Channel Estimation Filter Using Sinc-Interpolation for UTRA FDD Downlink

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Abstract: This paper introduces a new Channel Estimation (CE) filter called Polyphase SInc-Interpolation (PSII)-filter for downlink coherent Rake-combining in a DS-CDMA mobile environment. In this case some UTRA FDD scenarios are taken for showing the feasibility of this approach. It uses known periodically time-multiplexed pilot symbols for interpolating the channel coefficients in between. Main advantage is the aptitude of the PSII-CE especially for high mobile environments. Unfortunately this advantage is paid by the use of at least six pilot sequences resulting in a three slot delay and poor noise reduction capabilities. Some simulation results of this PSII-filter compared with other well known CE-filters are presented.

Keywords: CDMA, UTRA FDD, Channel-Estimation, Sinc-Interpolation, CE, UMTS

1 Introduction

UMTS is the 3rd generation mobile cellular communication system of the (near) future. Upon others it defines a FDD-Wide-Band-CDMA scheme using a coherent Rake-receiver as shown in figure 1. The Signal y_{ID} after the I&D-Operation is denoted as[1]

$$y_{ID,l}(i) = \frac{1}{SF} \sum_{j=i\cdot SF}^{(i+1)\cdot SF-1} x_l(j)\cdot\frac{1}{\sqrt{2}}\cdot c^*_{scr}(j)\cdot c_{ch}(j) \quad (1)$$

with $c_{ch}(j)$ is the real valued channelization code also known as OVSF-code with spreading factor SF and c_{scr} denotes the complex scrambling code.

The received signal after maximum ratio combining for a L_R-finger Rake-receiver is denoted by

$$y_{MRC}(i) = \sum_{l=0}^{L_R} y_{ID,l}(i)\cdot\hat{h}^*_l(i) \quad (2)$$

with $\hat{h}_l(i)$ is the estimated channel coefficient for the l-th finger.

In UMTS-FDD downlink a Dedicated Physical Channel (DPCH) consists of user data also called Dedicated Physical Data Channel (DPDCH), time-multiplexed with control information within the so called Dedicated Physical Control Channel (DPCCH) as depicted in figure 2 or in [1]. The DPCCH itself is divided in Transmit Power Control (TPC) Symbols, Transport Format Combination Indicator (TFCI), and a pilot sequence. Fifteen slots with $M_{slot} = 2560$ chips using this structure are combined in one Frame of 10 ms length. The pilot sequence changing for every slot within a frame is used for channel estimation by correlating the incoming signal with the pilot sequence.

[1](*) denotes conjugate complex

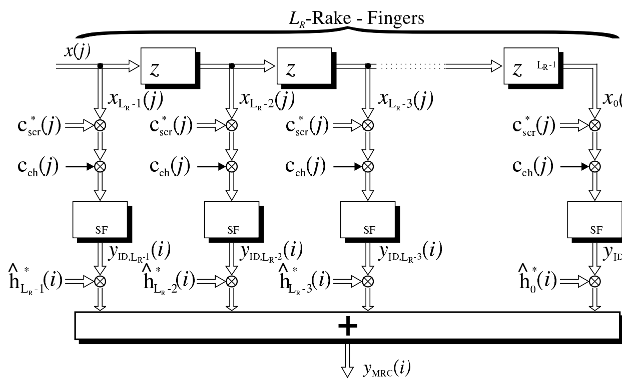

Fig. 1: Rake-receiver

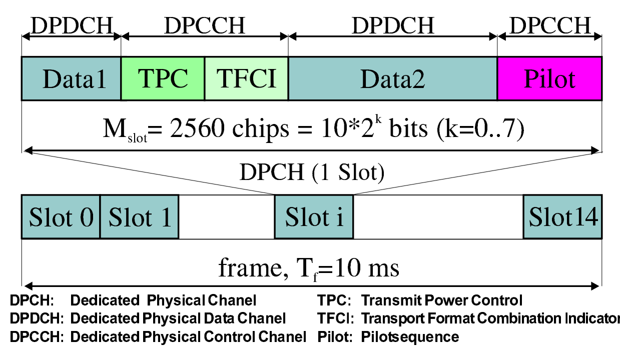

Fig. 2: Slot and frame structure

Denote

$$N_D = N_{slot} - N_{pilot} \quad \text{with} \quad N_{slot} = M_{slot}/SF \quad (3)$$

and N_{pilot} denotes the number of pilot symbol used in

DPCCH we can estimate a channel coefficient in centre of the pilot sequence in the ν-th slot:

$$\hat{h}_{l,\nu} = \frac{1}{N_{pilot}} \sum_{k=N_D}^{N_{slot}-1} y_{ID,l}(k) \cdot \frac{1}{\sqrt{2}} \cdot p_\nu^*(k - N_D). \quad (4)$$

