This is the accepted version of a paper presented at ISIT 2013 IEEE International Symposium on Information Theory; Istanbul, Turkey, 7-12 July 2013.

Citation for the original published paper:

Ergodic interference alignment with noisy channel state information.
http://dx.doi.org/10.1109/ISIT.2013.6620293

N.B. When citing this work, cite the original published paper.

Permanent link to this version:
http://urn.kb.se/resolve?urn=urn:nbn:se:kth:diva-121338
Ergodic Interference Alignment with Noisy Channel State Information

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Abstract—We investigate the time-varying Gaussian interference channel (IC) in which each source desires to communicate to an intended destination. For the ergodic time-varying IC with global perfect CSI at all terminals, it has been known that with an interference alignment technique each source-destination pair can communicate at half of the interference-free achievable rate. In practice, the channel gains are estimated by transmitting known pilot symbols from the sources, and the channel estimation procedure is hence prone to errors. In this paper, we model the channel estimation error at the destinations by an independent additive Gaussian noise and study the behavior of the ergodic interference alignment scheme with the global noisy CSI at all terminals. Toward this end, we present a closed-form inner bound on the achievable rate region by which we conclude that the achievable degrees of freedom with global perfect CSI can be preserved, if the variance of channel estimation error is proportional to the inverse of the transmitted power.

I. INTRODUCTION

Characterizing the capacity region of interference channels (ICs) has attracted much interest for decades, e.g. the two-user IC has been the subject of extensive research. Although certain achievable rate regions and outer bounds on the capacity region of the two-user IC have been proposed [1], [2], the exact capacity region is still unknown except for certain special cases (e.g. when interference is either very weak or strong [3], [4]). Furthermore, extension of the results on the two-user ICs to general $K$-user ICs is not straightforward. Recently, via a novel interference management technique referred to as interference alignment [5], [6], it has been shown that ICs may not be interference limited in high signal-to-noise ratio (SNR) regime. Through properly designing the transmitted signals, the received interference signals at each destination can be aligned such that they occupy only a sub-space of the received signal space. Consequently, a $K$-user time-varying (or frequency-selective) IC can achieve the sum-rate of \( \frac{K}{2} \log(\text{SNR}) + o(\log(\text{SNR})) \), where

\[ \lim_{\text{SNR} \to \infty} \frac{o(\log(\text{SNR}))}{\log(\text{SNR})} = 0 \]

This achievable sum-rate linearly scales with the number of users at high SNR and is substantially higher than that of the time-division-multiple-access (TDMA) scheme, which is only $\log(\text{SNR}) + o(\log(\text{SNR}))$. Furthermore, when the channel gains are ergodic time-varying and symmetrically distributed (e.g. Rayleigh fading channels), the ergodic interference alignment scheme has been developed in [7] which achieves the sum-rate of $\frac{K}{2} E[\log(1 + 2|h|^2 \text{SNR})]$. This sum-rate is achieved by exploiting the time variations of the channel and retransmission of the same symbols over properly chosen time slots. This implies that IC under time-varying channel environments is not interference limited at any SNR.

To achieve the performance promised by the aforementioned schemes, however, global channel state information (CSI) is assumed to be perfectly known at all the sources and destinations. Since acquiring such perfect CSI is a challenging problem, references [7]–[11] have investigated the cases in which each destination perfectly knows CSI of its incoming channel gains. Thus, it provides either the quantized version or the un-coded version of the channel gains to the other terminals through digital feedback signals or analog feedback signals, respectively. It has been shown that if the number of feedback bits of the digital feedback signals [7]–[10] or the transmission powers of the analog feedback signals [11] properly scale with transmission power, the outstanding performance of interference alignment is still achievable. Furthermore, it has been shown in reference [10] that even with a limited number of feedback bits, the throughput of applying interference alignment can still be larger than that achieved by TDMA.

