SPEED ESTIMATION FOR WCDMA BASED ON THE CHANNEL ENVELOPE DERIVATIVE

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ABSTRACT

Reliable estimates of the speed or, equivalently, the perceived channel Doppler spread, are of great importance in mobile cellular systems for optimization of wireless receiver algorithms such as channel estimation, handoff algorithms etc. In this paper we propose a novel speed estimation algorithm based on the derivative of the received channel envelope. We characterize the effects of additive white noise analytically and verify against simulation results. Simulations illustrate high reliability, low complexity, and good robustness of the approach for various propagation conditions. The performance of the algorithm is demonstrated for WCDMA in the speed range from 0 to 250 km/h.

1. INTRODUCTION

In the literature methods for speed estimation have been based on many approaches, e.g. the channel auto-correlation [5, 9], wavelet expansion [6], simplified high/low speed estimation [7], eigenvalue decomposition [10], two antenna Doppler modelling [8], maxima of the received signal envelope [12], antenna diversity [13], and higher order statistics [10, 11]. Some of the methods are very complex, while others only perform well under special propagation conditions. In this paper we propose a novel approach which utilizes the probability density function (pdf.) of the channel envelope derivative.

2. ESTIMATION OF THE SPEED

Estimation of the UE (User Equipment) speed can be done by estimating the maximum Doppler frequency of a mobile channel. The maximum Doppler frequency, f_m, is related to the speed of the UE, *v*, as

$$f_m = \frac{v \cdot f_c}{c} \tag{1}$$

where c is the speed of light and f_c is the carrier frequency. For WCDMA for UMTS (region 2) the carrier frequency band is around 2.14 GHz [4]. Clearly, in the case of a fading channel where the speed of the UE (or the speed of objects in the channel) is low, we have a small f_m.

The proposed speed estimation method is based on the derivative of the received channel envelope, as estimated from the demodulated CPICH symbols available in WCDMA for UMTS. In order to describe the statistical time varying nature of the received envelope, $r(t)$, we assume the real, i(t), and imaginary, q(t), parts of the received symbols to be stationary IID Gaussian processes, i(t), q(t) $\in$ N(0,b_0) with variance b_0. This assumption is commonly used for a flat fading signal or for individual multipath components [1], and leads to $r(t)$ being Rayleigh distributed. That is, if we let $h(t)$ denote the complex channel estimates as a function of time t, the received envelope $r(t)$ is denoted by

$$\begin{aligned} h(t) &= i(t) + j \cdot q(t) \\ r(t) &= |h(t)| = \sqrt{i^2(t) + q^2(t)} \end{aligned} \tag{2}$$

3. THE PROPOSED ESTIMATOR

In [2] it is shown that the marginal pdf. of the derivative, $\dot{r}$, of the envelope, r, of a fading channel tap with respect to time is given by ([2], Equation (1.3-34))

$$p(\dot{r}) = \frac{1}{\sqrt{2\pi b_2}} e^{-\dot{r}^2/2b_2} \tag{3}$$

Note that the time-dependencies of $\dot{r}$ are dropped for notational convenience. b_2 is the moment as specified in [2]. Equation (3) is recognized as a Gaussian distribution with variance, $\sigma^2\{\dot{r}\} = b_2$, which can be linked to the maximum Doppler frequency, f_m by:

$$\begin{aligned} \sigma^2\{\dot{r}\} &= b_2 = (2\pi f_m)^2 b_0 / 2 \\ b_0 &= \frac{1}{2} E\{r^2\} \end{aligned} \tag{4}$$

which is taken from [2]. By inserting (1) into (4), and the obtained expression into (3) we obtain:

$$\hat{v} = \frac{c}{2\pi f_c} \sqrt{\frac{2\sigma^2\{\dot{r}\}}{b_0}} \quad (5)$$

which is the proposed estimator. Note that the estimator is independent of the shape of the power spectrum, in contrast to most of the existing estimation approaches, since it is based on (3) which is the pdf. of the derivative of the channel envelope.

