Pilot-assisted Channel Estimation for Broadband ANC
without Relay Feedback

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Abstract   Broadband analog network coding (ANC) has been recently introduced to further increase the network capacity by exploiting the broadcasting nature of the wireless channel. However, channel state information (CSI) knowledge is required for self-information removal and signal detection. Recently, a pilot-assisted channel estimation (PACE) scheme has been presented for broadband ANC; however, feedback of the channel estimates from the relay to the users is required. In this paper, we present a PACE scheme for broadband ANC, based on orthogonal frequency-division multiplexing (OFDM) radio access, without feedback from the relay. In the first time slot, the users transmit their respective pilots to the relay and then, in the second time slot, the relay amplifies and forwards the received pilot signals to both users. Each user can then estimate all the CSI it needs for self-information removal and signal detection, without requiring any feedback from the relay. The bit error rate (BER) performance of broadband ANC using the proposed PACE is evaluated by computer simulation. It was shown that the proposed PACE scheme can achieve almost the same BER performance as the conventional PACE scheme, while eliminating the feedback channel requirement.

Keywords   Broadband　ANC，OFDM，channel estimation

1. Introduction

Next generation wireless communication networks will be expected to provide very high capacity for the increasing broadband services such as multicasting, video conferences, video on demand, etc. In wired networks, network coding [1] has been widely used to increase the network capacity. The same concept can be used in wireless networks to exploit the broadcasting nature of wireless transmission and further increase the network capacity [2]. It has been shown that for bi-directional wireless communication, network coding at the physical layer (PNC) can double the network capacity [3], [4]. Narrowband analog network coding (ANC) was introduced in [5] as a simplification of PNC where the signals from the users are mixed in the wireless medium.

It was shown in [6] assuming the perfect knowledge of channel state information (CSI) that broadband ANC has the ability to overcome the frequency-selectivity of the broadband wireless channel. Coherent detection and self-information removal in the broadband ANC scheme require accurate channel estimation (CE). A straightforward approach would be allocating four time slots for pilot-assisted channel estimation, since the users'

signals interfere in the same time slot. However, this might greatly decrease the capacity benefit of ANC. In [7], a two-slot pilot-assisted channel estimation scheme is introduced. The users' pilot signals are designed in such a way that the interference problem is avoided. However, the scheme assumes ideal feedback from the relay to the users, which is practically difficult to achieve and very costly.

In this paper, we propose a two-slot pilot-assisted CE scheme for broadband ANC based on orthogonal frequency-division multiplexing (OFDM) radio access that does not require feedback from the relay. In the first stage, both users transmit their pilot signals to the relay. In order to avoid interference in the first time slot [7], the pilot signal of one of the users is cyclically shifted [8] to allow the signals to be separated at the destination, for channel estimation. The relay simply amplifies and forwards the received combined pilot signals, with the corresponding CSIs being estimated at the users. We evaluate the performance of broadband ANC with the proposed CE scheme by computer simulation and compare it to the perfect CSI case and the CE scheme introduced in [7]. We show that the proposed PACE scheme can achieve similar BER performance to that

of the conventional PACE scheme, with a slight degradation caused by noise enhancement, while costly feedback from the relay is not required.

The rest of the paper is organized as follows. In Section 2, we present the network model. The proposed CE scheme is presented in Section 3. Section 4 shows the results of the computer simulation and discussions. We summarize our findings in Section 5.

2. Network model

We consider a two-way relay network, with users U_0 and U_1 outside each other's coverage area, that communicate through the relay R, as shown in Fig.1. The communication between the users and the relay is done using time division duplex (TDD) in two slots: in the first time slot U_0 and U_1 transmit their signals to the relay and in the second time slot the relay broadcasts the received signal to the users through the amplify-and-forward (AF) protocol. In this paper, we assume that the channel between the users and the relay does not change during the two slots.
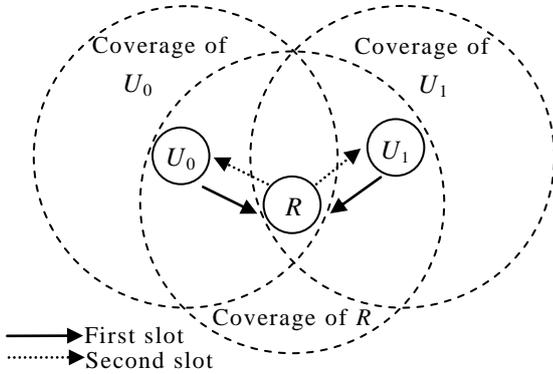

Fig. 1 Network model.

