Channel Estimation with Lowpass Filter for UTRA FDD Downlink

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Abstract- This paper introduces a new Channel Estimation (CE) filter called Polyphase Lowpass-Interpolation (PLI)-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. The idea of this approach is to find a compromise between noise reduction and adaptation to the time variant channel coefficients using a Remez lowpass design. This is paid by the use of at least six pilot sequences resulting in a three slot delay. Some simulation results of this PLI-filter compared with other well known CE-filters are presented.

I. 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} which can be detected after the I&D-Operation is represented as[1]

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

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

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

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

where $\hat{h}_l(i)$ 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 – the so called Dedicated Physical Control Channel (DPCCH) as depicted in fig. 2 or in [1].

The DPCCH itself is divided in Transmit Power Control (TPC) symbols, a Transport Format Combination Indicator (TFCI), and a pilot sequence. Fifteen slots with

[1] (*) denotes conjugate complex

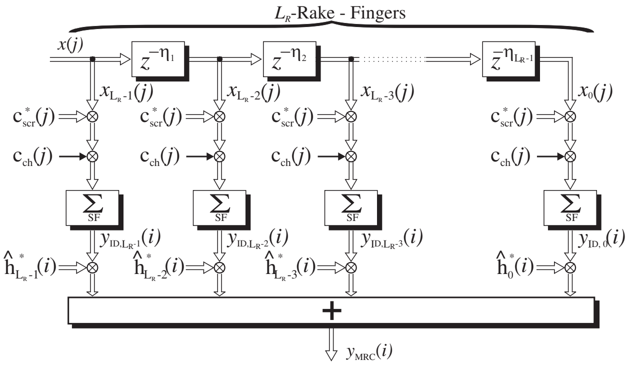

Figure 1: Rake-receiver

$M_{slot} = 2560$ 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. Besides of this pilot symbol aided scheme, 3GPP also provides pilot channels for CE but this is not viewed any further within this paper.

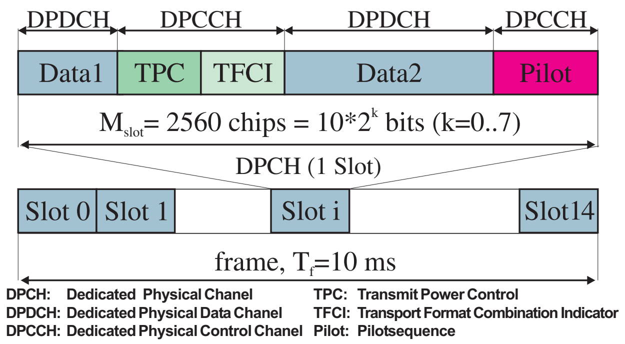

Figure 2: Slot and frame structure

Let

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

where 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 by correlation.

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

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

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

The estimates $\hat{h}_{l,\nu}$ are processed further on, using CE-filters. In general signal processing has to cope with noise and the time variant change of channel coefficients. Unfortunately both effects are contradictory to each other. In order 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. In linear interpolation, there is no noise reduction at the position of the pilot and up to three dB in the middle between two pilot sequences.

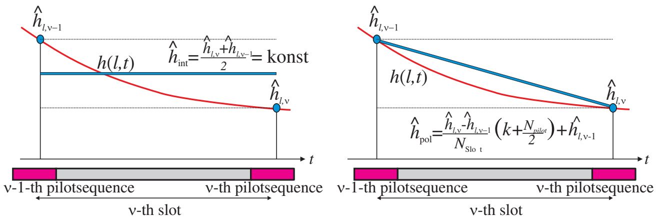

Figure 3: Classical CE, (a): Averaging, (b): Linear interpolation

In order to examine the performance of these two filters a bit further the Mean Squared Error (MSE) between the "real" channel coefficient $h_\nu(i)$ and its estimation $\hat{h}_{xxx,\nu}(i)$ with xxx denotes the estimation scheme, can be expressed as:

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

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

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

with $i \in \{0, \ldots, 2560/SF - 1\}$. The estimate $\hat{h}_{xxx}$ is the estimate of the considered CE xxx and $h_\nu(i) = \sum_{k=i}^{i+SF-1} h_\nu(k)$ where $h_\nu(k)$ 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.

