TIME PREDICTION OF NON FLAT FADING CHANNELS

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ABSTRACT

In this paper three different techniques to predict the channel transfer functions of non flat fading channels in multiple antenna systems are investigated. The prediction is done into time direction. In time division duplex systems, it is beneficial to have the full channel state information. A receiver can achieve the channel state information for one certain time instant based on the reception of a–priori known training signals. This estimated channel state information can be outdated when receiving signals at another time instant or especially when transmitting a signal back to the previously transmitter at another time instant. One possibility to cope with the problem of outdated channel state information is to retransmit a–priori known training signals in order to update the channel state information but this decreases the spectral efficiency.

This paper shows techniques to perform the prediction based on the estimation of the physical path parameters of the electromagnetic waves and based on filtering the outdated channel state information. The performance of the presented techniques is evaluated by means of measurement results.

Index Terms— prediction, antenna array

1. INTRODUCTION

To have the channel state information (CSI) of the channels over which the received signals propagate and of the channels over which the transmitted signals will propagate for different time instances is beneficial in order to achieve a high spectral efficiency. However, the properties of the mobile radio channel vary over time. For this reason the CSI of the channels is different for different time instants. In order to obtain the CSI for one certain time instant it is appropriate to estimate the channel transfer functions (CTF) based on a–priori known training signals. For estimating the CTF for another time instant it is necessary to estimate the CTFs based on a–priori known training signals again. But this decreases the spectral efficiency. Hence, it is desirable to perform a prediction of the already known CTFs in order to obtain the CSI of the channels over which the later received signals propagate or of the channels over which the transmitted signals will propagate.

In former contributions different techniques for predicting the CTF in SISO systems were investigated. Mainly two different approaches were explored. The first approach [1, 2, 3, 4] is based on the physical path parameters which have to be estimated, thus, this approach can be physically motivated. The second approach is based on filtering the already known CTFs [3, 4, 5]. Here, it is assumed that the samples of the CTF can be modeled as an autoregressive process, thus, this approach seems not to be physically motivated. In most cases flat fading channels were assumed or the prediction was performed for every frequency separately. In this contribution, the prediction is not limited to frequency flat fading channels and the prediction is done jointly for all frequencies.

The remainder of the paper is organized as follows: In Section 2 the underlying channels models will be introduced. In Section 3 the proposed prediction algorithms will be investigated. Section 4 is dedicated to the equivalence of both channel models. The measurement results are presented in Section 5. Finally, in Section 6 we draw our conclusion.

2. CHANNEL MODELS

We consider a scenario where the moving MS is equipped with 1 antenna and the BS is equipped with M antennas. The directional scenario is depicted in Figure 1. It is assumed that D plane waves impinge at the receiver antennas. The D propagation paths will have different complex attenuations, directions of arrival (DOAs), times of arrival (TOAs), and Doppler frequencies. The spatial distances among the antennas of the antenna array are in the range of some wavelengths, thus it is assumed that the complex attenuation, the DOAs, and the Doppler frequencies are equal for all M antennas. The TOAs are defined as the times which are needed for the propagation of the D electromagnetic waves from the MS at the start time \( t_s \) to the RP at the receiver side. With these assumptions the band limited CTF at the center frequency \( f_0 \) in the equivalent low–pass domain between the m-th receive antenna and the transmit antenna can be described by a superposition of D complex exponential functions

\[
H^{(m)}(f, t) = \text{rect} \left( \frac{f}{B} \right) \cdot \sum_{d=1}^{D} \alpha^{(d)} \cdot e^{-j2\pi(f+f_0)\tau^{(d)}} \cdot e^{-j2\pi(f+f_0)(m,D)\tau^{(d)}} \cdot e^{j2\pi f_A t_d^{(d)}(t+t_s)}. \tag{1}
\]
Here $\alpha^{(d)}$ is the complex weight of the $d$-th propagation path and it is assumed that the attenuations of the individual propagation paths do not depend on the frequency $f$ within the considered bandwidth $B$. $\tau^{(d)}$ is the delay of the $d$-th propagation path. The difference of the TOA of the $d$-th propagation path at the RP and at the $m$-th antenna is described by $\tau_{Nx}^{(m,d)}$. For the sake of simplicity it is assumed that $f$ is limited to the range $[-B/2,B/2]$, thus the rectangular function rect(·) in (1) can be neglected in the following. Furthermore, the narrowband assumption is made, thus
\[
e^{-j2\pi(f_{0})\tau_{Nx}^{(m,d)}} \approx e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}
\]
holds and the CTF of (1) results in
\[
H^{(m)}(f, t) = \sum_{d=1}^{D} H^{(d)}_{A}(f, t) \cdot e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}
\]
with
\[
H^{(d)}_{A}(f, t) = \alpha^{(d)}(t) \cdot e^{-j2\pi(f+f_{0})\tau^{(d)}_{A}} \cdot e^{j2\pi f_{d}^{(d)}(t+t_{s})}.
\]
Here $H^{(d)}_{A}(f, t)$ is the directional CTF and depends neither on the DOAs nor on the geometry of the antenna array. The factor $e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}$ is the steering factor and depends on the direction of the antenna array, and on the DOAs [6]. The path specific delays $\tau_{Nx}^{(m,d)}$ can be easily calculated with the knowledge of the DOAs [6].