Since the derivative of the envelope is equivalent to a digital filter of infinite length [14], we have to do an approximation of the envelope derivative $\dot{r}$ in order to use the proposed estimator in practice. We here choose a simple nearest neighbor approximation, i.e. a 2 tap FIR filter

$$\hat{\dot{r}}(t_n) = \frac{r(t_n) - r(t_{n-1})}{\Delta T} \quad (6)$$

The envelope of the channel estimates are only available as digital samples $r(t_n) = n\Delta T$, where ΔT is the time-spacing between the samples of the channel estimates – for CPICH in WCDMA there are 10 pilot symbols per slot, yielding a time-spacing between each pilot symbol of $\Delta T = 66{,}7\mu\sec$ [4].

To apply (5) only three simple estimates are required,

$$E\{\hat{\dot{r}}^2\} \cong \frac{1}{N}\sum_{n=1}^{N} \hat{\dot{r}}(t_n)^2, \quad E\{\hat{\dot{r}}\} \cong \frac{1}{N}\sum_{n=1}^{N} \hat{\dot{r}}(t_n), \text{ and}$$
$$E\{r^2\} \cong \frac{1}{N}\sum_{n=1}^{N} r(t_n)^2 \quad (7)$$

which we obtain using maximum likelihood estimators over a finite length observation window, assuming the noise to be AWGN.

We note that the estimators in (7) are simple flat FIR filters of the squared envelope derivative, the envelope derivative, and the squared envelope, respectively. Hence we can estimate the variance, $\sigma^2\{\dot{r}\}$, and the moment, b_0, as:

$$\begin{aligned} \sigma^2\{\dot{r}\} &= E\{\hat{\dot{r}}^2\} - \left(E\{\hat{\dot{r}}\}\right)^2 \\ \hat{b}_0 &= E\{r^2\}/2 \end{aligned} \quad (8)$$

and use these directly in (5).

In the simulations presented later we have accumulated the sums for a certain observation window and computed a single speed estimate at the end of the observation window.

Next we analytically characterize the effects of AWGN on the estimator.

4. THE PROPOSED ESTIMATOR ANALYSED WITH ADDITIVE WHITE NOISE

In order to obtain analytical results of the expected performance of the estimator we assume the envelope, r, to be contaminated with an AWGN term ε, which is uncorrelated with r.

$$r_{\text{noise}}(t_n) = r(t_n) + \varepsilon(t_n), \quad E\{r \cdot \varepsilon\} = 0 \quad (9)$$

Since the mean envelope is constant we can neglect $E\{\dot{r}_{noise}\}$, and get

$$\hat{v} = \frac{c}{2\pi f_c}\sqrt{\frac{2\sigma^2\{\dot{r}_{\text{noise}}\}}{b_0}} \approx \frac{c}{\pi f_c}\sqrt{\frac{E\{\dot{r}_{\text{noise}}^2\}}{E\{r_{\text{noise}}^2\}}} \quad (10)$$

where $\sigma^2\{\dot{r}_{noise}\} = E\{\dot{r}^2_{noise}\} - E\{\dot{r}_{noise}\}^2 \approx E\{\dot{r}^2_{noise}\}$. Now, the exercise is to find the two average values in (10). First the denominator

$$E\{r_{\text{noise}}^2\} = E\{(r+\varepsilon)^2\} = R_r(0) + R_\varepsilon(0) \quad (11)$$

where the auto-correlation, $R_x(\tau) = E\{x(t)x(t+\tau)\}$, of the envelope and noise has been introduced. Next the numerator, where the derivative is substituted by the first order approximation shown in (6). Introducing the short-hand notation $r(t_n) = r(n)$ and $\varepsilon(t_n) = \varepsilon(n)$ the mean of the squared derivative is calculated as:

$$\begin{aligned} E\{\dot{r}_{\text{noise}}^2\} &\approx (\Delta T)^{-2} E\{(r(n) + \varepsilon(n) - r(n-1) - \varepsilon(n-1))^2\} \\ &= (\Delta T)^{-2}\begin{pmatrix} R_r(0) + R_r(0) - 2R_r(\Delta T) \\ + R_\varepsilon(0) + R_\varepsilon(0) - 0 \\ + 0 \end{pmatrix} \end{aligned}$$

where it has been exploited that the auto-correlation of the noise is zero for non-zero arguments, and the cross-correlation between the envelope and noise is assumed to be zero. The average value of the squared derivative reduces to

$$E\{\dot{r}_{\text{noise}}^2\} \approx 2(\Delta T)^{-2}\left(R_r(0) - R_r(\Delta T) + R_\varepsilon(0)\right) \quad (12)$$

