A Novel Multipath Interference Cancellation Scheme for RAKE Channel Estimation

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Abstract

A novel channel estimation scheme is proposed for a RAKE receiver operating in the downlink of a mobile communication scenario. The approach is an extension of the well known Wiener channel estimator and partly cancels multipath interference, with the additional advantage of being able to cope with physically closely spaced multipaths which arise, for example, in an indoor scenario. It is demonstrated by means of simulation that the algorithm improves the quality of the channel estimates significantly. Little additional computational effort is necessary to implement the algorithm, making it an ideal candidate for improving reception capabilities of the mobile user equipment.

1. Introduction

The mobile radio channel is characterized by multipath propagation where a number of reflected or scattered radio rays arrive at the receiving end. Each of the rays, as seen by the receiver, is characterized by a distinct phasor and time-delay. The RAKE receiver is suggested by ETSI/3GPP [4] as a low-complexity solution for fast time-to-market, and will be the receiver of choice for the first wave of UMTS/3GPP handsets. In each finger of the RAKE receiver, the attenuation, the phase shift, and the propagation delay of the selected multipath have to be compensated for, a task called synchronisation. Any realizable receiver follows the concept of synchronised detection [1] for which a channel estimate or a sampled version thereof must be formed and subsequently used for detection as if it were the true known channel. This paper is concerned with phasor estimation for a RAKE receiver operating in the downlink of a mobile communication scenario, i.e. the mobile user equiment. Usually, when performing phasor estimation in a RAKE finger, it is assumed that other multipaths do not amount to much interference and can be neglected. However, this is only true, when multipaths are spaced apart exact multiples of chip periods, otherwise performance of the receiver is degraded. Furthermore, in indoor scenarios, multipaths can be spaced very closely, down to one chip or less. The resulting model mismatch then becomes fatal and leads to severe performance drops. The channel estimation scheme presented here is able to cope very efficiently with these type of channels. A recently developed new class of timing-error-detectors [6, 7] can be used to deliver the high resolution timing estimates that the novel channel estimator relies on. After reviewing the system model, a novel phasor estimation algorithms appropriate for a RAKE receiver in the downlink of a multipath propagation scenario is developed. Finally, the performance and the advantages of the algorithm are discussed.

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2. System Model

The system we are concerned with consists of a CDMA transmitter sending a complex valued data sequence $\{a_n\}$. These data symbols are spread by the spreading factor N_c using the effective spreading sequence $\{d_\nu\}$ which is assumed to be complex valued as well. The spread sequence is transmitted using a pulse shaping filter $g(t)$ which in the case of 3GPP is a root-raised cosine filter with a rolloff factor of 0.22. The resulting baseband-equivalent transmit signal is given by

$$s(t) = \sum_n a_n \sum_{\nu=0}^{N_c-1} d_\nu g(t - nT - \nu T_c) \qquad (1)$$

where T denotes the symbol rate and T_c is the chip rate. In order to incorporate effective spreading sequences with a periodicity longer than one symbol (e.g. resulting from additional scrambling) d_ν may be replaced by $d_{\nu+N_c n}$. The signal from the transmitter travels through a multipath propagation channel with N_P independent paths (WSSUS model). Each of these paths is characterized by its delay τ_l and fading coefficient c_l, resulting in the channel impulse response $h(t,\tau) = \sum_{l=0}^{N_p-1} c_l(t)\,\delta(\tau - \tau_l)$. The received signal subject to white Gaussian noise $m(t)$ can therefore be written as

$$r(t) = \sum_{l=0}^{N_p-1} c_l(t) \sum_n a_n \sum_{\nu=0}^{N_c-1} d_\nu g(t - nT - \nu T_c - \tau_l) + m(t) \tag{2}$$

In the first stage of signal processing in the receiver the signal is filtered with a filter matched to the pulse forming filter in the transmitter. Thus, in combination with the following RAKE receiver, a channel matched filter is formed. The signal at the output of the pulse matched filter is given by

$$z(t) = \sum_{l=0}^{N_p-1} c_l(t) \sum_n a_n \sum_{\nu=0}^{N_c-1} d_\nu R_g(t - nT - \nu T_c - \tau_l) + \tilde{m}(t) \tag{3}$$

