the square detectors at time t_k can be expressed as follows ...

$$v_{ik} = r_{ik}(r_{ik})^* \quad (4.78)$$

where ...

$$r_{ik} = r_i(t_k) = X_i(t_k) + n_i(t_k) \quad (3.18)$$

whose probability density function is given as [109] ...

$$p(v_{ik}) = \frac{1}{2\sigma_n^2} e^{-\frac{(R_{ik}^2 + v_{ik})}{2\sigma_n^2}} I_0\left(\frac{\sqrt{v_{ik}} R_{ik}}{\sigma_n^2}\right) \quad (4.79)$$

Fig. 4. 5. Filter bank detection of noncoherently orthogonal 4 level CPFSK signals

where $I_k(x)$ is the k^{th} order modified Bessel function of first kind and $R_{ik}^2 = |E\{r_{ik}\}|^2 = |X_{ik}|^2$, $X_{ik} = Y_i(t_k)$.

In the following analysis, we assume that the receive filters are ideal brickwall filters of bandwidth W . Therefore, the impulse responses of these filters have WT_s zeros every T_s seconds. Hence, the output of the IDF over one symbol period can be represented by a sum of WT_s samples, which are statistically independent of each other. Assuming a modulation index of $2n$, where n is an integer, each band pass filter has a bandwidth of $2n/T_s$. The output of the IDF can be expressed as the sum of $2n$ samples, whose probability density is given as follows [109].

$$p_Y(y_{ik}) = \frac{1}{2\sigma_n^2} \left(\frac{y_{ik}}{F_{ik}} \right)^{\frac{2n-1}{2}} e^{-\frac{(F_{ik}^2 + y_{ik})}{2\sigma_n^2}} I_{2n-1} \left(\sqrt{y_{ik}} \frac{F_{ik}}{\sigma_n^2} \right) \quad (4.80)$$

where F_{ik} is given by the following equation.

$$F_{ik} = \sum_{j=0}^{WT_s} R_i(t_k + j/W) \quad (4.81)$$

where F_{ik} is the sum of $2n$ samples of R_{ik}^2 over a symbol period. Note that, F_{ik} is a function of the transmitted symbol vector. To simplify the representation, we do not explicitly indicate the dependency. However, we reinstate this dependency at an appropriate time.

For $n = 1$, which is the minimum possible value to satisfy the condition of orthogonality, the above expression becomes ...

$$p_Y(y_{ik}) = \frac{1}{2\sigma_n^2} \left(\frac{y_{ik}}{F_{ik}} \right)^{\frac{1}{2}} e^{-\frac{(F_{ik}^2 + y_{ik})}{2\sigma_n^2}} I_1 \left(\sqrt{y_{ik}} \frac{F_{ik}}{\sigma_n^2} \right) \quad (4.82)$$

Assuming that the y_{ik} for $i = 1, 2, \dots, M$ are independent random variables, their joint probability function is given as ...

$$p_Y(y_{1k}, y_{2k}, \dots, y_{Mk}) = \left(\prod_{i=1}^M \left[\frac{1}{2\sigma_n^2} \left(\frac{y_{ik}}{F_{ik}^2} \right)^{\frac{1}{2}} I_1 \left(\sqrt{y_{ik}} \frac{F_{ik}}{\sigma_n^2} \right) \right] \right) e^{-\frac{\sum_{i=1}^M (F_{ik}^2 + y_{ik})}{2\sigma_n^2}} \quad (4.83)$$

For sequence detection using the Viterbi algorithm, the cumulative branch metric at time kT_s can be expressed as ...

$$\Gamma_k = \Gamma_{k-1} + \gamma_k \quad (4.84)$$

where the branch metric at time kT_s , γ_k , expressed as the log likelihood function from (4.83), is given as ...

$$\gamma_k = \sum_{i=1}^M \left\{ \frac{1}{2} \ln[y_{ik}] - \ln[F_{ik}] - \frac{(F_{ik}^2 + y_{ik})}{2\sigma_n^2} + \ln \left[I_1 \left(\sqrt{y_{ik}} \frac{F_{ik}}{\sigma_n^2} \right) \right] \right\} \quad (4.85)$$

The sequence with the largest cumulative metric is selected as the most probable transmitted sequence. Due to the monotonic nature of $I_1(x)$, the branch metric of (4.85) can be simplified as follows [109], [139].

$$\gamma_k = \sum_{i=1}^M \left\{ \frac{1}{2} \ln[y_{ik}] - \ln[F_{ik}] - \frac{(\sqrt{y_{ik}} - F_{ik})^2}{2\sigma_n^2} \right\} \quad (4.86)$$

when compared to the exponential term, the terms $y_i^{1/2}$ and F_i^{-1} contribute negligible variation in the likelihood function, therefore these terms can be neglected resulting in a suboptimal receiver metric.

$$\gamma_k = \sum_{i=1}^M (\sqrt{y_i} - R_i)^2 \quad (4.87)$$

However, due to the negative sign in (4.86), the sequence with the smallest cumulative metric is selected as the most probable transmitted sequence. The receiver, as shown in Fig. 4.6 for $M = 4$, consists of the filter bank detector followed by a square-root operator and a memory based equalizer.

