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Polynomial Predictive Filters: Implementation and Applications

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Abstract

In this thesis, smoothness of sampled real-world signals is exploited through the application of polynomial predictive filters. The principal reason for employing the polynomial signal model is principally twofold: firstly, assuming that the sampling rate is adequate, all real-world signals exhibit piecewise polynomial-like behavior, and secondly, polynomial-based signal processing is computationally efficient. By definition, polynomial predictive filters provide estimates of future values of polynomial-like signals. Thus, the potential applications of this research include a vast number of different delay sensitive operations on measurements like temperature, position, velocity, or power, especially in control engineering field.

The polynomial-based predictive signal processing is a well-known technique, but polynomial-predictive filters have had severe drawbacks, which have hindered their application; their white noise attenuation is generally low, or they exhibit considerable passband gain peaks, rendering them unattractive for most applications. It has been possible to design IIR polynomial predictors, which exhibit applicable magnitude response properties, but the severe problem with them, as well as with the FIR polynomial predictors, has been that they have generally not been implementable in low-precision fixed-point environments because of their coefficient quantization sensitivity. In this thesis, coefficient quantization error-free designs of both FIR and IIR polynomial predictors are presented, thus providing methods for overcoming the above drawbacks and design problems.

Polynomial differentiators are closely related to polynomial predictors; they are derived in a similar fashion, have design problems of a similar nature, and have applications in the control field. Both of these two filter types are discussed in this thesis; the proposed design methods are applicable to both of them.

The implementation aspects of polynomial predictors and differentiators investigated here are also connected to the practical requirements of the application, namely delay alleviation in closed loop transmitter power control of multiuser mobile communications systems. Particularly, if predictive received power level estimation is implemented in handheld mobile terminals, this application specifies the implementation criteria as requirements on low imposed computational burden, low power consumption, and compact hardware size. All these criteria are met by providing the desired functionality using a small number of fixed-point arithmetic operations. Taking into account the results presented in this thesis, polynomial prediction fulfills these criteria.

In this thesis, digital filter design methodologies are advanced by first-time introduction of exact low-degree polynomial prediction and discrete time differentiation in low-precision fixed-point computing environments, with, for example, 8 or 16 bits. Polynomial prediction is shown advantageous in the closed loop transmitter power control system application, and in comparisons with more complex and flexible predictors, it is shown to be a highly efficient method for this particular application.

This thesis is seen as contributing to advances in practical polynomial predictor and differentiator design methods, and thereafter studies the application of polynomial predictors in mobile communications system transmitter power control. This research will be of interest to signal processing, control, and communications engineers and researchers alike.

Preface

This work has been carried out at Helsinki University of Technology (HUT), Espoo, Finland, both in Institute of Intelligent Power Electronics, and in Signal Processing Laboratory (formerly Laboratory of Signal Processing and Computer Technology (STI)). I am in great dept to many people, both in my scientific and social lives. First and foremost, I would like to express my sincere gratitude to my supervisor Professor Seppo Ovaska (Institute of Intelligent Power Electronics, HUT), and to Professor Iiro Hartimo (Signal Processing Laboratory, HUT). I have been very fortunate to work for Seppo and Iiro for six years. They have provided me with a free and just, with one word, fatherly environment, within which it has been good to grow and do research. They have guided me and paved my way in all possible ways.

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Espoo, November 22, 2000
Jarno Tanskanen

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List of Publications

This thesis consists of an introduction and the following publications, which are referred to by [P1] – [P9] in the text.

[P1] J. M. A. Tanskanen and S. J. Ovaska, "Coefficient sensitivity of polynomial-predictive FIR differentiators: Analysis," in *Proc. 42nd IEEE Midwest Symposium on Circuits and Systems*, Las Cruces, NM, USA, Aug. 1999, pp. 405–408.

[P2] J. M. A. Tanskanen and S. J. Ovaska "Coefficient sensitivity of polynomial-predictive FIR differentiators: Design for short word length," in *Proc. 42nd IEEE Midwest Symposium on Circuits and Systems*, Las Cruces, NM, USA, Aug. 1999, pp. 520–523.

[P3] J. M. A. Tanskanen and V. S. Dimitrov, "Round-off Error Free Fixed-Point Design of Polynomial FIR Predictors and Predictive FIR Differentiators," Institute of Intelligent Power Electronics, Helsinki University of Technology, Espoo, Finland, Helsinki University of Technology Institute of Intelligent Power Electronics Publications, Publication 4, Aug. 2000 [electronic publication]. Available from <http://wooster.hut.fi/publications/tanskanen/round_off_error_free_predictors.pdf>. Manuscript submitted to *Digital Signal Processing, A Review Journal*, Aug. 2000.

[P4] J. M. A. Tanskanen, "Coefficient quantization error free fixed-point IIR polynomial predictor design," in *Proc. 2000 IEEE Nordic Signal Processing Symposium*, Kolmården, Sweden, June 2000, pp. 219–222.

[P5] J. M. A. Tanskanen, O. Vainio, and S. J. Ovaska, "Adaptive general parameter extension for tuning FIR predictors," in *Proc. 2nd IFAC Workshop on Linear Time Delay Systems*, Ancona, Italy, Sept. 2000, pp. 42–47.

[P6] J. M. A. Tanskanen, A. Huang, T. I. Laakso, and S. J. Ovaska, "Prediction of received signal power in CDMA cellular systems," in *Proc. 45th IEEE Vehicular Technology Conference*, Chicago, IL, USA, July 1995, pp. 922–926.

[P7] J. M. A. Tanskanen, J. Mattila, M. Hall, T. Korhonen, and S. J. Ovaska "Predictive closed loop power control for mobile CDMA systems," in *Proc. 47th IEEE Vehicular Technology Conference*, Phoenix, AZ, USA, May 1997, pp. 934–938.

[P8] J. M. A. Tanskanen, A. Huang, and I. O. Hartimo "Predictive power estimators in CDMA closed loop power control," in *Proc. 48th IEEE Vehicular Technology Conference*, Ottawa, Ontario, Canada, May 1998, pp. 1091–1095.

[P9] X. M. Gao, X. Z. Gao, J. M. A. Tanskanen, and S. J. Ovaska, "Power prediction in mobile communication systems using an optimal neural-network structure," *IEEE Transactions on Neural Networks*, vol. 8, pp. 1446–1455, Nov. 1997.

The above publications are available from URL:
http://www.hut.fi/Yksikot/Kirjasto/Diss/2000/isbn9512256312/.

List of Abbreviations

CDMA	Code-Division-Multiple-Access
CIR	Carrier-to-Interference Ratio
COSSAP	Communications System Simulation and Analysis Program
DPCCH	Dedicated Physical Control Channel
DS	Direct-Sequence
FDD	Frequency Division Duplex
FIR	Finite Impulse Response
GP	General Parameter
GSM	Global System for Mobile communications
HUT	Helsinki University of Technology
IEE	The Institution of Electrical Engineers
IEEE	The Institute of Electrical and Electronics Engineers, Inc.
IFAC	International Federation of Automatic Control
IIR	Infinite Impulse Response
LMS	Least Mean Square
LOS	Line-Of-Sight
MSE	Mean-Square-Error
MUD	Multi-User Detection
PFP	Polynomial FIR Predictor
PPFD	Polynomial-Predictive FIR Differentiator
SIR	Signal-to-Interference Ratio
SNR	Signal-to-Noise Ratio
STI	Laboratory of Signal Processing and Computer Technology
WCDMA	Wideband Code-Division-Multiple-Access

List of Symbols

Notation of the introduction part may differ from that used in the publications [P1] – [P9] in which the used notation is defined publication by publication.

ω	normalized digital angular frequency
Δ	quantization step
$\Delta f\,(t)$	Doppler shift
Δp	power control step size
$\Delta t(t)$	propagation delay
γ	adaptation gain factor
$\beta(n)$	general parameter (GP)
$\varepsilon(n)$	quantization error
$\varphi(t)$	phase
σ_0^2	truncation noise variance
σ^2	total average roundoff noise power
ω_F	frequency point in normalized frequency for magnitude response value evaluation
ω_G	frequency point in normalized frequency for group delay value evaluation
$\cdot'$	corresponding quantity with a different value
$\cdot_i$	subscript "*i*" denoting the inphase component of a complex-valued baseband-equivalent signal
$\cdot_Q$	subscript "*Q*" denoting a quantized value
$\cdot_q$	subscript "*q*" denoting the quadrature component of a complex-valued base-band-equivalent signal
$\cdot^{\mathrm{T}}$	transpose
$\hat{\cdot}$	estimate
$A(m)$	IIR filter denominator coefficient
$A(t)$	signal amplitude
$\mathbf{b}$	IIR extension feedback coefficient vector (row vector)
$b(k)$	IIR extension feedback coefficient
$B(k)$	IIR filter numerator coefficient
d	total control loop delay in samples
$\mathbf{e}$	quantization error feedback filter coefficient vector (row vector)
$e(k)$	quantization error feedback filter coefficient
$\mathrm{E}[\cdot]$	expectation
$F(\omega_F)$	value of the magnitude response of an infinite precision coefficient filter at the frequency ω_F
f_c	carrier frequency
$F_Q(\omega_F)$	value of the magnitude response of a quantized coefficient filter at the frequency ω_F
$G(\omega_G)$	value of the group delay of an infinite precision coefficient filter at the frequency ω_G
g_i	constraint for the filter coefficients for the *i*th degree polynomial input signal
$G_j(e^{j\omega})$	transfer function from the *j*th summation node to the filter output
$g_j(n)$	impulse response due to $G_j(e^{j\omega})$

$G_Q(\omega_G)$	value of the group delay of a quantized coefficient filter at the frequency ω_G
$\mathbf{h}$	FIR filter coefficient vector (row vector)
$h(k)$	FIR filter coefficient
$H(z)$	transfer function of a filter
$\mathcal{H}, \mathcal{H}_8$	quantized coefficient space, 8-bit quantized coefficient space
i	degree of a polynomial
I	maximum degree of a polynomial
$\mathbf{I}$	unit vector of length N (column vector)
j	summation node index in a filter structure, square root of –1
k	feed forward filter coefficient index, discrete time instant modifier
m, m_j	feedback filter coefficient index, number of multipliers feeding into the jth summation node
n	discrete time instant
N	filter length
N_e	quantization error feedback filter length
NG	white noise gain
p	forward prediction step in samples
$P(\omega)$	roundoff noise power spectrum at filter output
Q_j	arithmetic operation result quantizer associated with the jth summation node
$\mathbf{S}(n)$	vector of length N whose elements are all equal to $\sum_{k=1}^{N} y(n-k+1)$ (column vector)
t	continuous time instant
T_p	power control sampling period
u_i	polynomial coefficient of ith degree polynomial term, i = 0, 1, ..., I
$\mathbf{v}$	vector of length I + 1, $\mathbf{v} = [1, -p, ..., (-p)^I]^\mathrm{T}$ (column vector)
$v, \vec{v}$	mobile speed, velocity
$\mathbf{W}$	$(I + 1)$ by N matrix, $\mathbf{W} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 0 & 1 & \cdots & N-1 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 1 & \cdots & (N-1)^I \end{bmatrix}$
$w_1(n), w_2(n)$	samples whose values belong to an $\mathcal{H}$
$y(n)$	filter input sample
$\mathbf{y}(n)$	vector of N latest consecutive filter input samples (row vector)
$y_i(t)$	inphase component of a complex-valued baseband-equivalent signal
$y_q(t)$	quadrature component of a complex-valued baseband-equivalent signal

Terminology

augmented FIR filter	an FIR filter, extended to an IIR filter with a specially designed feedback (in this thesis, to shape the magnitude response of the FIR filter)
conventionally quantized coefficient filter	a filter, originally designed with infinite precision, whose coefficients are truncated or rounded (in the simulations, magnitude truncation of coefficients in two's complement form is used)
feedback extended FIR filter	c.f. augmented FIR filter
ideally quantized coefficient filter	a quantized coefficient filter that exactly fulfills the design criteria
infinite precision	computational precision of Matlab in a Pentium III PC
polynomial FIR predictor	an FIR filter designed to predict a polynomial input signal of a given maximum polynomial degree I by a given number p samples ahead
polynomial-predictive FIR differentiator	an FIR filter designed to predict the discrete time derivative of a polynomial input signal of a given maximum polynomial degree I by a given number p samples ahead

1 Introduction

This research started with the goal of increasing user capacity of mobile code-division-multiple-access (CDMA) communications systems through alleviating closed loop transmitter power control delays. For this effect, it was first assumed that since radio channel behavior is correlated over a period of time, channel fading history could be used to predict radio channel behavior, say 1 ms into the future. This was achieved by predictive filtering whose input was a finite history of received power level estimates, e.g., from 3 to 20 samples taken at 1 kHz sampling rate, and a predictive filter was to provide one-sample-ahead predicted received power level, or fading present in the radio channel. The conclusion was that the applied method is applicable in fine-tuning timing of a closed loop transmitter power control system. Also, constraints such as computational complexity, power consumption and required hardware size, imposed on the signal processing methods by application in handheld mobile user terminals, were to be considered. This called for simple and effective signal processing methods, executable also in low-precision fixed-point environments. Thereafter, the work concentrated on devising methods for designing coefficient quantization robust polynomial predictors. Polynomial-predictive differentiator is a filter type very closely related to polynomial predictors, and thus polynomial-predictive differentiators were also considered in our work. This goal was achieved better than anticipated; not only do the new design methods produce coefficient quantization robust filters, but filters that *exactly* meet the design criteria, i.e., possess the desired prediction and/or differentiation properties under coefficient quantization.

This work is in the interest of signal processing, communications, and control engineers. It introduces and evaluates a set of straightforward and practical methods, and brings the concept of exactness into the field of polynomial model based fixed-point signal processing.

In the present chapter, the fundamental concepts appearing in this thesis are presented using references, and also the author’s work is briefly discussed with regard to these fundamentals. Chapter 2 describes the polynomial-predictive and differentiative filters, as found in the literature before the author’s work, and Chapter 3 describes the related essentials of mobile communications systems. The electronic databases, and especially the new Internet tools (e.g., the IEEE Xplore™ [25]), have greatly enhanced the usability of scientific literature, and therefore it is not necessary to write a comprehensive literature survey on the several engineering disciplines involved in this thesis; only a few principal references per topic are adequate to serve as starting points. Chapter 4 describes the author's work and results in his original publications, and Chapter 5 gives the author's conclusions. In writing the introduction part of the thesis, the intention has been that a practicing engineer should be able to quickly grasp the main contributions as well as the relevant essentials of communications and signal processing without having to face numerous equations.

