A Channel Estimation Scheme for Amplify-and-Forward OFDM Relay Networks

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***Abstract*—The channel state information (CSI) of all links should be available at all nodes for cooperative strategies design in relay networks. However, many channel estimation methods for amplify-and-forward (AF) relay networks only estimate a compound CSI of source-relay-destination (S-R-D) link. In this paper, we propose a channel estimation scheme for destination to disintegrate the CSIs of S-R and R-D links from the cascaded S-R-D link in AF relay networks. The proposed scheme estimates the CSI of R-D link firstly by utilizing some pilot signals inserted at the relay node. Then, the CSI of S-R link is estimated based on the received pilot signals and the pre-estimated CSI of R-D link. The channel estimation error of the proposed scheme is also analyzed. Especially, least square (LS) algorithm and minimum mean square error (MMSE) algorithm for the CSI estimation of S-R link are compared. We verify that the mean square error of LS estimation for S-R link is nonconvergent but that of MMSE estimation is convergent. At the end of this paper, simulation results are given to support the theoretical results.**

I. Introduction

The transmission with the aid of relay station has recently emerged as a powerful method in wireless networks due to its significant performance gains in terms of link reliability, spectral efficiency, system capacity, and transmission range [1]. The potential application areas of the relay technology include cellular, ad-hoc, WLAN and hybrid networks. A typical relay model consists of a source node (S), a relay node (R), and a destination node (D). In consequence, there are three types of independent links: from source to destination (S-D), from source to relay (S-R), and from relay to destination (R-D). In order to fully exploit the benefit of relay-aided transmissions, the channel state information (CSI) of all links should be available at source and relay nodes for cooperative transmission schemes design. However, many existing cooperative transmission protocols in relay networks have assumed that the CSIs of all links were pre-known, and have not mentioned how to obtain these CSIs [2]–[4].

Channel estimation problems in decode-and-forward (DF) relay systems basically consist of individual estimation of S-R link at relay node and R-D link at destination node. Therefore, the traditional pilot patterns and estimation algorithms designed for direct links can be used in DF relay systems. On the other hand, many channel estimation methods in amplify-and-forward (AF) relay systems are aimed at obtaining a cascaded channel consisting of S-R and R-D links at destination [5]–[7]. However, the CSIs of S-R and R-D links cannot be disintegrated from the compound CSI estimation of S-R-D link in those works. One way for destination to obtain the CSIs of S-R and R-D links has been proposed in [8], which made a strong assumption that the link condition of S-R was constant in a certain period and had been known by destination in fixed source and fixed relay networks. Another scheme in [9] requires the relay be equipped with a channel estimator and feed-forwarding a quantized version of the CSI estimation of S-R link to the destination node. It needs strong signal processing capability at relay node and will increase system overhead due to the existence of feed-forward channels.

In this paper, in order to obtain the CSIs of S-R and R-D links in OFDM-based AF relay networks, we propose a channel estimation scheme without the requirement of feed-forward channels. On the OFDM symbols used for channel estimation, the pilot signals sent from source are only located on given subcarriers, and some subcarriers reserved for relay pilot signals remain silent. The pilot signals inserted at relay node only undergo the channel fading of R-D link, and the CSI of R-D link could be obtained at first. Then, the CSI of S-R link is estimated based on the received pilot signals and the pre-estimated CSI of R-D link. In the CSI estimation of S-R link, it is verified that the CSI obtained by least square (LS) estimator is rather inaccurate for the channel estimation mean square error (MSE) is nonconvergent. In contrast, the MSE of minimum mean square error (MMSE) estimator is convergent and better performance can be obtained. Finally, simulation results are given to validate the analysis results.

The remainder of the paper is organized as follows. In section II, the system model and the proposed pilot pattern are described. Then, the proposed channel estimation algorithm and estimation error analysis are introduced in section III. In section IV, simulation results and discussions are given. Finally, we conclude this paper in section V.

II. System Model and Pilot Scheme

We consider a three-node AF relay network, each node equipped with one antenna. The relay node operates in the time division half-duplex mode that cannot receive and transmit simultaneously. In the first phase, the source node broadcasts information to both the relay and destination nodes. In the second phase, the source node keeps silent, while the relay node amplifies the signals received from the source node with a pre-defined amplification factor, and forwards them to the destination node.

