Intelligent Reflecting Surface-Enhanced OFDM: Channel Estimation and Reflection Optimization

Authors: Beixiong Zheng, *Member*, *IEEE*, and Rui Zhang , *Fellow*, *IEEE*

Published in: *IEEE Wireless Communications Letters* (Volume: 9, Issue: 4, April 2020)

無線通訊系統概論 - 期末報告

學生: Q36114221 蘇沛錦

報告日期: 2022/12/19 (Mon.)

Outline

- Introduction
- System Model & Transmission Protocol
- Channel Estimation & Reflection Optimization
- Simulation Results
- Conclusion

Introduction

- Prior works on IRS mainly focus on the design of reflection coefficients under the assumption of perfect channel state information (CSI). (It is difficult to realize in practice)

- The joint design of practical channel estimation and reflection optimization under imperfect CSI tailored to the IRS-aided system is practically challenging due to massive number of passive elements without transmitting/receiving capabilities.

- It is more cost-effective to estimate the concatenated user-IRS-AP channels at the **AP** with properly designed IRS reflection pattern based on the received pilot signals sent by the user and reflected by the IRS.

Introduction

- The ON/OFF-based channel estimation method (**Prior works**) [1]
  - The direct channel is estimated with all sub-surfaces turned OFF and the reflecting link is estimated with one out of M sub-surfaces turned ON sequentially.

- **Main drawbacks** of this method for IRS channel estimation:
  1. This requires separate amplitude control (in addition to phase shift) of each IRS element.

     $\Longrightarrow$ It is practically costly to implement the ON/OFF switching of the massive IRS elements frequently.
  2. Only a small portion of IRS elements is switched ON at each time.

     $\Longrightarrow$ Degrade the channel estimation accuracy

[1] Y. Yang, B. Zheng, S. Zhang, and R. Zhang, "Intelligent reflecting surface meets OFDM: Protocol design and rate maximization," 2019. [Online]. Available: https://arxiv.org/abs/1906.09956.

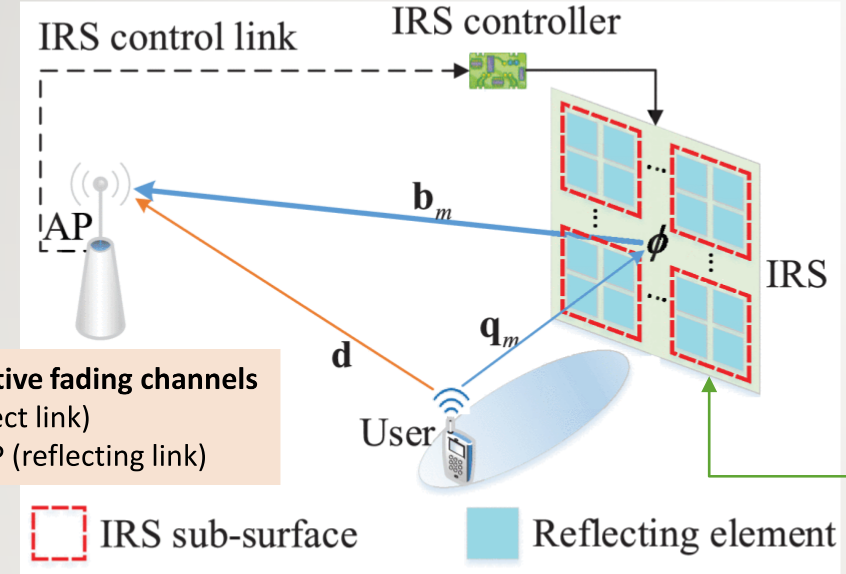
Introduction

- The authors consider a practical wideband IRS-enhanced OFDM system under frequency-selective fading channels, for which a practical transmission protocol is proposed to execute channel estimation and reflection optimization successively.

- A novel phase-shift pattern satisfying the unit-modulus constraint is designed for the IRS to facilitate the concatenated user-IRS-AP channel estimation at the **AP** based on the **uplink** pilot signals from the user. (All of IRS elements are switched ON with $|\phi_m| = 1$)

- Based on the estimated CSI, the reflection coefficients are then optimized to maximize the strongest time-domain path channel gain.

System Model



Frequency-selective fading channels
- user→AP (direct link)
- user→IRS→AP (reflecting link)

The IRS composed of *K* reflecting elements is divided into *M* sub-surfaces, e.g., $\bar{K} = \frac{K}{M} = 4$.

An illustration of the IRS-enhanced OFDM communication in the **uplink**.