1) First time slot

The data-modulated symbol sequence of the jth user U_j is represented by $\{d_j(n); n=0\sim N_c-1\}$ for $j\in\{0,1\}$. An N_c-point inverse fast Fourier transform (IFFT) is applied to $\{d_j(n); n=0\sim N_c-1\}$ in order to generate the respective OFDM signals. An N_g-sample guard interval (GI) is inserted, and the signals from the users are transmitted through the frequency-selective fading channel. The GI length is assumed to be longer than the maximum time delay of the channel.

The frequency-domain received signal $\{R_r(n);$ $n=0\sim N_c-1\}$ at the relay can be expressed as

$$R_r(n) = \sum_{j=0}^{1}\sqrt{2P_t}\,d_j(n)H_j(n) + N_r(n), \qquad (1)$$

where $P_t(=E_s/T_cN_c)$, $H_j(n)$, and $N_r(n)$ denote the transmit signal power of the users, the channel gain between user U_j and the relay, and the additive white Gaussian noise (AWGN) at the nth frequency with single-sided power spectral density N_0, respectively. E_s and T_c denote the symbol energy and the sampling period of IFFT, respectively.

2) Second time slot

The relay terminal normalizes the received signal shown in Eq. (1) by the factor $G=\sqrt{1/E\!\left[R_r(n)^2\right]}$, so that the average energy is unity, and broadcasts it with transmit power P_r. The frequency-domain representation of the received signal $\{$ $R_j(n)$; $n=0\sim N_c-1\}$ at user U_j's receiver can be written as

$$R_j(n) = \sqrt{2P_r}\,G\cdot R_r(n)H_j(n) + N_j(n), \qquad (2)$$

where $N_j(n)$ denotes the AWGN at the user side with single-sided power spectral density N_0.

The jth user U_j removes its self-information. The frequency-domain signal $\{$ $\tilde{R}_j(n)$; $n=0\sim N_c-1\}$ after self-information removal can be expressed as

$$\tilde{R}_j(n) = R_j(n) - d_j(n)H_j^{(0)}(n), \qquad (3)$$

where $H_j^{(0)}(n)$ denotes the channel between user U_j and U_j via the relay, given by

$$H_j^{(0)}(n) = 2\sqrt{P_rP_t}\,G\cdot H_j^2(n). \qquad (4)$$

One-tap zero forcing frequency domain equalization (ZF-FDE) is then applied as

$$\hat{R}_j(n) = \tilde{R}_j(n)W_j(n), \qquad (5)$$

where $W_j(n)$ is the equalization weight for the nth subcarrier, given by

$$W_j(n) = \frac{\left\{H_j^{(1)}(n)\right\}^*}{\left|H_j^{(1)}(n)\right|^2}, \qquad (6)$$

where $H_j^{(1)}(n) = 2\sqrt{P_rP_t}\,GH_0(n)H_1(n)$ is the channel gain of the channel between user U_0 and U_1 and $(\cdot)^*$ denotes the complex conjugate operation.

The self-information removal given by Eq. (3) and the ZF-FDE given by Eq. (6) require CSI. In practice,

the equivalent channel gains, $\{H_j^{(0)}(n); n = 0 = N_c - 1\}$ and $\{H_j^{(1)}(n); n = 0 = N_c - 1\}$, are replaced by the respective channel gain estimates, $\{\breve{H}_j^{(0)}(n); n = 0 = N_c - 1\}$ and $\{\breve{H}_j^{(1)}(n); n = 0 = N_c - 1\}$, respectively.

3. Channel estimation

Accurate channel estimation is crucial to broadband ANC transmissions, due to the use of CSI in both self-information removal and FDE. The straightforward approach for CE for a TDD bi-directional relay network would be a four time slot scheme, to separate the pilot signals of different users. However, in a fast fading environment, where we need an increased pilot insertion rate in order to improve the tracking ability, this would significantly decrease the capacity of the ANC scheme. A two-slot CE scheme that uses cyclically shifted pilot signals to separate the signals from the users has been introduced for broadband ANC systems [7], but it requires feedback of the channel estimates from the relay to the users. Perfect feedback poses some difficult implementation problems, and accurate feedback can be very costly.