One sample of the CTF of (3) in the frequency domain can be written as
\[
H^{(m)}_{w,v}(f_{0}, t_{s}) = \sum_{d=1}^{D} \hat{h}^{(d)}_{w,v}(f_{0}, t_{s}) \cdot e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}
\]
with
\[
\hat{h}^{(d)}_{w,v}(f_{0}, t_{s}) = \hat{a}^{(d)} \cdot e^{-j2\pi(F_{w}+f_{0})\Delta\tau_{v}} \cdot e^{j2\pi f_{d}^{(d)}(T_{v}+t_{s})},
\]
where $F$ is the sampling distance along the frequency axis and $T_{v}$ is the sampling distance along the time axis. Furthermore, for the index $w$ holds $w = -\frac{1}{\Delta\tau_{v}}$, $\ldots, \frac{1}{\Delta\tau_{v}}$ and for the index $v$ holds $v = 0 \ldots V−1$. The combination of all $W \cdot V$ samples of the CTF for one antenna pair leads to
\[
H^{(m)}(f_{0}, t_{s}) = \left( H^{(m)}_{-\frac{1}{\Delta\tau_{v}},0} \cdots H^{(m)}_{\frac{1}{\Delta\tau_{v}},0} \cdots H^{(m)}_{\frac{1}{\Delta\tau_{v}},V-1} \right)^{T}
\]
(7).

The combination of all $M$ channel vectors leads to the $W \cdot V \times M$ channel matrix
\[
\mathbf{H}(f_{0}, t_{s}) = \left( \mathbf{H}^{(1)}(f_{0}, t_{s}) \cdots \mathbf{H}^{(M)}(f_{0}, t_{s}) \right).
\]
(8)

Sampling the directional CTFs of (4) leads to the directional channel vector $\mathbf{h}_{A}^{(d)}(f_{0}, t_{s})$. The combination of all $D$ directional channel vectors leads to the $W \cdot V \times D$ directional channel matrix $\mathbf{H}(f_{0}, t_{s})$. The combination of the steering factors $e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}$ for all DOAs for all $M$ antennas results in the $M \times D$ steering matrix $\mathbf{A}(f_{0})$. With these matrices the channel matrix can be written as
\[
\mathbf{H}(f_{0}, t_{s}) = \mathbf{H}_{A}(f_{0}, t_{s}) \mathbf{A}^{T}(f_{0}).
\]
(9)

In addition to this channel model which is based on the physical description of the propagation paths it is also possible to describe a sample of the CTF as a linear combination of $Q$ other known samples of the same CTF. In this case the $(w, v + \Delta v)$-th samples of the $m$-th CTF can be written as
\[
\begin{pmatrix}
H^{(m)}_{W-1} \cdot e^{j2\pi f_{d}^{(d)}T_{v}} \\
\vdots \\
H^{(m)}_{W-1} \cdot e^{j2\pi f_{d}^{(d)}T_{v} - (Q-1)}
\end{pmatrix}
\mathbf{P}.
\]
Here, the $(w, v + \Delta v)$-th sample only depends on $Q$ preceding samples of the same frequency index $w$. In general, it is possible to describe the $(w, v + \Delta v)$-th sample also in dependency of samples of other frequency indices.
Assuming $\Delta v = 1$, in some contributions this channel model is called autoregressive channel model. At a first glance, these two channel models seem to be fundamentally different. Nevertheless, both can be used in order to predict the CTF.