The noise term $\tilde{m}(t)$ represents both the noise filtered by the pulse matched filter and the interference from other users. The combined transmit and receive filter pulse form is denoted by $R_g(t)$, being the effective pulse form at the output of the pulse matched filter, i.e. $R_g(t) = \int_{-\infty}^{\infty} g^*(\tau) g(t+\tau)\,d\tau$. The matched filter output signal is used as input for the RAKE receiver which operates on samples from the matched filter output taken at an arbitrary rate $1/T_S$ (at least Nyquist sampling). In each of the M branches of the RAKE the samples are processed by means of interpolation and decimation [1] in order to obtain intermediate samples $z_{l,k}$ with a rate of $1/T_c$ and to compensate for the delay τ_l. Alternatively, instead of interpolation and decimation one could adapt the code phases in order to compensate the delays in the individual paths. The resulting time discrete signal for branch l of the RAKE can be expressed as:

$$\begin{aligned} z_{l,k} &= a_n c_{l,n} \cdot \sum_{\nu=0}^{N_c-1} d_\nu \cdot R_g(kT_c - nT - \nu T_c) + \\ a_n &\cdot \sum_{\substack{i=0 \\ i \neq l}}^{N_p-1} c_{i,n} \sum_{\nu=0}^{N_c-1} d_\nu \cdot R_g(kT_c - nT - \nu T_c - \tau_l + \tau_i) + \bar{m}_{l,k} \end{aligned} \tag{4}$$

Since Nyquist pulses are used for transmission, after pulse matched filtering and compensation of the path delay, only one transmitted symbol a_n contributes to the resulting signal originating from path l. As path delay compensation can only be done for one path at the same time this is not exactly true for the other paths. But in this framework it is sufficient to approximate this effect by an additional noise term contributing to $m_{l,k}$. Furthermore, the fading coefficients are assumed to be constant over one symbol interval.

The next step required for detecting the symbol is the despreading. In order to do so, the signal $z_{l,k}$ is now multiplied with the spreading sequence $\mathbf{d}$, before summing the signal over one symbol period. Therefore, the despread signal, denoted with $y_{l,n}$, simplifies to an expression very familiar to the matched filter output known from traditional (non-spread) frequency nonselective single-carrier communication systems, apart from some additional unwanted multipath interference.

$$\begin{aligned} y_{l,n} &= \sum_{k=nN_c}^{(n+1)N_c-1} d_{k \bmod N_c} z_{l,k} = a_n c_{l,n} + \\ &\underbrace{a_n \sum_{i,k,\nu} c_{l,n} d_{k \bmod N_c} d_\nu R_g(kT_c - nT - \nu T_c - \tau_l + \tau_i)}_{\text{multipath interference}} + \tilde{m}_{l,n} \end{aligned} \tag{5}$$

3. Phasor Estimation

The time-variant fading coefficients $\{c_{l,n}\}$ are modelled as complex-valued random processes. The random fading process is assumed to be white-sense stationary (WSS). Furthermore, the N_p fading processes are assumed to undergo mutually uncorrelated scattering (US), which is plausible since individual paths can often be attributed to distinct physical scatterers. The power spectral density $S(e^{j\omega})$ of a channel coefficient $c_{l,n}$ is given by the so-called Jakes spectrum. It is strictly bandlimited to the (normalized to symbol rate) Doppler frequency λ, and its autocorrelation is given by $R(n) = \rho_l \cdot J_0(2\pi\lambda \cdot n) = \rho_l \cdot \alpha(n)$, where ρ_l is the average process power of the lth path, f_0 is the carrier frequency, c is the speed of light, v is the velocity of the mobile handset, and $J_0(\cdot)$ is the Bessel function of the first kind of order 0. Based on the system model given by equation (5), the optimal channel estimator is given by the conditional expected value given all available observations $\mathbf{y}$ and the corresponding symbol sequence $\mathbf{a}$.