- N sub-carriers
- Maximal delay spread L taps
- Received OFDM symbol $\mathbf{y} \triangleq [Y_0, Y_1, \ldots, Y_{N-1}]^T$

CP of length L_{cp} is assumed
$L_{cp} \geq L$ (Prevent ISI)

$$\mathbf{y} = \mathbf{X}\left(\sum_{m=1}^{M} \mathbf{q}_m \phi_m \odot \mathbf{b}_m + \mathbf{d}\right) + \mathbf{v} \quad (1)$$

$\mathbf{X} = \text{diag}(\mathbf{x})$ $\quad \mathbf{x} \triangleq [X_0, X_1, \ldots, X_{N-1}]^T$

$\mathbf{d} \triangleq [D_0, D_1, \ldots, D_{N-1}]^T \in \mathbb{C}^{N\times 1}$

$\mathbf{q}_m \in \mathbb{C}^{N\times 1}$ $\quad \mathbf{b}_m \in \mathbb{C}^{N\times 1}$

$\mathbf{v} \triangleq [V_0, V_1, \ldots, V_{N-1}]^T \sim \mathcal{N}_c(\mathbf{0}, \sigma^2 \mathbf{I}_N)$ (AWGN vector)

$$\phi_m = \beta_m e^{j\varphi_m}, \quad m = 1, \ldots, M \quad (2)$$

$\beta_m \in [0, 1]$ $\quad \varphi_m \in (0, 2\pi]$

System Model

- To maximize the reflection power of the IRS and simplify its hardware design, the authors fix $\beta_m = 1, \forall m = 1, \ldots, M$ and only adjust the phase shift φ_m for both channel estimation and reflection optimization in this letter.

$$\mathbf{g}_m \triangleq [G_{m,0}, G_{m,1}, \ldots, G_{m,N-1}]^T = \mathbf{q}_m \odot \mathbf{b}_m$$

$$(1) \quad \Longrightarrow \quad \mathbf{y} = \mathbf{X}\left(\sum_{m=1}^{M} \phi_m \mathbf{g}_m + \mathbf{d}\right) + \mathbf{v} \quad (3)$$

$$\mathbf{G} = [\mathbf{g}_1, \mathbf{g}_2, \ldots, \mathbf{g}_M]$$

$$\boldsymbol{\phi} \triangleq [\phi_1, \phi_2, \ldots, \phi_M]^T$$ (phase-shift vector)

$$\Longrightarrow \quad \mathbf{y} = \mathbf{X}\underbrace{(\mathbf{G}\boldsymbol{\phi} + \mathbf{d})}_{\mathbf{h}} + \mathbf{v} \quad (4)$$

- The superimposed channel frequency response (CFR) of the direct link and the reflecting link

$$\mathbf{h} = [H_0, H_1, \ldots, H_{N-1}]^T$$

Transmission Protocol

- Based on the $(M+1)$ consecutive pilot symbols and their pre-designed reflection states, i.e., $\{\mathbf{X}^{(i)}, \boldsymbol{\phi}^{(i)}\}_{i=0}^{M}$, the AP can estimate the CSI of $\mathbf{G}$ and $\mathbf{d}$.
- With the estimated CSI of $\mathbf{G}$ and $\mathbf{d}$, we optimize the IRS phase-shift vector $\boldsymbol{\phi}$ for data transmission in the second sub-frame.

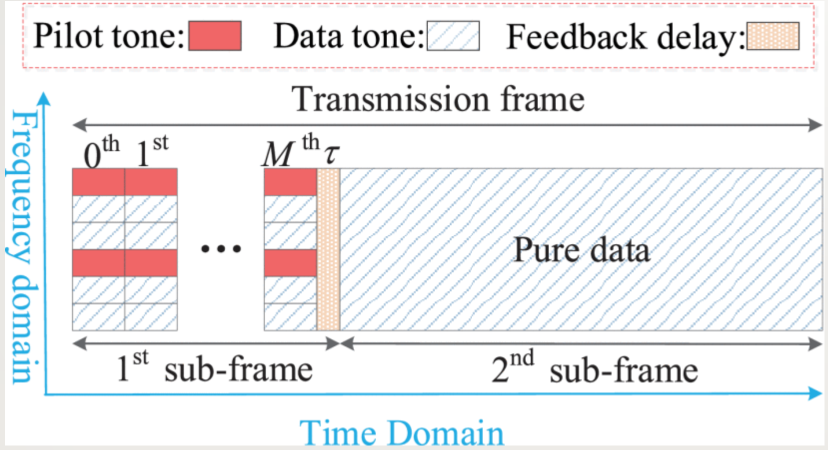
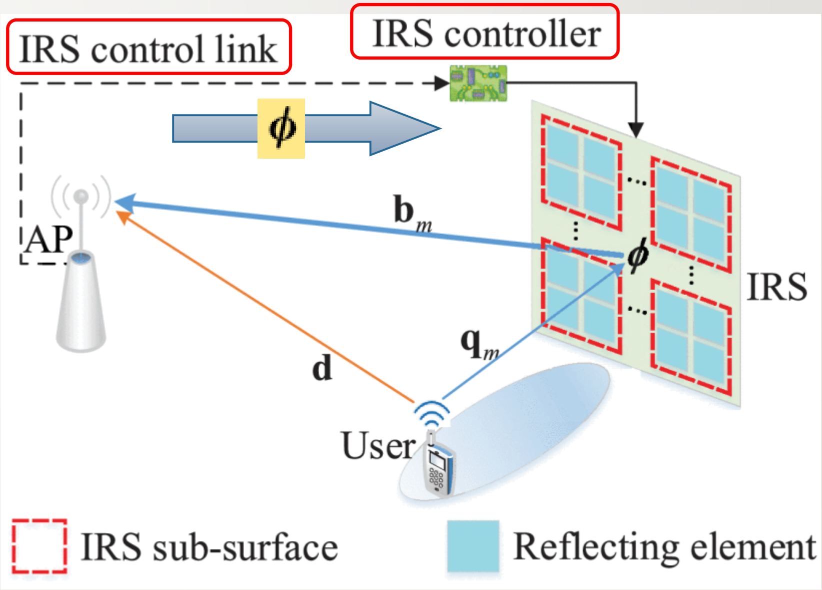