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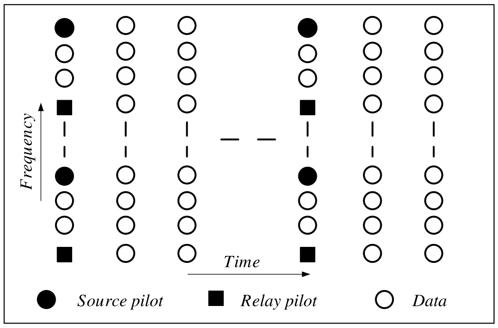

Fig. 1. An example of pilot scheme.

In this paper, we only focus on the transmission of pilot signals on the cascaded link S-R-D. In order to obtain the CSIs of S-R and R-D links respectively, some pilot signals are inserted at relay node. An example of pilot arrangement is shown in Fig.1. At source node, on the OFDM symbols used for channel estimation, a N_p-length pilot symbol is uniformly inserted into N subcarriers with the pilot space $L = N/N_p$, where N is the total number of subcarriers. Assume the pilot signals transmitted from source are $X_{s,i}$, with unit power $E[X_{s,i}X^*_{s,i}] = 1$. The index $i = mL + i_0$ $(m = 0, 1, \cdots, N_p - 1)$ means a source pilot signal is located on the ith subcarrier, where i_0 is the index of the first pilot subcarrier. The subcarriers reserved for relay pilot signals transmit nulls at source node.

The received pilot signals from the source at relay node can be expressed as

$$Y_{r,i} = H_{sr,i}\sqrt{P_s}X_{s,i} + n_{sr,i}, \tag{1}$$

where P_s is the transmission power at source, $H_{sr,i}$ is the flat fading coefficient at the ith subcarrier of S-R link, with $H_{sr,i} \sim \mathcal{CN}(0, \sigma^2_{sr})$, and $n_{sr,i}$ is the Gaussian white noise on the ith subcarrier, with $n_{sr,i} \sim \mathcal{CN}(0, N_{sr})$. We keep average power P_r transmitted from relay node, and the pre-defined amplification factor G is given by

$$G = \sqrt{\frac{P_r}{\sigma^2_{sr}P_s + N_{sr}}}. \tag{2}$$

The received information on the subcarriers reserved for relay pilot signals are noise. Abandon that and insert new pilot signals on these subcarriers. It is assumed the pilot inserted at relay node are $X_{r,j}$, with unit power $E[X_{r,j}X^*_{r,j}] = 1$. $j = mL+j_0$ $(m = 0, 1, \cdots, N_p-1)$ is the index of relay pilot subcarrier, where N_p is the relay pilots length, $L = N/N_p$ is the pilot space, and j_0 is the index of the first relay pilot subcarrier. In order to keep the transmitted power being P_r at relay node, the pilot signals $X_{r,j}$ are amplified by a factor $\sqrt{P_r}$. Therefore, the pilot signals transmitted from relay node are $\sqrt{P_r}X_{r,j}$.

At destination node, the received pilot signals from source and relay nodes are derived as

$$Y_{d,i} = H_{rd,i}H_{sr,i}G\sqrt{P_s}X_{s,i} + H_{rd,i}Gn_{sr,i} + n_{rd,i}, \tag{3}$$

$$Y_{d,j} = H_{rd,j}\sqrt{P_r}X_{r,j} + n_{rd,j}, \tag{4}$$

where $H_{rd,i}$ and $H_{rd,j}$ are the flat fading coefficient at the ith and the jth subcarriers of R-D link, with $H_{rd,i}, H_{rd,j} \sim \mathcal{CN}(0, \sigma^2_{rd})$, $n_{rd,i}$ and $n_{rd,j}$ are Gaussian white noise, with $n_{rd,i}, n_{rd,j} \sim \mathcal{CN}(0, N_{rd})$.