With the pilot sequence $p_\nu(k)$ defined as:

$$p_\nu(k) \begin{cases} \neq 0 & : \quad 0 \leq k < N_{pilot} \\ = 0 & : \quad \text{otherwise} \end{cases} \quad (5)$$

Then, these estimates $\hat{h}_{l,\nu}$ are processed further using CE-filters. In general all further signal processing has to cope with noise and the time variant change of channel coefficients. Unfortunately both effects are contradictory to each other. To antagonise noise averaging is mandatory but this will deteriorate the performance in a high mobile environment. Therefore a compromise has to be found. There are two standard linear CE-filters discussed in [2] which will be used as reference in this paper.

The averaging filter also shown in figure 3a) tries to reduce the noise and is not following the channel coefficient within a slot at all. The linear interpolation-filter is just drawing a line between two neighbouring estimates to follow the change of the time variant channel coefficient of one Rake-finger as shown in figure 3b). Therefore noise reduction of the linear interpolation CE depends on the position within the slot. From no reduction at the position of the pilot up to three dB in the middle between two pilot sequences.

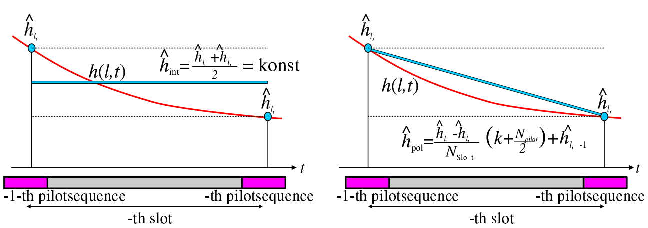

Fig. 3: Classical CE, (a): Averaging, (b): Linear interpolation

To examine the performance of these two filters a bit further the Mean Squared Error (MSE) between the "real" channel coefficient and its estimation can be expressed with:

$$MSE_\nu(i) := \mathrm{E}\left\{ \left| \hat{h}_{xxx,\nu}(i) - h_\nu(i) \right|^2 \right\} \quad (6)$$

In this case expectation is substituted by time average due to ergodicity.

$$MSE(i) = \lim_{N \to \infty} \frac{1}{N} \sum_{\nu=0}^{N-1} \left| \hat{h}_{xxx,\nu}(i) - h_\nu(i) \right|^2 \quad (7)$$

with $i \in [0, 2560/SF - 1]$ and $\hat{h}_{xxx}$ is the estimate of the considered CE and with $h_\nu(i) = \sum_{k=i}^{i+SF-1} h_\nu(k)$ and $h_\nu(k)$ is the real channel coefficient calculated for every chip .

Using a one tap Rayleigh channel at a velocity v with an E_B/N_0 of 8 dB and the UMTS-transmission slot format #8 (SF=128, 2 pilot symbols per sequence) the MSE depends on the position within slot. This is depicted for the two slot averaging CE in figure 4.

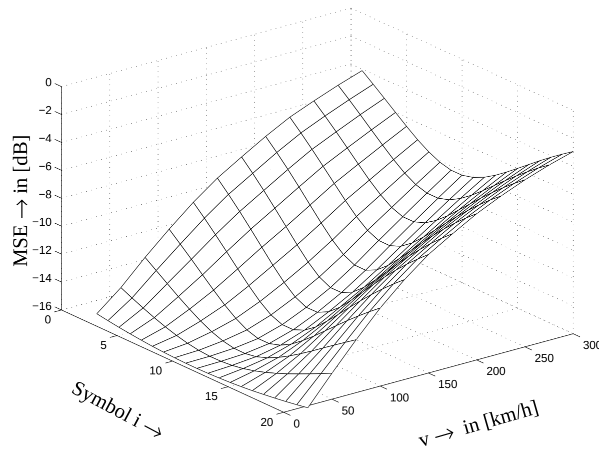

Fig. 4: MSE of averaging CE

A global minimum in the middle between two pilot sequences can be seen in figure 4 according to the intersection of the estimated and the "real" channel coefficient in figure 3a. Especially for higher velocities this method degrades fast, but for lower velocities it does effectuate a 3 dB gain against a single slot CE. In figure 5 the MSE for the linear interpolation is displayed.