Nevertheless, the channel estimation at each destination in general is not perfect. Thus, the available CSI at each terminal is subject to some estimation errors. Such errors can potentially degrade the performance of the network. To the best of our knowledge, it has been unknown how accurate the channel estimation is required to be to attain similar performance as

The research leading to these results has received funding from the Swedish Foundation for Strategic Research through RAMCOORAN project.
the one which is achievable by applying interference alignment based on perfect CSI. Thus, we investigate an achievable rate region of IC in this case, when the ergodic interference alignment scheme is applied. Our result shows that, when the variance of channel estimation error is fixed, an IC is interference limited at high SNR, i.e. simply increasing SNR would not improve the achievable sum-rate. It reveals that, however, if the variance of the channel estimation error is proportional to the inverse of transmission power, then the achievable degrees of freedom region is the same as the one when perfect CSI is available.

This paper is organized as follows. Section II describes the system model and the transmission scheme. An achievable rate region with noisy CSI is derived in Section III. Numerical evaluations are presented in Section V. Finally, Section VI concludes the paper.

II. TIME-VARYING IC WITH NOISY CSI

This paper considers a time-varying wireless IC composed of $K$ sources and $K$ destinations, as illustrated in Fig. 1. The sources and destinations are denoted by $S_k$ and $D_k$ ($k \in \{1, 2, \ldots, K\}$), respectively. The source $S_k$ intends to communicate to $D_k$ and choose message $m_k$ independently with uniform distribution from the set $\mathcal{M} = \{1, 2, \ldots, 2^nR_k\}$, where $R_k > 0$ is the code rate. It encodes its message $m_k$ into a length $n$ codeword $X_k^{(n)}$ which satisfies the power constraint

$$\frac{1}{n} \sum_{t=1}^{n} |X_k^t|^2 \leq P.$$  

Since all sources share the wireless transmission medium, at time slot $t$, the destination $D_k$ receives its desired message from the corresponding source $S_k$ over the direct link with corresponding channel gain $h_{kk}^t$, as well as interference signals transmitted from all other sources $S_l$ ($l \in \{1, 2, \ldots, K\}$, $l \neq k$) over interference links with channel gains $h_{kl}^t$. Therefore, the channel output at $D_k$ is

$$Y_k^t = h_{kk}^t X_k^t + \sum_{l=1, l \neq k}^{K} h_{kl}^t X_l^t + Z_k^t,$$

where $Z_k^t$ denotes a zero-mean additive white Gaussian noise (AWGN), i.e. $Z_k^t \sim \mathcal{CN}(0, 1)$. The channel gains are ergodic time-varying and have independent and identical distribution across different time slots. At time slot $t$, the channel gains are independently drawn from a complex Gaussian distribution, i.e. $h_{kl}^t \sim \mathcal{CN}(0, 1)$.

We assume that only imperfect estimation of global CSI is available at each terminal. To obtain this, at the beginning of each time slot, each destination estimates its incoming channels. This estimation is subject to some errors in general. Next, all destinations broadcast their estimations to all the other terminals through orthogonal feedback channels. For instance, each destination sends out the quantized versions of the estimated channel gains with sufficiently high quantization resolution. The feedback channels are assumed to be error-free.

The channel estimation is modeled according to

$$h_{kl}^t = \tilde{h}_{kl}^t + \varepsilon_{kl}^t, \quad \forall l, k \in \{1, 2, \ldots, K\},$$