Illustration of the proposed transmission protocol.



Channel Estimation

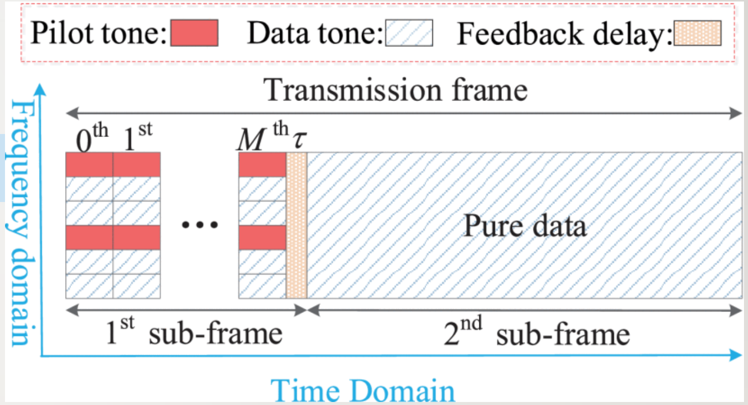
- Least-square estimation of CFRs on the pilot tones $\mathcal{P}$:

$$\mathbf{r} \triangleq \left[R_0, R_1, \ldots, R_{N_p-1}\right]^T$$
$$= \mathbf{X}_{\mathcal{P}}^{-1}\mathbf{y}_{\mathcal{P}} = \mathbf{h}_{\mathcal{P}} + \mathbf{X}_{\mathcal{P}}^{-1}\mathbf{v}_{\mathcal{P}} \qquad (8)$$

N_p pilot tones

$$\mathbf{X}_{\mathcal{P}} = \mathrm{diag}(\mathbf{x}_{\mathcal{P}})$$

$\mathbf{x}_{\mathcal{P}}$ (pilot sequence)



(comb-type pilot scheme)

- N_p-point **IDFT** of $\mathbf{r}$:

$$\tilde{\mathbf{r}} \triangleq \left[r_0, r_1, \ldots, r_{N_p-1}\right]^T$$

Time-Domain

$$\hat{\mathbf{r}} = \sqrt{\frac{N}{N_p}}[\tilde{\mathbf{r}}]_{1:L}$$

(the estimate of the superimposed channel impulse response)

- N-point **DFT** on $\hat{\mathbf{r}}$:

Padded with $(N - L)$ zeros

$$\hat{\mathbf{h}} = \sqrt{\frac{1}{N}}\mathbf{F}_N\left[\hat{\mathbf{r}}^T, \mathbf{0}_{1\times(N-L)}\right]^T = \mathbf{h} + \bar{\mathbf{v}} \qquad (9)$$

$\mathbf{F}_N$ ($N \times N$ **DFT** matrix with $[\mathbf{F}_N]_{\imath,\jmath} = e^{-j\frac{2\pi\imath\jmath}{N}}$ for $0 \leq \imath,\jmath \leq N-1$)

$\bar{\mathbf{v}}$ (noise vector distributed as $\mathcal{N}_c(\mathbf{0}, \mathbf{F}_N\tilde{\mathbf{I}}\mathbf{F}_N^H)$ with $\tilde{\mathbf{I}} = \begin{bmatrix} \frac{\sigma^2 N}{N_p P_t}\mathbf{I}_L & \mathbf{0}_{L\times(N-L)} \\ \mathbf{0}_{(N-L)\times L} & \mathbf{0}_{(N-L)\times(N-L)} \end{bmatrix}$)