Define by γ the noise ratio between the power of the noise and the power of the envelope $\gamma = R_\varepsilon(0)/R_r(0)$, and introduce the correlation coefficient $\rho_r(\tau) = R_r(\tau)/R_r(0)$. The estimator is then approximated by

$$\hat{v} \approx \frac{c}{\pi f_c \Delta T}\sqrt{\frac{2(1 - \rho_r(\Delta T) + \gamma)}{1+\gamma}} \quad (13)$$

We carry on by approximating the auto-correlation coefficient of the envelope, $\rho_r(\tau)$, see e.g. [2] (Equation 1.3-16) or [3] (page 193)

$$\rho_r(\tau) = \frac{\pi}{4} + \left(1 - \frac{\pi}{4}\right) J_o^2(2\pi f_m \tau) \quad (14)$$

Using this and a Taylor-expansion to the second order of the squared Bessel function, $J_0^2(\tau) \approx 1 - \tau^2/2$, we obtain

$$\hat{v} \approx \frac{c}{\pi f_c \Delta T} \sqrt{\frac{(4-\pi)(\pi f_m \Delta T)^2 + 2\gamma}{1+\gamma}} \tag{15}$$

From (15) several observations can be made. First, if setting the noise ratio, γ, to zero the estimator reduces to $\hat{v} = v \cdot \sqrt{4-\pi} \approx 0.93v$. The reason for the non-ideal result lies primarily in the approximation made in (14).

Second, in case of $\gamma > 0$ both the numerator and the denominator of the fraction within the square-root become positively biased due to the noise. Note also that the quality of the estimator is closely related to the ratio of $(\pi f_m \Delta T)^2$ and 2γ (here we might ignore the factor $4-\pi$ which is close to one). When $(\pi f_m \Delta T)^2$ and 2γ is small the estimator has little bias. The result is interesting, since is shows that for a certain noise-ratio the estimator is significantly biased under a certain speed which can be derived from the noise-ratio.

The approximation in (15) is obtained by analyzing the effect of real WGN added to the envelope, rather than complex WGN added to the channel $h(t)$. Still the results and observations are important, since corresponding real envelope AWGN can be derived from the complex channel AWGN.

Fig. 1 shows the predicted speed estimate as a function of speed for geometry factors G ={-3, 0, 3, 6, 9} dB. For e.g. G = -3 dB we observe a saturation of the speed estimator at 65 km/h. Furthermore, we see that the estimator has a linear part with a slope close to, but not exactly one. Some of this mismatch is due to the approximations made in the derivation of (15).

As observed from (15) and seen in Fig. 1 the bias of the proposed estimator should be reduced and it's slope adjusted to slope one, in order to obtain a better estimate. In the next section two possible such enhancements are presented.

5. ENHANCEMENTS OF THE ESTIMATOR

We aim to estimate the speed in the range from 0 to 250 km/h, so we should exploit that the spectrum of the fading channel is band-limited. We know from (15) that a reduction of the noise results in a bias decrease. In order to reduce the noise power we could lowpass filter the envelope of the channel with a cutoff frequency set to the corresponding signal-bandwidth for the maximum speed, e.g. by using a Butterworth lowpass IIR filter. Since we also like to be immune to any potential frequency offset we will eliminate a phase rotation by filtering the envelope rather than the complex channel estimate.

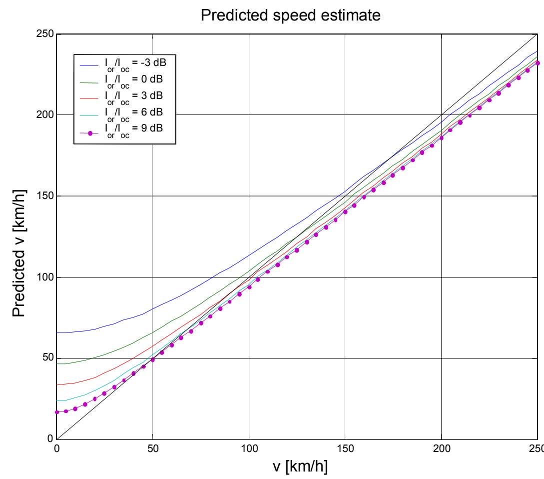

Fig. 1: Predicted behavior of the speed estimator as function of the true speed for five different geometry factors and Ec/Ior = -10dB.