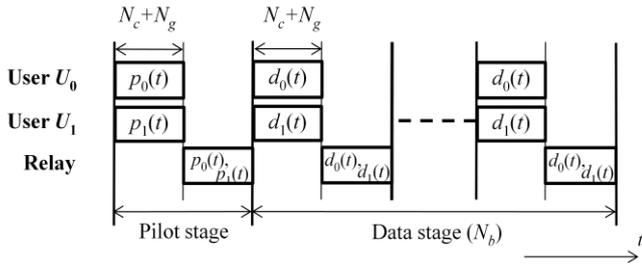We propose a channel estimation scheme that addresses the above mentioned problems.

Fig. 2 Transmission frame structure.

Our proposal is a pilot-assisted channel estimation scheme with time-domain multiplexed (TDM) pilots. The transmission frame structure of the users and the relay is shown in Fig. 2. Both the pilot and data frames are divided into two time slots, each of length $N_c + N_g$ samples. In the first time slot of the pilot stage, both users U_0 and U_1 transmit their respective pilot signals, $p_0(t)$ and $p_1(t)$, to the relay. During the second time slot the relay broadcasts through an AF

protocol the received superimposed pilot signals of the two users. Channel estimation is performed only at the users' side, and the estimates of the equivalent channels (one from user j to the relay and back, and the one from the other user through the relay to user j) are used in the following N_b data frames for self-information removal and equalization. The two pilot stage time slots are described below.

1) **First pilot time slot**

As shown in Fig. 2, during the first time slot, users U_0 and U_1 transmit their pilot signals to the relay through a frequency-selective fading channel. The received pilot signal $\{R_{r,p}(n); n=0\sim N_c-1\}$ at the relay can be expressed in frequency domain as

$$R_{r,p}(n) = \sum_{j=0}^{1} \sqrt{2P_t} P_j(n) H_j(n) + N_r(n), \qquad (7)$$

where $P_j(n)$ is the frequency domain representation of the pilot signal from user j.

In order to avoid the problem of the two channel impulse responses overlapping, the pilot signal $\{p_1(t)\}$ of U_1 is cyclically shifted by θ samples relative to user U_0's pilot signal $\{p_0(t)\}$, so that $p_1(t)=p_0((t-\theta) \bmod N_c)$ [8]. Thus the pilot signal of user U_1 can be expressed in frequency domain as

$$P_1(n) = P_0(n)\exp\left(-j2\pi\theta \frac{n}{N_c}\right) \qquad (8)$$

for $n=0\sim N_c-1$, and the pilot signal received by the relay can be written as

$$R_{r,p}(n) = \sqrt{2P_t} P_0(n)\left\{H_0(n) + H_1(n)\exp\left(-j2\pi\theta \frac{n}{N_c}\right)\right\} \qquad (9)$$
$$+ N_r(n).$$

2) **Second pilot time slot**

The relay simply normalizes the received superimposed pilot by G and broadcasts it with power P_r. After GI removal and N_c-point FFT, the frequency-domain received signal $\{R_{j,p}(n); n=0\sim N_c-1\}$ at user U_j's receiver can be represented as

$$R_{j,p}(n) = \sqrt{2P_r} G \cdot R_{r,p}(n).H_j(n) + N_j(n). \qquad (10)$$

This can be rewritten as

$$R_{j,p}(n) = P_0(n)\left\{\begin{array}{l} H_j^{(0)}(n) \\ + H_{1j}^{(1)}(n)\exp\left(-j2\pi\theta \frac{n}{N_c}\right) \end{array}\right\} + \tilde{N}_j(n), \qquad (11)$$

where $\tilde{N}_j(n) = \sqrt{2P_r} GN_r(n)H_j(n) + N_j(n)$ denotes the composite noise.