Channel Estimation

- Estimation of the superimposed CFR during the transmission of the i-th pilot symbol

$$\hat{\mathbf{h}}^{(i)} = \mathbf{G}\boldsymbol{\phi}^{(i)} + \mathbf{d} + \bar{\mathbf{v}}^{(i)} = \vec{\mathbf{G}}\vec{\boldsymbol{\phi}}^{(i)} + \bar{\mathbf{v}}^{(i)} \qquad (11)$$

$\boldsymbol{\phi}^{(i)}$ (pre-designed IRS reflection state) $\quad \vec{\mathbf{G}} = [\mathbf{d}, \mathbf{G}] \quad \vec{\boldsymbol{\phi}}^{(i)} = \begin{bmatrix} 1 \\ \boldsymbol{\phi}^{(i)} \end{bmatrix}$

- By stacking $\hat{\mathbf{h}}^{(i)}$ with $i = 0, 1, \ldots, M$ into $\hat{\mathbf{H}} = [\hat{\mathbf{h}}^{(0)}, \hat{\mathbf{h}}^{(1)}, \ldots, \hat{\mathbf{h}}^{(M)}]$, we can obtain

$$\hat{\mathbf{H}} = \vec{\mathbf{G}}\boldsymbol{\Theta} + \bar{\mathbf{V}} \qquad (12)$$

$\boldsymbol{\Theta} = [\vec{\boldsymbol{\phi}}^{(0)}, \vec{\boldsymbol{\phi}}^{(1)}, \ldots, \vec{\boldsymbol{\phi}}^{(M)}]$ (IRS reflection pattern matrix) $\quad \bar{\mathbf{V}} = [\bar{\mathbf{v}}^{(0)}, \bar{\mathbf{v}}^{(1)}, \ldots, \bar{\mathbf{v}}^{(M)}]$ (noise matrix)

- Base on (12), the CSI of $\mathbf{G}$ and $\mathbf{d}$ is estimated as

$$\left[\hat{\mathbf{d}}\ \hat{\mathbf{G}}\right] = \hat{\mathbf{H}}\boldsymbol{\Theta}^{-1}. \qquad (13)$$

Channel Estimation

- Generally, $\boldsymbol{\Theta}$ is a **full-rank** matrix, i.e., $\mathrm{rank}(\boldsymbol{\Theta}) = M+1$ under the reflection amplitude constraint of $|\phi_m^{(i)}| = 1$ ($\forall m = 1,\ldots,M$ and $\forall i = 0,1,\ldots,M$).
- The mean square error (MSE) of channel estimation on a sub-carrier basis:

$$
\begin{aligned}
\varepsilon &= \frac{1}{N}\cdot\mathbb{E}\left\{\left\|\left[\hat{\mathbf{d}}\ \hat{\mathbf{G}}\right]-\left[\mathbf{d}\ \mathbf{G}\right]\right\|_F^2\right\}\\
&= \frac{1}{N}\cdot\mathbb{E}\left\{\left\|\bar{\mathbf{V}}\boldsymbol{\Theta}^{-1}\right\|_F^2\right\}\\
&= \frac{1}{N}\cdot\mathrm{tr}\left\{\left(\boldsymbol{\Theta}^{-1}\right)^H\mathbb{E}\left\{\bar{\mathbf{V}}^H\bar{\mathbf{V}}\right\}\boldsymbol{\Theta}^{-1}\right\}\\
&\overset{(a)}{=} \frac{\sigma^2 NL}{N_pP_t}\cdot\mathrm{tr}\left\{\left(\boldsymbol{\Theta}^H\boldsymbol{\Theta}\right)^{-1}\right\}. \qquad (14)
\end{aligned}
$$

The equality of (a) holds as $\mathbb{E}\{\bar{\mathbf{V}}^H\bar{\mathbf{V}}\} = \frac{\sigma^2N^2L}{N_pP_t}\mathbf{I}_{M+1}$

To minimize the variance of the channel estimation error, the matrix $\boldsymbol{\Theta}$ is required to satisfy $\boldsymbol{\Theta}^H\boldsymbol{\Theta} = (M+1)\mathbf{I}_{M+1}$.

$\Longrightarrow$ $\boldsymbol{\Theta}$ is an **orthogonal matrix**.