$$\hat{c}_{l,n} = E\left[c_{l,n} | \mathbf{y}, \mathbf{a}\right] \tag{6}$$

In [1], it is shown, that this optimal channel estimate can be generated by low-pass filtering maximum-likelihood channel estimates $\hat{\mathbf{c}}_{ML;l}$ with a Wiener filter $\mathbf{w}$ of bandwidth λ, i.e. we have that

$$\hat{c}_{l,n} = \mathbf{w}^H \cdot \hat{\mathbf{c}}_{ML;l} \tag{7}$$

Here, we constrain the Wiener filter to a fixed number of N_w co-efficients. The Wiener filter therefore reduces to a single, time-invariant N_w-tap FIR filter $\mathbf{w} = (w_0, w_1, \ldots, w_{N_w-1})^T$. In essence, it can be said that the phasor estimation algorithm relies on ML channel estimates, which are simply postprocessed (filtered) by a fixed FIR filter.

3.1. Maximum Likelihood Channel Estimation with Multipath Interference Cancellation

Usually, in a CDMA system the assumption is made that there is no unwanted multipath interference, and in that case the despreader output reduces to $y_{l,n} = a_n c_{l,n} + \tilde{m}_{l,n}$. If this assumption holds, the maximum likelihood (ML) channel estimates are simply given by $\hat{c}_{ML;l,n} = a_n^* \cdot y_{l,n} = c_{l,n} + \bar{q}_{l,n}$ [1]. Unfortunately, this assumption only holds in the special case of the additive white Gaussian noise (AWGN) channel, where the signal is not subject to multipath propagation. If, however, multipath propagation is present, the interfering multipaths cannot be neglected. Let's assume that ML phasor estimates are formed using the despreader output of equation (5) and a simple complex conjugate multiplication with a known symbol a_n. The multipath interference then introduces additional terms in the ML phasor estimate which are given by

$$\sum_i \sum_k \sum_\nu c_{l,k} d_{k \ mod \ N_c} d_\nu g(kT_c - nT - \nu T_c - \tau_{l,i}) = \sum_i R_g(-\tau_{l,i}) c_{i,n} + \tilde{q}_{l,n} \quad (8)$$

The first term is directly dependent on the raised cosine function $R_g(t)$, the timing delays τ_l and the phasors $c_{i,n}$ of the unwanted multipaths. This term manifests itself as a systematic error, being a random process of bandwidth λ given by the Doppler Spectrum $S(e^{j\omega})$. This interference cannot be suppressed by the Wiener filter and has to be cancelled beforehand, or otherwise, a significant performance degradation of the RAKE receiver may result. The second term $q_{l,n}$, however, is a function of $d_{k \ mod \ N_c} d_\nu$ for $k \ mod \ N_c \neq \nu$. Under the assumption of long scrambling sequences with good autocorrelation properties, this (stochastic) error will behave approximately like white Gaussian noise and can be suppressed, at least partially, by the Wiener filter. Therefore, it is now possible to write the ML phasor estimates as follows

$$\begin{aligned} \hat{c}_{ML;l,n} &= c_{l,n} + \sum_{\substack{i=0 \\ i \neq l}}^{N_p-1} R_g(-\tau_l + \tau_i) c_{i,n} + q_{l,n} \\ &= \sum_{i=0}^{N_p-1} R_g(-\tau_l + \tau_i) c_{i,n} + q_{l,n} \end{aligned} \quad (9)$$

where $q_{l,n} = \bar{q}_{l,n} + \tilde{q}_{l,n}$ is assumed to be AWGN of power σ_q^2. Rewriting this in matrix form for all RAKE fingers, we get

$$\underbrace{\begin{pmatrix} \hat{c}_{ML;0,n} \\ \vdots \\ \hat{c}_{ML;N_p-1,n} \end{pmatrix}}_{\hat{\mathbf{c}}_{ML;n}} = \mathbf{G} \underbrace{\begin{pmatrix} c_{0,n} \\ \vdots \\ c_{N_p-1,n} \end{pmatrix}}_{\mathbf{c}_n} + \mathbf{q} \quad (10)$$

where the matrix $\mathbf{G}$ is given by

$$\mathbf{G} = \begin{pmatrix} R_g(-\tau_0 + \tau_0) & \cdots & R_g(-\tau_0 + \tau_{N_p-1}) \\ \vdots & \ddots & \vdots \\ R_g(-\tau_{N_p-1} + \tau_0) & \cdots & R_g(-\tau_{N_p-1} + \tau_{N_p-1}) \end{pmatrix} \quad (11)$$