The reverse modulation is applied to $R_{j,p}(n)$ to remove the pilot as

$$\breve{H}_j(n) = \frac{R_{j,p}(n)}{P_0(n)}$$

$$= H_j^{(0)}(n) + H_j^{(1)}(n)\exp\left(-j2\pi\theta\frac{n}{N_c}\right) + \breve{N}_j(n), \quad (12)$$

where $\breve{N}_j(n) = \tilde{N}_j(n)/P_0(n)$. The Fourier transform $\{\breve{h}_j(\tau); \tau = 0\sim N_c-1\}$ of $\{\breve{H}_j(n); n = 0\sim N_c-1\}$ is obtained by taking an N_c-point IFFT as

$$\breve{h}_j(\tau) = h_j^{(0)}(\tau) + h_j^{(1)}((\tau-\theta)\bmod N_c) + \breve{n}_j(\tau), \quad (13)$$

where $h_j^{(0)}(\tau)$ and $h_j^{(1)}(\tau)$ denote the desired impulse responses of the channel between user U_j and U_j via the relay and of the channel between U_0 and U_1 via the relay, respectively. The third term is the noise. Note that due to the fact that the pilot of user U_1 is cyclically shifted by θ samples, the impulse response of the channel from this user has a delay of θ samples, which separates it from the impulse response of the channel from user U_0 and allows us to estimate both channels at the same time.
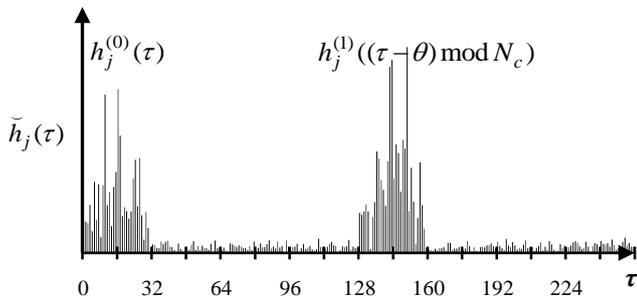

Fig. 3   Estimated channel impulse response.

An example of an actual estimate of the channel impulse response at the receiver when N_c=256 subcarriers, N_g=32 samples, θ=128 samples, for a multipath channel with the number of sample-spaced paths L=16 is shown in Fig. 3. It can be clearly seen from the figure that the channel impulse responses from the two users are completely separated in the delay time domain. Note that the delay spread of the equivalent channels increases compared to the original channel impulse response, and the GI size N_g and the cyclic shift of the pilot signal θ must be set accordingly, in order to avoid overlapping the two channel impulse responses. Since the equivalent

channels that are estimated can be expressed as the convolution of two sample-spaced L-path channels, the delay spread will become as high as double, and in order to avoid overlapping, we must have $2L \leq N_g \leq \theta \leq N_c/2$ for the case of FFT sample-spaced time delays.

A delay time domain window is used to separate the two channel impulse responses estimates as described in [7], by taking

$$\breve{h}_j^{(0)}(\tau) = \begin{cases} \breve{h}_j(\tau), & \text{for } \tau = 0 \sim N_g - 1 \\ 0, & \text{elsewhere} \end{cases} \quad (14)$$

and

$$\breve{h}_j^{(1)}(\tau) = \begin{cases} \breve{h}_j(\tau + \theta), & \text{for } \tau = 0 \sim N_g - 1 \\ 0, & \text{elsewhere.} \end{cases} \quad (15)$$

Finally, an N_c-point FFT is applied to both channel impulse response estimates, $\{h_j^{(0)}(\tau); \tau = 0 \sim N_g - 1\}$ and $\{h_j^{(1)}(\tau); \tau = 0 \sim N_g - 1\}$, to obtain the estimates of the channel gains $\{\breve{H}_j^{(0)}(n); n = 0 = N_c - 1\}$ and $\{\breve{H}_j^{(1)}(n); n = 0 = N_c - 1\}$, respectively.

Table 1   Simulation parameters

Transmitter U_0 ,U_1	Data modulation	QPSK
	Block size	$N_c = 256$
	GI	$N_g = 32$
Channel	L = 16-path block Rayleigh fading (uniform power delay profile)	
Relay R	Protocol	Amplify-and-forward
Receiver U_0 ,U_1	FDE	ZF
	Channel estimation	Pilot-assisted (Chu seq.)