Fig. 4. 6. Suboptimal sequence detection of filter bank detected noncoherently orthogonal CPFSK signals

To obtain F as a function of the transmitted symbol sequence I , the equalizer, following the filter bank, is trained similar to the previous scheme.

$$F_{ik}(I_k) = F_{i(k-1)}(I_k) + \mu e_{ik} \quad (4.88)$$

where $e_{ik} = \sqrt{y_{ik}} - R_{i(k-1)}(I_k)$ and y_{ik} represents the output of i^{th} filter at time kT_s , i varies from 1 to M . In the data detection phase, the cumulative metric at time kT_s for all the possible transitions is found as follows.

$$\Gamma_k^{(l \rightarrow m)} = \Gamma_{k-1}^{(l)} + \sum_{i=1}^M \left| \sqrt{y_{ik}} - F_{i(k-1)}(S^{(l \rightarrow m)}) \right|^2 \quad \text{for } l = 1, 2, \dots, l_M \quad (4.89)$$

This receiver configuration is represented as FBME($a, [b]$), where a and b represent the number state of the MLSE and number addresses of the M dimensional RAM.

From the two approaches presented in Sections 4. 4 and 4. 5, clearly the receiver structures derived in Section 4. 5 are of lower complexity. However, if the channel dispersion spans many more than 2 or 3 symbols, the memory and training requirements may still become impractical. In the following sections we evaluate the proposed system of Section 4. 5 in different channel conditions. Since the optimal receivers of Section 4. 4 involve significant computational complexity for the present situations, we will not address these receivers further.

4. 6 System performance evaluation in a two path quasi-static channel

Before testing the performance of the proposed receivers using measured channel responses in the millimeter wave frequency band, their performance is evaluated in a two path quasi-static channel. The systems are compared with a standard system with respect to system transmission bandwidth and performance. The performance of the proposed systems is evaluated using computer simulations. Four level CPFSK ($M = 4$) transmission is considered for the system evaluations. The channel considered here consists of two paths, each complex Gaussian distributed, separated by a time delay of τ_c . The channel is assumed to be quasi-static.

A: Scheme - I: DDME

The differential detector followed by a memory based equalizer is shown in Fig 4.4. The RAM consists of 16 addresses and the MLSE consists of 4 states. The decision depth

of the Viterbi processor is 15 symbols. The symbol error rate (SER) is averaged over 1000 randomly selected channels. In these simulations, there is no receive filter. We assume perfect clock recovery. The equalizer is initially trained using a preamble of 1600 symbols (which is 100 times the number of addresses). In fact, as mentioned in Chapter 1, seventeen symbols, which cover all the states, are enough for training. During the detection phase, the contents of the RAM are frozen.

B: Scheme - II: FBME

The receiver structure for noncoherent orthogonal signals, shown in Fig. 4.6, is modelled on the computer using ideal brick wall filters². The modulation index should be an even number for the orthogonal condition to be valid [109]. Here, there are 4 RAMs, each having 16 addresses. The MLSE consists of 4 states. The equalizer is trained using a preamble of 1600 symbols. During the detection phase, the contents of the RAM are frozen. The other parameters are as given in scheme-I, i.e., DDME.

C: Reference scheme:

For comparison, we also considered a system with QPSK transmission and combined coherent demodulation and equalization, as shown in Fig. 4.7. The equalizer for this case is a fractional $T_s/2$ spaced decision feedback equalizer (DFE). The receiver filter is a square root raised cosine (SRRC) filter with roll-off factor, β , 0.35. The equalizer considered here consists of $5 T_s/2$ spaced forward taps and one backward tap, denoted as $DFE(5,1)$. The five forward taps are close to optimal and the performance of the system does not change significantly with additional taps. Here, we assume perfect clock recovery. Note that, the realization of the reference system is a real challenge using existing technology. Therefore, practical realization of the reference system is currently unattractive at the frequencies and bandwidth of interest here.

D: Results

In this section, we present the performance results for the proposed receivers and compare them with the performance of the reference system. Fig. 4.8 shows the SER performance against the modulation index, h , for the differential detection based scheme, DDME. Here, the detection delay, τ , is $T_s/32$ and the inter-path delay, τ_c , is T_s . The results are shown for two values of average E_b/N_0 , 17dB and 26 dB. If the requirement is to

2. Here, we evaluated the system performance using correlator detectors also and found that the performance is the same for both these types.

Fig. 4. 7. Reference system block diagram

Fig. 4. 8. DDME system SER performance as a function of modulation index

contain the transmission bandwidth without sacrificing the system performance significantly, the region between about $h = 0.4$ and $h = 0.85$ is a reasonable compromise. The average SER is not a smooth function of the modulation index, h . This is expected because the distance properties of the CPFSK signals exhibit similar phenomenon [5]. Note that for lower values of the modulation index, the Gaussian assumption made in Section 4.5.1 (page 142) is not valid. This is the main reason for the improvement in system performance for large values of the modulation index.

Fig. 4.9 shows the system error rates as a function of the detection delay, τ . It is clearly evident from this figure that larger delays increase the channel delay spread further and exceed the equalizer capabilities. If the equalizer states and RAM memory are increased to span 3 symbols, i.e. 16 states and 64 addresses, then the system performance is improved even for larger detection delays. Therefore, there is an optimum delay for good system performance and low complexity. For an equalizer spanning 2 symbols, the optimum detection delay is between $2T_b/64$ and $4T_b/64$.