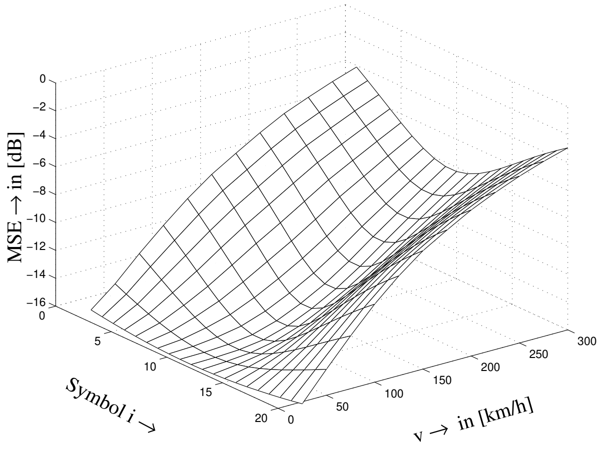

Figure 4: MSE of averaging CE

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 yields a 3 dB gain compared to a single slot CE. In figure 5 the MSE for the linear interpolation is depicted.

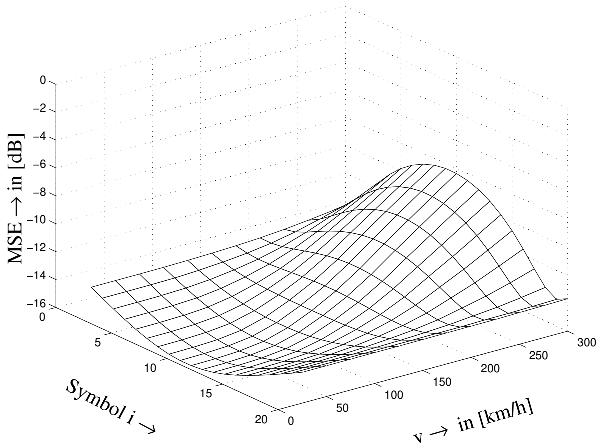

Figure 5: MSE of linear interpolation CE

At the edges (near the pilot sequences) and for low velocities this method provides a three dB loss against the averaging CE, as expected, it reaches the same MSE as the averaging CE in the middle between two pilot sequences. On the other hand it provides 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).

II. LOWPASS-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[2].

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

In order to reconstruct an original signal out of a sampled version of it, a Sinc-interpolation (ideal lowpass) is needed as done in [3]. In general a Sinc-filter designed for very high velocities up to 405 km/h is not really necessary and has poor noise reduction features. It is more advantageous to trade the aptitude for very high mobile environment in a better noise reduction capability. In order to achieve this goal a Remez design with a lower cut-off-frequency is taken. This cut-off-frequency can be chosen freely in certain boundaries. Like the WMSA channel estimation filter introduced by [4], which is quite often uses within todays planned UMTS mobile phones, pilot sequences (l_p) of up to six slots are used to obtain the channel coefficients. More pilot sequences would be more preferable but this would also introduce additional delay, and more filter coefficients.

The maximum number of filter coefficients are needed for a spreading factor of four.

$$N = \frac{M_{slot}}{SF} \cdot l_p = 2560/4 * 6 = 3840 \tag{9}$$

This amount may be too high for a Remez design. This obstacle can be overcome by a Remez-design for a larger spreading factor and a Sinc-Interpolation to get the missing coefficients[3]. Because we will only use polyphases of this now called mother-filter, it is important to take a Remez design without any saltus. Unfortunately the filter-design itself is limited due to the restriction to of six slots, leaving not much space for a proper design.

In this case four exemplary designs are taken with following parameters displayed in table I . All designs have a transition bandwidth B_T of about 900 Hz. This parameter may be changed but there is not much room to do so. As a rough estimate for this value, the estimation of the transition bandwidth of a FIR-filter using Hamming windowing can be taken. In [5] this value is estimated as

$$\Omega_T = 2 \cdot pi \cdot B_T = 8 \cdot \pi / N. \tag{10}$$

Resulting in a transition bandwidth B_T of 1 kHz.