This dissertation consists of analysis and development of polynomial FIR predictors (PFPs) [19] and polynomial-predictive FIR differentiators (PPFDs) [52][55] for applications in fixed-point environments [24, pp. 348–358, 723–725][36, pp. 338–343][59, pg. 188], and of applying polynomial FIR predictors in transmitter power control of mobile communications systems, particularly in direct-sequence code-division-multiple-access (DS CDMA) systems [20, pp. 25–27][33][41, pp. 39–46]. The work started with the goal of improving mobile transmitter power control accuracy. It was considered that being able to predict the behavior of a radio channel would benefit the transmitter power control function. From the literature it is evident that improvement in the power control function would contribute to the maximum number of active simultaneous users in a mobile CDMA communications system, i.e., to system capacity [37, pp. 172–173][53, pg. xvii]. The first task was to find out if Rayleigh fading signals, used in modeling radio channel fading under certain frequently occurring conditions [12, pp. 132–134, 367][27, pp. 67–76][40, pp. 98–99, 120–121], were predictable by means of polynomial prediction. Simulations showed that properly sampled, for example at the power control rate ≈ 1 kHz, Rayleigh fading (at urban mobile speeds, e.g., $v \leq 50$ km/h, with the carrier frequency 1.8 GHz used in the simulations) is well predictable by polynomial FIR predictors, and that computationally efficient prediction can be applied in efforts to alleviate closed loop power control delays. Also, minimum mean-square-error (MSE) optimized power prediction of complex valued signals was derived and simulated.

The predictive mobile power control research directly called for fixed-point polynomial FIR predictors; it was previously not well known how these filters would perform under coefficient quantization, although the PFP coefficient quantization problem had been recognized in [17]. At this point, it was decided to also take the closely related filter family, i.e., polynomial-predictive FIR differentiators, into the coefficient quantization investigation. It soon became evident that these filters suffered greatly from coefficient quantization. As a result, a method for designing quantized coefficient polynomial FIR predictors and polynomial-predictive FIR differentiators was devised. The new filter design method yields polynomial FIR predictors and polynomial-predictive FIR differentiators that exactly fulfill the design criteria under coefficient quantization; they are thus coefficient quantization error free, and called ideally quantized coefficient filters. The polynomial-predictive FIR differentiators and polynomial FIR predictors are closely related; both families are similarly derived from similar sets of linear constraints on filter coefficients. The existence of such linear constraints was the key feature that allowed for coefficient quantization error free design of these filters.

Short (e.g., filters of lengths $N < 32$) polynomial FIR predictors and polynomial-predictive FIR differentiators were known to have poor white noise attenuation properties, whereas long ideally quantized coefficient fixed-point filters were found to be difficult or computationally impossible (for $N \geq 32$) to design because of long computational times required. For example, finding an ideally quantized coefficient filter of length $N = 32$ using the search algorithm described in [P3] would take of the order of 100 000 years on a 166 MHz Pentium processor! To improve applicability of these filters, a previously known feedback extension for magnitude response shaping [38][57] was adopted. It was shown possible to design coefficient quantization error free feedback extended fixed-point polynomial predictors and predictive differentiators based on the corresponding FIR filters, and criteria for such filter designs were formulated. As the coefficient quantization problem had now been shown solvable, the next concern was possible roundoff noise because of the low precision fixed-point arithmetic used [24, pp. 349, 356–357, 440][36, pp. 346–348]. Depending on the application and the selected filter, possibility of roundoff noise problems was identified. As a cure, a simple quantization error feedback [29][30][32][36, pp. 421–425] was found effective, and worth recommending for alleviating roundoff noise problems. The applied quantization error feedback is computationally very efficient, and does not affect the overall transfer function of the filter.

Inspired by the time-varying nature of the mobile radio channel, while remembering the power consumption and silicon area requirements of mobile terminals, a simple fine-tuning scheme was attached to the polynomial FIR predictors. The applied general parameter (GP) adaptation was shown to yield extended applicability ranges for the polynomial FIR predictors. The principal idea of the GP method is that a single adaptive general parameter (GP) is added to each filter coefficient, and updated in a least-mean-squares (LMS) like adaptation [2][5] (for LMS, c.f. [18, pp. 365–327]). Despite the simple computations and considerable potential for adaptive signal processing, the GP method is quite unknown. The method was also applied to extending applicability of sinusoidal FIR prediction. Also, a stability criterion for the GP adaptation was derived.

This dissertation is given impetus by the general smooth nature of adequately sampled signals, such as several measurements, like position, velocity, temperature, and power; and by the delays inherently caused by signal propagation, acquisition and processing. Real-world signals in general, if adequately sampled, offer themselves for piece-wise modeling as noisy polynomials of low degree. Thereafter, polynomial signal model can be used to predict (or extrapolate) such signals, or their discrete time derivatives, in order to alleviate the signal processing, acquisition and propagation delays. In this dissertation, the research is presented in time-reversed order to allow the reader to get acquainted with the employed signal processing tools before considering their applications. The abovementioned filter development work and results are presented first [P1] – [P5], and followed by the predictive mobile power control application [P6] – [P9]. The dissertation answers several previously unanswered questions concerning polynomial predictors and polynomial-predictive differentiators. Some aspects, like existence of IIR limit cycles [44, pp. 549–554][59, pp. 188–191] and possible over-flow oscillation problems [24, pp. 425–416, 432–433][36, pp. 377–381][59, pp. 191–192] still remain open questions, which should be determined for each design and application separately as necessary by the filter designer.

The nature of this dissertation is mostly of engineering type; it promotes predictive signal processing, which is based on time domain characteristics of the input signals. Besides, it is a collection of methods with illustrative examples for filter designers aiming at delay minimization while preserving the imposed computational load at the minimum. The main topics covered by this dissertation are:

1. Analysis of coefficient quantization errors in polynomial FIR predictors and predictive differentiators.
2. Design of polynomial FIR and IIR predictors and differentiators for low-precision, fixed-point implementation.
3. General parameter-based fine-tuning of polynomial and sinusoidal FIR predictors.
4. Predictability analysis of Rayleigh fading power signals.
5. Application of polynomial FIR predictors to closed loop control of transmitter power in cellular communications systems.

2 Polynomial Prediction and Differentiation

Generally, p-step-ahead linear prediction based on N past signal samples and M past predictor outputs is given [44, pp. 102–103, 828] by

$$\hat{y}(n+p) = \sum_{k=1}^{N} B(k)y(n-k+1) - \sum_{m=1}^{M} A(m)\hat{y}(n-m+1) \tag{1}$$

where n is discrete time instant, $B(k)$ and $A(m)$ are IIR predictor coefficients of the feed forward and feedback paths, respectively, $y(n)$ is predictor input signal sample, and hat denotes an estimate. In addition to providing for the prediction (1), it is often desired to minimize the white noise gain of the filter in order to maximize attenuation of wide-band noise present beyond the frequency band of the primary signal. This white noise gain can be expressed as [36, pg. 397]

$$NG = \frac{1}{2\pi}\int_{0}^{2\pi}\left|H\left(e^{j\omega}\right)\right|^2 d\omega \tag{2}$$

where $H(\mathrm{e}^{j\omega}) = B(\mathrm{e}^{j\omega}) / [1 + A(\mathrm{e}^{j\omega})]$ is the filter transfer function. For FIR filters, the white noise gain can be written as

$$NG = \sum_{k=1}^{N}\left|h(k)\right|^2 \tag{3}$$

with the predictive FIR filtering given by

$$\hat{y}(n+p) = \sum_{k=1}^{N} h(k)y(n-k+1)\,. \tag{4}$$

Here we have reserved the notation $B(k)$, $k = 1, \ldots, N$, for coefficients of the feed forward part of an IIR filter, while $h(k)$, $k = 1, \ldots, N$, refer to pure FIR coefficients.

In this thesis, the filter input signal $y(n)$ is assumed to be piecewise modelable by low degree polynomials contaminated by white Gaussian noise. A natural signal can often be piecewise modeled by low-degree polynomials, provided that the sampling rate is sufficiently high so that a low-degree polynomial, i.e., of $I = 0$, 1, or 2, can be fitted to a number samples, say, $N = 3$, ..., 50, of the primary signal with a desired accuracy. Polynomials of degrees $I = 0$, 1, 2, (and $I = 3$) can be considered sufficient for modeling in most applications. The higher degree polynomial predictors ($I \geq 3$) generally exhibit high passband peak gain (>> 1) or poor white noise attenuation, and are, therefore, not practicable in most applications. Also, for the higher degree polynomial predictors, it is very difficult to design feedback extensions which are used to shape magnitude responses of the polynomial predictors in order to increase practicable applicability [17]. In exact polynomial prediction, for the maximum input signal polynomial degree of one, a constant or ramp can be exactly fitted to a number ($N \geq 2$) of consecutive samples of the signal, or for the maximum input signal polynomial degree of two, a constant, ramp, or parabola to a number ($N \geq 3$) of consecutive samples.

With the selected maximum filter length of $N = 100$ adopted work, the predictor window would contain 0.1 s of channel fading history at 1 kHz mobile power control rate [P6][P9] (approximately 1.6 kHz in [P7][P8]). According to [43, pg. 765], channel coherence time can be approximated by the reciprocal of the Doppler spread; with 1.8 GHz carrier frequency, at 5 km/h the coherence time is of the order of 0.1 s, and of the order of 0.01 s at 50 km/h, which justify the selection of the $N = 100$ ($N = 50$ in [P6]) as the maximum applied filter length. Also from the real-time implementation point of view, it is advantageous not to apply unnecessarily long filters.

2.1 Polynomial FIR Predictors (PFP)

Polynomial FIR predictors [19] are derived in time domain using a polynomial signal model. Assuming a piece-wise degree polynomial signal contaminated by white noise, and an FIR type filter, the optimum predictor coefficients $h(k)$, $k = 1, 2, \ldots, N$, (4) for one-step-ahead prediction $p = 1$ can be calculated in closed form [19] starting with a set of linear constraints on filter coefficients:

$$g_0 = \sum_{k=1}^{N} h(k) - 1 = 0\,, \tag{5}$$

$$g_1 = \sum_{k=1}^{N} k h(k) = 0\,, \tag{6}$$

$$g_2 = \sum_{k=1}^{N} k^2 h(k) = 0\,, \tag{7}$$

$$\vdots$$

$$g_I = \sum_{k=1}^{N} k^I h(k) = 0\,. \tag{8}$$

The constraints (5)–(8) give prediction of each of the polynomial degrees $i = 0, 1, \ldots, I$, and from them closed form solutions for the FIR coefficients for low-degree polynomial input signals be can solved by the method of Lagrange multipliers [7, Chapter 1]. After fulfilling the constraints (5)–(8), the remaining degrees of freedom are used to minimize the noise gain (3).

Closed form solutions for FIR coefficients for one-step-ahead prediction of the first, second, and third degree polynomial input signals can be found in [19]. For example, with the highest polynomial input signal component degree of two, $I = 2$, we have to fulfill the constraints (5), (6), and (7), and the infinite precision coefficients for such one-step-ahead $p = 1$ polynomial FIR predictors are given by [19]

$$h(k) = \frac{9N^2 + (9 - 36k)N + 30k^2 - 18k + 6}{N^3 - 3N^2 + 2N},\ k \in [1, 2, ..., N]. \tag{9}$$

Additional insight in PFP design is gained by noting that the design problem is a least-squares filter design problem [23]. With a polynomial input signal

$$y(n) = \sum_{i=0}^{I} u_i n^i,\quad n = ..., -1, 0, 1, ... \tag{10}$$

with polynomial coefficients u_i, $i = 0, 1, ..., I$, constraints on the filter coefficients can be written as

$$\sum_{i=0}^{I} u_i (n + p)^i = \sum_{i=0}^{I} u_i \sum_{k=1}^{N} h(k)(n - k + 1)^i . \tag{11}$$

Observing (11) at time $n = 0$ without losing generality, from (11) it can be identified that

$$\begin{aligned} p^i = \sum_{k=1}^{N} h(k)(-k + 1)^i \Leftrightarrow p^i = (-1)^i \sum_{k=1}^{N} h(k)(k - 1)^i \\ \Leftrightarrow \sum_{k=1}^{N} (k - 1)^i h(k) = (-p)^i,\ i = 0, 1, ..., I . \end{aligned} \tag{12}$$

Eq. (12) is a set of linear equations, which can be written as

$$\mathbf{W}\mathbf{h}^{\mathrm{T}} = \mathbf{v} \tag{13}$$

where $\mathbf{W} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 0 & 1 & \cdots & N-1 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 1 & \cdots & (N-1)^I \end{bmatrix}$ and $\mathbf{v} = [1, -p, ..., (-p)^I\,]^{\mathrm{T}}$. In (13) there are N unknowns and $I + 1$ equations, thus, if $N >> I$, there are degrees of freedom left after providing for prediction that can be used for noise gain minimization. Minimizing the noise gain (3) with the linear constraints (13), gives us the solution [23]

$$\mathbf{h}^{\mathrm{T}} = \mathbf{W}^{\mathrm{T}}\left(\mathbf{W}\mathbf{W}^{\mathrm{T}}\right)^{-1}\mathbf{v} . \tag{14}$$

2.2 Polynomial-Predictive FIR Differentiators (PPFD)

Polynomial-predictive FIR differentiators [52][55][56] are a family of filters derived analogously to polynomial FIR predictors. The linear constraints on filter coefficients are now given by

$$g_0 = \sum_{k=1}^{N} h(k) = 0\,, \tag{15}$$

$$g_1 = \sum_{k=1}^{N} (N-k)h(k) = 1\,, \tag{16}$$

$$g_2 = \sum_{k=1}^{N} (N-k)^2 h(k) = 2(N-1+p)\,, \tag{17}$$

$$\vdots$$

$$g_I = \sum_{k=1}^{N} (N-k)^I h(k) = I(N-1+p)^{I-1}\,. \tag{18}$$

The constraints (15)–(18) provide for prediction and differentiation of the polynomial degrees $i = 0, 1, \ldots, I$, and the closed form solutions for the FIR coefficients for low-degree polynomial input signals are again obtained by the method of Lagrange multipliers [7, Chapter 1]. After fulfilling the constraints (15)–(18), the remaining degrees of freedom are used to minimize the noise gain (3).

The closed form solution for the FIR coefficients for the second degree polynomial input signals, with prediction step p as a parameter, is given in [55]. Again, using the case with the highest polynomial input signal component degree of two $I = 2$, and one-step-ahead prediction $p = 1$, as an example, we have to fulfill the constraints (15), (16) and (17), and use the remaining degrees of freedom to minimize the noise gain (3), yielding the coefficients [55]

$$h(k) = \frac{6\left[(30N+30)(k-1)^2 + (-32N^2+38)(k-1) + 6N^3 - 11N^2 - 9N + 14\right]}{(N-2)(N-1)N(N+1)(N+2)},\ k \in [1,\ 2,\ \ldots,\ N]. \tag{19}$$

For definitions of differentiation, see [10, pp. 69–72, 78–80][36, pp. 913–936].