P_t (the total transmission power at the user)

Reflection Optimization

- Based on the estimated CSI of $\mathbf{G}$ and $\mathbf{d}$, we aim to optimize $\boldsymbol{\phi}$ for maximizing the average achievable rate in (5) subject to the IRS reflection amplitude constraint.

$$\text{(P1)}: \max_{\boldsymbol{\phi}} \quad C(\boldsymbol{\phi}) = \frac{1}{N+L_{cp}} \sum_{n=0}^{N-1} \log_2\left(1+\frac{P_t \hat{W}_n(\boldsymbol{\phi})}{N\Gamma\sigma^2}\right) \quad (5)$$

$$\text{s.t.} \quad |\phi_m| = 1, \quad \forall m = 1, \ldots, M \quad (6)$$

$$\hat{W}_n(\boldsymbol{\phi}) = \left|\sum_{m=1}^{M} \phi_m \hat{G}_{m,n} + \hat{D}_n\right|^2, \quad n = 0, \ldots, N-1$$

(the estimated channel gain of the n-th subcarrier)

$$\Gamma \geq 1$$

(the achievable rate gap due to a practical modulation and coding scheme)

- Problem (P1) is non-convex and thus difficult to solve optimally.

Reflection Optimization

- Alternatively, we consider to maximize the rate **upper bound** of (5)

Based on the Jensen's inequality

$$(5) \Longrightarrow \quad C(\boldsymbol{\phi}) \leq \frac{N}{N+L_{cp}} \log_2\left(1 + \frac{1}{N}\sum_{n=0}^{N-1} \frac{P_t \hat{W}_n(\boldsymbol{\phi})}{N\Gamma\sigma^2}\right) \qquad (16)$$

and formulate the following optimization problem

$$(\text{P2}): \max_{\boldsymbol{\phi}} \sum_{n=0}^{N-1} \left| \sum_{m=1}^{M} \phi_m \hat{G}_{m,n} + \hat{D}_n \right|^2 \qquad (17)$$

$$\text{s.t.} \quad |\phi_m| = 1, \quad \forall m = 1, \ldots, M \qquad (18)$$

- Although the **semidefinite relaxation (SDR)** method in [1] can be applied to solve problem (P2) sub-optimally. The SDR method has a complexity order of $\mathcal{O}((M+1)^6)$, which is practically costly for large values of M.

[1] Y. Yang, B. Zheng, S. Zhang, and R. Zhang, "Intelligent reflecting surface meets OFDM: Protocol design and rate maximization," 2019. [Online]. Available: https://arxiv.org/abs/1906.09956.

Reflection Optimization

- The authors propose the strongest-CIR maximization (SCM) method to solve problem (P2) suboptimally by exploiting the time domain property.

➢ channel impulse response (CIR)

Based on the Parseval's theorem

$$(17) \quad \Longrightarrow \quad \sum_{l=0}^{L-1}\left|\sum_{m=1}^{M} \phi_m \hat{g}_{m,l} + \hat{d}_l\right|^2 \qquad (19)$$

Time-Domain

$\hat{g}_{m,l}$ (the l-th tap of the estimated CIR for the cascaded reflecting link)

$\hat{d}_l$ (the l-th tap of the estimated CIR for the direct link)

$L \leq L_{cp} \ll N$
(in typical wireless environment)

① Find the strongest CIR gain with respect to the tap index l

$$\check{l} = \arg \max_{l \in \{0,\ldots,L-1\}} \left|\sum_{m=1}^{M} |\hat{g}_{m,l}| + |\hat{d}_l|\right|^2 \qquad (20)$$

② Align the reflection phase shifts to the strongest CIR as

$$\check{\varphi}_m = -\angle \hat{g}_{m,\check{l}} + \angle \hat{d}_{\check{l}}, \qquad m = 1, \ldots, M \qquad (21)$$

Simulation Settings

- A uniform square array for the IRS ($K = 12 \times 12 = 144$ reflecting elements with half wavelength spacing)
- Path loss exponents

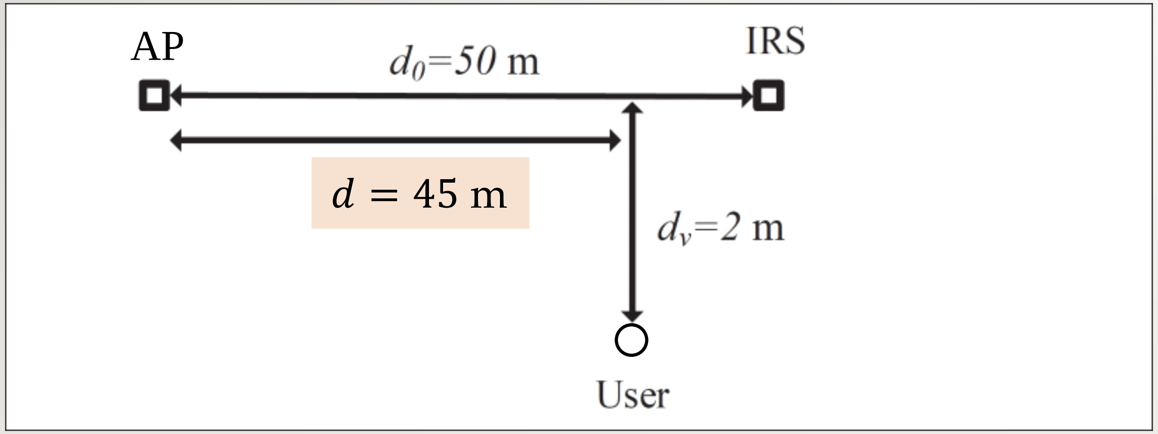
user → AP	user → IRS	IRS → AP
3.5	2.4	2.2