The number of coefficients is always 6 times the number of symbols (2560/SF) within a slot. The frequencies for passband or stopband are freely elegible according to the used mobility-scenario.

The resulting vector g_{rm} of the Remez filter design has to be normalised, to assure a gain of one at 0 Hz.

$$g(i) = \frac{g_{rm}(i)}{\sum_{i=0}^{N-1} g_{rm}(i)} \cdot \frac{2560}{SF} \tag{11}$$

[2] 2 GHz carrier frequency

[3] This can be done by DFT, zero padding and IDFT

TABLE I

Design parameters

Example	f_0	f_p	f_s	B_T
I	450 Hz	1 Hz	900 Hz	899 Hz
II	500 Hz	50 Hz	950 Hz	900 Hz
III	550 Hz	100 Hz	1000 Hz	900 Hz
IV	750 Hz	300 Hz	1200 Hz	900 Hz

One possible resulting filter (Example I) $g(i)$ in time domain is depicted in figure 6 and in frequency domain in figure 7.

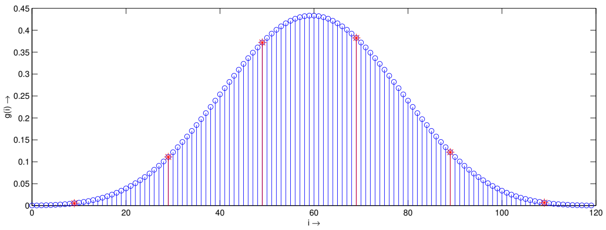

Figure 6: Polyphase mother filter $g(i)$ for $f_0 = 450$ Hz with $SF = 128$ and one polyphase $g_\mu(9)$

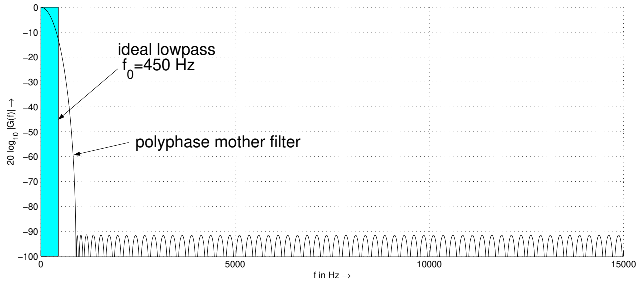

Figure 7: Polyphase mother filter, Example I ($SF = 128$) in frequency domain $[0, f_n]$

Polyphases i of $g_\nu(i)$ are know taken as follows.

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

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

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

In order to analyse this PLI-CE the MSE of an example design using UTRA FDD slot structure #8 is shown in figure 8.

For lower velocities the MSE is relatively flat indicating good behaviour. The MSE for PLI-CE in this case and all other cases made for low velocities are not dependent on the position within the slot. A design for high velocities

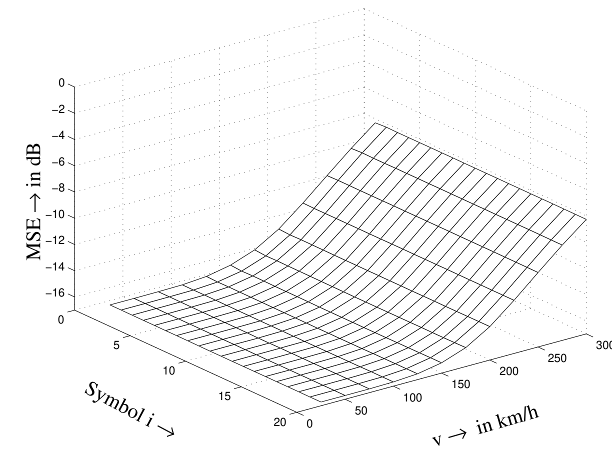

Figure 8: MSE of an example design of PLI-CE (Example II), $f_0 = 500$ Hz, E_B/N_0=8 dB

changes this behaviour, but the effects are low compared with averaging or linear interpolation CE. For higher velocities in respect of the design of the PLI-CE Filter this CE degrades fast. In figure 9 the MSE is taken as a function of velocity and cutoff frequency f_0 and was averaged all symbols of one slot.