2.2.1 Polynomial Differentiator Applications

The research on polynomial differentiators arose from the research of polynomial predictors; polynomial differentiators were not applied by the author. Generally, differentiators can be applied in designs in which the measurements do not provide the desired information directly but only after discrete time differentiation. For example, in motion control applications it is practicable to employ a velocity or position sensor, and obtain the acceleration information via differentiation [39].

2.3 Magnitude Response Shaping Feedback Extension

Short polynomial FIR predictors (PFPs) and polynomial-predictive FIR differentiators (PPFDs) possess poor if any noise attenuation, and are thus problematic in real applications since noise degrades prediction accuracy. The noise gain of PFPs is approximately given by [23]

$$NG \approx (I+1)^2 / N\,, \tag{20}$$

which is illustrated in Fig. 1. From Fig. 1 it is seen that the noise gain increases with increasing polynomial degree and decreases with increasing filter length.

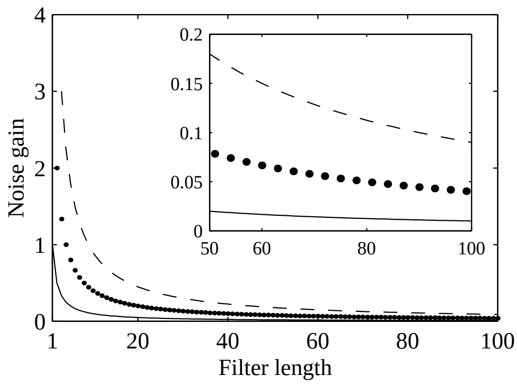

Fig. 1. PFP noise gain approximations (20) for $I = 0$, $N = 1, ..., 100$ (solid), $I = 1$, $N = 2, ..., 100$ (dotted), and $I = 2$, $N = 3, ..., 100$ (dashed).

The poor noise attenuation performance can be alleviated by employing a cascade of two or more filters as in a prefiltering approach [31], or with a cascade of polynomial estimators and predictors [38]. The cascade structures allow for more efficient designs than increasing the order of a single filter alone; for example, the prefilter can be a short polynomial IIR estimator to provide for delayless noise reduction, followed by a short polynomial predictor. The noise gain problem can also be alleviated by magnitude response shaping feedback for polynomial predictors [38][55][57], extending the basis FIRs into IIR filters with improved noise attenuation properties. The feedback extension proposed in the above papers preserves exactly the prediction and/or differentiation properties of the underlying basis FIR filters, and is studied in this thesis for implementations in fixed-point environments.

Also, in the design of the feedback extension (Fig. 2), fixed-point effects must be taken into account since roundoff noise generation may not be tolerable in short coefficient word length fixed-point environments. The roundoff noise problem has been identified in [15]. Also, improper design of a quantized coefficient feedback extension affects directly the prediction and/or differentiation properties.

Effectively, the feedback extension (Fig. 2), computes a weighted average of the filter input and its predicted value [38]. In the first summation node from the left in Fig. 2, the sum is given by

$$y(n)[1-b(k)] + \hat{y}(n)b(k),\ k \in [1,2], \tag{21}$$

and likewise in the second summation node. The fact that the value of the coefficient $b(k)$ in (21) implies the value of the coefficient $1 - b(k)$ exactly, ensures that the feedback enhances magnitude response of the basis FIR predictor without affecting the steady-state prediction properties.

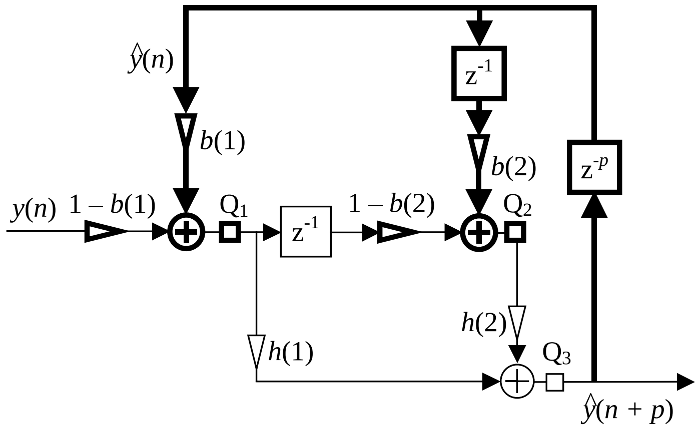

Fig. 2. A feedback extended FIR predictor of length $N = 2$ with MAC result quantizers Q_j, $j = 1, 2, 3$. Operators associated with the feedback extension are drawn with thick lines.

2.4 Fixed-Point Effects

Considering applications in mobile communications systems, especially in handheld user terminals, computational complexity, power consumption, hardware size and cost are to be carefully considered. This directly motivates research on fixed-point effects in polynomial predictive filtering.

2.4.1 Roundoff Noise Analysis

Roundoff noise analysis is presented thoroughly in [26, pp. 382–387], and discussed with feedback extended polynomial predictors in [15]. In a quantized two's complement number space with magnitude truncation and quantization step Δ, truncation error takes a random value with mean $-\Delta/2$ and variance of

$$\sigma_0^2 = \Delta^2 / 12 . \tag{22}$$

For the analysis, roundoff noise power spectrum at the filter output is given by

$$P(\omega) = \sigma_0^2 \sum_{j=1}^{S} m_j \left| G_j\left(e^{j\omega}\right) \right|^2 \tag{23}$$

where S is number of summation nodes, $G_j(e^{j\omega})$ is the transfer function from the jth summation node to the filter output, and m_j multipliers feed into the jth summation node. Now, the total average roundoff noise power is given as

$$\sigma^2 = \sigma_0^2 \sum_{j=1}^{S} m_j \sum_{n=0}^{\infty} \left| g_j(n) \right|^2 \tag{24}$$

where $g_j(n)$ is the impulse response due to $G_j(e^{j\omega})$.

In this thesis, the above roundoff noise analysis is used to analyze roundoff noise contributions of the individual multiply-accumulate (MAC) operation output quantizers Q_j, $j = 1, 2, 3$, seen in the feedback extended FIR structure in Fig. 2. Roundoff noise analysis is performed in order to locate the noisiest quantizer for roundoff noise suppression.

The above general roundoff noise analysis assumes that the roundoff error process is a white noise process with uniform distribution of error $\varepsilon(n)$ within

$$-\Delta < \varepsilon(n) \leq 0. \tag{25}$$

For polynomial input signals $y(n)$ which are noiseless this assumption clearly does not hold, but the roundoff error $\varepsilon(n)$ is itself piecewise polynomial. With a quantized input signal $y_Q(n)$,

$$y(n) - y_Q(n) = \varepsilon(n), \varepsilon(n) \text{ polynomial while } y_Q(n) \text{ constant.} \tag{26}$$

Computed over a long period of time, the average roundoff noise power is still given by (22), provided that the signal dynamics span several quantization steps making the error distribution uniform in (25). In practice the measured signals are always noisy, and the assumption holds if the noise present at the filter input is itself sufficient to give rise to the quantization noise being white with average power given by (22) even with a constant primary signal. Thus, for the analysis presented in [P4] to hold, either a noisy input must be assumed, or $G_j(e^{j\omega})$ in (23) should also take into account the quantization error spectrum. Let us note that with noiseless polynomial input signals, there would be nothing to gain from feedback extended polynomial FIR predictor implementations since they are aimed at noise reduction, and it would be practicable to apply short PFPs instead. On the validity of the discussed quantization error model, see also [6].

2.4.2 Quantization Error Feedback

Error feedback [26, pp. 418–422][36, pp. 421–425] is an efficient method for alleviating effects resulting from quantization of arithmetic operations in recursive systems. Multiplication and accumulation (MAC) may be done with double precision, while MAC results are available only as single precision numbers, while the bits of the higher precision part can usually be made available for error feedback within the arithmetic logic unit.

In Fig. 3, error feedback is illustrated with a single exemplary quantization error feedback filter installed, similar structures could be installed on all MAC output quantizers Q_j, $j = 1, 2, 3$. In Fig. 3, quantization error is calculated over the MAC results quantizer Q_1, and an error feedback filter of length $N_e = 2$, with a coefficient vector $\mathbf{e} = [e(1)\ e(2)]$, is installed. Design of optimal error feedback filter that minimizes the quantization noise power in LMS sense is rigorously performed in [30], where the solution is shown to be a special case of Wiener filtering.

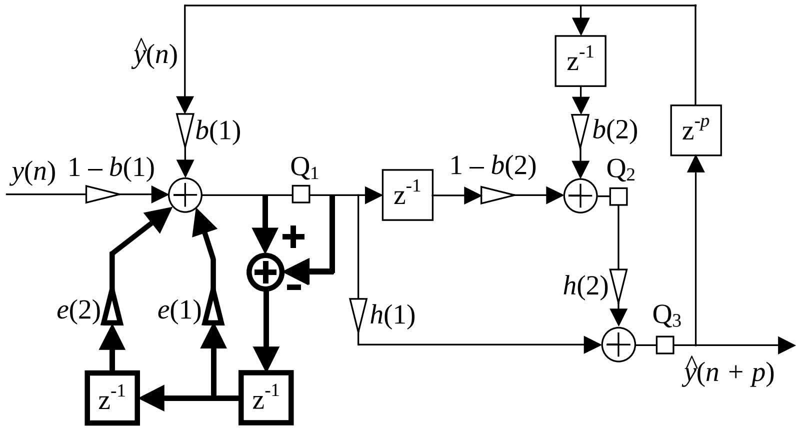

Fig. 3. A feedback extended FIR predictive filter with an exemplary error feedback of length $N_e = 2$ installed (drawn with thick lines) and all the MAC output quantizers Q_j, $j = 1, 2, 3$, shown.

Notation in Fig. 3:	PFP coefficients	$h(1)$, $h(2)$
	feedback extension coefficients	$b(1)$, $1 - b(1)$, $b(2)$, $1 - b(2)$
	error feedback filter coefficients	$e(1)$, $e(2)$
	MAC results quantizers	Q_1, Q_2, Q_3
	prediction step	p
	filter input sample estimate	$\hat{y}(n)$

2.5 Genetic Algorithms in Digital Filter Design

Genetic algorithms, a form of evolutionary computation, [9, pp. 31–99][14] are suitable for solving problems whose solution space can be coded in a form of a string, or a chromosome, and the nature of the solution space is unknown or analytically difficult to define. The main principle is that the chromosomes evolve from one generation to another through operations which are drawn from the evolution process of life. The operations employed in this thesis are crossover, mutation, fitness evaluation, selection, and reproduction. In filter design, a straightforward method is to consider coefficient vectors as chromosomes with fixed-point number elements, and perform genetic operations as per coefficient or on the coefficient boundaries only. A fitness function is to be defined according to the filter design specifications.

From the literature, it seems that genetic algorithms are not widely used for digital filter design. Since genetic algorithms are designed for flexible self-guiding search in a large unknown solution space, they can be expected to find improved fixed-point filters in the sense defined in the fitness function, if such exist. As with any other optimization-type filter design methods, no other assumptions on the quality of the resulting filters can be made. Filter design with genetic algorithms has been proposed, for example, for generating computationally efficient FIRs [48][58], for design and optimization of IIR filter structures [49], and for optimizing feedback extended polynomial predictors [16]. The family of filters considered in [16], is also of concern in this thesis. While [16] addresses feedback extension design, in this thesis, coefficient quantization robust designs are searched for.

2.6 General Parameter Adaptive FIRs

As real-world signals are seldom stationary, especially not those encountered in multiuser mobile communications systems, it is desirable to introduce a fine-tuning extension to polynomial FIR predictors and polynomial-predictive FIR differentiators. Here, motivated by the small power consumption and die size requirements of mobile communications applications, a fairly unknown but simple and computationally efficient adaptation scheme, general parameter (GP) adaptive filtering, is adopted.

The general parameter (GP) adaptation [5] consists of a single GP that is updated and added to each FIR coefficient. GP extended FIR filtering is given by

$$\sum_{k=1}^{N}[\beta(n)+h(k)]y(n-k+1)=\hat{y}(n+p) \tag{27}$$

where $\beta(n)$ is GP which is adapted according to the input signal and prediction error as

$$\beta(n+1)=\beta(n)-\gamma[\hat{y}(n)-y(n)]\sum_{k=1}^{N}y(n-k+1) \tag{28}$$

where $\gamma > 0$ is adaptation gain factor. GP adaptation (28) is seen to take the input signal statistics into account in the form of average over the filtering window. Writing the input-output-relation (27) as

$$\beta(n)\sum_{k=1}^{N}y(n-k+1)+\sum_{k=1}^{N}h(k)y(n-k+1)=\hat{y}(n+p), \tag{29}$$

GP adaptation is seen to act as a variable gain on average input signal amplitude, which is added as a bias to the output of the basis FIR. In contrast to convergence of the LMS algorithm [18, pp. 393–394], the single GP does not provide full adaptation, but rather filter fine-tuning. A computationally efficient GP extended FIR structure is shown in Fig. 4.

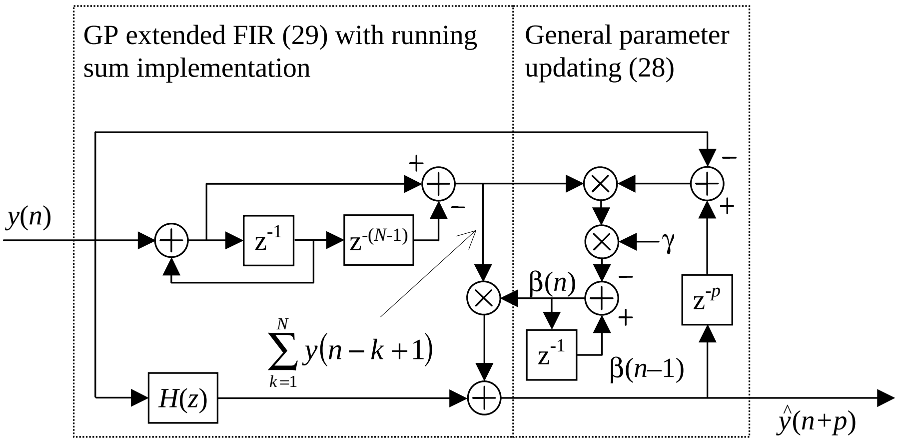

Fig. 4. Structure of the GP extended FIR.