where $\tilde{h}_{kl}^t$ and $\varepsilon_{kl}^t$ denote estimated channel gain and estimation error, respectively. We denote the estimated channel matrix of the considered network at time slot $t$ as $\tilde{H}^t$. All terminals only know $\tilde{H}^t$, thus, the estimation error can be interpreted as noise which degrades available CSI at terminals. The estimation error is independent of the estimated channel gain and we have $\varepsilon_{kk}^t \sim \mathcal{CN}(0, \sigma_{\varepsilon,1}^2)$ and $\varepsilon_{kl}^t \sim \mathcal{CN}(0, \sigma_{\varepsilon,II}^2)$ ($\forall l \neq k \in \{1, 2, \ldots, K\}$). Also, $\tilde{h}_{kk}^t \sim \mathcal{CN}(0, 1 - \sigma_{\varepsilon,1}^2)$ and $\tilde{h}_{kl}^t \sim \mathcal{CN}(0, 1 - \sigma_{\varepsilon,II}^2)$ ($\forall l \neq k \in \{1, 2, \ldots, K\}$). The parameters $\sigma_{\varepsilon,1}^2$ and $\sigma_{\varepsilon,II}^2$ indicate the variance of error for direct link estimation and interference links estimations, respectively. If $\sigma_{\varepsilon,1}^2, \sigma_{\varepsilon,II}^2 \rightarrow 0$, then the model reduces to the case that perfect CSI is available at terminals, and when $\sigma_{\varepsilon,1}^2, \sigma_{\varepsilon,II}^2 \rightarrow 1$ it represents that no CSI is available terminals. The parameters $\sigma_{\varepsilon,1}^2$ and $\sigma_{\varepsilon,II}^2$ can potentially have different values corresponding to different accuracy of channel estimation of the direct links and that of the interference links.

We apply the ergodic interference alignment transmission scheme proposed in [7], but now only the estimated channel gains are available at all terminals. Thus, if estimated channel gains at time slots $t$ and $t_p$ ($t_p > t$) satisfies $h_{kl}^t = h_{kl}^{t_p}$ and $\tilde{h}_{kl}^t = -\tilde{h}_{kl}^{t_p}$ ($\forall k \in \{1, 2, \ldots, K\}$, $k \neq l$), then $S_k$ at time $t_p$ retransmits the codeword which was transmitted at time $t$, i.e. $X_{k}^{t_p} = X_{k}^t$. To avoid measure zero event, this channel pairing can be performed based on quantized version of the estimated channel gain with sufficiently fine quantizer [7]. Therefore, $D_k$ receives the following signals at the corresponding time slots

$$Y_k^t = h_{kk}^t X_k^t + \sum_{l=1, l \neq k}^{K} h_{kl}^t X_l^t + Z_k^t,$$

$$Y_k^{t_p} = h_{kk}^t X_k^t + \sum_{l=1, l \neq k}^{K} h_{kl}^t X_l^t + Z_k^{t_p}.$$ 

The destination $D_k$ combines the received signals in (4) and (5) to obtain the following signal

$$\gamma_k^{t} = Y_k^{t} + Y_k^{t_p} = \left(2\tilde{h}_{kk}^t + \varepsilon_{kk}^t + \varepsilon_{kk}^{t_p} \right)X_k^t + \sum_{l=1, l \neq k}^{K} \left(\tilde{h}_{kl}^t + \varepsilon_{kl}^t \right)X_l^t + \left(Z_k^t + Z_k^{t_p}\right).$$

Next, it decodes the observed channel output $\{\gamma_k^{t}\}_{t=1}^{n/2}$ to an estimate $\hat{m}_k$ of the transmitted message.

**Definition 1:** A rate tuple $(R_1, R_2, \ldots, R_K)$ is achievable if for all $\epsilon > 0$ and sufficiently large code length $n$, channel encoding and decoding functions exist such that

$$R_k > R_k - \epsilon, \quad k \in \{1, 2, \ldots, K\},$$

$$\Pr\left(\bigcup_{k=1}^{K} \hat{m}_k \neq m_k\right) < \epsilon.$$
Theorem bounded as each destination has perfect knowledge of its incoming channel region in (8) is $R$ and $\mathbf{R}$.