Fig. 4.10 shows the system performance as a function of the received SNR for $h = 0.45$ and $h = 0.85$. The transmission bandwidth requirements for $h = 0.45$ and $h = 0.85$ are shown in Fig. 4.11. The power density spectrum is estimated using Rader's method [110], which involves estimation of the auto-correlation function. The curves obtained using the theoretical equations, given in [109], plotted for comparison are overlapping on the simulation results. As shown in Figs. 4.10 and 4.11, the system transmission bandwidth can be reduced but with a performance penalty. Furthermore, DDME with $h=0.85$ gives comparable performance to the reference system at low SNRs. At larger SNRs, the conditional received signal power averaged over a symbol period, $\xi_1(I_k)$, as indicated in (4.69), is not negligible. Therefore, DDME receiver structure considered here is no longer optimum. Fig. 4.12 shows the system performance for various values of channel delay at an average E_b/N_0 of 17 dB. For comparison, the average SER results for the reference system are also shown. These results show that the proposed receiver performs very near to the reference scheme at low SNR. In addition, memory based equalizers are well-known to combat any degree of nonlinearity.

Fig. 4.13 shows the performance of the FBME receiver configuration. Here, the modulation index is 2, which is the minimum requirement for orthogonality. Each of the receive filters has a bandwidth of $1/T_b$, where T_b is the bit period. For comparison, we show the error rates for a system without an equalizer in the presence of one and two path quasi-static channels. The system performance is 2.5 - 3.0 dB inferior to the reference system performance at low SNRs. System performance is significantly improved relative to the flat fading case. Fig. 4.14 shows the spectral properties of the CPFSK signal for $h = 2$ and

Fig. 4. 9. System performance against the normalized detection delay (DDME)

Fig. 4. 10. System SER performance as a function of receiver SNR (DDME)

Fig. 4. 11. Spectral properties of the signalling schemes under consideration

Fig. 4. 12. System SER as a function of channel path delay (DDME)

Fig. 4. 13. System SER performance of FBME

compares it with the reference system. Fig. 4.15 shows the SER as a function of the channel delay, τ_c , at a SNR per bit of 17 dB. The important observation is that the system performance is unaffected by the channel delay. Even for a delay slightly more than $T_b/8$, the equalizer exploits the time correlation in the received signal.

From Figs. 4.10 - 4.13, it is clear that for a two-path quasi-static channel DDME is significantly better at low SNRs when compared to FBME. However, FBME performs better at high SNRs and is less sensitive to channel delay. Selection between the two proposed receivers depends on the particular system requirements.

The performance results of the three systems described above are compared under ideal conditions, without considering any nonlinearity in the process of transmission. System bandwidth and SNR requirements are listed in Table 4. 1. As noted above, the reference scheme is based on QPSK transmission with filtering and combined coherent demodulation and equalization. Linearized power amplifiers and carrier recovery are difficult to achieve at millimeter wave frequencies. By comparison, DDME and FBME are based on CPFSK, a constant envelope modulation and can employ power efficient class-C amplifiers. Further, the receivers in DDME and FBME use noncoherent demodulation and do not require carrier recovery. For personal communications, which are power limited, and considering the available technology in the millimeter wave frequency band, the proposed schemes are an attractive alternative. Results for DDME and FBME presented in this section are reported in [24], [26], [30].

4.7 Comparison with Commutation signalling

As mentioned in Sections 1. 1 and 1. 2, an alternative to equalization, commutation signalling can be used to defeat ISI [137]. A performance comparison of DDME and FBME with a system employing commutation signalling (CS) and DPSK modulation is presented in this section. Here the performance results reported in [80] for CS/ DPSK are used. In CS, similar to direct sequence spread spectrum systems (DS-SS), the information symbols are modulated using orthogonal or near to orthogonal codes. The number of codes used in a system depends on the expected channel dispersion. For example, if the channel ISI is expected to be over 4 symbols, any consecutive four information symbols are transmitted using four different codes which are near to orthogonal. In [80], codes of size 5 were used to obtain three different code words. More details are available in [80] and the references thereof. In [80], the performance results for CS/ DPSK were presented in the presence of a three path quasi-static channel in terms of average SER as a function of the channel delay. The channel is assumed to be consisting of two paths, $2 T_b^3$ apart and the third path as a variable. Therefore, a similar evaluation is carried out for DDME and FBME systems also.

As mentioned previously, the RAM should have M^{L+1} locations and the MLSE M^L states. Since the maximum channel dispersion assumed here is 2 symbol periods (i.e., four bit periods), we require 64 addresses and 16 states. The performance results are presented in Fig. 4.16. The comparative transmission bandwidth requirements are also listed. From these results, we conclude that the performance of the proposed systems is very close to CS/ DPSK. Since there are only three codes used in CS/ DPSK, this system cannot defeat ISI involving the fourth symbol, during which one of the three codes are repeated. Another observation we can make from these results is that unlike CS/ DPSK, the proposed systems are not sensitive to the inter-path delay of the channel.

4.8 System performance using measured channel responses

In practice, at high data rates of the order of 20 Mbps and higher, the channel dispersion spans many more than one symbol period, say q symbol periods. As we have seen in the previous sections, the memory requirements for the equalizer increase with an increase in the channel dispersion. In a typical indoor environment, an approximate exponential power delay profile is expected, i.e. the effect from the distant past symbols is relatively low on the present symbol. The well-known capture properties [92] of FSK signalling may assist in suppressing the residual ISI arising from the distant past symbols. Therefore, we use an equalizer, which spans only the initial q_M ($< q$) symbols out of q symbols, where q_M is typically 2 or 3. In this section, we examine the system performance using experimentally obtained channels.