For a carrier frequency of 2.17 GHz and a maximum speed of 250 km/h the maximum Doppler frequency of the channel is approximately 500 Hz, by (5). This means that when lowpass filtering the squared envelope we must use a cut-off frequency of around 1000 Hz, if we want to avoid cropping the spectrum of the fading process. This follows directly from the sampling theorem.

The effect of lowpass filtering on the proposed estimator can be seen in Fig. 2 for the simulation parameters listed in TABLE I, i.e. relevant parameters for 3GPP WCDMA FDD [4]. As seen a 4th order Butterworth IIR lowpass filter is used for the envelope filtering.

TABLE I:
SIMULATION SETUP

Ec/Ior_{CPICH}	-10 dB
Geometry factor (Ior/Ioc)	-3 dB
Carrier frequency	2.14 GHz
Symbol rate	15 kHz
Envelope cut-off frequency, 4th order Butterworth IIR low pass filter	Sweep from 300 – 1400 Hz in steps of 100 Hz
Radio channel	1 tap Rayleigh fading

As seen in Fig. 2 the cut off frequency of the lowpass filter directly affects the performance of the proposed estimator as expected, e.g. if a low cut off frequency is applied in the filtering of the envelope the bias in the low speed region is reduced and the maximum speed of the proposed estimator is reduced. This observation can be used to enhance the performance of the estimator for specific speed regions. A further step in enhancing the performance is to correct the slope of the estimator output.

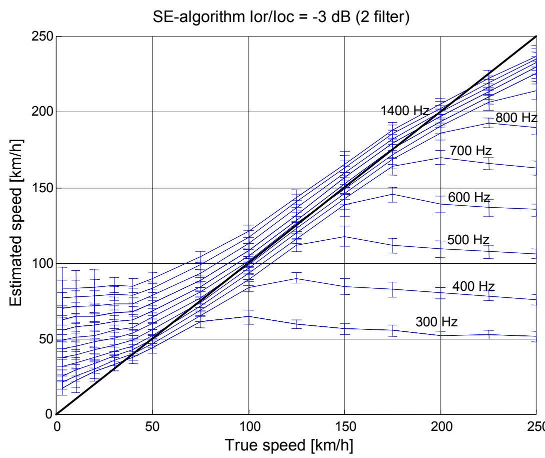

Fig. 2: Speed estimator performance after lowpass filtering the envelope at various cut off frequencies.

In Fig. 3 is shown a system overview for the proposed estimator. Due to the lowpass filter and due to the noise effect discussed analytically in section 4 the linear part of the estimator curve will have a slope which is too low. This artefact could be corrected e.g. by applying a linear mapping of the speed estimate, with slope and offset parameters estimated from simulations based on the actual selection of filter and sampling parameters.

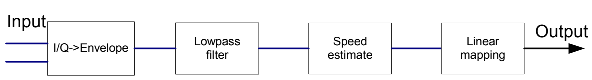

Fig. 3: Speed estimator system overview.

6. SIMULATION RESULTS

In Fig. 4 is shown the performance for 3GPP fading case 2 with a geometry factor of 9 dB.

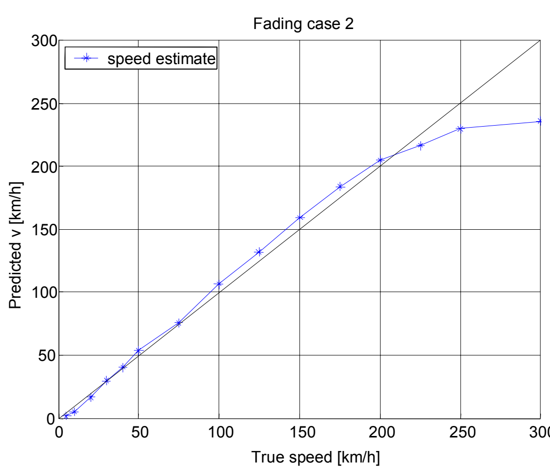

Fig. 4: Estimated speed as function of true speed. The x-axis denotes the actual speed and the y-axis denotes the estimated speed.

