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Intelligent Reflecting Surfaces for Wireless Communications:

Fundamentals, Designs, and Open Issues

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Outline of Part 1

Introduction to intelligent reflecting surface (IRS)

- Active and passive beamforming

System modeling and optimization

- Derive system model, optimization of IRS

Narrowband channel modeling

- Pathloss, array responses, correlation

Channel estimation

Wideband modeling and optimization

Competing technologies and weaknesses

My research is sponsored by

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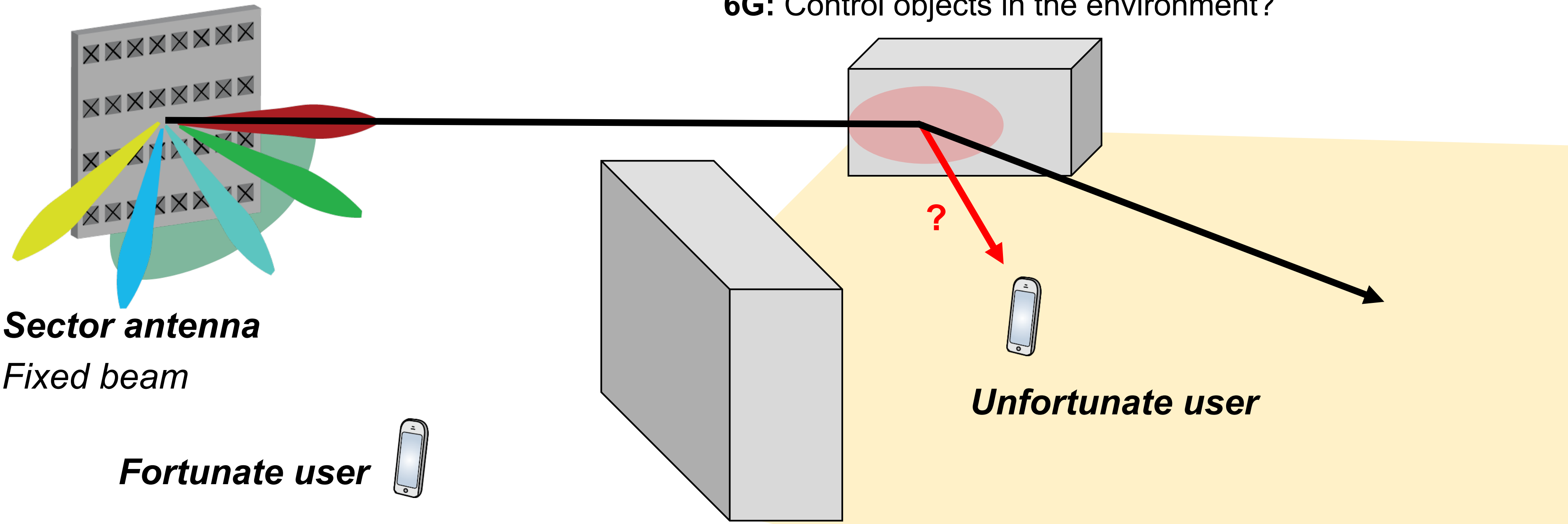
Project Grant

INTRODUCTION

Evolution of Wireless Infrastructure

5G: Adaptive multi-user beamforming

6G: Control objects in the environment?