4.8.1 Channel measurements

The experiments were conducted in an open office at Bell-Northern Research (BNR) Ltd., Ottawa by Bultitude et. al [33]. The measurement area is a furnished open-concept office with 1.22 m high semi-permanent partitions dividing many small work spaces. The top view of the office floor is as shown in Fig. 4.17. The false ceiling is at a height of 4 m above the floor level. The experiments are conducted using two types of transmitter antennas; omni directional and horn. The horn antenna has a gain of 23 dB and a 3 dB beam width of 12 deg. The receiver, representing the portable, is always equipped with an omni directional antenna. The transmitter and receiver are placed at a height of 2.26 m and 1.27 m above the floor level, respectively. The position of the transmitter, representing the base station, is placed 1 m from the side wall as shown in the figure. The channel frequency responses were obtained using the swept frequency technique [135].

3. Since the results reported in [80] are expressed in terms of T_b , the bit duration, it is convenient to represent the results for the proposed systems also in terms of T_b . Note that, $T_b = T/2$ for the proposed systems.

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PM-1 3 1/2" x 4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

Fig. 4. 14. Spectral properties of the signalling schemes under consideration

Fig. 4. 15. System SER as a function of channel path delay

Table 4.1 System comparison under ideal conditions

System	99% Bandwidth	Required average E_b/N_0 for an Average SER of 1×10^{-3}
DDME: $h=0.85$	$4.4/T_s$	18.1
DDME: $h=0.45$	$3.3/T_s$	21.5
FBME	$8.0/T_s$	20.4
QPSK: $\beta = 35\%$	$1.36/T_s$	17.9

Fig. 4. 16. Comparison of CS/ DPSK and DDME, FBME schemes over three path quasi-static channel

Fig. 4. 17. Top-view of the office floor showing the receiver locations for horn to omni measurements

The experiments for the horn to omni case were repeated at 207 different locations within the room as shown in Fig. 4.17 at a transmission frequency of 40 GHz. The main lobe of the horn antenna is directed towards the receiver. The channel frequency responses at locations lying within the 3 dB beam width of the horn antenna are measured. The directions of the beam and the respective locations are indicated in Fig. 4.17. For the omni to omni case the measurements were repeated at 193 different locations as shown in Fig. 4.18. These measurements include both line-of-sight (LOS) and shadowed cases. The channel impulse responses are obtained from the reported frequency responses, which consist of 800 points over a frequency band of 800 MHz, after applying a Hamming window and interpolation. In extracting the impulse responses, the amplitude levels below a certain power level or threshold are considered as noise. This threshold level is a variable from location to location. The threshold level is decided first by conducting some experiments on the capture properties of the FSK signals. It was observed that the delayed paths whose power is 25 dB down from the main path do not affect the system performance. These experiments were carried out by considering a two path fixed channel. These experiments were repeated for various channel inter-path time delays. Therefore, when extracting the channel responses, those paths which are 30 dB (another 5 dB is added to the safe level) below the power of the main path, are considered as zero level. The impulse responses indicate that typically the delay spread is less than 200 ns but is as large as 200 ns (which corresponds to 32 bit periods at a data rate of 160 Mbps) in some cases. This is expected because of the dimensions of the room where the measurements were carried out. The root mean squared (RMS) value of the delay spread which is defined (in discrete time representation) by the following equation is calculated and plotted in Fig. 4.19.

$$\tau_{RMS} = \sqrt{\langle \tau_c^2 \rangle - \langle \tau_c \rangle^2} \quad (4.90)$$

$$\langle \tau_c^k \rangle = \frac{\sum_n \tau_{cn}^k \lambda_n^2}{\sum_n \lambda_n^2} \quad (4.91)$$

where λ_n , τ_{cn} are the amplitude and delay of the n^{th} path, respectively. Due to the finite number of channel measurements considered, the uncertainty in the prediction of RMS delay spread with 99% confidence is indicated in the table, adjacent to Fig. 4.19. The uncertainty is calculated following references [43], [142]. The results show that the RMS delay spread is not much different using omni-directional or directional antenna at the base station. The similarity between the horn and omni responses is probably a function of narrowness of the horn beam width and perhaps also the physical dimensions of the environment [59]. From hereon we consider only horn to omni measurements for system

Fig. 4. 18. Top-view of the office floor showing the receiver locations for omni to omni measurements

Fig. 4. 19. RMS delay spread obtained from the measured channels

performance evaluation. The propagation loss as a function of location is illustrated in Fig. 4.20. Details on these calculations can be obtained from [33].

4.8.2 Performance without power control [29], [30]

Using the obtained impulse responses, the receiver structures of Fig. 4.4 and Fig. 4.6 are simulated. In the system evaluations, the outage probability is considered as the performance criterion. Here, the outage probability is defined as the probability that at a particular location the achievable error rate is greater than a specified error rate. The simulations are run for four different data rates 20 Mbps, 40 Mbps, 80 Mbps and 160 Mbps. The measured channel impulse responses were used as is, without scaling, so that the results are influenced by spatial variation of the received signal power. To calculate the E_b/N_0 at each location, a fixed amount of noise is added to the channel output. The receive filters are ideal brick wall filters and the delay involved in sequence detection is 20 symbol durations.

For the suggested receiver structures, to cover the channel dispersion of 200 ns, the required RAM memory and the MLSE states are enormous as the data rate is increased above 20 Mbps. For a channel dispersion spanning τ_r sec, the required RAM memory and MLSE states are given as M^{L+1} and M^L respectively, where $L = \lceil \tau_r d_r / k \rceil$ where $k = \log_2 M$, M and d_r are the size of the symbol alphabet and the data rate in bits/sec, respectively, and $\lceil x \rceil$ represents the smallest integer greater than x . For example, for four level CPFSK (i.e., $M = 4$) at a data rate of 160 Mbps, the RAM memory and the MLSE states become respectively 4^{17} and 4^{16} for a delay spread of 200 ns, which are impractical. Here, we use an equalizer which spans only $q_M = 2$ symbol periods (i.e. 64 addresses and 16 states) for all the data rates. Therefore, for a data rate of 20 Mbps, the equalizer spans the full channel dispersion. For more than 20 Mbps, the ISI outside the equalizer span tests the capture properties of FSK. Fig 4.21 shows the results on outage probability for three different rates as a function of symbol error rate (SER).