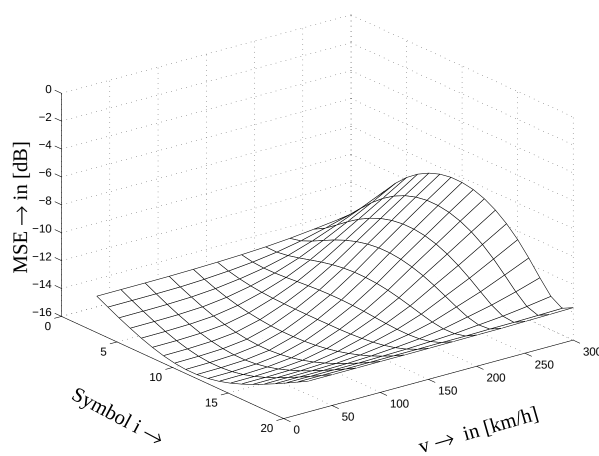

Fig. 5: MSE of linear interpolation CE

While at the edges (near the pilot sequences) and for low velocities this method provides a three dB loss against the averaging CE, it does reach it in the middle between two pilot sequences as expected. On the other hand it does provide a much better behaviour in a highly mobile environment. For higher velocities the advantage of 3 dB noise reduction in the middle will be countered by ascending nonlinearity of the real channel coefficient resulting in a global maximum at the centre, that can also be deduced from figure 3b).

2 Sinc-Interpolation

Looking on the channel estimation done so far from another point of view, we are trying to reconstruct the time variant channel coefficient from a set of samples. Using the 3GPP UTRA FDD DL specification [1], the sampling theorem is fulfilled till doppler-shift of 750 Hz. Therefore the theoretical upper bound for the maximum velocity is 405 km/h.

$$f_{d,max} = \frac{1}{2} \cdot f_s = \frac{1}{2} \frac{15}{10\text{ ms}} = 750\text{ Hz} \tag{8}$$

To reconstruct an original signal out of a sampled version of it, a Sinc-interpolation (ideal lowpass) is needed. Like the WMSA channel estimation filter introduced by [3] pilot sequences of up to six slots are used to obtain the channel coefficients. The reconstruction (mother) filter is therefore chosen as:

$$g(i) = \frac{\sin(\pi \cdot (\frac{SF}{2560} \cdot (i-0.5) - 3)}{\pi \cdot (\frac{SF}{2560} \cdot (i-0.5) - 3)} \tag{9}$$

with

$$0 \leq i < 6 \cdot \frac{2560}{SF} \qquad \text{and} \qquad g(i) = 0 \qquad \text{otherwise.}$$

Note that the delay of half a chip is only needed if the number of pilot symbols is even. To overcome problems associated with the leakage effect, $g(i)$ is additionally weighted using a Hamming window.

$$f^{hm}(i) = 0.54 + 0.46 \cdot \cos\left(\frac{2 \cdot \pi \cdot i}{6 \cdot 2560/SF - 1}\right) \tag{10}$$

$$g_{hm}(i) = g(i) \cdot f^{hm}(i) \tag{11}$$

with

$$0 \leq i < 6 \cdot \frac{2560}{SF} \text{ and } g_{hm}(i) = 0\,,\; f^{hm}(i) = 0 \text{ otherwise.}$$

The filter $g_{hm}(i)$ in time domain is depicted in figure 6 and in frequency domain in figure 7.