Two estimators are used in parallel with filters that have been optimized to cover the two ranges from 0-100 km/h and 100-250 km/h respectively. A simple threshold level is used to separate the two ranges.

Fig. 5 shows an accelerating UE. The estimator yields a speed estimate every 0.5 seconds. We observe that the estimator works very well.

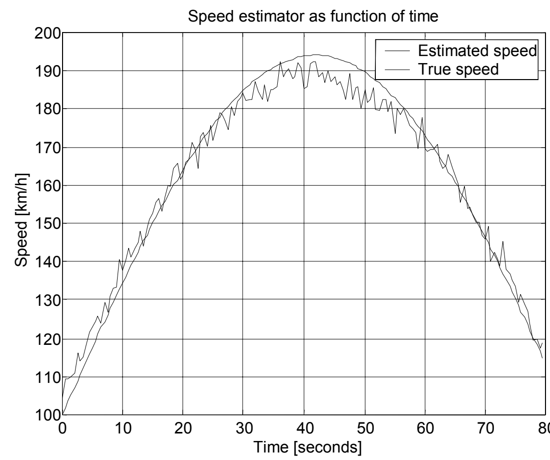

Fig. 5: The true and estimated speed as functions of time with an estimator update interval of 0.5 seconds.

Fig. 6 shows the performance of the estimator with one fading tap in the presence of a Rice component – the performance is shown for Rice factors between -15 and 10 dB. As seen the accuracy of the algorithm is reduced as the Rice factor is increased. This dependency of the power spectrum shape was not expected, and occurs from the derivative approximation done in (6). For Rice factors of -15, -10, and -5 dB the performance is virtually not affected.

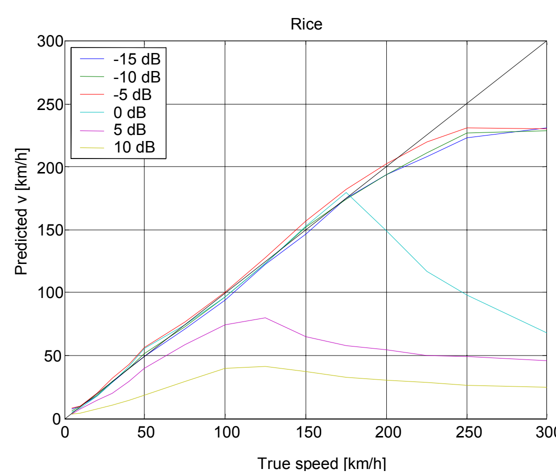

Fig. 6: The true and estimated speed as functions of various Rice factors. The x-axis denotes the actual speed and the y-axis denotes the estimated speed.

For a Rice factor of 0 dB the performance is fine up to approximately 175 km/h, but then the speed estimator starts to fail. The reason is that the low speed region estimate for 0 dB Rice factor has declined below the threshold which determines whether we use the low- or high-speed estimator, hence the break. This effect depends on the speed range combining method, the chosen speed ranges, and the amount of speed ranges.

For a Rice factor of 5 dB the estimator fails at around 100 km/h. For very dominant Rice components (e.g. a Rice factors > 10dB) the estimate is biased towards zero, which is an acceptable behaviour, since a radio channel with a dominant Rice component acts like a static channel.

7. CONCLUSION

In this paper we have proposed a reliable speed estimator, which estimates the speed corresponding to the maximum Doppler shift. As seen in the performance plots the proposed estimator has very good accuracy from 0 to approximately 200 km/h. A limited bias of 5-10 km/h is observed for the full speed-range. Typically, the proposed estimator starts to saturate when the speed exceeds 200 km/h. However, this problem can be solved by combining two or more of the proposed estimators, where each is optimised to a specific range.

The proposed algorithm has a low complexity, hence it is feasible for DSP-implementation in the UE. The core of the algorithm is easy to implement, since it only requires filtering, the calculation of a difference and a few sums in order to compute the average values for a certain observation window.

Our experience is that the algorithm works fine for radio channels close to a single radio tap as well as for multi-tap propagation conditions such as the 3GPP test cases. The proposed estimator has also been found to work well in case of frequency shifts, multi-path interference, and saturation – the latter is not reported in this work.

8. REFERENCES

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