For FBME, 47, 100, 193 and 206 locations out of 207 offer better than 1×10^{-4} SER performance for data rates of 160, 80, 40 and 20 Mbps respectively at an average E_b/N_0 of 30 dB. For DDME, 150 locations offer better than 1×10^{-4} SER performance for a data rate of 20 Mbps. For 20 Mbps, the performance is mainly dictated by spatial variations in the received signal power. For data rates above 20 Mbps, DDME is observed to perform poorly. From these results it is evident that the DDME receiver does not show any significant capture properties to suppress the ISI lying outside the equalizer span.

Fig. 4. 20. Propagation loss as a function of location

Fig. 4.21. Outage probability performance of DDME and FBME for different data rates

Therefore, in the subsequent sections, only FBME type of receivers are further evaluated. The outage probability shown previously does not give a complete representation of the system performance. To isolate the effects of spatial variation, which can be improved further by employing some type of micro diversity or power control, the system performance is interpreted as a function of location. Figs. 4.22, 4.23, 4.24, and 4.25 show the system SER as a function of portable location for data rates of 20, 40, 80 and 160 Mbps, respectively. As can be seen, at the locations far from the transmitter (i.e. near to the wall) the system performance is significantly bad.

4.8.3 Performance with power control [29], [30]

In the above evaluation, the effects of signal spatial variation were considered. In this section, we evaluate the possibility of sending high data rates assuming a constant E_b/N_0 at each location, which could be achieved with power control or some form of micro-diversity. This evaluation also gives additional insight into the ISI suppression properties of FSK signals. This is examined by defining a parameter, α , as the signal power lying within the equalizer span (which is 2 symbol periods) to the signal power lying outside the equalizer span. To isolate the effect of spatial variation, we assume that the E_b/N_0 at all the locations is 27 dB. Fig. 4.26 shows the SER at each location as a function of α . The scattered behavior of this plot shows the effect of the distribution of the signal power lying outside the capabilities of the equalizer. The upper bound shows that if approximately $1/100^{th}$ or less of the signal power is lying outside the equalizer span, irrespective of how the power is distributed, the SER is always better than 1×10^{-4} at an E_b/N_0 of 27 dB. These results also show that for some of the channels the performance is better than 1×10^{-4} for α as low as 0 - 5 dB. This is observed whenever the signal power, lying outside the equalizer span, is equally distributed among all the symbols. The 260 points that are lying on the X-axis have a SER of 1×10^{-6} or less. Of the 260 responses, 31, 70 and 159 are for 160, 80 and 40 Mbps, respectively. Therefore, 12%, 27%, 61% of the channels have a SER better than 1×10^{-6} for 160, 80 and 40 Mbps, respectively.

Using these results, at an E_b/N_0 of 27 dB, we plot the number of locations with better than a specified SER as a function of α as shown in Fig. 4.27. To obtain these results, we counted the number of locations with $\alpha > \alpha_d$ shown in Fig. 4.26, and the fraction of these locations with a SER better than the specified error rate are plotted against α_d . It is evident that 99% of the locations are better than 1×10^{-4} SER when α is greater than approximately 11.5 dB. From these results we can evaluate the feasibility of high data

Fig. 4.22 Performance of FBME: MLSE(16, 64) as a function of location at 20 Mbps

Fig. 4. 23. Performance of FBME-MLSE(16, [64]) as a function of location at 40 Mbps

Fig. 4. 24. Performance of FBME-MLSE(16, [64]) as a function of location at 80 Mbps

Fig. 4. 25. Performance of FBME-MLSE(16, [64]) as a function of location at 160 Mbps

Fig. 4. 26. System performance as a function of equalizer span

Fig. 4. 27. System performance as a function of α for a fixed received signal power

transmission using FBME in the environment where the measurements were carried out. Fig. 4.28 shows that approximately 90% of the locations offer better than 1×10^{-4} symbol error rate for data rates above 40 Mbps (if the E_b/N_0 at each of the locations is 27 dB).

4.8.4 Decision feedback sequence estimation (DFSE) without power control

In this section, we attempt to improve the system performance by modifying the proposed FBME receiver using decision feedback sequence estimation. The demodulator output at time kT_s is a nonlinear function of the linear product of the signal waveform vector, $z_k(t) = \{z_k(t), z_{k-1}(t), \dots, z_{k-L}(t)\}$, and the discrete time channel impulse response represented as $c_k = \{c_{k,0}, c_{k,1}, \dots, c_{k,L}\}$ at time kT_s as shown below.

$$y_k = f(z_k(t) \bullet c_k) \quad (4.92)$$

which can also be expressed as follows.