Polyphases of g_{hm} are know taken as follows.

$$g_{hm,\mu}(i) = g_{hm}(i + \mu \cdot 2560/SF) \quad \mu\ in\ [0,5] \tag{12}$$

With this an estimate $h_{psii(\nu,l)}(i)$ for the ν-th slot can be calculated.

$$h_{psii(\nu,l)}(i) = \sum_{\mu=0}^{5} h_{\nu-\mu+2,l} \cdot g_{hm,\mu}(i) \tag{13}$$

The MSE of the PSII-CE using UTRA FDD slot structure #9 is shown in 8.

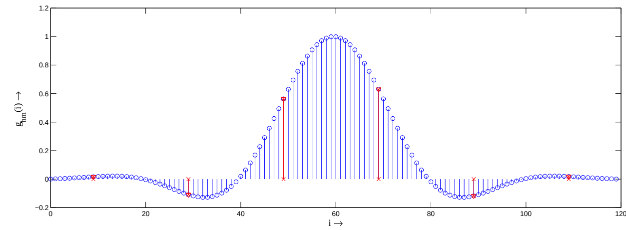

Fig. 6: Polyphase mother filter $g_{hm,\nu}(i)$ with $SF = 128$ and one polyphase $g_{hm,\mu}(9)$

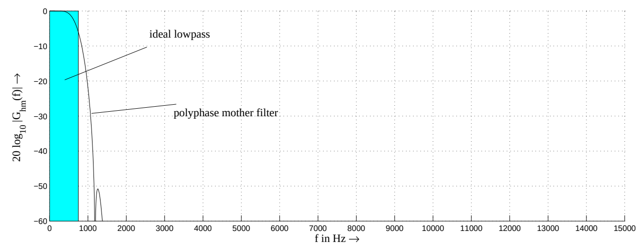

Fig. 7: Polyphase mother filter ($SF = 128$) in frequency domain $[0, f_n]$ and ideal lowpass with $f_g = 750$ Hz

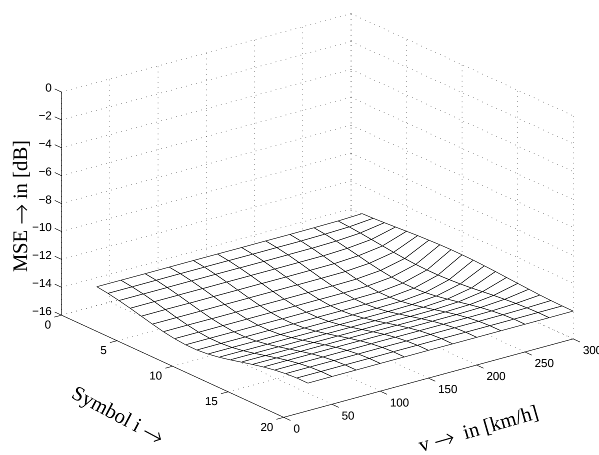

Fig. 8: MSE of PSII-CE

Because of the non perfect lowpass characteristic, and the non ideal sampling of the channel coefficients, performance of PSII-CE degrades with velocities higher than 300 km/h on the one hand and with very high SF on the other. For lower velocities the MSE is relatively flat indicating good behaviour, nevertheless for velocities lower than 130 km/h the MSE of the linear interpolation filter for an $E_B/N_0 = 8$ dB and a spreading factor of 128 is slightly better due to the better noise reduction in the centre of a slot.

3 Simulation Results

In this section simulation results are presented. The slot formats taken for simulation are shown in table 1

Slot Format	SF	DPCH	DPDCH	Pilot
		in Symbols per Slot		
#8	128	20	17	2
#13	32	80	70	8
#16	4	640	624	16

Table 1: Used slot formats within the simulations

or the complete list can be found in [1]. A Vehicular-A channel is taken, described in [4] with a defined number of delays for the channel-taps. This knowledge is used in the CE instead of searching for the best suitable taps by correlation for some time, due to this only a correlation is done for the six defined taps. This is not a significant drawback, because the correlation over the whole time-axis will provide all possible taps but will increase simulation time without giving new insights of the feasibility of the CEs.