- The path loss at the reference distance of 1 meter (m) is set as 30 dB for each individual link.
- System parameters
  - 150 OFDM symbols in one transmission frame
  - $N = 64$ sub-carriers
  - a CP of length $L_{cp} = 8$
  - $\Gamma = 9$ dB, $\sigma^2 = -80$ dBm
- Frequency-selective Rician fading channels ($L = 6$ taps for both direct link & reflecting link)
  - ✓ The first tap is set as the deterministic **line-of-sight (LoS)** component.
  - ✓ The remaining taps are non-LoS components following the Rayleigh fading distribution.

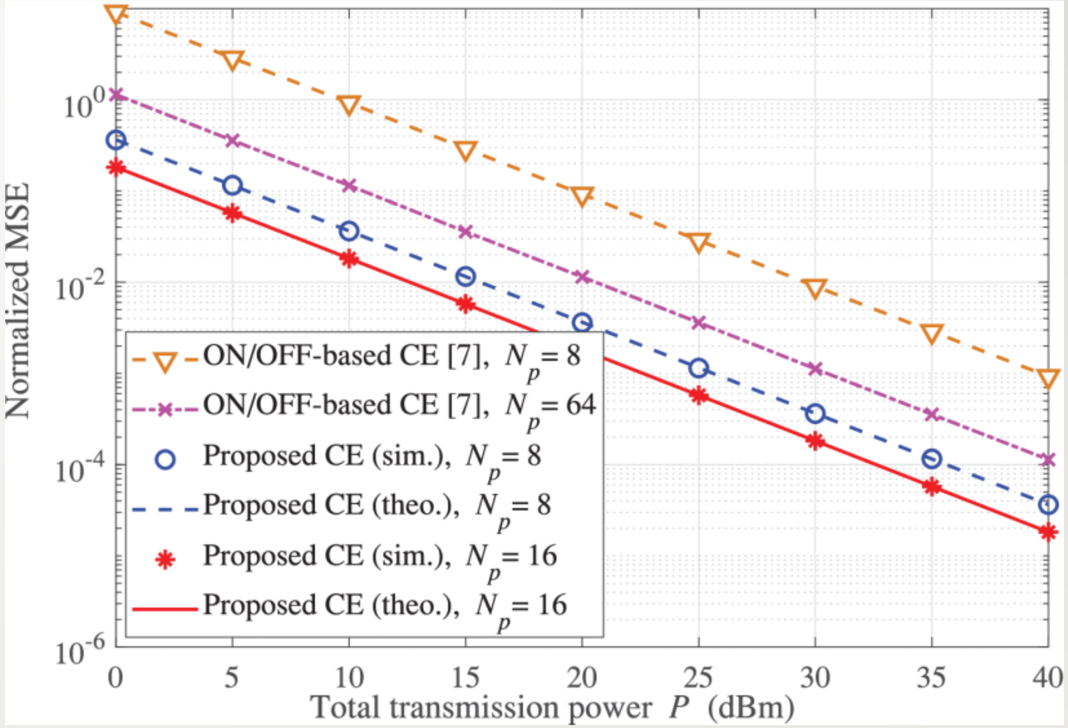
η (the ratio of the total power of non-LoS components to that of LoS component)

Simulation Results

$M = 12$
$\eta = 0.5$



Simulation setup. (Reproduced from [2])

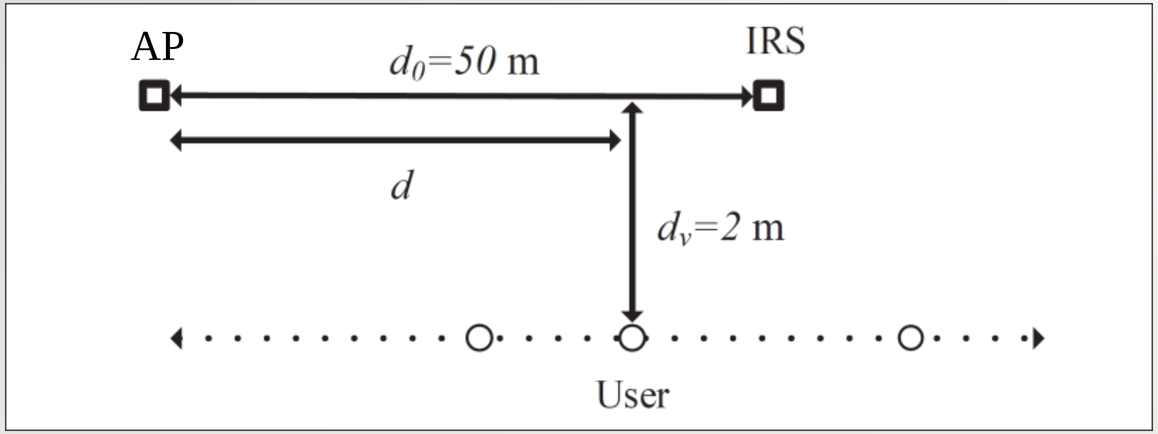


Normalized MSE versus transmit power P_t .