$$y_k = f\left(\sum_{i=0}^L c_{k,i} z_{k-i}(t)\right) \quad (4.93)$$

However, due to the approximate exponential decaying nature of the channel impulse responses, the nonlinear products between all the distant past symbols are not dominant. Using this argument, the equalizer is slightly modified to support higher data rate transmissions with more reliability. A decision feedback sequence estimator is configured as shown in Fig. 4.29. Here, we use additional RAMs of 16 addresses each to account for the nonlinearity between the present and the distant past symbols, which are not covered by the main RAM. The main RAM learns the nonlinearity among the present and past $q_M = 2$ symbols and the subsequent RAMs learn the nonlinear products between the present and q_{M+1}^{th} to q^{th} symbols. The MLSE still has 16 addresses as in the previous case. However, the past detected symbols with a decision depth of q_M symbol durations are used to subtract the nonlinearity between the present and the past symbols. Therefore the MLSE branch metric is modified as follows:

$$\Gamma_k^{(l \rightarrow m)} = \Gamma_{k-1}^{(l)} + \sum_{i=1}^M \left| \sqrt{y_{ik}} - F_{i(k-1)}(S^{(l \rightarrow m)}) - \sum_{j=1}^{q_{ii}} F_{i(k-1)}^{(j)}(\hat{I}_{n-q_M}) \right|^2 \quad (4.94)$$

Fig. 4. 28. Data rate handling capability for FBME with MLSE

Fig. 4. 29. Equalizer configuration for DFSE

where $F_{i(k-1)}^{(j)}$ are the extra RAMs designated for learning the nonlinearity between the present symbol and the past symbols which are not covered by the main RAM and $q_{ex} = q - q_M$. $\hat{I}$ represent the reconstructed symbols. Note that, the main RAM covers only the first q_M past symbols. The detected symbols are used to determine the state of the excess RAMs. Obviously, there will be error propagation, which affects the receiver performance. This type of receiver is denoted as DFSE($a, [b]$), where a represents the number states of the sequence estimator and b represents the total number of memory locations. Note that, these two parameters, a and b respectively equal to M^{q_M} and $M^{q_M+1} + q_{ex}M^2$.

As can be seen from Fig. 4.30, a DFSE type receiver improves the system performance significantly for data rates of 80 Mbps and 40 Mbps compared to the MLSE configuration (i.e., without the extra RAMs). The proposed DFSE configuration offers better than 1×10^{-4} SER at 142 and 201 locations out of 207 locations for 80 Mbps and 40 Mbps, respectively. Therefore, 42 more locations corresponding to 80 Mbps and 8 more locations corresponding to 40 Mbps are improved by using DFSE in the place of MLSE.

Figs. 4.31, 4.32 show the system SER for DFSE as a function of portable locations for 40 and 80 Mbps, respectively. The locations at which the performance has been improved by employing DFSE can be identified by comparing with Figs. 4.23, and 4.24. These figures show that at some locations the system performance is worsened by employing DFSE. This can be attributed to the error propagation due to the past erroneous decisions.

Implementation:

Here, we give implementation details of DFSE type of receiver structure for both DDME and FBME. MLSE receiver can be considered as a special case of DFSE receiver when q_{ex} is zero. Similar to the structure for MLSE-PSP, proposed in Chapter 3, the structural representation of the DFSE is shown in Fig. 4.33. The number of MLSE states are represented by p , which is equal to M^{q_M} .

Computational complexity:

The number of computations per symbol involved in DFSE is shown in Tables 4.2 and 4.3 for DDME and FBME, respectively. For example, for four level CPFSK signalling, assuming multiplications constitute the major part of the computational complexity, for

Fig. 4.30. Outage probability performance of FBME for MLSE and DFSE

- 1 - MLSE(16, [64]), 80 Mbps
- 2 - DFSE(16, [160]), 80 Mbps
- 3 - MLSE(16, [64]), 40 Mbps
- 4 - DFSE(16, [96]), 40 Mbps
- 5 - MLSE(16, [64]), 20 Mbps

Fig. 4. 31. Performance of FBME-DFSE(16, [64]) as a function of location at 40 Mbps

Fig. 4. 32. Performance of FBME-DFSE(16, 164) as a function of location at 80 Mbps

Fig. 4. 33. Structural representation of memory based DFSE

Table 4.2 Computational complexity/ symbol for DDME

Function/ Operation	real multiplications	real additions	comparisons
Training mode	$q_{ex}+1$	$2q_{ex}+1$	0
Decision mode (i) Sequence detection	ρM	$2\rho M+q_{ex}$	$\rho(1 \text{ out of } 4)$
(ii) Tracking	$q_{ex}+1$	$2q_{ex}+1$	0

Table 4.3 Computational complexity/ symbol for DDME

Function/ Operation	real multiplications	real additions	comparisons
Training mode	$q_{ex}+M$	$2Mq_{ex}+M$	0
Decision mode (i) Sequence detection	ρM^2	$2\rho M^2+q_{ex}$	$\rho(1 \text{ out of } 4)$
(ii) Tracking	$q_{ex}+M$	$2Mq_{ex}+M$	0

FBME - MLSE, i.e. $q_{ex} = 0$, with $q_M = 2$, the number of real multiplications required per symbol is 4 in the training phase and 260 in the detection phase. For a DFSE type receiver, the number of real multiplications is increased by the number of excess RAMs used. These figures are independent of data rate. The number of times a RAM is accessed during sequence detection depends on the implementation structure. For example, in Fig. 4.33, a single RAM (main) is shown to store the nonlinearity between the present and the q_M immediate past symbols for simple representations. However, if p RAMs are accessed in parallel, each with M addresses, each corresponding to one trellis state, the access speed can be reduced by a factor of p . Since this is done by parallel processing, there will be an increase in the power consumption. However, we then require only four accesses during a symbol period.