In this paper, it is assumed perfect synchronization at all nodes. The channel between any two nodes is quasi-stationary fading in that it is constant within one frame but may vary from frame to frame.

III. Channel Estimation

A. The CSI estimation of R-D link

The pilot signals transmitted from relay node only undergo the channel fading of R-D link. Thus, the CSI of R-D link can be obtained firstly. In the estimation algorithm design, take into account two prohibitive factors in realistic cases: 1) the real time correlation function of channel is hardly obtained and 2) high complexity algorithms should be avoided due to the hardware implementation difficulty.

The LS channel estimation of R-D link can be expressed as

$$\widetilde{H}_{rd,j} = \frac{Y_{d,j}}{\sqrt{P_r}X_{r,j}} = H_{rd,j} + \frac{n_{rd,j}}{\sqrt{P_r}X_{r,j}}. \tag{5}$$

The channel estimation MSE of LS estimation is given by

$$\widetilde{MSE}_{ls} = E\left[\left|H_{rd,j} - \widetilde{H}_{rd,j}\right|^2\right] = \frac{1}{SNR_{rd}}, \tag{6}$$

where $SNR_{rd} = P_r/N_{rd}$ is the signal-to-noise ratio (SNR) of R-D link. Then, the linear interpolation is adopted to estimate the CSIs of the other subcarriers by using the channel information at pilot subcarriers. The channel estimation at subcarrier k between two pilot subcarriers, $j < k < j + L$, using linear interpolation is given by

$$\begin{aligned}\widetilde{H}_{rd,k} &= \widetilde{H}_{rd,j+l} \qquad (0 < l < L)\\ &= \frac{l}{L}(\widetilde{H}_{rd,j+L} - \widetilde{H}_{rd,j}) + \widetilde{H}_{rd,j}.\end{aligned} \tag{7}$$

The MSE after linear interpolation has been derived in [10] as equation (8), where L is the pilot space, $R_f[l]$ is the frequency domain correlation of two subcarriers with space l, and $\Re$ denotes the real part of a complex number. The equation (8) shows that the MSE after linear interpolation is a function of pilot space, channel statistics and SNR.

The estimation by LS and linear interpolation is rough and always cannot meet the realistic requirements. In order to improve estimation accuracy, the noise could be constrained at time domain via IFFT operation. The significant channel taps at time domain are concentrated into a subregion [11]. By zeroing the terms out of this subregion that corresponds to noise, only the significant region is retained. More accurate CSI can be obtained by transforming the noise reduced signal back into the frequency domain via FFT operation. The step can be expressed as

$$\widehat{\mathbf{H}}_{rd} = \mathbf{F}\mathbf{A}\mathbf{F}^H\widetilde{\mathbf{H}}_{rd}, \tag{9}$$

where $\widetilde{\mathbf{H}}_{rd} = [\widetilde{H}_{rd,0}\ \widetilde{H}_{rd,1}\ \cdots\ \widetilde{H}_{rd,N-1}]^T$ is the CSI estimated after linear interpolation, $\mathbf{F}$ is the $N \times N$ Fourier

$$\underline{MSE} = \frac{1}{3}\left(5+\frac{1}{L}\right)\left(1+\frac{1}{L}\right)R_f[0] + \frac{1}{3}\left(2+\frac{1}{L^2}\right)\widetilde{MSE}_{ls} - \frac{4}{L}\sum_{l=0}^{L-1}\left(1-\frac{1}{L}\right)\Re\{R_f[l]\} + \frac{1}{3}\left(1-\frac{1}{L^2}\right)\Re\{R_f[L]\}, \quad (8)$$

transform matrix. Considering that it is difficult to know the accurate paths delay in realistic cases, we use the fact that the maximum paths delay is smaller than cyclic prefix (CP) length, and choose the CP length as the useful subregion. Furthermore, $\mathbf{A}$ can be written as

$$\mathbf{A} = \begin{bmatrix} \mathbf{I}_{L_{cp}} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} \end{bmatrix}, \quad (10)$$

where $\mathbf{I}_{L_{cp}}$ is the identity matrix, L_{cp} is the CP length. The processing is equivalent to keep the channel taps and the noise in the subregion, and the estimation MSE is given by

$$MSE_{rd} = \frac{L_{cp}}{N}\underline{MSE}, \quad (11)$$

where $\underline{MSE}$ is given by equation (8).