3 Mobile CDMA Transmitter Power Control

In code-division-multiple-access (CDMA) mobile communications systems [41][53], including wideband CDMA (WCDMA) systems [20], power control is regarded as crucial in maximizing user capacity of the communications system [13][20, pp. 34–35][37, pp. 172–173]. Improper power control may even result in a network failure [41, pg. 316], while an efficient power control allows the system to accommodate more active simultaneous users [20, pp. 34–35] since the operating conditions for the receiver are optimized in the sense that the received individual users' SNRs are balanced above an adequate quality level, so that no user is received better than the others at the expense of the other users. With RAKE receivers [42], an improvement in power control operation can be expected to increase user capacity of a CDMA system through decreased other user interference. Current WCDMA systems employ a received signal-to-interference (SIR) based power control [4][20, pp. 34–37, 109–110][45, pp. 10–11, 34–37], while application simulations in this work concentrate on received power level based power control used in earlier digital communications systems, for example in [3][33], and for GSM (Global System for Mobile communications) [46, pp. 78–79, 209, 260–262].

The choice of received power level as the control variable offers the following advantages: Effects of predictive filtering could be made clearly visible in simulations, whereas in received SIR based power control simulations, the complex relationship between radio channel variations, interference conditions, and received SIR, should have been taken into account. Also otherwise the simulator would grow more complex with the SIR based power control, and unnecessarily complicate the interpretation of the results concerning the effects of prediction. On the other hand, this choice of the control variable means that the results presented in this thesis can be viewed mainly as qualitative encouragement for applying the proposed methods in SIR based power control systems, and no quantitative performance improvement expectations can be drawn for SIR based systems. In SIR based power control the individual users' received power levels must be estimated, and thus also a SIR based power control system, e.g., such as in Fig. 9, can be designed to take advantage of the predicted radio channel fading information.

A RAKE receiver [20, pp. 30–34][42][53, pg. 89] takes advantage of the multipath propagation in a mobile radio channel. Principally, multipath propagation means that several copies of the radio transmission are received at the receiver antenna possibly after different delays due to the transmission arriving at the antenna via different propagation paths, c.f. Fig. 5. A RAKE receiver distinguishes and combines a few, generally 3 to 6, strongest paths with sufficient delay separation. In the RAKE receiver, there is one receiver correlator structure, i.e., finger, for each path it receives. On the other hand, a RAKE receiver alone does not take into account other detected users in order to improve reception via interference cancellation [41, pp. 317–336][47, pg. 953], while multiuser detectors (MUDs) [54] do so. For systems employing multiuser detection (MUD) receivers instead of RAKE receivers, power control requirements are not so strict [41, pg. 316], or the power control that maximized the user capacity is not the one which results in received SIR balancing [54, pp. 360–361].

In [50] a six-finger RAKE receiver for WCDMA communications systems is implemented on a single fixed-point signal processor. Polynomial predictive filtering could be easily inserted into that implementation, either to operate on the output levels of each receiver finger, or on the total output signal level after path combining.

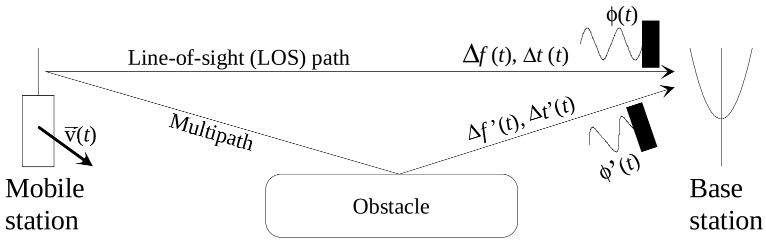

Fig. 5. An uplink multipath propagation scene. Notation: mobile velocity $\vec{v}(t)$, propagation delays $\Delta t(t)$, $\Delta t'(t)$, Doppler shifts $\Delta f(t)$, $\Delta f'(t)$, and arrival phases $\phi(t)$, $\phi'(t)$.

3.1 Baseband-Equivalent Signal Model

All communications signal processing in this thesis is preformed in baseband-equivalent signal domain [12, pp. 36–38][47, pp. 92–95]. Generally, baseband-equivalent signal is the complex envelope of the modulated bandpass signal, i.e., with carrier frequency f_c, signal amplitude $A(t)$, and phase $\varphi(t)$, the received bandpass signal centered at the carrier frequency f_c is given as

$$y(t) = A(t)\cos[2\pi f_c t + \varphi(t)] \tag{30}$$

which can be written as

$$y(t) = A(t)\cos\varphi(t)\cos 2\pi f_c t - A(t)\sin\varphi(t)\cos 2\pi f_c t \tag{31}$$

from which the complex components of the baseband-equivalent signal model are identified as

$$\begin{cases} y_i(t) = A(t)\cos\varphi(t) \\ y_q(t) = A(t)\sin\varphi(t) \end{cases}. \tag{32}$$

The real and imaginary components in (32) are called inphase and quadrature components, respectively. Now, magnitude of the complex signal $y_i(t) + jy_q(t)$, or envelope of the received signal, is given by

$$r(t) = \sqrt{y_i(t)^2 + y_q(t)^2}\ . \tag{33}$$

In the simulations in this thesis, input to the predictors is either $y(n) = r(n)$, or $y(n) = y_i(n)$ and $y(n) = y_q(n)$. In the latter case, there are two predictors operating independently.

3.2 Mobile Radio Channel and Multiuser Interference

In a mobile radio channel [27, pp. 11–132][40][47, pp. 91–185], radio transmission propagation from a transmitter to a receiver takes place through any path that it can find. These may include a line-of-sight (LOS) path, and several reflected and/or scattered paths. A multipath scene with one LOS path and one reflected path is illustrated in Fig. 5. The multipath phenomenon results in fast fading of the received signal strength. Occasionally, received paths can sum up destructively, causing fast and deep fades to occur in the received signal strength. If the multipath environment is such that no significant LOS path exists [40, pg. 99], and the scatterers are not fixed with respect to the transmitter and/or receiver (which is generally the case when the mobile user terminal is moving) [43, pg. 761], the received signal amplitude distribution is well described by Rayleigh distribution [47, pp. 5–10]. Therefore, the received signal amplitude $r(t)$ in (33), is in this thesis assumed to be a fast fading Rayleigh distributed signal, such as illustrated in Fig. 6. Power spectrum of the received Rayleigh fading signal is set to model the spectrum seen by an omnidirectional vertical monopole antenna [11][27, pg. 75].

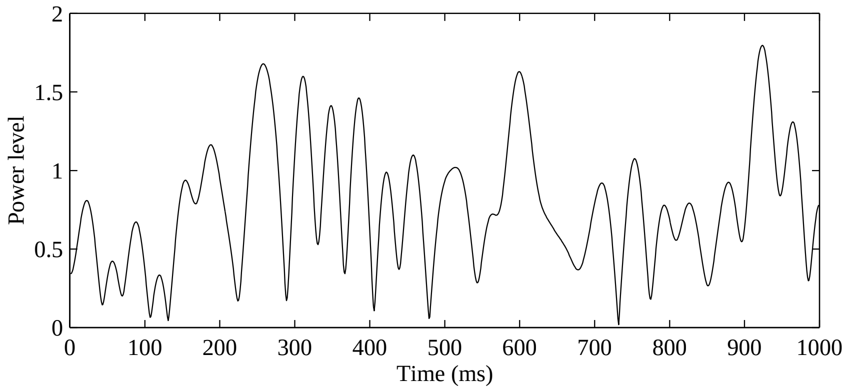

Fig. 6. A Rayleigh fading signal at the mobile speed of 10 km/h with 1.8 GHz carrier frequency.

A wideband transmission channel is characterized by the phenomenon of frequency selective fading, i.e., fading depends on the frequency within the band; on frequencies with sufficient frequency separation, fading is independent. In other words, transmission channel is wider than the coherence bandwidth. Basically, a wideband transmission channel is a channel that is wider than the coherence bandwidth [40, pg. 236][60, pp. 130–131, 141–142]. This provides fading resistance to a wideband communications system, since it is probable that only small portions of channel bandwidth are in deep fades at any moment

[60, pg. 389]. This means, as already mentioned, that simulations in this thesis correspond to received power level prediction in a narrowband CDMA system (or to difficult conditions in a WCDMA system). Simulation of wideband channels is described in [28][40, pp. 243–252], and is basically a time-varying FIR filter with each coefficient drawn from independent fading (e.g., Rayleigh fading) processes, with the smallest delay separation of the taps inversely proportional to the transmission bandwidth [43, pp. 795–797].

Considering predictability of the employed Rayleigh fading signals, signal-to-error power ratios predicting two differently produced Rayleigh fading signals at several mobile speeds with first and second degree PFPs are shown in Fig. 7, along with the signal-to-error power rations caused by the delay of one sample at the filter input. The Rayleigh fading signals for Fig. 7(a) are produced by the Jakes' model [27, pp. 70–76], and for Fig. 7(b) from two independent Gaussian noise processes [43, pp. 45–47] which are shaped according to the antenna geometry [11]. Especially the signal-to-error power ratio gains seen at lower speeds clearly suggest that the Rayleigh fading model employed is well predictable by the means of polynomial FIR prediction. For real-life noisy signals, either longer PFPs or other means of noise reduction are necessary in conjunction with polynomial prediction. Predictability is also the issue in [P6].

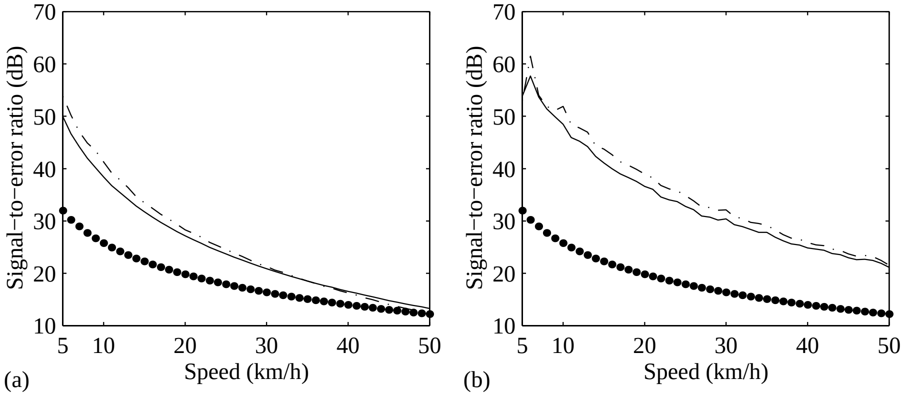

Fig. 7. Signal-to-error power ratios predicting two Rayleigh fading models with the first degree (solid) and second degree (dash-dot) PFPs, along with the signal-to-error power ratios resulting from the one sample delay at the filter input (dotted). In (a) Jakes' Rayleigh fading generators are employed, and in (b) noise shaping Rayleigh fading generators.

In this research, only urban mobile speeds up to 50 km/h are considered since it is assumed that transmission errors are randomized at higher mobile speeds, and can thus be corrected without accurate power control [45, pg. 54]. At urban mobile speeds, long fades would cause several consecutive transmission errors, and accurate power control must be applied to counteract the fades.

3.3 Closed Loop Power Control

In this thesis, purpose of the transmitter power control [12, pp. 370–380, 437–441][20, pp. 34–37] is to counteract radio channel fading, such as shown in Fig. 6, so that the signals of different users are received at an equal and constant level. Schematically, a power level based closed power control loop is presented in Fig. 8 [3]. In Fig. 8, base station estimates the received power level, which is the transmitted power level affected by radio channel fading ("channel variation"), compares the estimate to a preset received power level setpoint ("desired received power level"), and forms a 1-bit (up/down, i.e., bang-bang) power control command accordingly. The power control command is sent to the mobile station via a return radio channel, and is subject to error in transmission. All the loop delays, including processing, acquisition, and propagation delays, are modeled by a single "total loop delay" element. The mobile station receives the power control command, interprets it according to the preset incremental power control step size Δp (e.g., $\Delta p = 1$ dB), and corrects the current transmitter power setting accordingly. While [3] considers closed loop power control based on the received power level, in its continuation part [4], SIR based power control system is studied. Also, a power control system based on combined received power level and SIR has been proposed [61].

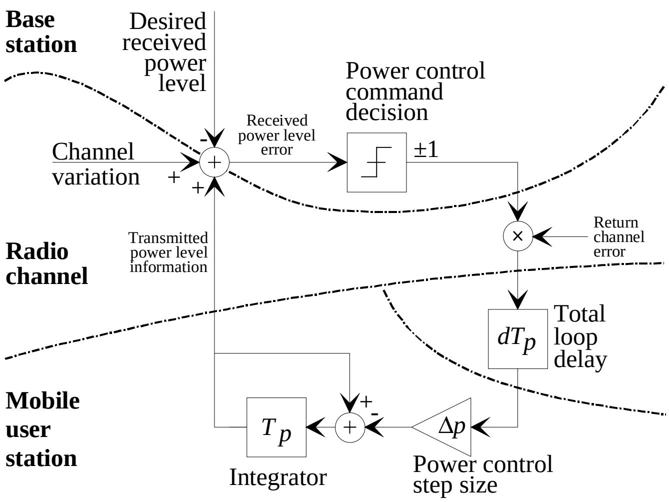

Fig. 8. Schematic view of a closed uplink power control loop. T_p denotes sampling period, d total loop delay in samples, and Δp is the incremental power control step size. Adopted from [3].

A received SIR based mobile CDMA power control system [8, Chapter 2.6.3] is illustrated in Fig. 9. Of Fig. 9, the part of interest in this thesis are the three blocks (colored red): "Received DPCCH power estimation" (DPCCH (Dedicated Physical Control Channel)), "Total downlink received interference estimation", and "SIR calc." in the mobile station; and the corresponding structure in the base station with the block "Total uplink received interference estimation", since it is the received DPCCH power estimation block in which the methods described in this thesis can be applied. Also, it would be possible to apply prediction on the calculated SIR. In Fig. 9, for example, uplink power control loop (colored blue) goes as follows: Starting at the "data source" of the mobile station, after setting the transmitter power, the data is transmitted through the "uplink radio channel" in which the transmission is affected by the wide-band radio channel fading ("WB fading generation"), and also the interference from other active users is added to the signal. At the base station "RAKE receiver", data is detected. For the received SIR calculation, received power level is estimated ("Received DPCCH power estimation") and also the "total uplink received interference estimation" is performed. Based on the received power level and interference estimates, received SIR is calculated, compared to a preset SIR target, and an incremental power control command is formed accordingly. SIR target is modified by an outer loop power control based on transmission quality estimates in order to guarantee a required quality of service under changing conditions and service requirements. The power control command is sent to the mobile station in the downlink channel in which the command is subject to possible transmit power control bit errors. After receiving the transmit power control command, the mobile station corrects its transmission power accordingly. In Fig. 9, the points where predictive filtering could be applied in straightforward fashion are marked with yellow circles. The downlink power control in [8, Chapter 2.6.3.2] is the same as the uplink power control [8, Chapter 2.6.3.1], as presented in Fig. 9. A more up-to-date description of the third generation WCDMA power control function can be found in [1, Chapter 5] but the differences have no implications for the main points of interest in this thesis.