4.9 Summary and conclusions

In this chapter, optimal/ suboptimal receivers for noncoherently demodulated CPFSK signals were derived considering two types of noncoherent demodulators: (1) differential detector and (2) filter bank detector for noncoherently orthogonal signals. Compared to optimal/ suboptimal noncoherent sequence detection strategies derived in Section 4.4, these proposed receivers are less complex. A comparison of these two proposed receivers with a known reference system was carried out by considering performance and transmission bandwidth in a simple two path quasi-static channel. Though the proposed receivers require more transmission bandwidth compared to the reference system, the proposed receivers are preferred where nonlinear distortion and complexity are the main concerns as expected in the millimeter wave frequency band.

Performance comparison with a commutation signalling based system with DPSK modulation in a three path quasi-static channel shows that the proposed systems offer significantly better performance.

We also evaluated the performance of the proposed systems using measured channel responses. The channel measurements were carried out by Bultitude et. al. in a large open-concept office room, which represents a worst case scenario. The reported frequency responses were processed further to get the channel impulse responses. By using these measured channel responses, we include the effects of shadowing and signal spatial variations in the system simulations. The temporal fading of the channel is expected to be very slow because of the high transmission frequencies considered in this work. The results presented for FBME system with the equalizer spanning only 2 symbol intervals give acceptable performance up to a data rate of 40 Mbps. To support higher data rate transmission, either directional antennas with less than 12 deg beam width at transmitter

and/or receiver or an equalizer which spans more than 2 symbol periods is required or both. However, increasing the equalizer span increases the computational complexity. Therefore, we recommend the use of highly directive antennas for future work.

Chapter 5

Conclusions

In this thesis, adaptive equalization is considered as a pivotal signal processing technique to improve the performance of indoor wireless cellular systems. Low computational complexity and acceptable error rate performance are defined as the main objectives. Two distinctly different transmission scenarios are considered in this work: (1) low rate data transmission using the existing digital cellular radio standards at 900 MHz and (2) high rate data transmission above 20 Mbps in the millimeter wave frequency band (30 - 300 GHz). A major constraint imposed throughout this work due to the requirement of low complexity is that the nature of adaptive algorithm employed is limited to the simple LMS or its modified forms, such as NLMS.

The following conclusions can be derived from the results presented in the previous chapters. The conclusions are presented separately for narrow band and wide band systems in Chapters 3 and 4, respectively.

Narrowband systems:

System simulation results presented in Sections 3.3 - 3.9 show that delayed diversity transmission from the base station and equalization at the portable receiver improves the system performance in a typical indoor environment. As noted in Section 3.3, the main factors that affect the system performance are: (1) the transmission delay between the transmitting antennas (2) the equalization strategy used at the portable receiver and (3) the channel acquisition and tracking algorithms. The optimum transmission delay between the antennas is concluded as around a symbol period, considering the system performance and computational complexity. This is true for all the equalization methods considered in this thesis. In subsequent evaluations, we considered a full rate user with the IS-54 standard frame structure and diversity transmission from the base station with a relative transmission delay of a symbol period.

The channel tracking algorithm has been restricted to the simple LMS and its modified forms, such as the normalized LMS (NLMS) algorithm during the tracking phase to reduce the computational complexity. In the initial evaluations, the RLS algorithm is used for faster acquisition of the channel during a limited training period. The performance of various equalization strategies indicates that MLSE-PSP, Bayesian DFE and bi-directional DFE are the most promising equalization strategies at a maximum Doppler frequency of 6 Hz, which corresponds to a portable speed of 2 m/s at 900 MHz. Without

FEC coding, MLSE-PSP always outperforms all the other equalization methods, both in terms of error rates as well as computational complexity. However, in the presence of coding, less complexity is involved in extracting the soft decisions from a DFE or a Bayesian DFE than from a MLSE. Due to this advantage, Bayesian DFE achieves an FER of 2×10^{-2} at a lower SNR for the FEC coded system.

We defined two system configurations, one without FEC coding and the other with the recommended FEC coding. Note that compared to an uncoded system, the coded system has the following differences:

- (1) The equalizer operates at a 3 dB lower energy per symbol.
- (2) The user slot consists of half the information symbols compared to an uncoded system.

Therefore, to obtain a coding gain, the finite interleaver and FEC decoder must compensate for the performance degradation of the equalizer due to the 3 dB increase in noise relative to the uncoded system. Furthermore, coding should give a gain to account for the reduction in information rate by a factor of 2.

The recommended error correction coding for IS-54 standard frame structure is shown to provide marginal gains with respect to system frame error rates. This is true even at a Doppler frequency as low as 2 Hz. This comparison is based on the SNR per information symbol required to attain a frame error rate of 2×10^{-2} . Note that, this type of comparison might not do justice to the equalization, since the symbol energy is 3 dB less in the coded system. Therefore, the equalizer eliminates the ISI less efficiently compared to an uncoded system because of the presence of more noise per symbol. The marginal gains obtained can be improved by using soft decision decoding at the portable receiver, however, with an increase in the computational complexity.

Since our objective is to employ existing North American digital cellular standard system in an indoor environment, the equalizers evaluated above should also give acceptable performance in an outdoor environment. Note that this should be done using only the LMS algorithm. Computer simulation results show that bi-directional DFE and Bayesian DFE are significantly worse than MLSE-PSP in an outdoor environment, where the channel variations can exceed 90 Hz corresponding to a vehicle speed of 108 KMPH. Furthermore, MLSE-PSP is shown to acquire the channel blindly using the LMS algorithm within 30 symbol periods even at a vehicle speed of 108 KMPH.