B. The CSI estimation of S-R link

Assume the CSIs of R-D link at the subcarriers which carry source pilot signals have been estimated as

$$\widehat{H}_{rd,i} = H_{rd,i} + e_{rd}, \quad (12)$$

where e_{rd} is the channel estimation error. For convenience, assume $e_{rd} \sim \mathcal{CN}(0, \sigma_e^2)$, $\sigma_e^2 = MSE_{rd}$, and e_{rd} is independent with $H_{rd,i}$. The distribution of $\widehat{H}_{rd,i}$ can be expressed as $\widehat{H}_{rd,i} \sim \mathcal{CN}(0, \widehat{\sigma}_{rd}^2)$, where $\widehat{\sigma}_{rd}^2 = \sigma_{rd}^2 + \sigma_e^2$. The CSI of S-R link is estimated based on the received pilot signals as equation (3) and the pre-estimated channel information of R-D link as equation (12).

1) LS estimation: The LS estimation of S-R link is given by (13), whose MSE can be calculated as

$$\begin{aligned}
\widetilde{MSE}_{ls} &= E\left[\left|H_{sr,i} - \widetilde{H}_{sr,i}\right|^2\right] \\
&= E\left[\left|\frac{H_{rd,i}Gn_{sr,i} + n_{rd,i} - e_{rd}H_{sr,i}G\sqrt{P_s}X_{s,i}}{\widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}}\right|^2\right] \\
&= E\left[\left|\frac{A}{B}\right|^2\right], \quad (14)
\end{aligned}$$

where $A = H_{rd,i}Gn_{sr,i} + n_{rd,i} - e_{rd}H_{sr,i}G\sqrt{P_s}X_{s,i}$ and $B = \widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}$. The zero mean Gaussian random variables $H_{rd,i}$, $H_{sr,i}$, $n_{sr,i}$, $n_{rd,i}$ and e_{rd} are independent with each other. It could be derived that random variable A is independent with random variable B. And the equation (14) can be expressed as

$$E\left[\left|\frac{A}{B}\right|^2\right] = E\left[|A|^2\right]E\left[\frac{1}{|B|^2}\right], \quad (15)$$

where

$$E\left[|A|^2\right] = \sigma_{rd}^2G^2N_{sr} + N_{rd} + \sigma_e^2\sigma_{sr}^2G^2P_s, \quad (16)$$

$$E\left[\frac{1}{|B|^2}\right] = \frac{1}{G^2P_s}E\left[\frac{1}{\left|\widehat{H}_{rd,i}\right|^2}\right]. \quad (17)$$

Define a variable $z = |\widehat{H}_{rd,i}|^2$, where the variable $\widehat{H}_{rd,i} \sim \mathcal{CN}(0, \widehat{\sigma}_{rd}^2)$. The power variable z is exponentially distributed, whose probability density function is given by

$$P(z) = \frac{1}{\widehat{\sigma}_{rd}^2}\exp\left(-\frac{z}{\widehat{\sigma}_{rd}^2}\right) \quad z \geq 0. \quad (18)$$

The expectation in equation (17) can be calculated as follows

$$\begin{aligned}
E\left[\frac{1}{\left|\widehat{H}_{rd,i}\right|^2}\right] &= E\left[\frac{1}{z}\right] = \int_0^{+\infty}\frac{1}{z}P(z)dz \\
&= \frac{1}{\widehat{\sigma}_{rd}^2}\int_0^{+\infty}\frac{1}{z}\exp\left(-\frac{z}{\widehat{\sigma}_{rd}^2}\right)dz \\
&= \frac{1}{\widehat{\sigma}_{rd}^2}\lim_{a\to 0^+}Ei_1\left(\frac{a}{\widehat{\sigma}_{rd}^2}\right) = +\infty, \quad (19)
\end{aligned}$$

where $Ei_1(x)$ is a kind of form of exponential integral function $Ei(x)$ and its definition is

$$Ei_1(x) = \int_1^{+\infty}\frac{e^{-xu}}{u}du \qquad (0 < x < +\infty). \quad (20)$$

From the previous mathematical deductions, we can see the MSE of LS estimation is nonconvergent. Consequently, The LS estimation of S-R link based on equation (3) and (12) is rather inaccurate.