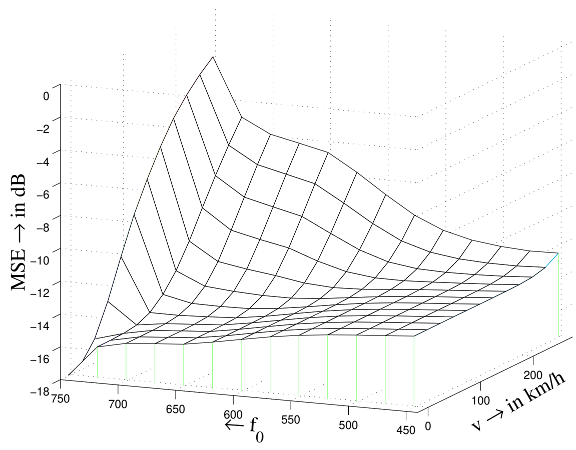

Figure 9: Over one slot averaged MSE of PLI-CE, E_B/N_0=8dB

As long as the design can cope the time variance of the channel, MSE will rise slightly with rising parameter f_0 because of lesser noise reduction. The difference in the MSE between the best fit design and the design for maximum velocity may extend two dB for velocities lesser 200 km/h. For velocities not higher than 50 km/h the gain is up to 4 dB. If the design can not compensate the time variance it will become absolutely useless. In order to verify this thought some simulation will back them up.

III. Simulation Results

In this section simulation results are presented. The slot formats taken for simulation are shown in table II. The complete list of slot formats can be found in [1]. A Vehicular-A channel with fixed delays for the channel-taps is taken, as described in [6].

TABLE II

Used slot formats within the simulations

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

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. Power control and channel coding are not in use. Next to the following simulation results the classical AWGN-curve is displayed. In figure 10 a Vehicular-A channel for a velocity of 60 km/h and slot format # 8 is taken. For this velocity PLI-CE with $f_g = 450$ Hz provides the best results and is about $\frac{3}{4}$ dB better than the follow up. Due to minor time variant changes of the channel, noise reduction is the main problem. Therefore averaging over two slots does have good results together with PLI-CE with $f_g = 500$ Hz and $f_g = 550$ Hz. It can be seen that only for very high E_b/N_0 time variance has an effect on the performance. CEs capable for high mobile environment are not suited for this scenario due to their bad noise reduction abilities.

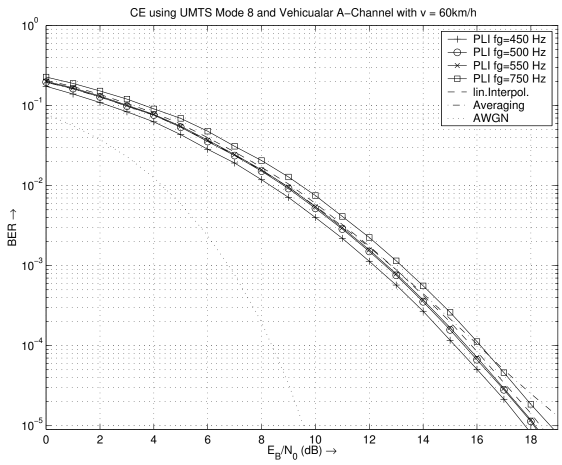

Figure 10: Vehicular-A slot format # 8, v=60km/h

In figure 11 velocity is doubled. With a velocity of 120 km/h time variance is much stronger and averaging and linear interpolation are getting nearly the same results for small E_b/N_0 while PLI-CE with $f_g = 500$ Hz and $f_g = 550$ Hz are a little bit better than both classical CEs. Due to good noise reduction capabilities PLI-CE with $f_g = 450$ Hz does reach the best BER in a noisy environment.
For better E_b/N_0 the time variance becomes the dominating issue. Therefore PLI-CE with $f_g = 500$ Hz, 550 Hz and a little worse linear interpolation can cope time variance very

well. PLI-CE with $f_g = 450$ Hz cannot adept to time variant effects as good as PLI-CE with higher f_g. Time variance is also too high for averaging to provide good results.