From the results presented in Chapter 3, it can be concluded that the IS-54 standard portables equipped with the above concluded equalizers can be used for indoor wireless applications with diversity transmission. However, all the base stations should be equipped with dual antennas with one symbol period delay between the antennas.

Wideband systems:

From the previous literature, it is concluded that a constant envelope modulation with noncoherent detection is preferred for high rate data transmission in the millimeter wave frequency band in an indoor environment. Continuous phase modulation schemes, which preserve the constant envelope property even after nonlinear distortions, are considered for this work. For the expected channel dispersion, which varies from 10 ns to 200 ns, and at the data rates considered in this work, from 20 Mbps - 160 Mbps, severe intersymbol interference is the main concern. Some type of signal processing technique, such as adaptive equalization or commutation signalling or anti-multipath modulation, is required to cope with this situation. In this work, adaptive equalization is considered. However, some of the equalization techniques for CPFSK signals, reported in the earlier literature, require high processing speeds. Here, we considered equalization of CPFSK signals, without the prior knowledge of the carrier phase. This can be performed in two different ways: (1) considering noncoherent demodulation and equalization as a single operation and (2) considering equalization of noncoherently demodulated, ISI corrupted CPFSK signals. The analytical results indicate that for the first scheme, considering demodulation and equalization as a single operation, the optimum receiver is very complex. In the second scheme, equalization of the demodulated signals, optimum/ suboptimal sequence detection strategies are derived. This analytical work is done for two types of demodulation schemes: (1) differential detection with fractional symbol detection delays and (2) filter bank detection for noncoherently orthogonal signals. The suboptimal sequence detectors consist of a noncoherent demodulator followed by a memory based equalizer for both the differential detector as well as the filter bank detector. These two receivers are respectively represented as DDME and FBME. The performance of the obtained receivers is evaluated using computer simulations using two types of channel models: (1) a two path quasi-static channel and (2) experimentally obtained channel responses, measured in a large open-office. In the case of a two path channel, a well-known system with QPSK transmission and a conventional complex tap DFE is considered for comparison. Notice that, this reference system may not be attractive to implement at the transmission frequencies and data rates considered in this work. Both the proposed systems are shown to be comparable to the reference system. System performance and transmission bandwidth requirements are optimum for CPFSK signals with $h = 0.85$ and a detection delay of $3T_b/64$. FBME is shown to be unaffected by the channel path delays. For a channel

dispersion spanning τ_r sec, the required RAM memory and MLSE states for both DDME and FBME are M^L and M^{L-1} respectively, where $L = \lceil k\tau_r/d_r \rceil$, $k = \log_2 M$ and M and d_r are the size of the symbol alphabet and the data rate in bits/sec respectively, and $\lceil x \rceil$ represents the smallest integer greater than x . However, FBME requires M RAMs and DDME requires only one RAM.

The proposed systems, with a sub-optimal MLSE, are also examined in a typical indoor environment using measured channel frequency responses. The measured frequency responses were provided by Bultitude et. al. The large open-concept office considered for the experiments represents a worst case environment. The system performance is evaluated at four different transmission rates 20, 40, 80 and 160 Mbps. The performance evaluation is done in two different cases: (1) without power control, i.e. with spatial variation in the received signal power and (2) with power control. The results indicate that the FMBE with sub-optimal memory based MLSE gives acceptable performance up to a data rate of 40 Mbps. We considered an SER of better than 1×10^{-4} as an acceptable performance. A location based variable data rate system can be suggested from the results presented. Data rate can be selected depending on the channel dispersion (say τ_c sec) at a particular location. If the equalizer span is fixed to NT_s sec, T_s is adjusted such that α , the signal power within the equalizer span to the signal power outside the equalizer span, is equal to or greater than 11.5 dB (as concluded in 4.8.3). However, for this scheme, we require some type of power control mechanism.

5.1 Recommendations for further work:

In the case of narrowband systems, we assumed perfect timing recovery at the portable receiver. PSP can be used to recover the optimum sampling time, as shown in [74]. It is of interest to investigate the performance of MLSE-PSP receiver with combined timing, channel tracking and sequence detection¹.

The computational complexity of the MLSE-PSP receiver can be reduced by employing reduced state sequence estimation (RSSE) [51], [39] or decision feedback sequence estimation (DFSE) [44] methods. In the case of PSP, the error propagation, which is the main concern in the conventional RSSE/DFSE structure, does not affect the system performance significantly since in the feedback process, all the surviving sequences are considered. A limiting form of this approach would be a four state MLSE-PSP.

1. In Chapter 3, the effect of carrier recovery on the system performance is investigated. Note that, the carrier phase recovery is performed in the process of adaptive equalization.

In the case of wide band systems, we concentrated our work on differential detection and filter bank detection type of demodulators. Another well known noncoherent demodulation method is the limiter discriminator (LD) or limiter discriminator integrator (LDI). An optimum sequence detection strategy, similar to the work presented in this thesis, can also be derived for LD or LDI type of demodulators.

The results presented in Chapter 4 suggest that horn type directional antennas of 12 degrees beamwidth at the base station do not reduce the delay spread of channel sufficiently to support high speed data transmission. There are two alternatives to improve the system performance in this situation: (1) highly directive antennas with less than 12 deg beam width can be used to reduce the RMS delay spread, and thus increase the probability of higher data transmission with the sub-optimal MLSE spanning only two symbol periods and (2) directional antennas can be employed at both the base station and the portable (similar to Motorola's ALTAIR system). For evaluating the system performance using directive antennas at both the base station and the portable, the approach adopted in [81] is recommended.

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