Consider the ultimate situation that $SNR_{rd} \to +\infty$. It also means there is no noise on the R-D link and $n_{rd} = 0$. Based on the assumption, it is easily to obtain the perfect estimation of R-D link that is $\widehat{H}_{rd,i} = H_{rd,i}$. The received pilot signals given by equation (3) can be modified as

$$Y_{d,i} = H_{rd,i}H_{sr,i}G\sqrt{P_s}X_{s,i} + H_{rd,i}Gn_{sr,i}. \quad (21)$$

The LS estimation is given by

$$\widetilde{H}_{sr,i} = \frac{Y_{d,i}}{H_{rd,i}G\sqrt{P_s}X_{s,i}} = H_{sr,i} + \frac{n_{sr,i}}{\sqrt{P_s}X_{sr,i}}, \quad (22)$$

$$\widetilde{MSE}_{ls} = E\left[\left|\frac{n_{sr,i}}{\sqrt{P_s}X_{sr,i}}\right|^2\right] = \frac{1}{SNR_{sr}}, \quad (23)$$

where $SNR_{sr} = P_s/N_{sr}$ is the SNR of S-R link. The analysis result indicates that the LS estimation of S-R link goes to the LS estimation of general direct link when there is no noise on the R-D link.

$$\widetilde{H}_{sr,i} = \frac{Y_{d,i}}{\widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}} = \frac{H_{rd,i}H_{sr,i}G\sqrt{P_s}X_{s,i} + H_{rd,i}Gn_{sr,i} + n_{rd,i}}{\widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}} = H_{sr,i} + \frac{H_{rd,i}Gn_{sr,i} + n_{rd,i} - e_{rd}H_{sr,i}G\sqrt{P_s}X_{s,i}}{\widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}}. \tag{13}$$

2) MMSE estimation: To overcome the estimation problem brought by LS algorithm on S-R link, we propose the MMSE estimation that takes into account the noise statistics.

Assume the noise statistics $n_{sr,i} \sim \mathcal{CN}(0, N_{sr})$ and $n_{rd,i} \sim \mathcal{CN}(0, N_{rd})$ are pre-known at destination node. Rewrite the equation (3) as follows

$$\begin{aligned} Y_{d,i} &= H_{rd,i}H_{sr,i}G\sqrt{P_s}X_{s,i} + H_{rd,i}Gn_{sr,i} + n_{rd,i} \\ &= H_{sr,i}X + n, \end{aligned} \tag{24}$$

where $X = H_{rd,i}G\sqrt{P_s}X_{s,i}$ and $n = H_{rd,i}Gn_{sr,i} + n_{rd,i}$. Given the observation $Y_{d,i}$, the MMSE estimation of $H_{sr,i}$ is given by

$$\begin{aligned} \overline{H}_{sr,i} &= E\left[H_{sr,i}Y_{d,i}^*\right] E\left[Y_{d,i}Y_{d,i}^*\right]^{-1} Y_{d,i} \\ &= \frac{E\left[|H_{sr,i}|^2\right] X^*}{E\left[|H_{sr,i}|^2\right] |X|^2 + E\left[|n|^2\right]} Y_{d,i} \\ &= \frac{\sigma_{sr}^2 X^*}{\sigma_{sr}^2 |X|^2 + N} Y_{d,i}, \end{aligned} \tag{25}$$

where $N = E[|n|^2] = \sigma_{rd}^2 G^2 N_{sr} + N_{rd}$. In realistic cases, only the estimated value $\widehat{H}_{rd,i} = H_{rd,i} + e_{rd}$ can be obtained. Therefore, the equation (25) should be modified as

$$\overline{H}_{sr,i} = \frac{\sigma_{sr}^2 \widehat{X}^*}{\sigma_{sr}^2 \left|\widehat{X}\right|^2 + N} Y_{d,i}, \tag{26}$$