3. PREDICTION AND ESTIMATION OF THE MODEL PARAMETERS

3.1. Propagation path based prediction

With equation (3) and (4) the band limited CTF at the $m$-th antenna in the equivalent low–pass domain at time $t + \Delta t$ reads
\[
H^{(m)}(f_{0}, t + \Delta t) = \sum_{d=1}^{D} H^{(d)}_{A}(f_{0}, t) \cdot e^{j2\pi f_{d}^{(d)}(t+t_{s})} \cdot e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}}.
\]
(11)
Sampling the $M$ CTFs of (11) leads to the channel matrix $\mathbf{H}(f_{0}, t_{s} + \Delta t)$, where one sample can be written as
\[
H^{(m)}_{w,v + \Delta v} = \sum_{d=1}^{D} L^{(d)}_{w,v}(f_{0}, t_{s}) \cdot e^{-j2\pi f_{0}\tau_{Nx}^{(m,d)}} \cdot e^{j2\pi f_{d}^{(d)}T_{v}} + t_{s},
\]
with $\Delta t = T_{v} \Delta v$. This channel matrix can be written as
\[
\mathbf{H}(f_{0}, t_{s} + \Delta t) = \mathbf{H}_{A}(f_{0}, t_{s} + \Delta t) \mathbf{A}^{T}(f_{0}).
\]
(13)
Thus, it is sufficient just to predict the directional channel matrix. This leads to
\[
\mathbf{H}(f_{0}, t_{s} + \Delta t) = \hat{\mathbf{H}}_{A}(f_{0}, t_{s}) \cdot e^{j2\pi f_{d}^{(d)}T_{v}} \cdot \mathbf{A}(f_{0})^{T}.
\]
(14)

In order to predict the CTFs respectively the directional channel matrix it is necessary to firstly estimate the model parameters. Regarding the first model (9) it is necessary to estimate the physical propagation path parameters of the $D$ electromagnetic waves. The physical path parameters can be estimated, e.g., employing the SAGE algorithm [7].

Secondly, it is necessary to estimate the directional channel matrix and the steering matrix at the center frequency $f_{0}$ at the start time $t_{s}$. With the knowledge of the $D$ DOAs the steering matrix can be computed [8]. In order to estimate the directional channel matrix we propose two possibilities. The first approach is based on the channel matrix $\mathbf{H}(f_{0}, t_{s})$ and is based on the previously estimated steering matrix $\mathbf{A}(f_{0})$. With the knowledge of these matrices it is possible
to compute a least squares estimate of the directional channel matrix [9]. The second approach to estimate the directional channel matrix is based on the knowledge of the $D$ TOAs, the $D$ Doppler frequencies, and the $D$ complex weights. Referring to equation (4) the directional CTFs can be easily reconstructed with the knowledge of these parameters. This approach is called reconstruction approach.

3.2. Filter based prediction

The prediction of the CTFs based on the filter based channel model does not require the knowledge of the physical path parameters. In this case only $Q$ preceding samples for each CTF for all frequency indices $W$ and the knowledge of the $Q$ filter coefficients are required. With the filter order $Q$ and $V$ known samples of each of the $M$ CTFs it is possible to predict $(V - Q + 1)$ samples of each CTF for one frequency index $w$

$$
\begin{pmatrix}
H_w^{(m)}(Q - 1 + \Delta v) \\
\vdots \\
H_w^{(m)}(V - 1 + \Delta v)
\end{pmatrix}
= \begin{pmatrix}
H_w^{(m)}(Q - 1) & \cdots & H_w^{(m)}(0) \\
\vdots & \ddots & \vdots \\
H_w^{(m)}(V - 1) & \cdots & H_w^{(m)}(V - 1 - (Q - 1))
\end{pmatrix}
\cdot \mathbf{p},
$$

where $\mathbf{p}$ is the vector containing the $Q$ filter coefficients. In order to predict more than $(V - Q + 1)$ samples of one CTF for all frequency indices $W$ this technique can be continued iteratively.

It is necessary to determine the filter coefficients $p_{w,q}$, $q = 0 \ldots Q$ in order to predict the CTFs. Based on (15) it is feasible to estimate $\mathbf{p}$ by predicting a-priori known samples of the CTFs [4]. As long as all samples of the CTFs in (15) are known it is feasible to perform a least squares estimation of the filter coefficients. Furthermore, due to the fact that $\mathbf{p}$ neither depends on the frequency index $w$ nor on the antenna index $m$ equation (15) can be extended for all frequencies and antennas. Thus, the quality of the least squares estimation of the filter coefficients can be increased. This technique to estimate the filter coefficients is known as the covariance method in the literature [10].

4. RELATION BETWEEN BOTH CHANNEL MODELS

The aforementioned channel models seem to be completely different. However, it can be shown that both models are related.