The Rake-receiver got six fingers but is only using taps for MRC that have at least 10% of the power of the most powerful tap, because very small taps are very hard to estimate due to crosstalk and wrong selection of delay[2] is also possible when using a long time correlation to find it but this is common to all CEs. Power control and channel coding are not in use. Beside the following simulation always the classical AWGN-curve is displayed. In figure 9 a Vehicular-A channel for a velocity of 120 km/h and slot format # 8 is taken. The velocity of 120 km/h is still much to high for the averaging to provide good results. The linear interpolation on the other hand does provide the best results because the interpolation itself is good enough to cope with the time variance of the channel. PSII-CE is nearly as good as the linear interpolation which benefits from the light noise reduction in the centre of each slot resulting in a better overall performance of the system with linear interpolation CE.

Using a higher velocity, in this case 300 km/h will change the performance especially for higher E_B/N_0 as can be seen in figure 10. While averaging is further degrading so is linear interpolation because this CE cannot compensate time variant effects any longer resulting in an error floor at about $2 \cdot 10^{-4}$. While for low E_B/N_0, noise is the dominating effect the PSII-CE does not distinguished itself from other CEs. As higher E_B/N_0 gets, as better becomes the performance of PSII-CE.

In figure 11 and 13 simulations for slot format # 13 and # 16 are taken for a velocity of 120 km/h. While in figure 12 and 14 the same slot formats are taken for a velocity of 300 km/h. In general the measures do not change. Averaging is out of touch for this kind of velocity, while linear interpolation is best for 120 km/h and PSII-CE does perform superior for 300 km/h. As higher the data rate gets (or as lower the spreading factor is set) the overall performance does naturally degrade.

In the case of slot format # 16 with its spreading factor $SF = 4$ interference becomes a problem because of the

[2]This problem will not occur here but in field testing

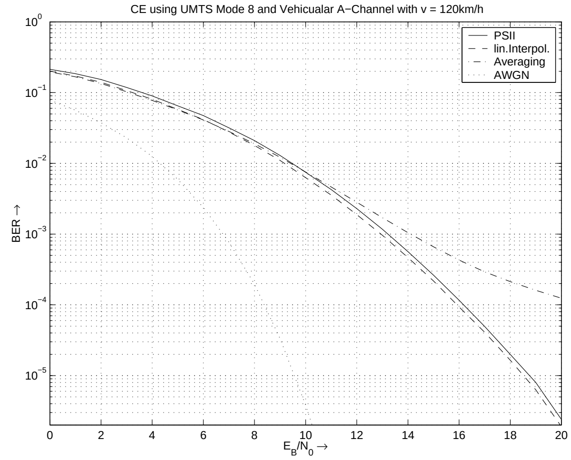

Fig. 9: Vehicular-A slot format # 8, v=120km/h

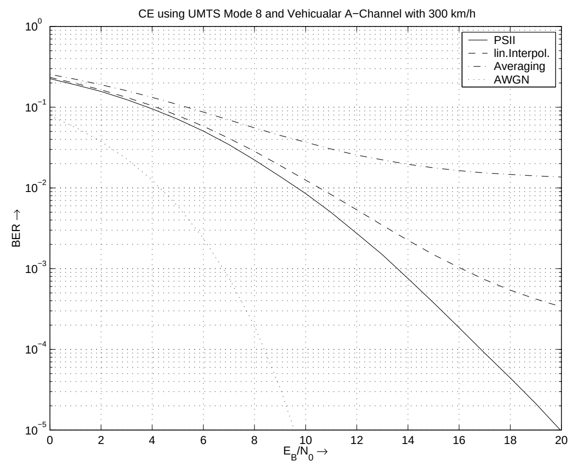

Fig. 10: Vehicular-A slot-format # 8, v=300km/h

lower separation capability of the Gold-Code. Meaning that noise is the dominating issue also for a velocity of 120 km/h therefore in a high noise environment averaging does have an advantage over PSII-CE. Nevertheless the performance is bad and all CE-schemes are to close together in these areas that by using a channel coding results may be different.