where $\widehat{X} = \widehat{H}_{rd,i}G\sqrt{P_s}X_{s,i}$. The MSE of MMSE estimation is given by equation (27), and

$$E\left[|C|^2\right] = \sigma_{sr}^2 \left[N^2 + W\left(W\sigma_e^2 + \widehat{\sigma}_{rd}^2 N\right)\right], \tag{28}$$

$$E\left[\left|\frac{1}{D}\right|^2\right] = E\left[\frac{1}{\left|\sigma_{sr}^2\left|\widehat{X}\right|^2 + N\right|^2}\right] = E\left[\frac{1}{|Wz+N|^2}\right], \tag{29}$$

where $W = \sigma_{sr}^2 G^2 P_s$ is a constant, $z = \left|\widehat{H}_{rd,i}\right|^2$ is exponentially distributed as shown in equation (18). The calculation of equation (29) is expressed as

$$\begin{aligned} E\left[\frac{1}{|D|^2}\right] &= \int_0^{+\infty} \frac{1}{|Wz+N|^2} \frac{1}{\widehat{\sigma}_{rd}^2} \exp\left(-\frac{z}{\widehat{\sigma}_{rd}^2}\right) dz \\ &= \frac{1}{\widehat{\sigma}_{rd}^2}\left(\frac{1}{WN} - \frac{\exp\left(\frac{N}{W\widehat{\sigma}_{rd}^2}\right) Ei_1\left(\frac{N}{W\widehat{\sigma}_{rd}^2}\right)}{W^2\widehat{\sigma}_{rd}^2}\right). \end{aligned} \tag{30}$$

The result shows the MSE of MMSE estimation is convergent and can be obtained by equation (28) and (30).

After estimating the channel information on the pilot subcarriers, the linear interpolation is adopted to obtain the CSIs of the others subcarriers as shown in R-D link estimation. Assume the CSIs of S-R link obtained by linear interpolation is $\overline{\mathbf{H}}_{sr} = [\overline{H}_{sr,0}\ \overline{H}_{sr,1}\ \cdots\ \overline{H}_{sr,N-1}]^T$. The estimation accuracy also can be improved in time domain via IFFT operation

$$\widehat{\mathbf{H}}_{sr} = \mathbf{F}\mathbf{A}\mathbf{F}^H\overline{\mathbf{H}}_{sr}, \tag{31}$$

where $\mathbf{F}$ is the $N \times N$ Fourier transform matrix, and $\mathbf{A}$ is given by equation (10). The estimation MSE is

$$MSE_{sr} = \frac{L_{cp}}{N}\underline{MSE}, \tag{32}$$

where $\underline{MSE}$ is the estimation error after linear interpolation that can be obtained by equation (8).

IV. Simulation Results and Discussions

In the simulations, the fading channels between S-R and R-D are Pedestrian B model with 6 taps at speed $3km/h$. The entire bandwidth of $5MHz$ is divided into 512 subcarriers, where 200 subcarriers are reserved as virtual subcarriers on both edges of the frequency spectrum. The subcarrier space is $\Delta f = 15kHz$ and the sampling rate is $7.68 \times 10^6 Hz$. The maximum path delay is located on the 28th subcarrier, and the coherence bandwidth can be calculated roughly

$$B_c = \frac{1}{\tau_{max}} = \frac{1}{\frac{28}{512}\frac{1}{\Delta f}} \approx 18\Delta f, \tag{33}$$

where τ_{max} is the maximum path delay. The CP length should be larger than $\tau_{max,sr} + \tau_{max,rd}$ in order to eliminate inter-symbol interference totally, and $CP = 64$ in the simulations. The channel power is normalized to 1, i.e, $\sigma_{sr}^2 = \sigma_{rd}^2 = 1$.