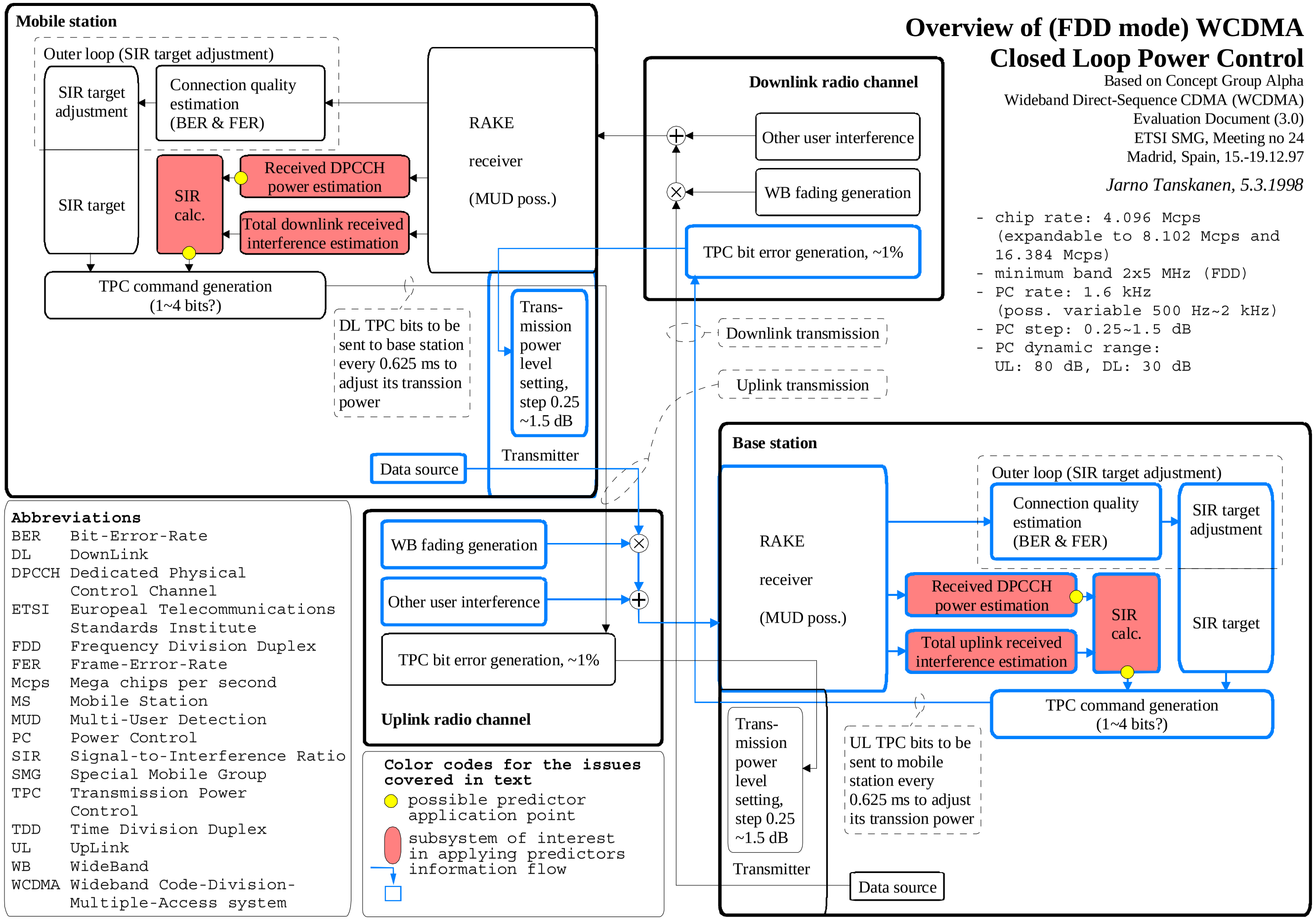

Fig. 9. Overview of a power control system for a WCDMA mobile communications system according to [8, Chapter 2.6.3]. The subsystems of interest in applying predictors are colored red, uplink power control loop blue, and the most obvious predictor application points yellow.

4 Summary of the Publications

4.1 Coefficient Quantization Errors in Polynomial FIR Predictors and Polynomial-Predictive FIR Differentiators

4.1.1 Coefficient Quantization Errors in Polynomial FIR Predictors

As the polynomial FIR predictors are designed to exactly fulfill the constraints (5)–(8), and to minimize noise gain (3) with infinite precision coefficients, it is clear that as the coefficients $h(k)$ are quantized, the constraints are generally not exactly fulfilled anymore [P3]. Nevertheless, there are a few polynomial FIR predictors whose infinite precision coefficients are numbers that belong to an 8-bit quantized coefficient space $\mathcal{H}_8$, and which are thus not affected by coefficient quantization to eight bits. Of the polynomial FIR predictors with lengths $N = 2, ..., 100$, polynomial degree $I = 1$, and prediction step $p = 1$, and $N = 3, ..., 100$, $I = 2$, $p = 1$, studied in this thesis, those that are coefficient quantization error free by their nature with 8-bit coefficients are listed in Table 1. Actually, 5 bits is enough for presenting the coefficients in Table 1.

Table 1. Amongst the filters studied in this thesis, the one-step-ahead predictive PFPs and PPFDs whose coefficients are not affected by coefficient quantization to 8 bits.

Filter type	Polynomial degree	Filters	
PFP	$I = 1$	$N = 2$, $\mathbf{h} = [-2\ 1]$	$N = 4$, $\mathbf{h} = [1\ 0.5\ 0\ -0.5]$
PFP	$I = 2$	$N = 3$, $\mathbf{h} = [3\ -2\ 1]$	$N = 4$, $\mathbf{h} = [2.25\ -0.75\ -1.25\ 0.75]$
PPFD	$I = 2$	$N = 3$, $\mathbf{h} = [2.5\ -4\ 1.5]$	

For exact prediction of polynomial signals, predictors are to exhibit two necessary properties set forth by the linear constraints on the filter coefficients; unity gain at zero frequency, and group delay $-p$ at zero frequency, c.f. Fig. 10 for the case $p = 1$, $N = 8$, $I = 2$. These two conditions are necessary for prediction of polynomials of any degree, and necessary and sufficient for prediction of first-degree polynomials. Generally, straightforward coefficient quantization destroys these properties, and even though errors in magnitude response or group delay at zero frequency may not be great in absolute value, the effect on the actual filtering performance may be devastating. For an example of low coefficient precision effects on magnitude responses and group delays, see Fig. 10 for conventionally quantized, and ideally quantized coefficient second-degree one-step-ahead PFP of length $N = 8$ with the coefficient precision of 8 bits [P3]. An example of output behavior of the filters in Fig. 10 is shown in Fig. 11. It is seen in Fig. 11, that conventional coefficient quantization has rendered the filter behavior unacceptable, whereas the ideally quantized coefficient filter performs exactly correctly as it should.

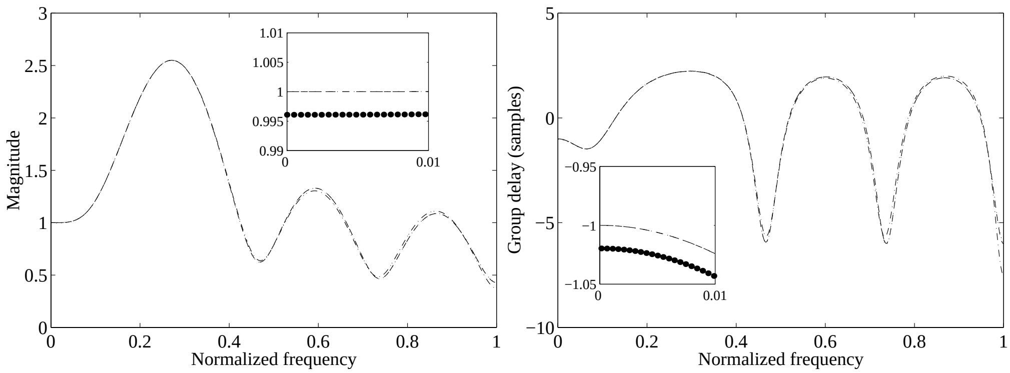

Fig. 10. Magnitude responses (left) and group delays (right) of the infinite precision (dashed), conventionally quantized (dotted), and ideally quantized (dash-dot; mostly overlaps dashed) coefficient second-degree one-step-ahead PFP of length $N = 8$ with the coefficient precision of 8 bits [P3].

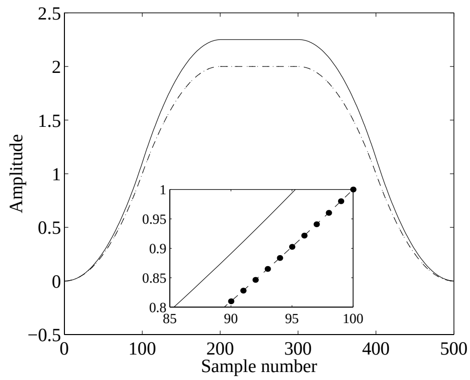

Fig. 11. Prediction of a piece-wise polynomial signal with second and zeroth degree polynomial sections; the desired filter output (dashed), prediction with the PFP of length $N = 8$ with truncated 8-bit coefficients (solid), and prediction with the corresponding ideally quantized 8-bit coefficient filter (dotted; overlaps dashed) [P3].

4.1.2 Publication [P1]: Coefficient Quantization Effects in Polynomial-Predictive FIR Differentiators

As the polynomial-predictive FIR differentiators (PPFDs) are designed to exactly fulfill the design constraints (15)–(18), and thereafter to minimize noise gain (3) with infinite precision coefficients, it is clear that if the coefficients $h(k)$ are rounded or truncated, the constraints are generally not met exactly anymore [P3]. Although, among the 98 PPFDs considered for Table 1, there exists one polynomial-predictive FIR differentiator whose exact coefficients are numbers that belong to the 8-bit quantized coefficient space $\mathcal{H}_8$, and are thus not affected by such coefficient quantization.

For exact predictive differentiation of polynomial signals, differentiators are to exactly exhibit the properties set forth by the linear constraints on the filter coefficients. For the predictive differentiation of the first degree polynomials, these are zero gain at zero frequency, and group delay $-p$ at zero frequency. These two conditions are necessary for predictive differentiation of polynomials of any degree, and necessary and sufficient for predictive differentiation of first-degree polynomials. Polynomial differentiators are also characterized by the ramp shaped magnitude response within the differentiation band, c.f. Fig. 12 for the case $p = 1$, $I = 2$. Generally, conventional coefficient quantization destroys these properties, c.f. Fig. 12 for an example of a PPFD with 8-bit coefficients, and even quantization to 16 bits may destroy the group delay properties of polynomial-predictive FIR differentiators (Fig. 13). Comparing Fig. 12 and Fig. 13 of PPFD $N = 16$, $I = 2$, $p = 1$, with coefficients truncated to 8 and 16 bits, respectively, it is demonstrated that differentiation property is here more robust against coefficient quantization with regard to the error at zero frequency than the prediction property. Although errors in magnitude response or group delay at zero frequency may not be great in absolute value, effects on the actual filtering performance may be devastating, c.f. Fig. 14 for an example of predictive differentiation of a polynomial signal with the infinite precision, conventionally quantized, and ideally quantized coefficient filters shown in Fig. 12. In Fig. 14, conventional coefficient quantization is again seen to yield unacceptable filter behavior.

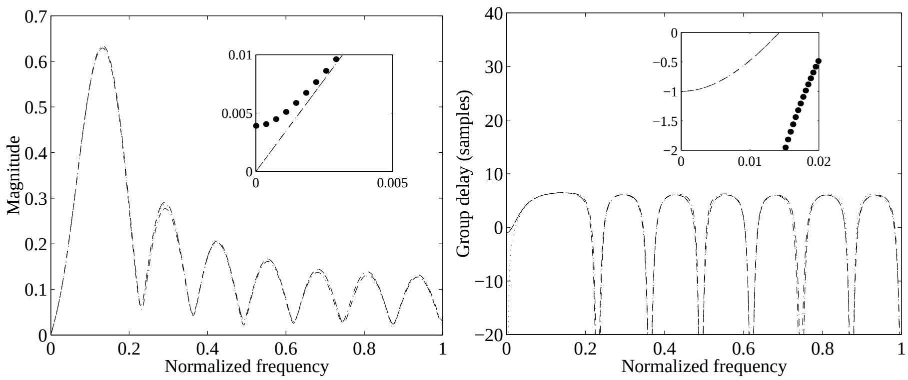

Fig. 12. Magnitude responses (left) and group delays (right) of the infinite precision (dashed), conventionally quantized (dotted), and ideally quantized (dash-dot; mostly overlaps dashed) coefficient second-degree one-step-ahead PPFD of length $N = 16$ with the coefficient precision of 8 bits [P3].

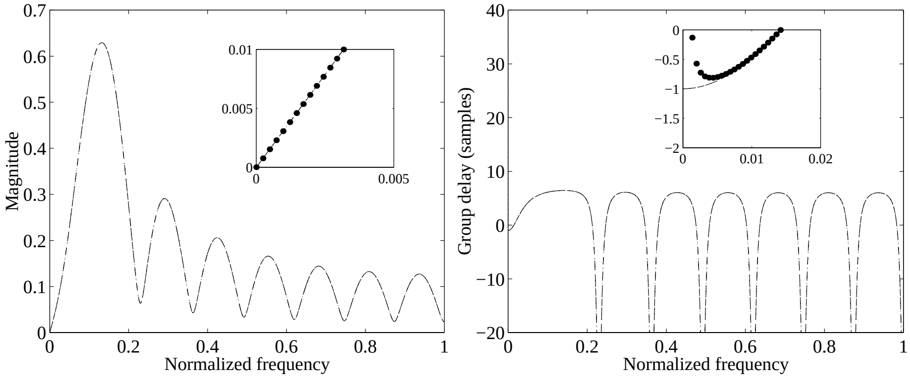

Fig. 13. Magnitude responses (left) and group delays (right) of the infinite precision (dashed), conventionally quantized (dotted), and ideally quantized (dash-dot; mostly overlaps dashed) coefficient second-degree one-step-ahead PPFD of length $N = 16$ with the coefficient precision of 16 bits [P3].

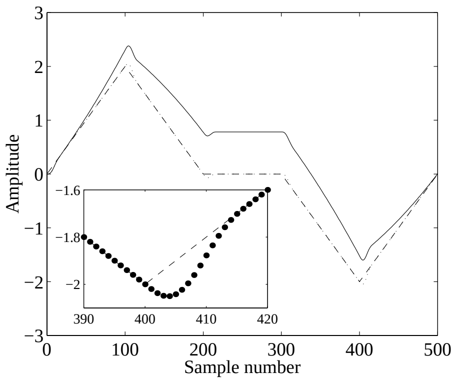

Fig. 14. Differentiation of the piece-wise polynomial signal seen in Fig. 11; the desired filter output (dashed), differentiation by the PPFD of length $N = 16$ with truncated 8-bit coefficients (solid), and differentiation by the ideally quantized 8-bit coefficient PPFD (dotted) [P3].