The minimum mean square error (LMMSE) [3] solution of the interference cancellation problem is thus given by

$$\underbrace{\begin{pmatrix} \hat{c}_{LMMSE;0,n} \\ \vdots \\ \hat{c}_{LMMSE;N_p-1,n} \end{pmatrix}}_{\hat{\mathbf{c}}_{LMMSE}} = \mathbf{C}_C \mathbf{G}^T \left(\mathbf{G}^T \mathbf{C}_C \mathbf{G} + \sigma_q^2 \cdot \mathbf{I} \right)^{-1} \hat{\mathbf{c}}_{ML;n} \quad (12)$$

Since the N_p fading processes are assumed to undergo mutually uncorrelated scattering, the autocorrelation matrix $\mathbf{C}_C$ can be written as

$$\mathbf{C}_C = \mathrm{diag}\left(\rho_0 \cdots \rho_{N_p-1} \right) \quad (13)$$

Furthermore, it is noticed, that knowledge of the multipath delays τ_l is required in order to perform the partial interference cancellation. These timing delays can be obtained from the timing-error-detector (TED), which is part of every RAKE finger. The multipath delays τ_l are extremely slowly varying processes, and for the purpose of channel estimation can safely be assumed constants. A new class of TED's, which was developed recently, is capable of delivering delay estimates even in the case of physically very closely spaced multipaths [6, 7].

3.2. Wiener Filtering

The resulting structure for channel estimation is illustrated in Figure 1. The required known symbols a_n^* are either generated by forming hard decisions after the RAKE combiner, or by inserting known pilot symbols. The ML channel estimates are postprocessed by the interference canceller, before they are filtered with the Wiener filter. The interference canceller obtains delay estimates from the TED. The corresponding Wiener filter is denoted in the illustration with $F(z)$. The Wiener phasor estimator is identical in form to the linear minimum mean square error (LMMSE) [3] estimator. In [1] it is shown, that the estimator and the corresponding MMSE are given by

$$\begin{aligned} \hat{c}_n &= \underbrace{\mathbf{c}_D^H \cdot (\mathbf{C}_D + N_0 \cdot \mathbf{I})^{-1}}_{\mathbf{w}^H} \cdot \underbrace{\begin{pmatrix} \hat{c}_{LMMSE,n-N_w+P} \\ \vdots \\ \hat{c}_{LMMSE,n-1+P} \end{pmatrix}}_{\hat{\mathbf{c}}_{LMMSE}} \\ \sigma_{c;W}^2 &= \mathbf{w}^H (\mathbf{C}_D + N_0 \cdot \mathbf{I}) \mathbf{w} - \mathbf{w}^H \mathbf{c}_D - \mathbf{c}_D^H \mathbf{w} + \rho \end{aligned} \quad (14)$$