Proposition 2: An inner bound on the achievable rate region in (8) is $R_k \geq R_k^C$ ($\forall k \in \{1, 2, ..., K\}$), where

$$R_k^C = \frac{1}{2} \mathbb{E} \left[ \log \left( 1 + \frac{2 \hat{h}_{kk}^2 P}{\sigma_{z,k}^2 + (K-1)\sigma_{\epsilon,H}^2 P} \right) \right] \quad (9)$$

Proof: The term $I \left( X_k; \hat{Y}_k | \mathbf{H} \right)$ in (8) can be lower bounded as

$$I \left( X_k; \hat{Y}_k | \mathbf{H} \right) \overset{(a)}{=} h \left( X_k | \mathbf{H} \right) - h \left( X_k | \mathbf{H}, \hat{Y}_k \right)$$
$$\overset{(b)}{=} h \left( X_k | \mathbf{H}, \hat{Y}_k \right) - h \left( X_k | \mathbf{H}, \mathbf{V}_k \right)$$
$$\overset{(c)}{=} h \left( X_k | \mathbf{H}, \mathbf{V}_k \right) - h \left( X_k - \hat{X}_k | \mathbf{H}, \mathbf{V}_k \right)$$
$$\overset{(d)}{=} \log 2\pi e P - h \left( X_k - \hat{X}_k | \mathbf{H}, \mathbf{V}_k \right)$$
$$\overset{(e)}{\geq} \log 2\pi e P - \log 2\pi e \sigma^2 \quad (10)$$

where $\sigma^2$ is the conditional variance of $\left( X_k - \hat{X}_k \right)$. In this equation (a) follows the definition of the conditional mutual information; (b) holds since the transmitted codeword is chosen independent of the noisy CSI; (c) follows the fact that $\hat{X}_k$ is a function of $\mathbf{H}$ and $\mathbf{V}_k$ which will be specified in the below; (d) follows the assumption that $X_k$ is a complex Gaussian random variable; (e) follows [12, Theorem 8.6.5] that shows the entropy of a random variable with given bounded variance is upper bounded by that of a random variable with Gaussian distribution. To obtain a tight bound on the achievable rate in (10), we choose $\hat{X}_k$ to be a minimum mean square error (MMSE) estimate of $X_k$; that is

$$\hat{X}_k = \frac{\mathbb{E} \left( X_k (\hat{Y}_k)^* | \mathbf{H}, \mathbf{V}_k \right)}{\mathbb{E} \left( (\hat{Y}_k)^* | \mathbf{H}, \mathbf{V}_k \right)} \mathbf{V}_k$$
$$= \frac{\left( \hat{h}_{kk} \right)^P \mathbf{V}_k}{1 + \left( \sigma_{z,1}^2 + (K-1)\sigma_{\epsilon,H}^2 + 2 \left| \hat{h}_{kk} \right|^2 \right) P} \mathbf{V}_k \quad (11)$$

which yields

$$\sigma^2 = \frac{P}{1 + \frac{2 |\hat{h}_{kk}|^2 P}{\sigma_{z,1}^2 + (K-1)\sigma_{\epsilon,H}^2 + 2 |\hat{h}_{kk}|^2}} \quad (12)$$

The details of the derivation of $\sigma^2$ are presented in Appendix A. The proof is completed by substituting (12) in (10). $\blacksquare$

We next characterize achievable degrees of freedom region with noisy CSI and present a sufficient condition on channel estimation error that preserves the achievable degrees of freedom region of interference alignment with global perfect CSI.

IV. ACHIEVABLE DEGREE OF FREEDOM REGION WITH NOISY CSI

The following corollary characterizes the achievable rate region at asymptotically high SNR region.

Corollary 1: When $\sigma_{z,1}^2$ and $\sigma_{\epsilon,H}^2$ are fixed, if $P \to \infty$, then an inner bound on the achievable rate region is

$$R_k^C = \frac{1}{2} \mathbb{E} \left[ \log \left( 1 + \frac{2 |\hat{h}_{kk}|^2}{\sigma_{z,1}^2 + (K-1)\sigma_{\epsilon,H}^2} \right) \right], \forall k \in \{1, ..., K\}. \quad (13)$$

Proof: The proof is completed by taking the limit of the lower bound on the achievable rates in (9) and using the monotone convergence theorem [13]. $\blacksquare$

This result implies that for fixed variances of channel estimation errors, the channel is basically interference limited, i.e. increasing SNR does not improve the achievable rates at high SNR. The following corollary, however, reveals that the achievable degrees of freedom with perfect CSI can be preserved, if the variance of channel estimation error properly decays as transmission power increases. First, we define achievable degree of freedom region.