In [P1], coefficient truncation sensitivity of polynomial-predictive FIR differentiators is investigated. As sensitivity criteria, absolute values of the errors of the magnitude response at zero frequency, and absolute values of the errors of the group delay values near zero frequency, are employed. Both magnitude response and group delay values at zero frequency are exactly set by the design criteria. Also, coefficient quantization effects on the noise gain (3) of the filters are observed. The second-degree $I = 2$ polynomial-predictive FIR differentiators of length $N = 3, 4, \ldots, 100$, are taken as examples. Two's complement magnitude truncation of coefficients to 8 and 16 bits is employed. The results show that 97 of the 98 filters are affected by the coefficient quantization, and for example, there are eight second-degree PPFDs within $N = 3, 4, \ldots, 100$, whose group delay error at zero frequency is less than one percent. Also it is observed that for the tested filters, the errors in magnitude response and group delay are correlated so that robustness of the magnitude response always implies also robustness of the group delay properties. Thereafter, the same 98 filters are converted into lattice structures [44, pp. 476–485], which are known to have low coefficient quantization sensitivity [44, pg. 502], and the quantization effects are observed. The results show that lattice structures are generally more robust to coefficient quantization than the corresponding direct-form FIR implementations, but the correlation between magnitude response and group delay errors is lost, which is a noteworthy aspect in designing coefficient quantization robust FIR filters also in general.

The conclusion of [P1] is that the coefficient quantization effects are significant, and have to be considered in the filter design process. Paying special attention to the group delay properties, lattice structures may be used to alleviate coefficient quantization effects.

4.1.2.1 Scientific Value of the Publication [P1]

The publication [P1] clearly shows that coefficient quantization is a problem with polynomial-predictive FIR differentiators, and it can be alleviated with careful filter selection or through lattice implementations. It is also noted that lattice implementation may result in filters whose prediction properties are not robust against coefficient quantization even if the magnitude response properties are.

4.2 Design of Fixed-Point Polynomial Predictors and Polynomial-Predictive Differentiators

4.2.1 Publication [P2]: Genetic Algorithm for PPFD Design

In publication [P2], quantized coefficient polynomial-predictive FIR differentiators are designed using a genetic algorithm. The devised genetic algorithm is described in detail in the publication. The algorithm is found capable of finding improved, thought not prefect, quantized coefficient PPFDs in most cases. Designing second-degree one-step-ahead predictive PPFDs of length $N = 40$ using the fitness function (34) as the criterion, the genetic algorithm succeeded in finding improved filters in 5 out of 9 runs, and for $N = 80$, all nine runs produced improved filters. In [16] optimizing polynomial-predictive IIR feedback coefficients with genetic algorithms is considered, whereas in this thesis, genetic algorithms are used in designing coefficient quantization robust PPFDs. The devised fitness function to be minimized, given by

$$\begin{aligned} e &= \sum_{\omega_F} \left[F_Q(\omega_F) - F(\omega_F) \right]^2 + \sum_{\omega_G} \left[G_Q(\omega_G) - G(\omega_G) \right]^2, \\ \omega_F &= \{0.0001, 0.001\} \text{ and } \omega_G = \{0.0001, 0.001\}, \end{aligned} \tag{34}$$

is not designed for any specific application, but it only pursues to promote emergence and growth of predictive differentiation properties in the population. In (34), $F(\omega_F)$ and $F_Q(\omega_F)$ are magnitude response values of the desired filter and the quantized coefficient candidate filter, respectively, at the normalized frequency points ω_F, and $G(\omega_G)$ and $G_Q(\omega_G)$ are group delay values of the desired filter and the quantized coefficient candidate filter, respectively, at the normalized frequency points ω_G. Therefore, the algorithm, and especially the fitness function, is to be modified according to the specific application. On a 500 MHz Pentium III, 200 generations of the genetic algorithm proposed in [P2], programmed with Matlab, takes approximately half an hour to run.

4.2.1.1 Scientific Value of the Publication [P2]

The publication [P2] demonstrates that is it possible to design quantized coefficient polynomial-predictive FIR differentiators that perform better than the corresponding conventionally quantized coefficient filters. A genetic algorithm for the filter design task is devised and presented.

4.2.2 Publication [P3]: Ideally Quantized Coefficient PFP and PPFD Design

In [P3], a design method for coefficient quantization error free, i.e., ideally quantized coefficient PFPs and PPFDs is devised, c.f. Fig. 10 for an example of designed ideally quantized coefficient PFPs, and Fig. 12 and Fig. 13 for examples of designed ideally quantized coefficient PPFDs. In most cases considered in [P3], ideal quantization is achieved with as low as 6-bit, and in some cases even with 4-bit coefficients.

Basic requirement for a PFP, or a PPFD, to be coefficient quantization free, is that its coefficients belong to a quantized coefficient space $h(k) \in \mathcal{H}$, and exactly fulfill the constraints (5)–(8) for PFPs, or (15)–(18) for PPFDs. At simplest, these quantized coefficient PFPs and PPFDs can be found by a straightforward search algorithm. An exhaustive search within a vicinity around the infinite precision coefficients for the desired filter degree, length, and prediction step, is successfully used in [P3]. After finding all the quantized coefficient solutions to the set of equations (5)–(8), or (15)–(18), within the selected region around the infinite precision solution, the solution with the minimum noise gain (3) is selected as the ideally quantized coefficient filter. The selected search region is ±2 quantization levels around the infinite precision coefficients. This means that the noise gain (3) is minimized only locally within the search region; global noise gain minimization would present us with an impossibly long computation time, since the whole quantized coefficient space $\mathcal{H}$ would have to be searched through. For example, with 8-bit coefficients and filter length $N = 10$ this would mean $(2^8)^{10} \approx 1.2 \cdot 10^{24}$ different filter candidates to test for the fulfillment of the constraints. As the infinite precision PFPs and PPFDs are designed for exact prediction and/or differentiation, and thereafter to minimize the noise gain, the corresponding ideally quantized coefficient filters necessarily exhibit higher noise gain than their infinite precision coefficient counterparts. In [P3], it is observed that this noise gain growth can be regarded negligible; noise gain growth of the ideally quantized coefficient PFPs and PPFDs, $N = 16$, $I = 2$, $p = 1$, with coefficient precisions 8, 10, 12, 14, and 16 bits, with respect to the noise gains of the corresponding infinite precision coefficient filters, are presented in Table 2 [P3]. In the worst case presented in Table 2, the noise gain growth is only 0.3 percents. The negligible noise gain growth also justifies the proposed search range of ±2 quantization steps around each infinite precision coefficient of the filter to find ideally quantized coefficients since it would not be worth the required computation time with an enlarged search region to find ideally quantized coefficient filters with even lower noise gain values; with a search range of ±4 quantization steps around each infinite precision coefficient, it would take of the order of 10 months to find the ideally quantized coefficient PFP of the length $N = 16$ which minimizes the noise gain within the search region, while with a search band of ±2, this takes approximately 47 minutes on a 166 MHz Pentium.

Table 2. Noise gain growth in percents of the approximate noise gains of the infinite precision (Inf. prec.) PFPs and PPFDs with $I = 2$, $p = 1$, and $N = 16$ with several coefficient precisions [P3].

Coefficient precision (bits)	8	10	12	14	16	Inf. prec.
Noise gain, PFP	0.3 %	0.02 %	$9 \cdot 10^{-4}$ %	$3 \cdot 10^{-4}$ %	$3 \cdot 10^{-6}$ %	0.7303571
Noise gain, PPFD	0.2 %	0.01 %	$7 \cdot 10^{-4}$ %	$3 \cdot 10^{-5}$ %	$2 \cdot 10^{-6}$ %	0.0535364

In [P3], the problem of solving the constraints on the filter coefficients is formulated as an integer programming problem. This is advantageous, since more efficient search algorithms can be devised based on integer programming theory. Also working with decimal numbers in a computer, it is hard to know if the values are truly exact or not, whereas all the solutions found in integers that fulfill the constraints (5)–(8) for PFPs, or (15)–(18), are exact for sure.

The devised search algorithm is suitable for finding relatively short ($N \leq 32$) PFPs and PPFDs due to the increased computation time required for finding longer filters. Because short PFPs and PPFDs are preferred basis filters for the magnitude response shaping feedbacks [38], this is not an actual drawback.

4.2.2.1 Scientific Value of the Publication [P3]

The results presented in [P3] allow application of the PPFDs and PFPs in fixed-point applications, which basically was not previously possible. This is possible via the devised and described ideally quantized coefficient PFP and PPFD design algorithm, which yields PFPs and PPFDs that exactly fulfill the design criteria, i.e., the constraints (5)–(7), *or* (15)–(17), *respectively, with quantized coefficients.*

4.2.3 Publication [P4]: Coefficient Quantization Error Free Feedback Extension Design

In the earlier literature, coefficient quantization had been identified as a problem with the feedback extended PFPs, but no final solution had been proposed until now. The feedback extension provides powerful means of shaping the PFP (or PPFD) magnitude response without affecting the desired prediction (and differentiation) properties. This exact property is destroyed by improper quantization of feedback extension coefficients if the quantization results in violation of the relation between the coefficients $b(k)$ and $1 - b(k)$, $k = 1, \ldots, N$. A feedback extended structure for an FIR of length $N = 2$ is show in Fig. 2, and is reproduced here as Fig. 15 for convenience with one MAC result quantization error feedback filter of length $N_e = 2$ installed. In [P4], fixed-point feedback extension design criteria are found and given for designing feedback extensions that exactly preserve the desired properties of the underlying FIR filters even under feedback coefficient quantization. Two examples of ideally quantized coefficient feedback extended first-degree one-step-ahead predictive PFPs of length $N = 2$ are shown in Fig. 16, along with their basis filter. For exact implementation of feedback extended PFPs seen in Fig. 16, 8-bit coefficient precision with 4 fractional bits is sufficient.

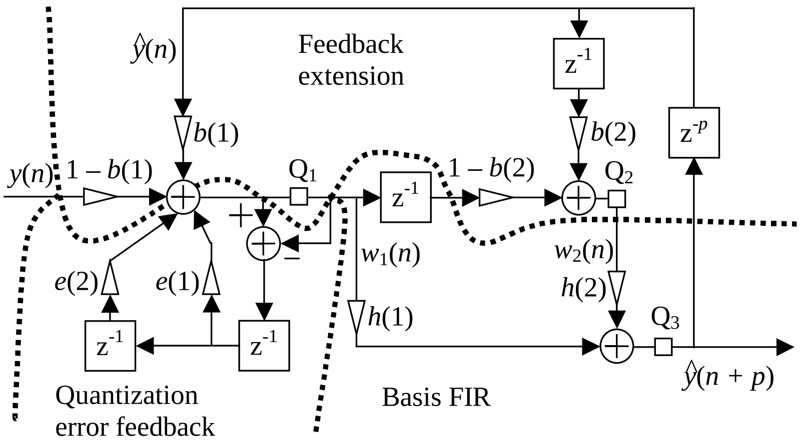

Fig. 15. Structure of a feedback extended PFP of length $N = 2$ with one quantization error feedback filter of length $N_e = 2$ installed.

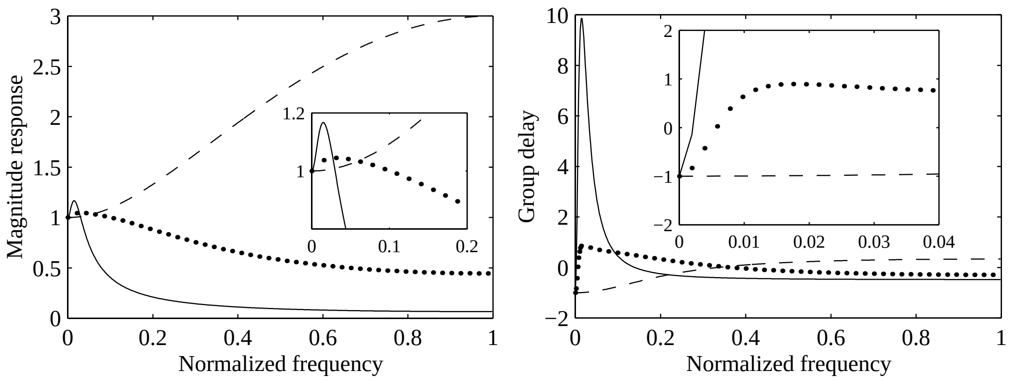

Fig. 16. Magnitude responses (left) and group delays (right) of coefficient quantization error free feedback extended one-step-ahead predictive first-degree PFPs with the feedback coefficients **b** = [0.6875 –0.9375] (dotted) and **b** = [0.9375 –0.9375] (solid), along with their basis filter basis FIR **h** = [2 –1] (dashed).

4.2.3.1 *Coefficient Quantization Error Free Augmented Form Fixed-Point Feedback Extension Design*

Taken that the coefficients of the selected basis FIR are ideally quantized, e.g., $h(k) \in \mathcal{H}_8$, $k = 1, \ldots, N$, and constraints for the filter coefficients are exactly fulfilled, the requirements for the coefficient quantization error free fixed-point feedback extension are that the feedback extension coefficients belong to the same quantized coefficients space, i.e.,

$$\{b(k)\,,\, 1 - b(k)\} \in \mathcal{H}_8. \qquad (35)$$

The latter condition in (35) is necessary because the effect of the feedback extension must be exactly that of weighted average only, in order to guarantee that the magnitude response and group delay properties at zero frequency are not affected. Observing that the two constraints on the feedback coefficients can be easily fulfilled, augmented form quantized coefficient feedback extension is seen to provide flexible design possibilities.