4. Simulation results

This section presents the computer simulation results for the bit-error rate (BER) performance of an ANC network such as the one described in Sect. 2, using the CE scheme introduced in Sect. 3. The parameters used in the simulation are summarized in Table 1. We assume an OFDM system with ideal coherent quadrature phase-shift keying (QPSK) modulation and demodulation, N_c=256 subcarriers, and N_g=32 samples. The propagation channel is FFT sample-spaced L=16-path block Rayleigh fading, having a uniform power delay profile. For the pilot signal we use a Chu-sequence given by $\{p_0(t)=\exp\{j\pi t^2/N_c\}; t = 0 \sim N_c - 1\}$ [9]. We can see from then network model of Fig. 1 that the tracking ability against fading in the channel estimation is

identical for the two PACE schemes. In order to compare the accuracy of the CE of the proposed scheme to that of the conventional scheme, we consider a quasi-static fading channel (i.e., the Doppler frequency $f_D \rightarrow 0$) for the computer simulation.

Figure 4 illustrates the BER performance as a function of the total average signal energy per bit-to-AWGN power spectrum density ratio E_b/N_0 (= $0.5 \cdot (P_t + P_r) T_c / N_0 \cdot (1 + N_g/N_c) \cdot (1 + 1/N_b)$). The power loss due to GI and pilot insertion is taken into consideration. We compare the performance of the broadband ANC system with the proposed CE scheme, denoted by "Proposed PACE (w/o feedback)"), to the case of perfect knowledge of CSI, denoted by "Perfect CSI" and the conventional CE scheme, introduced in [7], denoted by "Conventional PACE (w/ feedback)".
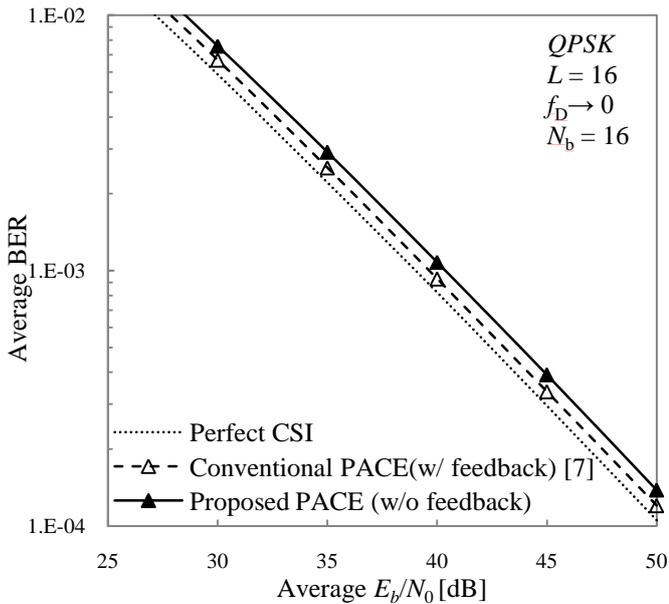

Fig. 4   BER performance.

As can be seen from the figure, the proposed PACE scheme achieves a satisfactory performance, while eliminating the need of a costly relay feedback. For example, for BER=10^{-3}, the E_b/N_0 degradation is 1.5dB in comparison to the perfect CSI case, and only 0.75dB in comparison with the CE scheme introduced in [7]. This degradation is due to the noise enhancement that occurs at the relay in the pilot stage. In the conventional scheme the channel is estimated both at the relay and at the users' side, and the pilot signals are not amplified and forwarded by the relay. In the proposed scheme the relay forwards the pilot signals along with the noise added at the relay, thus the noise component during the estimation process is enhanced. However, we can conclude that the noise enhancement does not severely affect the performance.

5. Conclusion

In this paper, we proposed a PACE scheme for broadband ANC systems that does not require feedback of CSI from the relay. The CE scheme is divided into two stages: first the users transmit their respective pilot signals to the relay, and then the relay amplifies and forwards the received combined signals back to the users, where the CSI required for self-information removal and coherent detection are estimated. Since each user can estimate all the CSI it needs for processing the data, there is no need for a costly feedback from the relay to the users.

The BER performance of an ANC system using the proposed CE scheme was evaluated by computer simulation, and compared to the case of perfect knowledge of CSI and the conventional CE scheme, that requires feedback from the relay. Our results show that the proposed scheme shows similar performance to the conventional CE scheme, with only a slight performance degradation due to noise enhancement at the relay during the pilot stage of the transmission, but achieves that without the need of feedback from the relay.

While the tracking ability of the proposed CE scheme is identical to that of the conventional scheme, the tracking problem has not been discussed in this paper. This is left as an interesting future work.

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