Definition 2: A tuple $(d_1, d_2, ..., d_K)$ denotes achievable degrees of freedom in which $d_k = \lim_{P \to \infty} \frac{R_k}{\log P}$, where $R_k$ is an achievable rate.

Corollary 2: Assuming that only noisy CSI is available at all terminals, if the variance of channel estimation error is
proportional to $P^{-\alpha}$ ($\alpha \in \mathbb{R}$), then the degrees of freedom region $(d_1, d_2, ..., d_K)$ is achievable, where

$$d_k = \begin{cases} 0 & \alpha \leq 0 \\ \alpha/2 & 0 < \alpha < 1 \\ 1/2 & \alpha \geq 1 \end{cases}, \quad k \in \{1, 2, ..., K\}. \quad (14)$$

**Proof:** Assume that $\sigma_{\varepsilon,1}^2 \propto P^{-\alpha}$ and $\sigma_{\varepsilon,II}^2 \propto P^{-\alpha}$. This is corresponding to $\sigma_{\varepsilon,1}^2 = aP^{-\alpha}$ and $\sigma_{\varepsilon,II}^2 = bP^{-\alpha}$, where $a$ and $b$ are constant values. If $0 < \alpha < 1$, then we have

$$d_k = \lim_{P \to \infty} \frac{\frac{1}{2} \log P \left( 1 + \frac{2|h_{kk}|^2 P}{1+(a+(K-1)b)P^{-\alpha}} \right)}{\log P}$$

$$= \lim_{P \to \infty} \frac{\frac{1}{2} \log P^{\alpha+\frac{1}{2}} \log \left( \frac{P^{-\alpha} + \frac{2|h_{kk}|^2 P^{1-\alpha}}{1+(a+(K-1)b)P^{-\alpha}}} \right)}{\log P}$$

$$= \frac{\alpha}{2} + \frac{1}{2} \mathcal{E} \left[ \lim_{P \to \infty} \frac{\log \left( P^{-\alpha} + \frac{2|h_{kk}|^2 P^{1-\alpha}}{1+(a+(K-1)b)P^{-\alpha}} \right)}{\log P} \right]$$

$$= \frac{\alpha}{2} \quad (15)$$

where $(a)$ follows the dominated convergence theorem [13]. We can similarly prove the achievable degrees of freedom for $\alpha \geq 1$ and $\alpha \leq 0$.

**V. Numerical Evaluation**

This section presents numerical evaluations of the lower bound given in Proposition 2 on the achievable rate of the considered network.

Fig. 2 shows the lower bound on the achievable rate per user of $K$-user ICs versus SNR. We observe that when the variance of channel estimation error is fixed, the achievable rate monotonically increases and at high SNR saturates. This observation confirms that IC in this case is interference limited which coincides with Corollary 1. Also, we can see that the achievable rate monotonically decreases as the number of the users increases.

Fig. 3 illustrates the sum-rate of a three-user IC for different variances of channel estimation error. It can be observed that the sum-rate increases as the variance of channel estimation error decreases. Indeed, the achievable sum-rate with noisy CSI approaches the one with perfect CSI, if the variance of channel estimation error becomes sufficiently small. This can be exploited to design a minimal channel estimator for the network: at any SNR, we can find the minimum required accuracy of the channel estimation to attain a desired transmission rate.

Fig. 4 shows the sum-rate of a three-user IC when the variance of channel estimation error is equal to $P^{-\alpha}$, where $P$ is the transmit power and $\alpha > 0$. We can see that, at high SNR, the sum-rate linearly scales with the power. This observation coincides with Corollary 2. The sum-rate has different behavior for $0 < \alpha < 1$ and $1 \leq \alpha$ at high SNR; when $0 < \alpha < 1$, the slope of the sum-rate versus SNR curve increases linearly with $\alpha$, however, when $1 \leq \alpha$ the curves have similar slopes which are the same as the one with perfect CSI. Furthermore, when $1 \leq \alpha$ we can see a gap between the achievable sum-rate with noisy CSI and the one with perfect CSI at high SNR. This gap decays as $\alpha$ increases.