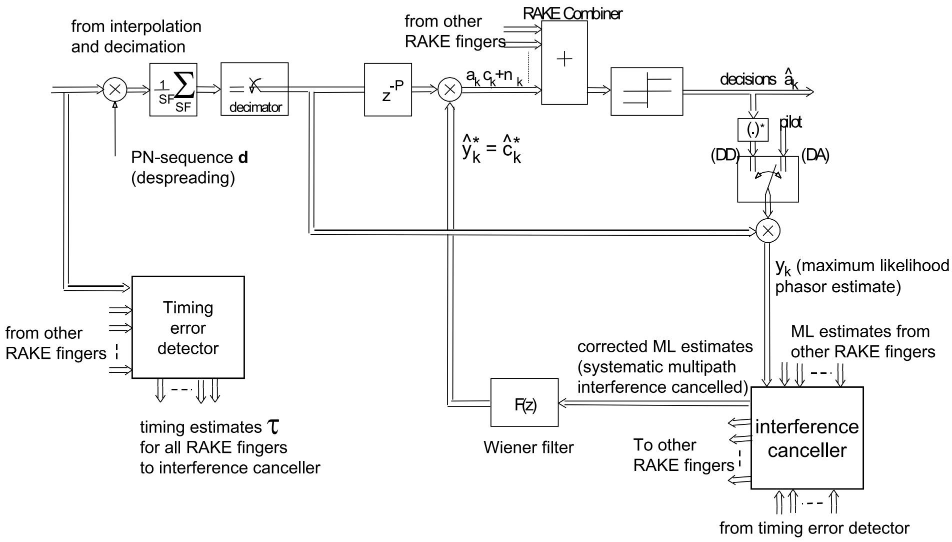

Figure 1. Phasor Estimation for a RAKE Receiver

where

$$\mathbf{C}_D = \rho \cdot \begin{pmatrix} \alpha(0) & \cdots & \alpha(N_w-1) \\ \vdots & \ddots & \vdots \\ \alpha(N_w-1) & \cdots & \alpha(0) \end{pmatrix} \quad (15)$$

$$\mathbf{c}_D = \rho \cdot (\alpha(N_w-1-P) \cdots \alpha(1-P))^T \quad (16)$$

are the channel tap autocorrelation matrix and the vector of channel tap autocorrelation samples. Note, that the factors $\alpha(m) = J_0(2\pi\lambda m)$ in equations 15 and 16, and the associated channel autocorrelation function, depend on the normalized Doppler shift λ. The variable P indicates, whether the estimator works as a predictor or a smoother. For $P > 0$, we have a Wiener smoother, which requires the knowledge of future ML phasor estimates. These future estimates can be obtained only, if a corresponding delay is introduced between the demodulator for the common pilot channel and the data stream [5]

4. Performance Analysis and Conclusion

The performance of the new channel estimator was tested by means of simulation. The simulation setup was as follows. We assumed a UMTS FDD system operating with a spreading factor of 4. The Wiener filter was employed as a one-step predictor with $N_w = 20$ taps. The velocity was chosen to be $v = 250 km/h$. Obviously, such a speed does not make any sense in an indoor scenario, but it helps to speed up simulation times, since the higher velocity makes the fading process much faster, and it still delivers representative results. The channel consisted of two equally strong paths, with two different channel spacings: A spacing of 0.5 chips, corresponding to an indoor environment, and a spacing of 1.5 chips, corresponding to an outdoor scenario.

The Figures 2 and 3 illustrate qualitatively the improvements which are achievable by the new channel estimation scheme. Figure 2 shows the power of the channel estimates vs time, when interference cancellation is turned off. This experiment assumed a multipath spacing of 0.5 chips and an SNR of $15dB$. Clearly, in such a scenario, the multipath interference is too strong, and the estimator is not able to follow the true channel. Figure 3 demonstrates that the new channel estimation scheme employing partial multipath interference cancellation is clearly outperforming the old scheme and tracks the true channel well. Finally, for the 1.5 spacing, the bit error ratio (BER) was simulated. The results are shown in Figure 4 and demonstrate that, in this scenario, the RAKE receiver employing the new channel estimation scheme offers considerable performance gains of approximately $3dB$.

It can be concluded that the new algorithm is a simple, yet powerful extension to Wiener channel estimation. Multipath interference is partially cancelled, which leads to substantial performance improvements. Since the calculation of the LS interference cancellation solution only depends on the very slowly varying timing estimates and the raised cosine function, it is sufficient to perform calculation of the estimator's coefficients on a very low rate. Therefore, the additional computational effort required is minimal, which makes the algorithm ideal for RAKE receivers in the mobile user equipment.