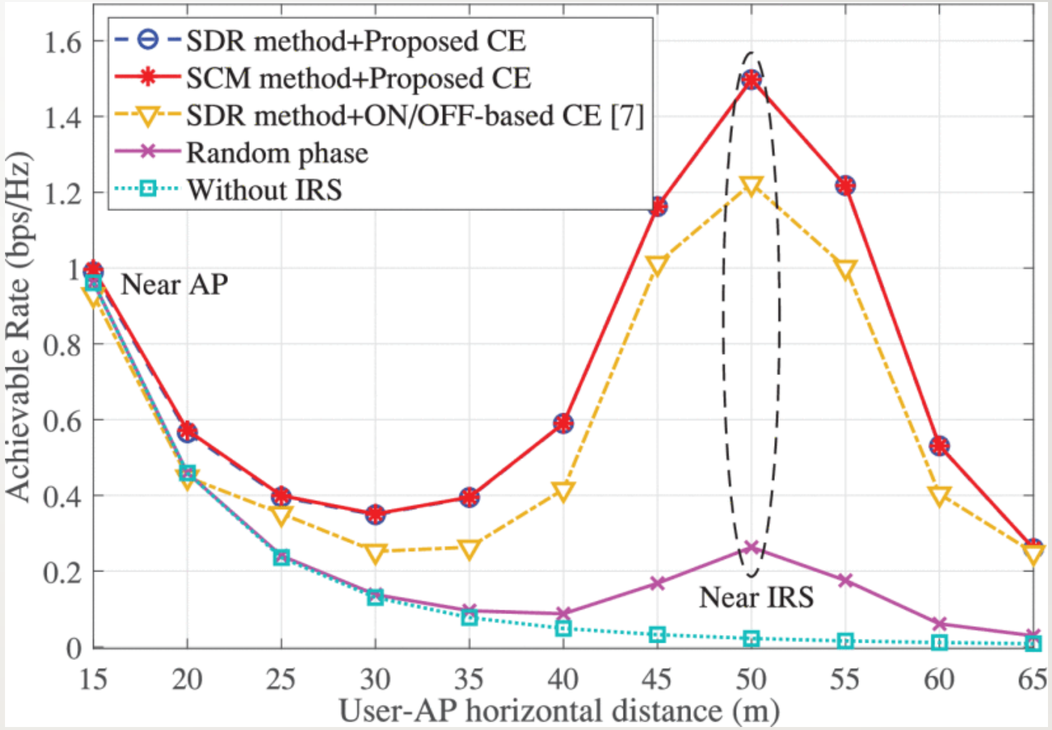
[2] Q. Wu and R. Zhang, “Towards smart and reconfigurable environment: Intelligent reflecting surface aided wireless network,” IEEE Comm. Mag., vol. 58, no. 1, pp. 106-112, January 2020.

Simulation Results

$$M = 12$$
$$N_p = 64$$
$$\eta = 0.5$$
$$P_t = 0 \text{ dBm}$$



Simulation setup. (Reproduced from [2])

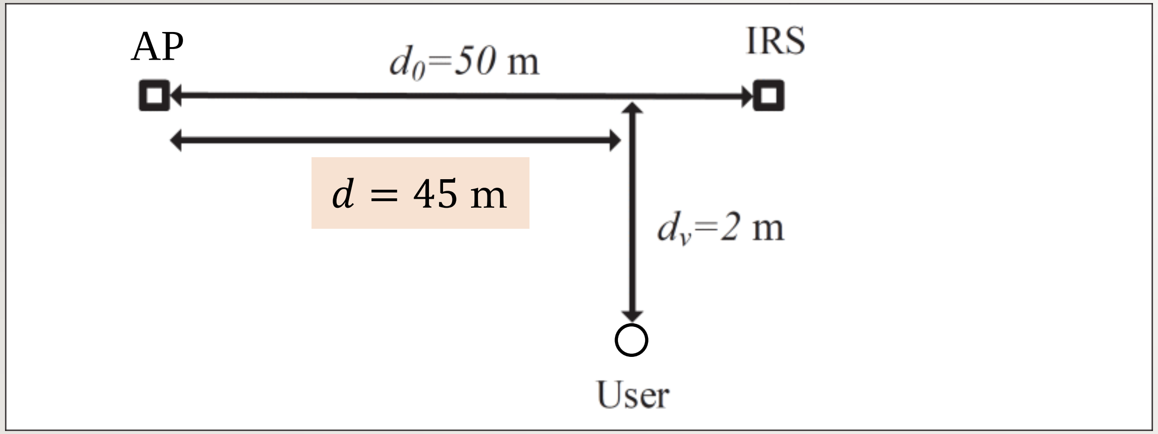


Achievable rate versus user-AP horizontal distance d .