Replacing the elements of the filter based channel model by the elements of the directional channel model of (12) in (10) leads to

$$
\sum_{d=1}^{D} p^{(d)}_{w,v} \cdot e^{-j2\pi f_{d}(m,d) T \cdot \Delta v} = \sum_{q=0}^{Q-1} p_{w,q} \cdot \sum_{d=1}^{D} \omega_{w,v}^{(m,d)} \cdot e^{-j2\pi f_{d}^{(m)} T \cdot q}.
$$

Equation (16) can be rewritten to

$$
\begin{pmatrix}
\omega^{(m,1)}_{w,v} & \cdots & \omega^{(m,D)}_{w,v}
\end{pmatrix}
\begin{pmatrix}
e^{j2\pi f_{1}^{(1)}(m) T \cdot \Delta v} \\
\vdots \\
e^{j2\pi f_{D}^{(D)}(m) T \cdot \Delta v}
\end{pmatrix}
= \begin{pmatrix}
e^{-j2\pi f_{1}^{(1)}(w) T \cdot 0} & \cdots & e^{-j2\pi f_{1}^{(1)}(w) T \cdot (Q-1)} \\
\vdots & \ddots & \vdots \\
e^{-j2\pi f_{D}^{(D)}(w) T \cdot 0} & \cdots & e^{-j2\pi f_{D}^{(D)}(w) T \cdot (Q-1)}
\end{pmatrix}
\cdot \mathbf{p},
$$

This equation must hold for any frequency index $w$, for any time index $v$, and for any antenna index $m$, thus, there are $W \cdot V \cdot M$ equations of the kind of equation (17). As long as the number of propagation paths $D$ is smaller than $W \cdot V \cdot M$ it is necessary to prove that the equality

$$
\begin{pmatrix}
e^{j2\pi f_{1}^{(1)}(1) T \cdot \Delta v} \\
\vdots \\
e^{j2\pi f_{D}^{(D)}(1) T \cdot \Delta v}
\end{pmatrix}
\begin{pmatrix}
e^{-j2\pi f_{1}^{(1)}(0) T \cdot 0} & \cdots & e^{-j2\pi f_{1}^{(1)}(0) T \cdot (Q-1)} \\
\vdots & \ddots & \vdots \\
e^{-j2\pi f_{D}^{(D)}(0) T \cdot 0} & \cdots & e^{-j2\pi f_{D}^{(D)}(0) T \cdot (Q-1)}
\end{pmatrix}
\cdot \mathbf{p}
$$

holds in order to show that the equality in (17) holds for all $W \cdot V \cdot M$ equations. Equation (18) is a system of $D$ linear equations with $Q$ unknowns. This system of linear equations has no solution in the case of $D > Q$. This means that the filter order must be equal or higher than the number of propagation paths. Obviously, the filter coefficients depend neither on $w$, $v$, nor on $m$. It is shown that the filter based channel model can also be motivated in a physical sense. This result can be used to increase the quality of the estimation of the filter coefficients by the covariance method because the filter coefficients are the same for all antennas and for all frequencies.

5. MEASUREMENT RESULTS

In order to get an idea of the performance of the presented techniques, we performed channel measurements in an indoor scenario using a vector network analyzer. The BS antenna could be moved on a circular path using a turntable for forming a virtual uniform circular array. The MS antenna was mounted on a linear motion table. Figure 2, the measurement setup is shown. The BS antenna was moved in 10° steps corresponding to $M = 36$ antenna positions. The radius of the turntable was 25 cm. The distance between two adjacent positions of the MS antenna was 1 cm, and the CTFs were measured at $W = 36$ different MS antenna positions. The center frequency was 2.45 GHz, the bandwidth was 100 MHz, and we measured the CTFs at $W = 201$ frequency points.

The first step in order to predict the CTFs is to estimate the parameters. In order to assess the proposed algorithms we only used the first $V = 18$ known time instants of the measured CTFs for estimating the physical path parameters and the filter coefficients, i.e., the uplink channel. The propagation path parameters were estimated by the SAGE algorithm. We assumed to have $D = 7$ propagation paths, thus, the power of the strongest path was approximately $14 \text{ dB}$
The results are shown in the Figure 3. The normalized mean square error is plotted versus the normalized distance $\Delta x/\lambda$, where $\lambda$ is the wavelength. It can be seen that it is not appropriate to directly use the CTFs of the last known time instant. The result of the filter approach is better than the results based on the directional channel model. This probably results from bad results of the SAGE algorithm and maybe from the fact that the made assumption concerning the directional channel model are not completely valid. Depending on the reason why this prediction was done, the results can be used either directly or the error can be further reduced by feeding back information from the MS to the BS with less information than it would be required for the complete feedback of the estimated CTFs.

6. CONCLUSION AND FUTURE WORK

This paper presents the principles of three techniques to predict the CTFs in multiple antenna systems based on the estimation of propagation path parameters and based on the estimation of filter coefficients of a predictor. The filter approach performs better than the approach based on the directional channel model, although it is possible to find filter coefficients in such a way that filter based channel model is equal to the directional channel model. Future work may include further measurement results and the prediction of the CTFs into time and frequency direction.

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8. REFERENCES