4 Conclusion

In general CE has to cope with noise and time variance of the channel. Unfortunately both problems are contradictory. Beginning with WMSA-CE by [3] on the time invariant and noisy side and ending up with the here proposed PSII-CE on the fast but low noise site. This paper shows the basic idea of the PSII-CE and provides some clues of the utilizability of such a scheme. This scheme may be improved by some points in the future. For example instead of looking for Sinc-Interpolation we may try some sort of remez-synthesis for a filter with a lower bandwidth meaning lower velocities but a better noise reduction. Another point is to use more

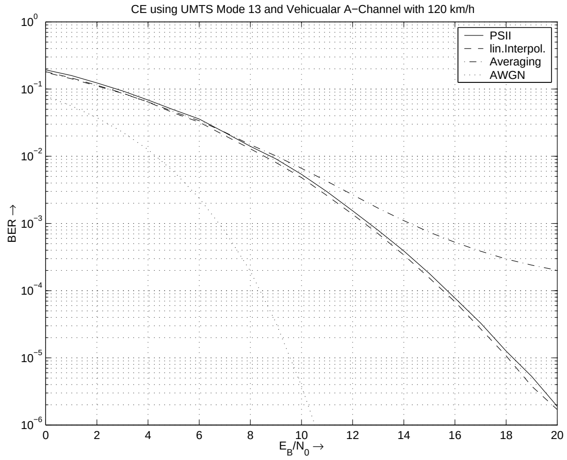

Fig. 11: Vehicular-A slot format # 13, v=120km/h

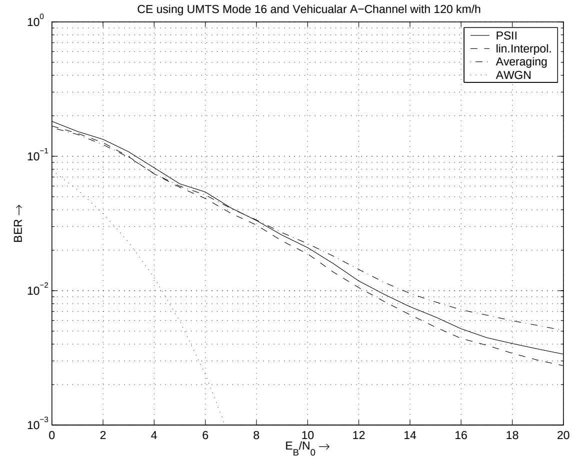

Fig. 13: Vehicular-A slot format # 16, v=120km/h

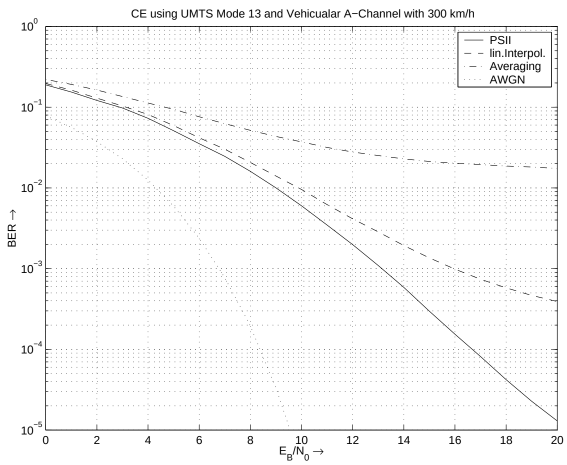

Fig. 12: Vehicular-A slot format # 13, v=300km/h

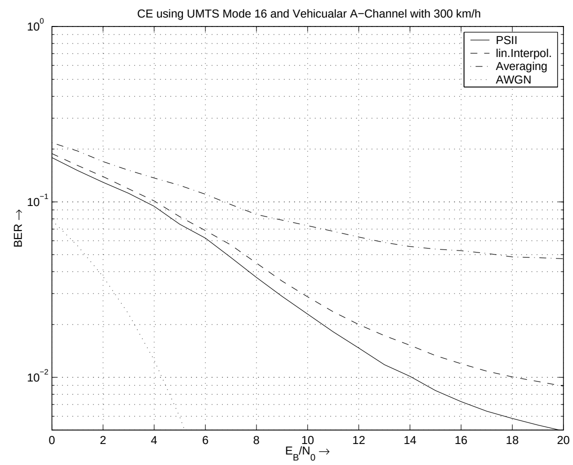

Fig. 14: Vehicular-A slot format # 16, v=300km/h

estimations instead of one $\hat{h}_l^*(i)$ for the whole pilot sequence. For example for every pilot symbol there may be one estimation, using this here proposed algorithm as described and averaging all N_{pilot} results at the end.

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