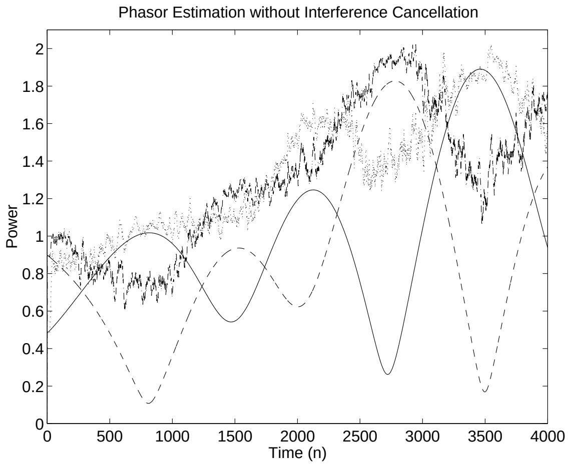

Figure 2. Phasor Estimation without Interference Cancellation

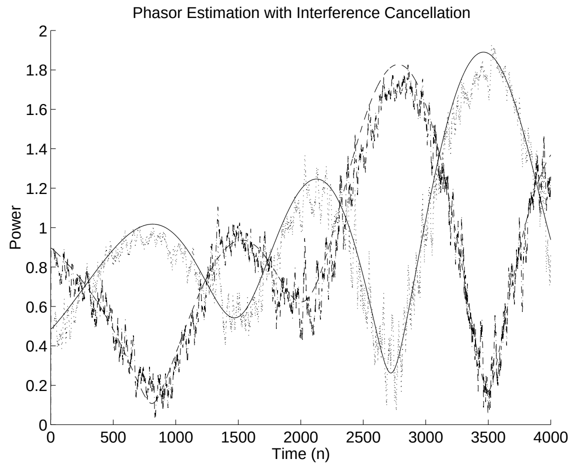

Figure 3. Phasor Estimation with Interference Cancellation

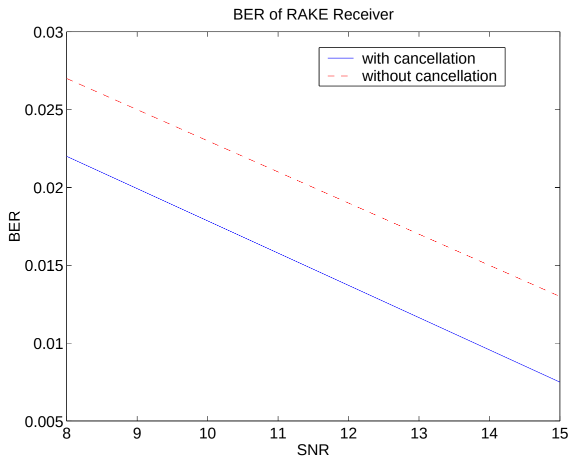

Figure 4. BER of RAKE Receiver

References

[1] Heinrich Meyr, Marc Moeneclaey and Stefan Fechtel, *Digital Communication Receivers: Synchronization, Channel Estimation and Signal Processing*, John Wiley and Sons, New York, 1998.

[2] John G. Proakis, *Digital Communications – Third Edition*, McGraw-Hill, New York, 1995.

[3] Steven M. Kay, *Fundamentals of Statistical Signal Processing – Estimation Theory*, Prentice Hall, Englewood Cliffs, New Jersey, 1993.

[4] 3GPP, *Technical Specification Group Radio Access Network; Physical layer - General description (3G TS 25.201 version 3.0.0*

[5] J. Baltersee, G. Fock, P. Schulz-Rittich, H. Meyr. "Performance Analysis of Phasor Estimation Algorithms for a FDD-UMTS RAKE Receiver", ISSSTA2000, Parsippany, New Jersey, September 2000.

[6] G. Fock, P. Schulz-Rittich, J. Baltersee, H. Meyr. "Multipath Resistant Coherent Timing-Error-Detector for DS-CDMA Applications", ISSSTA2000, Parsippany, New Jersey, September 2000.

[7] P. Schulz-Rittich, G. Fock, J. Baltersee, H. Meyr. "Low Complexity Adaptive Code Tracking with Improved Multipath Resolution for DS-CDMA Communications ovdf Fading Channels", ISSSTA2000, Parsippany, New Jersey, September 2000.