4.2.3.2 *Coefficient Quantization Error Free Direct Form Fixed-Point Feedback Extension Design*

Writing out the transfer function of the filter seen in Fig. 2 (or in Fig. 15 without the error feedback), we get [38]

$$
\begin{aligned}
H(z) &= \frac{[h(1)-h(1)b(1)]+[h(2)-h(2)b(1)-h(2)b(2)+h(2)b(1)b(2)]z^{-1}}{1-h(1)b(1)z^{-p}-[h(2)b(1)+h(2)b(2)-h(2)b(1)b(2)]z^{-p-1}} \\
&= \frac{B(1)+B(2)z^{-1}}{1-A(1)z^{-p}-A(2)z^{-p-1}}
\end{aligned}
\tag{36}
$$

in which the direct form IIR coefficients $B(k)$ and $A(m)$ can be identified. Now, calculating the coefficients $A(1)$, $A(2)$, $B(1)$, and $B(2)$, and quantizing them to yield $A_Q(1)$, $A_Q(2)$, $B_Q(1)$, and $B_Q(2)$, respectively, may well result in a situation that the system of equations

$$
\begin{cases}
A_Q(1) = h(1)b(1) \\
A_Q(2) = h(2)b(1)+h(2)b(2)-h(2)b(1)b(2) \\
B_Q(1) = h(1)-h(1)b(1) \\
B_Q(2) = h(2)-h(2)b(1)-h(2)b(2)+h(2)b(1)b(2)
\end{cases}
\tag{37}
$$

is not uniquely solvable any more, since the quantization may be such that $A_Q(1) \neq A(1)$, $A_Q(2) \neq A(2)$, $B_Q(1) \neq B(1)$, and/or $B_Q(2) \neq B(2)$, while $h(k)$ and $b(k)$, $k = 1, 2$, remain fixed to their original design values which lie in $\mathcal{H}$, for example in $\mathcal{H}_8$. This means, that the direct form implementation would not exactly correspond to the original feedback extended FIR, and the magnitude response and group delay properties at zero frequency would not be preserved in coefficient quantization. Therefore, it is necessary, that the direct form IIR coefficients are calculated starting with such basis FIR **h** and feedback extension coefficients **b**, that $A(m) = A_Q(m)$, and $B(k) = B_Q(k)$, in this example $\{k, m\} \in [1,2]$. For this it is necessary and sufficient that all the products in (36) making up the $B(k)$'s and $A(m)$'s after summations, are quantization error free, i.e., it is sufficient and necessary to require that

$$\{h(k), b(m), 1-b(m), h(k)b(m), h(k)b(m)b(m')\} \in \mathcal{H},\ \{k, m, m'\} \in [1,2], \tag{38}$$

with, for example $\mathcal{H} = \mathcal{H}_8$. As shown in [P4], it is possible to find such feedback extension coefficients **b** that (38) holds. Observing the constraints (38), and comparing them with the constraints of the augmented form quantized coefficient feedback extension structure (35), it is clear that the augmented form feedback extension structure offers much more freedom for the filter designer. Also, since the finite-size coefficient space $\mathcal{H}$ and the constraints (38) are together very restrictive, the resulting direct form IIR predictors posses impractical magnitude responses [P4] according to our experience. Thus, from a practical point of view, while it is possible to design ideally quantized coefficient feedback extended polynomial predictors as direct form IIRs, it is not worthwhile. Analogous coefficient quantization error free design constraints also apply to the fixed-point feedback extensions for the polynomial-predictive FIR differentiators.

4.2.3.3 Arithmetic Operation Result Quantization Error Feedback

In [P4], also effects of finite precision arithmetic are discussed. As an example, roundoff noise analysis is performed on a feedback augmented polynomial FIR predictor of length $N = 2$. Roundoff noise contributions of each summation node output quantization, Q_i, $i = 1, 2, 3$, in Fig. 15, are analyzed. It is observed that since the employed FIR coefficients are particularly convenient, $\mathbf{h} = [2\ {-1}]$, the MAC result feeding into the quantizer Q_3 in Fig. 15 does not produce any quantization errors, since with the quantized samples $w_1(n)$ and $w_2(n)$, c.f. Fig. 15, feeding into the filter $\mathbf{h} = [2\ {-1}]$, it is observer that

$$\{2\, w_1(n) - w_2(n)\} \in \mathcal{H}_8, \text{ since } \{w_1(n), w_2(n)\} \in \mathcal{H}_8. \tag{39}$$

Thereafter, the strongest noise source is identified through roundoff noise analysis presented in Chapter 2.4.1. The roundoff noise analysis showed that the quantizer Q_3 would contribute most to the roundoff noise power of the filter, but due to the just mentioned property, the roundoff noise contribution of Q_3 is actually zero, and most of the roundoff noise power is seen to be produced by the quantizer Q_1.

To alleviate the roundoff noise effects, an extremely simple error feedback is installed over identified single summation node associated with the quantizer Q_1, Fig. 15. The error feedback is shown effective in alleviating roundoff noise even though the feedback filter coefficients are chosen on the basis of cost effective computations as $\mathbf{e} = [1\ 0]$, or $\mathbf{e} = [1\ 1]$, and not through any rigorous optimization process. Also, only 1, 2, or 3 most significant bits of the quantization error are fed back through the quantization error feedback filter. In prediction of a low-frequency sinusoid signal, a feedback extended PFP with FIR and feedback extension coefficients $\mathbf{h} = [2\ {-1}]$, $\mathbf{b} = [0.9375\ {-0.9375}]$, respectively, with a tree-bit error feedback with $\mathbf{e} = [1\ 1]$, was able to decrease the MSE at the filter output by nearly 50 %. The simple error feedback was not effective for some feedback extended PFPs and input signals, though.

As noted in Section 2.4.1, the roundoff noise analysis used in identifying the noisiest summation node is not valid for the applied input signals (a *noiseless* ramp and sinusoid) in the examples presented in [P4]. Therefore, definite conclusions on the noisiest summation node cannot be made based on the presented analysis unless the input signal is noisy, as described in Section 2.4.1. This does not affect the quantization error free nature of the quantizer Q_3. Nevertheless, significant MSE decrease was obtained by attaching the quantization error feedback to the identified summation node. Thus, the examples given in [P4] fulfill their purpose as examples of effectiveness of the quantization error feedback. For the best results in an actual implementation, the roundoff noise contributions of individual summation nodes should be simulated or analyzed taking the actual (or anticipated) quantization noise power spectrum into account.

4.2.3.4 Scientific Value of the Publication [P4]

Publication [P4] gives the necessary and sufficient constraints, which the quantized coefficients of a frequency response shaping feedback for an ideally quantized coefficient PFP have to fulfill, in order for the feedback not to affect the prediction properties set forth exactly by the underlying FIR filter.

Roundoff noise analysis is preformed on the exemplary augmented FIR filter structure, and it is noted that the applied common method for roundoff noise analysis does not necessarily *yield satisfactory estimates, but also the results of the arithmetic operations with regard to signal characteristics and coefficient values must be analyzed. To alleviate roundoff noise problems, a simple error feedback is proposed.*

4.3 General Parameter Adaptive Polynomial FIR Predictors

4.3.1 Publication [P5]: Fine-Tuning FIR Predictors with General Parameter Adaptation

Generally, statistics of real-world signals tend to fluctuate over time, especially of the signals encountered in mobile communications systems. Therefore, there exists clear motivation for designing adaptive FIR predictors. In this thesis, simplicity of computations has been emphasized in many ways; short FIR filters form the basis of this thesis, feedback extension is applied to shortest possible filters, and roundoff error feedback is designed with short and computationally efficient error feedback filters. This is strongly motivated by the proposed application in mobile communications systems. Also, short basis FIRs for feedback extensions allow for more freedom for magnitude response shaping with the feedback extension coefficients; this has been indicated by the feedback optimization proposed in [17].

Following these lines, the adaptation employed in [P5] has a simple structure; it consists of a single general parameter (GP) that is added to each polynomial FIR predictor coefficient and adapted in an LMS-like adaptation, c.f. Chapter 2.6 for GP adaptive filtering in general, and Fig. 4 for the computational structure. GP adaptation, although not so well known, is shown capable for extending the applicability range of the polynomial FIR predictors, and in this work, also sinusoidal FIR predictors [22][51] are considered.

In [P5], the GP method is described, and applied to polynomial and sinusoidal FIR predictors. The polynomial predictors are applied in prediction of Rayleigh fading signals, which model fading present in mobile radio channels. The sinusoidal predictors are applied in prediction of sinusoids with frequencies differing from their nominal values. The Rayleigh fading prediction is intended for application in mobile communications system closed loop power control, and the sinusoidal prediction in power line frequency zero crossing detection. Also, a GP extended polynomial FIR predictor is tested in sinusoid prediction, c.f. Fig. 17. In [P5], the GP method is shown to have potential in expanding applicability ranges of fixed FIR filters designed for specific input signal statistics.

In [P5], stability conditions for GP filtering are derived; for stability it is required that the adaptation gain factor γ in (28) is within the limits

$$0 < \gamma \leq \frac{2}{\mathrm{E}[\mathbf{y}(n)\,\mathbf{S}(n)]} \tag{40}$$

where E[·] denotes expectation, $\mathbf{y}(n)$ is a row vector consisting of N latest input samples, and

$$\mathbf{S}(n) = \mathbf{I}\sum_{k=1}^{N} y(n-k+1). \tag{41}$$

where **I** is a unit column vector of length *N*. On the upper bound of (40), the GP update equation (28) is given by

$$\beta(n+1) = \beta(n) - \frac{1}{\sum_{k=1}^{N} y(n-k+1)}\left[\hat{y}(n) - y(n)\right] \tag{42}$$

which provides for the fastest stable adaptation possible in the sense of the largest change in filter coefficients. As the stability bound only guarantees that the prediction error will remain finite, employing the GP update on the stability bound (42) is generally dangerous as it may result in an oscillating error whose amplitude is large compared to the signal amplitude. Instead, an adaptation gain factor γ can be estimated form a stationary input signal using (40), and be employed in (28) on pg. 13. For this, upper bound of the stability region (40) can be observed over a period of time for a stationary input signal, and the gain factor γ can be set according to the found minimum upper bound of (40).

By its nature, GP adaptation does not yield convergence to constant coefficient values in any sense, except that bias error in coefficient is fully corrected by the GP adaptation. In contrast, for example in LMS adaptive filtering, the coefficients generally are expected to converge. Generally, flexibility of the adaptation of the GP method is insufficient to allow convergence to a fixed parameter value. In this thesis, GP adaptation is applied with FIR filters that are designed for specific time domain properties of the input signals. Having an input signal with different properties, GP adaptation is seen to track the input signal variation, c.f. Fig. 17 for GP behavior in GP extended polynomial prediction of a sinusoid signal. In Fig. 17, a periodic GP variation can be observed.

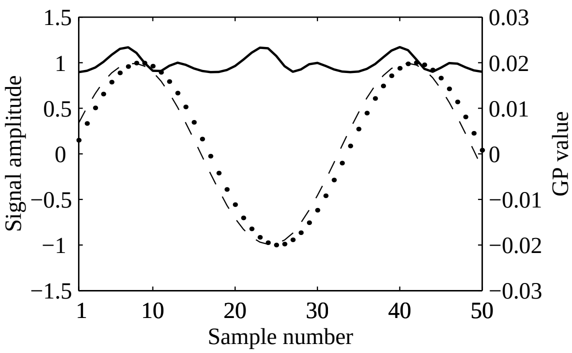

Fig. 17. One-step-ahead sinusoid prediction with a GP extended first-degree polynomial predictor (dashed), along with the filter input signal (dotted). Also shown is the GP value (solid).

To clearly demonstrate effects of GP filtering, the cases shown in [P5] are noiseless. For noisy situations, encountered in practice, filters with inherent noise attenuation capabilities are to be used as basis FIRs, or noise reduction is to be taken care of, for example, by prefiltering.

4.3.1.1 Scientific Value of the Publication [P5]

Publication [P5] promotes a simple and computationally efficient adaptive FIR filtering scheme; the general parameter method, in which a single adaptive parameter is added to each FIR coefficient. The stability bound for the adaptation is derived. The applied GP adaptation is shown to enhance the applicability range of basis polynomial or sinusoid FIR predictors.

4.4 Predictive Closed Loop Transmitter Power Control for Mobile CDMA Communications Systems

A received power level based closed power control loop is schematically illustrated in Fig. 18 [P6] for a single uplink transmission with predictive received power measurement (for computational schematics c.f. Fig. 8 on pg. 18 and for a WCDMA SIR based power control system block diagram Fig. 9 on pg. 20). The corresponding single user loop in a multiuser simulator is shown in Fig. 19 [P7]. The prediction in Fig. 18 can be done independently in the quadrature components after demodulation (32), pg. 15, or on the calculated received signal power. In the single closed loop power control model shown in Fig. 8, predictive filtering can operate on the “channel variation”, whereas in the WCDMA power control system in Fig. 9, predictive filtering can be applied, for example, on the output of the “Received DPCCH power estimation” block, or on the calculated SIR (“SIR cal.” in Fig. 9), both in the base station and in the mobile station. Also, predictive received power level and/or SIR estimation could be readily applied in a closed loop power control system, which combines these two control variables [61].

In a closed power control loop, critical factors are overall stability and delay. Although applying stable filtering inside a control loop does not imply that the resulting closed loop control system would remain stable, stability of the applied filter is naturally a necessary requirement. For accurate total loop delay compensation in real applications, the total loop delay should be estimated, and the prediction step should be adjusted to this. In practice, a predictor with an adaptive prediction step would be beneficial because of the changing radio channel. This could be achieved by the GP adaptation [5] attached to polynomial predictors [P5], or by other adaptive filtering methods. Also, a switched set of fixed polynomial predictors could be employed.

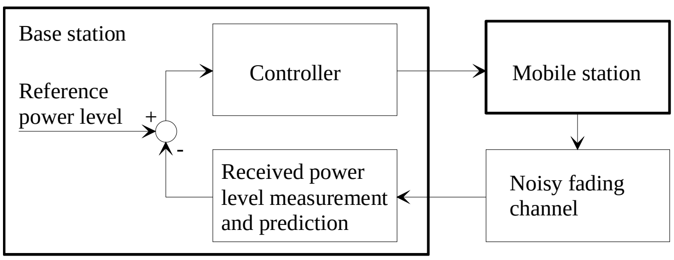

Fig. 18. Uplink closed loop power control in a CDMA system with received power level prediction.

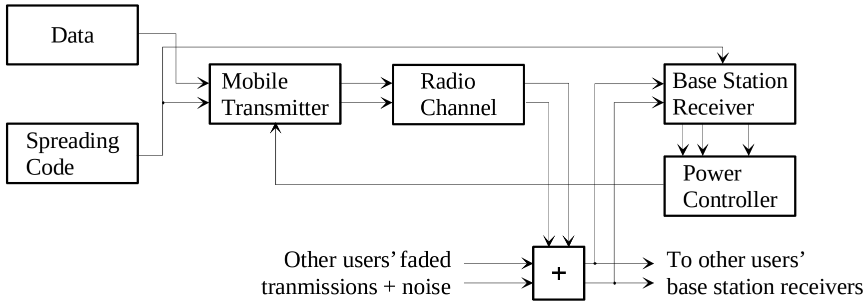

Fig. 19. Single user uplink connection in a multiuser simulator with a closed power control loop and connections to other users shown [P7].

4.4.1 Publication [P6]: Rayleigh Fading Prediction

In publication [P6], a predictive closed loop power control scheme for mobile communications systems is proposed, and one-step-ahead predictive first and second degree PFPs of lengths $N = 3, 4, \ldots, 50$, are applied in prediction of Rayleigh fading signals. The used Rayleigh fading signals model fast fading present in a radio channel at urban mobile speeds of 5 and 50 km/h. The results are given in a form of SNR gains with input signal SNR's of 10 dB and 0 dB; for this an SNR measure for complex signals was developed. The results show that the tested Rayleigh fading signals are well predictable by the means of low-degree polynomial predictors. These results can also be used as a starting point for selecting polynomial FIR predictors for applications in Rayleigh fading environments.