**VI. Conclusion**

We have investigated the achievable rate region of the Gaussian time-varying IC when only noisy estimations of the channel gains are available at all terminals. We have shown that when the variance of channel estimation error is fixed, IC is basically interference limited, i.e. increasing SNR would not improve the achievable rates at high SNR. However, if the variance of channel estimation error is proportional to $P^{-\alpha}$ ($\alpha \geq 0$), each user can achieve the degrees of freedom of
the estimation error of the MMSE estimator; where (12) can be derived as follows following substituting $\hat{X}_k$ given in (11); and (c) follows substituting $Y_k$ given in (6), and noting that $X_k$ is mutually independent of $Z_k^m$, $Z_k^-m$ and $X_l$ ($\forall l \in \{1, 2, \ldots, K\}, l \neq k$).

\begin{align}
\min\{\alpha/2, 1/2\}; \text{ indeed, if } \alpha = 1, \text{ then the achievable degrees of freedom are the same as those when global perfect CSI is available at all terminals. Therefore, channel estimation with certain accuracy is sufficient to attain the outstanding performance of the ergodic interference alignment scheme.}

\text{APPENDIX A}

\text{ THE PROOF OF PROPOSITION 2}

The variance in (12) can be derived as follows

\begin{align}
\sigma^2 &= \mathbb{E} \left[ (X_k - \hat{X}_k)(X_k - \hat{X}_k)^* \frac{1}{\mathbf{H}, Y_k} \right] \\
&= \mathbb{E} \left[ \left( X_k - \hat{X}_k, X_k - \hat{X}_k \right)^* \frac{1}{\mathbf{H}, Y_k} \right] \\
&= \mathbb{E} \left[ X_k \left( X_k - \hat{X}_k \right)^* \frac{1}{\mathbf{H}, Y_k} \right] - \mathbb{E} \left[ \left( X_k - \hat{X}_k \right)^* \frac{1}{\mathbf{H}, Y_k} X_k \right] \\
&= P - \mathbb{E} \left[ X_k \left( X_k - \hat{X}_k \right)^* \frac{1}{\mathbf{H}, Y_k} \right] \\
&= P - \mathbb{E} \left[ \frac{\hat{h}_{kk} X_k (Y_k)^*}{1 + \left( \sigma^2 + (K-1)\sigma^2_{\text{II}} + 2 \left( \hat{h}_{kk} \right)^2 \right) P} \frac{1}{\mathbf{H}, Y_k} \right] \\
&= P - \frac{2 \left( \hat{h}_{kk} \right)^2 P^2}{1 + \left( \sigma^2 + (K-1)\sigma^2_{\text{II}} + 2 \left( \hat{h}_{kk} \right)^2 \right) P} \\
&= \frac{P}{1 + \frac{2\left( \hat{h}_{kk} \right)^2 P}{1 + \left( \sigma^2 + (K-1)\sigma^2_{\text{II}} + 2 \left( \hat{h}_{kk} \right)^2 \right) P}} \\
\end{align}

where (a) follows the orthogonality of the estimated signal to the estimation error of the MMSE estimator; (b) follows the substitution of $\hat{X}_k$ given in (11); and (c) follows substituting $Y_k$ given in (6), and noting that $X_k$ is mutually independent of $Z_k^m$, $Z_k^-m$ and $X_l$ ($\forall l \in \{1, 2, \ldots, K\}, l \neq k$).

\text{REFERENCES}


Fig. 4: Achievable sum-rate in a three-user IC with noisy CSI, $\sigma^2_{\text{e,II}} = \sigma^2_{\text{e,s}} = P^{-\alpha}$. 

Achievable sum-rate [bits/channel use] 

SNR [dB]

SNR [dB]