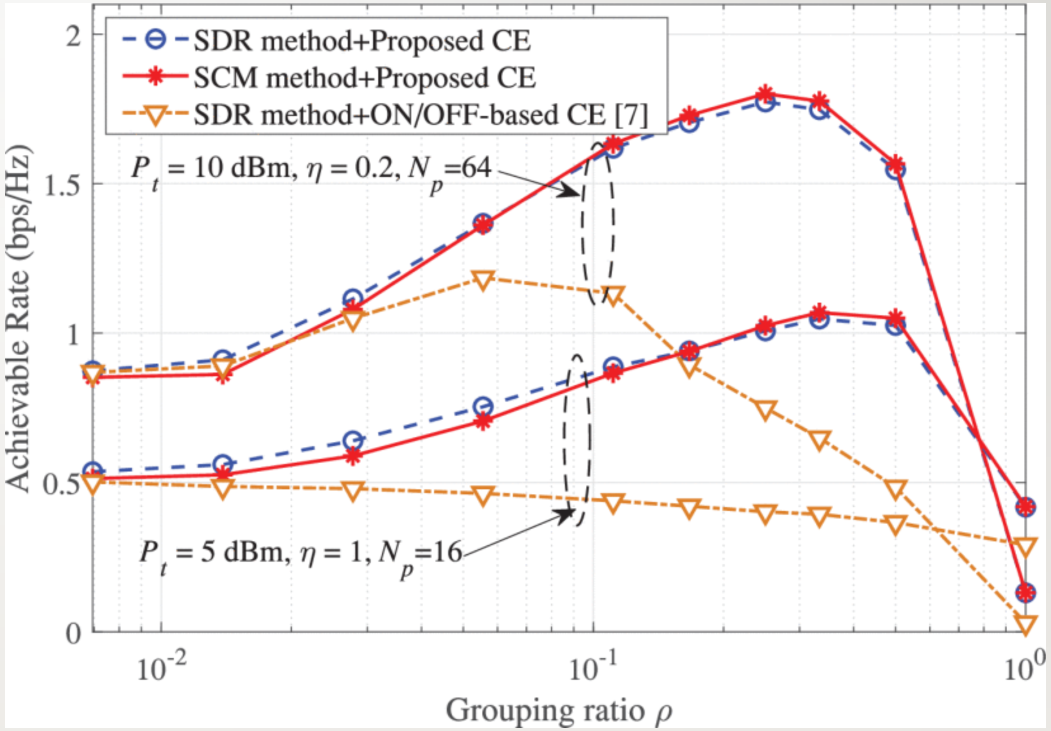
[2] Q. Wu and R. Zhang, "Towards smart and reconfigurable environment: Intelligent reflecting surface aided wireless network," IEEE Comm. Mag., vol. 58, no. 1, pp. 106-112, January 2020.

Simulation Results

$\rho \triangleq M/K$ (IRS grouping ratio)



Simulation setup. (Reproduced from [2])



Achievable rate versus IRS grouping ratio ρ .

[2] Q. Wu and R. Zhang, "Towards smart and reconfigurable environment: Intelligent reflecting surface aided wireless network," IEEE Comm. Mag., vol. 58, no. 1, pp. 106-112, January 2020.

Conclusion

- In this letter, authors have proposed a practical transmission protocol to execute channel estimation and reflection optimization for the IRS-enhanced OFDM system.

- Under the unit-modulus constraint, a novel reflection pattern is designed for channel estimation and optimized the reflection coefficients with a low-complexity SCM method.

- Simulation results have verified the superior performance of the proposed channel estimation and reflection optimization methods over the existing schemes.

References

1. Y. Yang, B. Zheng, S. Zhang, and R. Zhang, “Intelligent reflecting surface meets OFDM: Protocol design and rate maximization,” 2019. [Online]. Available: https://arxiv.org/abs/1906.09956.
2. Q. Wu and R. Zhang, “Towards smart and reconfigurable environment: Intelligent reflecting surface aided wireless network,” *IEEE Comm. Mag.*, vol. 58, no. 1, pp. 106-112, January 2020.

Thank you for listening.