4.4.1.1 Scientific Value of the Publication [P6]

Publication [P6] introduces the possibility of predicting Rayleigh fading signals using a low-degree polynomial signal model. It is shown that the approach is indeed beneficial for the task. The publication also gives guidelines for polynomial FIR predictor selection in Rayleigh fading prediction applications.

4.4.2 Publication [P7]: Predictive Power Control for CDMA Systems

In publication [P7], polynomial FIR predictors are applied to the same task as in [P6] but in a simulated multiuser CDMA environment with received power level based closed loop control. The simulator corresponds to the mobile CDMA communications system of Qualcomm [45, pp. 10–11, 34–37] in all the aspects essential in determining the effects of polynomial prediction, except that the Qualcomm system employs a received SIR based power control scheme. Selection of received power level or received SNR as the control variable only affects the role of predictive estimation. In the received power level based control, prediction of the actual control variable is provided directly, while in the SNR based power control, received power level prediction can be employed in calculating the estimate of the control variable.

For this research, a multiuser CDMA communications system was developed in COSSAP simulation environment, c.f. Fig. 19 for the block diagram of the single user power control loop in the constructed multiuser simulator [P7]. The simulator consists of a mobile transmitter, radio channel, base station receiver, and power controller modules for each active user. The simulator modules are described in the publication. Only uplink transmissions are considered since that is more crucial than downlink from the power control quality requirement point of view; since the transmissions originate from several independent mobile transmitters, power control is absolutely necessary in avoiding the situation in which nearby users would be received on high power levels prohibiting reception of faraway users who would be received at a low power level. On downlink on the other hand, not considering other cell interference, transmission originates from a single base station transmitter only, and thus no near-far effect [41, pp. 62–66] occurs at the mobile receivers. Still, as in the power control system presented in Fig. 9, both uplink and downlink power control systems may be identical, and benefit from predictive received power level (or received SNR) estimation. Also, the methods proposed in [P6]–[P9] are equally well applicable in uplink and downlink power control, whichever is considered by the communications system at hand. In the simulator, all the users are connected to a single common base station, and interfere with each other, for which effect non-orthogonal user-specific spreading codes are used in transmissions.

The simulation results are observed by relative comparisons of power consumption of a user, variance of the received power level at the base station, and the received bit-error-rate (BER), achieved with predictive and non-predictive power controllers. Relative power consumptions are observed as transmitted signal amplitudes integrated over each simulation run. The simulations show that predictive FIR filtering has potential for CDMA power control system applications. In most cases, when the received BER stays essentially constant, decrease of a few percents in variance of the received power level is observed while also the relative power consumption is decreased slightly, i.e., by 1 % or less. The effect of fine-tuning the closed loop control timing by a few bit durations with the polynomial FIR predictors is also demonstrated in [P7] where correctly predicted power-down commands are illustrated.

4.4.2.1 Scientific Value of the Publication [P7]

Publication [P7] demonstrates potential benefits of applying polynomial FIR predictors in a multiuser mobile CDMA communications system power control. This is observed as slight improvements in relative power consumption and received power level variance, both of which are to be minimized without sacrificing reception quality. Predictors are also illustrated to correctly predict power down commands, i.e., to fine-tune the closed loop control timing.

4.4.3 Publication [P8]: Optimum Power Prediction of Complex Signals

In a mobile communications receiver, a baseband equivalent signal after demodulation is complex-valued. Thus power prediction may be performed using two approaches: direct prediction of signal power, Fig. 20(a), or prediction of the complex components independently, Fig. 20(b). In [P6], polynomial FIR predictors of first and second degree are applied, and the two methods are compared in prediction of a complex-valued signal whose amplitude is Rayleigh distributed. Based on the results in [P6], polynomial FIR prediction in quadrature components is applied in multiuser simulations in [P7]. Power estimators that perform predictive filtering independently in the complex components, always yield positive power estimates at the power estimator output, Fig. 20(b), whereas with direct prediction of calculated signal power, Fig. 20(a), positive output cannot be automatically guaranteed. Thus, prediction in components is able to provide the power control system with valid received power level predictions also during those periods when the direct power prediction would yield negative received power levels, which should be interpreted as zero received power. Also, it is beneficial that as much as possible of the noise present in the complex components is removed before power calculation since the noise bandwidth doubles at squaring. Therefore, prediction in components is preferred over direct prediction of signal power. From the imposed computational load point of view, direct prediction of signal power is more efficient since it requires only one predictive filter as compared to two filters required for prediction in complex components. But, in direct power level prediction, predictor passband bandwidth should be designed for two times wider primary signal bandwidth than in the case of prediction in components, which calls for shorter predictors, which in turn exhibit poorer noise attenuation.

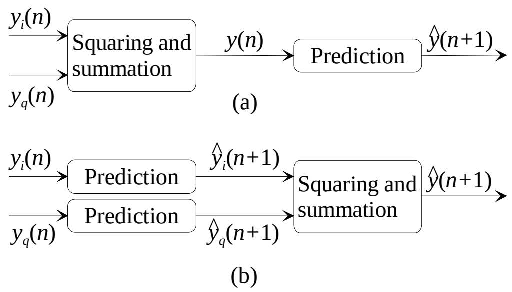

Fig. 20. Two schemes for prediction of power of a complex valued signal: (a) Direct prediction of calculated signal power (Hammerstein model). (b) Prediction in complex components before signal power calculation (Wiener model).

In publication [P8], a power estimator for complex signals which minimizes MSE at the outputs of the individual predictive filters is derived based on the Wiener model, Fig. 20 (b). In the Wiener model, powers of the complex components are first estimated independently and added thereafter. Implicit formula for global optimization of the power estimator is also given but since it cannot be solved in closed form, and since iterative numerical solution would require heavy computations and still exhibit uncertainty in the obtained filter coefficients, it is not elaborated further. Instead, a partially optimized power estimator is considered, and the solution is shown to be the Wiener-Hopf solution. Here the partial optimization means minimum MSE optimization of the individual filters in Fig. 20(b), in contrast to the global optimization in which the optimization (yielding minimized MSE at the filter output) is done with regard to the system output $\hat{y}(n+1)$. In [P8], the derived optimum predictive estimators are compared with polynomial FIR predictors in single user and multiuser CDMA system power control simulations, and also a non-predictive power control system is used as a reference. The optimum predictors are shown to yield results comparable to those with polynomial FIR predictors. Publication [P8] forms a publication pair together with [21]; in [21] a thorough treatment of the optimum Wiener model based power estimators are given, whereas [P8] provides a more communications simulation oriented view to the optimum prediction of power of complex valued signals.

4.4.3.1 Scientific Value of the Publication [P8]

In publication [P8], optimum power predictors, minimizing MSE at the outputs of the individual filters, are derived for complex valued signals. The derived optimum predictors are simulated in single user and multiuser CDMA power control simulations, and shown to yield comparable power control performance in the sense of decreasing variance of the channel output power level as compared to the corresponding power control employing polynomial prediction.

4.4.4 Publication [P9]: Received Power Level Prediction with Neural Networks

In publication [P9], a flexible neural network is designed for prediction of Rayleigh fading received power level in mobile communications systems. The simulation scenario is identical to that used in [P6] for predictability studies of Rayleigh fading signals by polynomial FIR predictors; carrier frequency is set to 1.8 GHz, received power level sampling frequency is 1 kHz, Jakes' Rayleigh fading received power level model with additive white noise to result in the SNR of 0 dB independently in the quadrature components is employed, and the mobile station is moving at urban speeds of 5 km/h or 50 km/h.

The neural network employed is a complex structure with learning capabilities, and it is tailored to the specific task at hand. The achieved SNR gains are listed in Table 3 for the neural network predictor [P9] and for the first-degree polynomial FIR predictors [P6]. The prediction is preformed both on inphase and quadrature components independently, and on the received power level. Noting the very high flexibility, learning capabilities, and computational complexity of the neural network predictor, it is clearly demonstrated by Table 3 that polynomial FIR prediction can be regarded as a highly efficient method for Rayleigh fading prediction; comparing the highest SNR gains achieved, at 5 km/h the neural network predictor provided for 3 dB higher SNR gain than the first-degree PFPs, and at 50 km/h, the neural network predictor could not outperform the PFPs, while the computational complexity of the neural network predictor was one or two orders of magnitude higher than that of the polynomial FIR predictors.

Table 3. Comparison of SNR gains and approximate computational complexities (appr. compl.) (arithmetic operations per output sample).

Predictor	Prediction of	5 km/h		50 km/h	
		SNR gain (dB)	appr. compl.	SNR gain (dB)	appr. compl.
Neural network	components	12	10000	7	10000
	power	11	5000	5	5000
Polynomial FIR	components	9	112 (N = 28)	8	120 (N = 30)
	power	5	100 (N = 50)	5	100 (N = 50)

4.4.4.1 Scientific Value of the Publication [P9]

Form the results in publication [P9], it can be concluded that even a complex neural computing structure with great learning capabilities does not produce significantly better received power level prediction than the polynomial FIR predictors. Thus, it is justified to regard polynomial FIR prediction as a highly efficient tool for Rayleigh fading received power level prediction.

4.5 Author's Contribution to the Published Work

The author's contribution to all the publications [P1] – [P9] has been essential. The work descriptions below have been agreed upon by all the co-authors of the publications.

[P1] The work for the publication [P1] was done by the author. The second author provided valuable insight to the research and comments on the manuscript.

[P2] The work for the publication [P2] was done by the author, including the genetic algorithm studies and devising the genetic algorithm for fixed-point filter design. The second author provided valuable comments on the manuscript.

[P3] The author was the main author responsible for the work, except for most of the Chapter 3. The original idea of designing coefficient quantization error free polynomial-predictive FIR differentiators and polynomial FIR predictors rests with the author. The second author was responsible for developing the method for designing the quantization error free filters, for designing the presented filters, and for writing most of the text of Chapter 3.

[P4] In [P4], the author was the sole contributor.

[P5] The author was responsible for most of the work for [P5]. The derivation of the normalized stability bound for the gain factor of the GP adaptation was done by the second author, from which result the author derived the GP adaptation rule at the stability boundary. The second author proposed also the running sum implementation of the GP extension. The author was responsible for all the simulations in [P5]. The original idea behind the research rests with the third author, who also provided valuable comments on the manuscript.

[P6] The work for the publication [P6] was done by the author, except for Chapter IV which is by the second author. The fourth author provided the original research topic, and along with the third author, provided valuable insight and comments on the publication.

[P7] The work for the publication [P7] was done by the author, including detailed communication system simulation studies, programming the COSSAP multiuser simulator, and designing the simulations. The second, third and fourth authors provided valuable comments on the simulator construction. The third and fourth authors also provided insight to the communications systems in general. The fifth author provided valuable comments on the manuscript, and on the signal processing matters.

[P8] The work for the publication [P8] was performed jointly by the author and the second author. The author was otherwise responsible for the publication, except for the work for parts of Chapter I and most of the Chapter II, including the derivations for the optimum power estimator, which were carried out by the second author. The author was also responsible for the simulations in the publication. The third author provided valuable comments on the research work and manuscript.

[P9] The author provided the fading channel model for simulations, defined the simulation experiments, and revised a draft of the publication. The publication was mainly written by the first author, and the second author assisted him with the simulations, and finalized the paper. The fourth author instructed the work.

5 Conclusions

5.1 Main Results

The main theme throughout this work is the applicability of polynomial model based signal processing with the application in the important power control problem in mobile CDMA communications systems. The application platform directly dictates also further research objectives, such as low computational complexity, preferably with low-precision fixed-point arithmetic, small die size, and low power consumption. The main results of the work presented in this thesis are:

1. Identification of coefficient quantization error problem with polynomial FIR predictors and polynomial-predictive FIR differentiators.
2. Coefficient quantization error free design of PFPs and PPFDs, and of magnitude shaping feedback extensions for these two FIR types.
3. Identification of possible roundoff noise problems with feedback extended PFPs, and PPFDs, and proposing a simple error feedback for roundoff noise alleviation.
4. A simple adaptive filtering scheme is shown effective in expanding the applicability of FIR predictors, and is thus proposed for mobile applications.
5. Rayleigh fading signals are shown to be predictable by the means of low-degree polynomial prediction.
6. PFPs are shown effective in fine-tuning closed transmitter power control loop timing in mobile communications CDMA systems.
7. Optimum power predictors for complex valued signals are derived.

To the best of our knowledge, items 2, 5, 6, and 7 were not presented in the literature before the author’s work, whereas items 1, 3, and 4 employ previously known techniques in a new context.

5.2 The Scientific Importance of the Author’s Work

In this chapter, a short summary of the scientific importance of the work, described as per publication in subchapters of Chapter 4, is presented.

Before the research reported in this thesis, it was not possible to apply PFPs or PPFDs in short word length fixed-point environments. This work provides straightforward methods for designing coefficient quantization error free PFPs and PPFDs. Also, the advantageous feedback extensions for shaping magnitude responses of PFPs and PPFDs, previously known from the literature, are in this thesis given the criteria that ensures that the quantized feedback coefficients do not disturb the exact prediction and/or differentiation properties of the underlying FIR filters. This means that now quantized coefficient FIR and feedback extended polynomial predictors and polynomial-predictive differentiators can be designed with zero error with regard to the design constraints. Of scientific importance is also the observation that even a simple error feedback can alleviate roundoff noise effects in feedback extended PFPs and PPFDs.

Use of computationally efficient prediction is introduced into the control system design with application in mobile CDMA closed loop transmitter power control. The closed loop control is inherently delay limited, and this work proposes a sound and simple means of alleviating effects of the delays, thus contributing to the user capacity of the CDMA communications systems.

5.3 Topics for Further Research

This thesis fills a previously apparent gap in the polynomial predictor and differentiator research; fixed-point design of polynomial predictors and differentiators, and also discussion on the truncation noise effects is given. The main aspects left to the filter designers are possible limit cycles and overflow problems of predictive IIR filters. Also, to further promote usage of polynomial based signal processing, the methods presented in the literature and in this thesis could be formulated into several application oriented design tools and publications, as in [35][34].

Looking into the direction set by the coefficient quantization error free design methods proposed in this thesis, also multiplierless predictor and differentiator designs would be worth searching for since they would offer prediction and differentiation capabilities at a greatly reduced computational cost.

Thereafter, looking at the features of the polynomial FIR predictors and differentiators that enabled coefficient quantization error free design, namely existence of linear constraints on filter coefficients, it would be interesting to know if more filter types could be described in a similar fashion, and finally designed to be coefficient quantization error free.

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