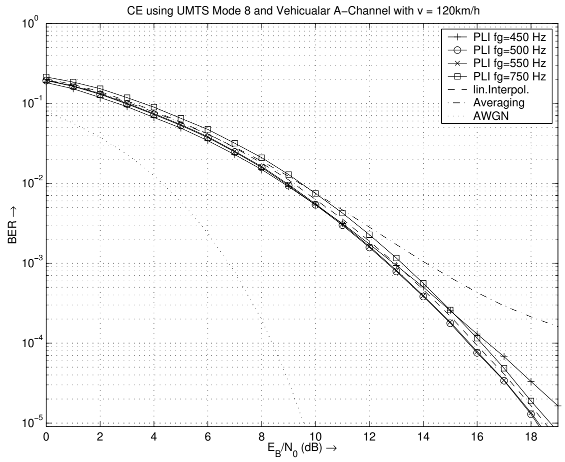

Figure 11: Vehicular-A slot format # 8, v=120km/h

In figure 12 a velocity of 180 Km/h is taken. In this case PLI-CE with $f_g = 500$ Hz and 550 Hz provide the best results because they do both, noise reduction on the one hand and adapting to the time variant channel on the other hand. This results in a small gain of about 0.5 dB. For small E_b/N_0 linear interpolation is worse because of the bad noise reduction and averaging is worse because of the poor adjustment to time variant channels. For high E_b/N_0 schemes made for a high mobile environment do provide good performance, while averaging results in an error floor. In this case PLI-CE with $f_g = 750$ Hz does perform very well for very low noise.

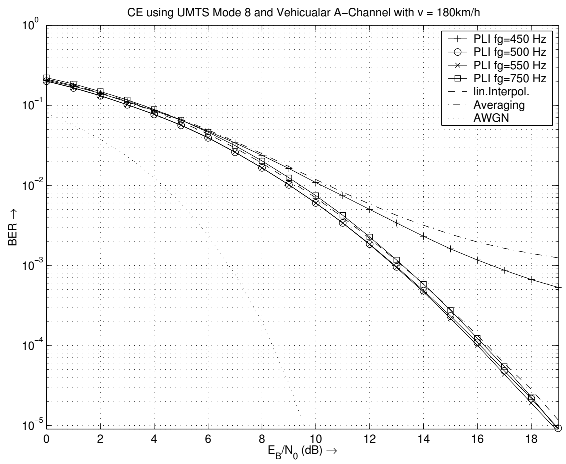

Figure 12: Vehicular-A slot format # 8, v=180km/h

In figure 13 the slot format has changed. In this case type # 13 with SF of 32 and a velocity of 120 km/h is taken. Besides a slightly changed BER the main circumstances for the single user case are the same as for the #8 case which is depicted in figure 11. All other simulation results for format # 23 do also correspond with the shown simulation results for format # 8.

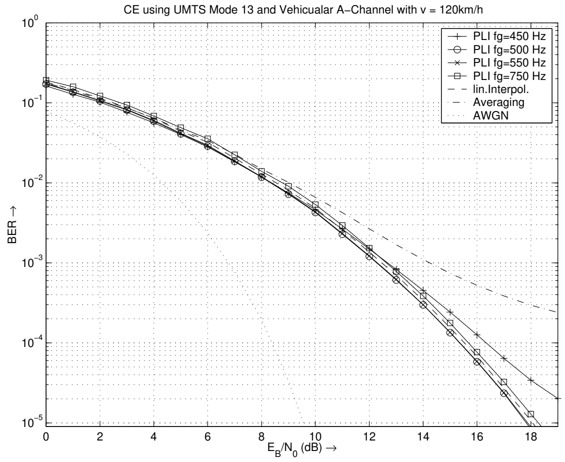

Figure 13: Vehicular-A slot format # 13, v=120km/h

IV. Conclusions

In general CE has to cope with noise and time variance of the channel. Unfortunately both problems are contradictory. The here proposed PLI-CE does provide a possibility to adapt to the problem. This paper shows the basic idea of the PLI-CE and provides some clues of the utilizability of such a scheme. This scheme may be improved by some points in the future. Especially the assumed restraints do have a hard impact to the performance and should be changed. The effects of power control on this scheme have to be examined.

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