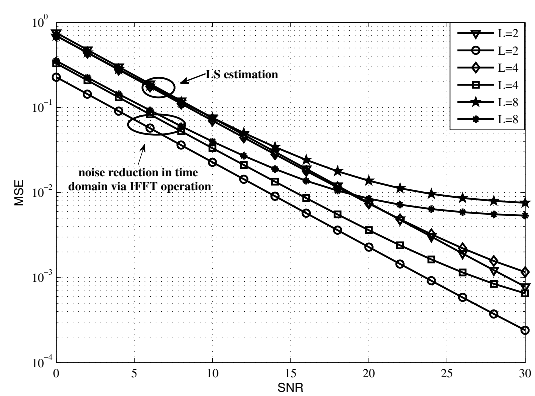

Fig. 2. Channel estimation MSE of R-D link.

$$\overline{MSE}_{mmse} = E\left[\left|H_{sr,i} - \overline{H}_{sr,i}\right|^2\right] = E\left[\left|\frac{H_{sr,i}N + \sigma_{sr}^2 G^2 P_s e_{rd} H_{sr,i} - \sigma_{sr}^2 \widehat{X}^* n}{\sigma_{sr}^2 \left|\widehat{X}\right|^2 + N}\right|^2\right] = E\left[\left|\frac{C}{D}\right|^2\right] = E\left[\left|C\right|^2\right] E\left[\left|\frac{1}{D}\right|^2\right], \quad (27)$$

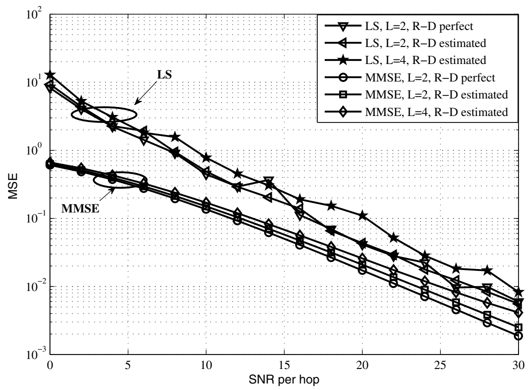

Fig. 3. Channel estimation MSE of S-R link.

And the SNR of S-R and R-D links are defined as follows: $SNR_{sr} = P_s/N_{sr}$, $SNR_{rd} = P_r/N_{rd}$.

Fig. 2 shows the MSE performances of R-D link with different pilot space, which is only influenced by SNR_{rd}. $L = K$ indicates the pilot space is K subcarriers. The ellipse with "LS estimation" means the channel is roughly estimated by LS and linear interpolation. The lines in the other ellipse show the MSE performances that the noise has been constrained in time domain via IFFT operation. As shown in the figure, the performance after noise constrained is much better than that of LS estimation. For instance, the MSE performance improvement is about 5 dB when the pilot space $L = 2$. With the pilot space increasing, The performance "floor" will appear when the SNR is high.

The MSE performance of S-R link is influenced by SNR_{sr} and SNR_{rd}. Fig. 3 shows the S-R link MSE performances comparison between LS and MMSE estimators. In the figure, "R-D perfect" and "R-D estimated" mean the CSIs of R-D link used for S-R link channel estimation are perfect and estimated by our method respectively. As the theoretical results that the MSE of LS estimator is nonconvergent, the MSE lines of LS estimator are not smooth though the CSI of R-D link is perfect. We nearly cannot obtain any channel information of S-R link in low SNR by LS estimator. For instance, when SNR pre hop is lower than 10 dB, the MSE of LS estimator is larger than channel normalized power 1. In contrast, the MSE lines of MMSE estimator are smooth. That is because the MMSE estimator considers the noise statistics and the MSE is convergent.

V. Conclusions

In this paper, a channel estimation scheme is proposed for destination to obtain the CSIs of S-R and R-D links in AF relay networks. The proposed scheme exploits pilot signals inserted at relay node to estimate the CSI of R-D link firstly. Then, the CSI of S-R link is obtained based on the received pilot signals and the pre-estimated CSI of R-D link. It has been verified in the paper that the S-R link MSE of LS estimator is nonconvergent and the CSI is hardly obtained in low SNR values. To solute the problem, we propose MMSE algorithm to estimate the CSI of S-R link, and the MSE of MMSE estimator also has been analyzed in the paper. The simulation results are given at the end of the paper to verify our analysis.

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