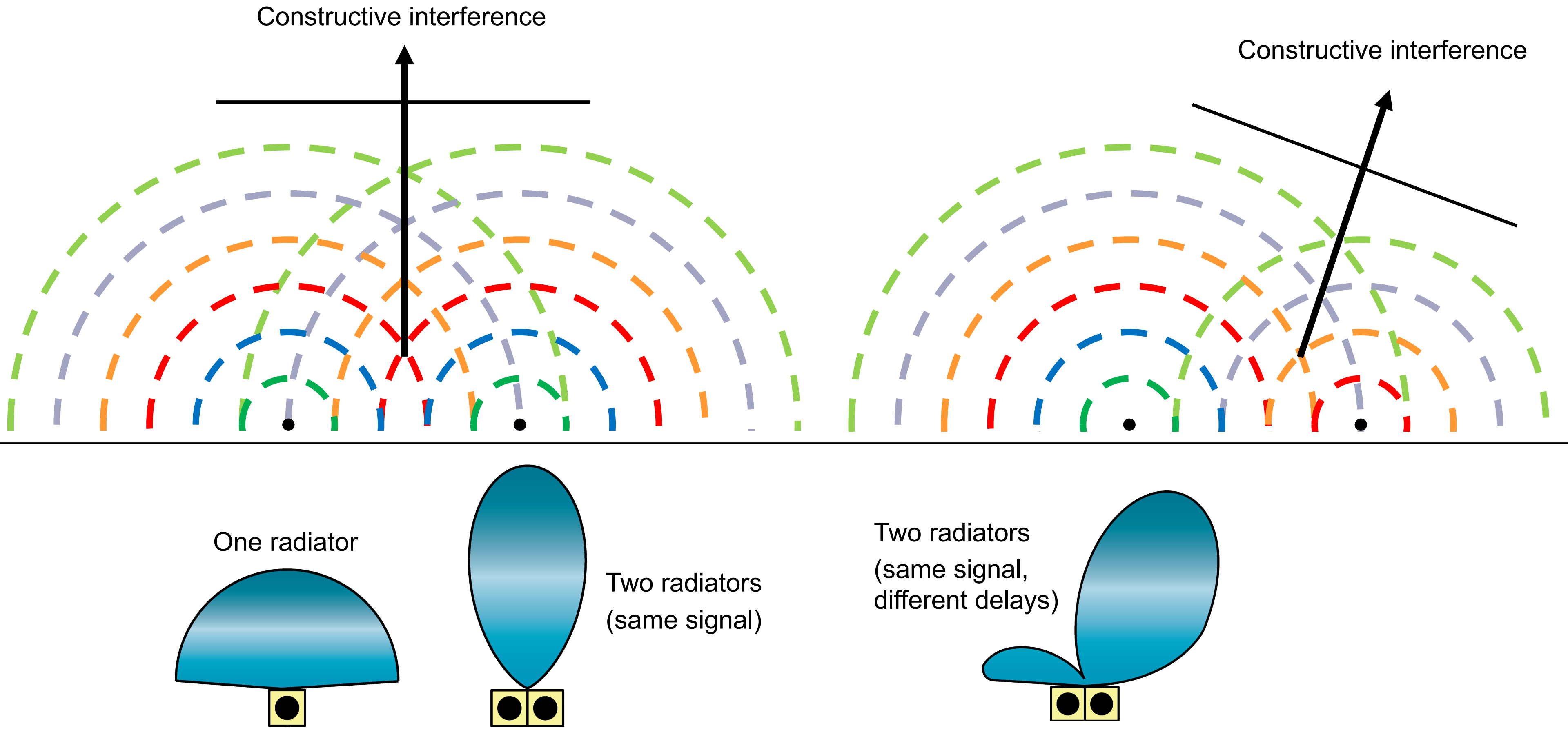


Reconfigurable metasurfaces

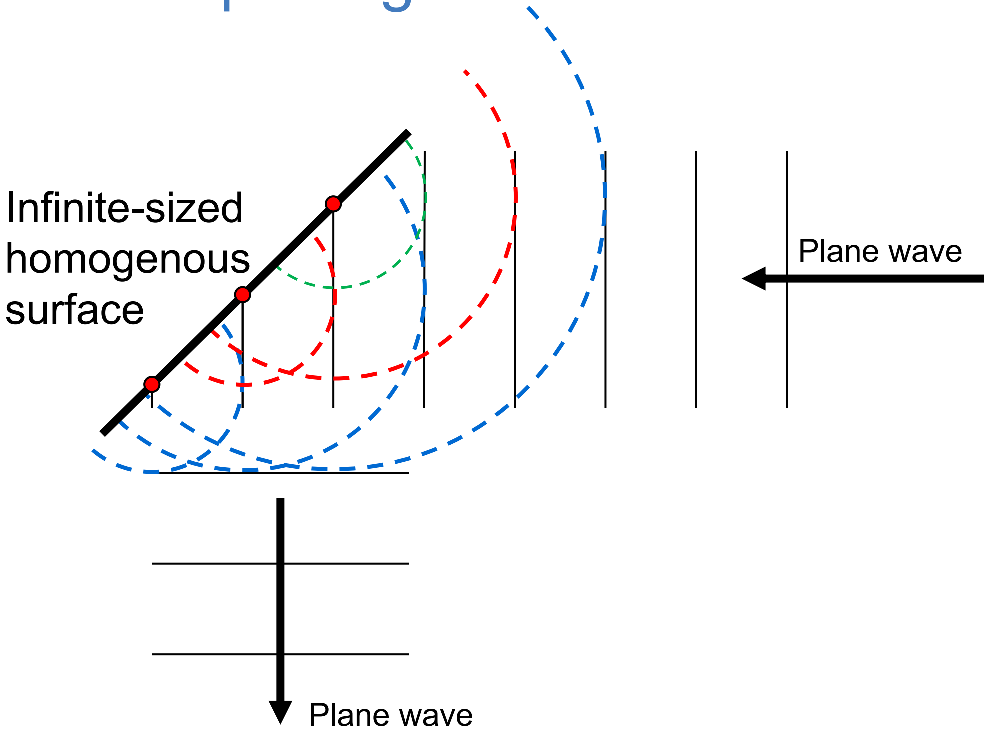
Intelligent reflecting surface (IRS)

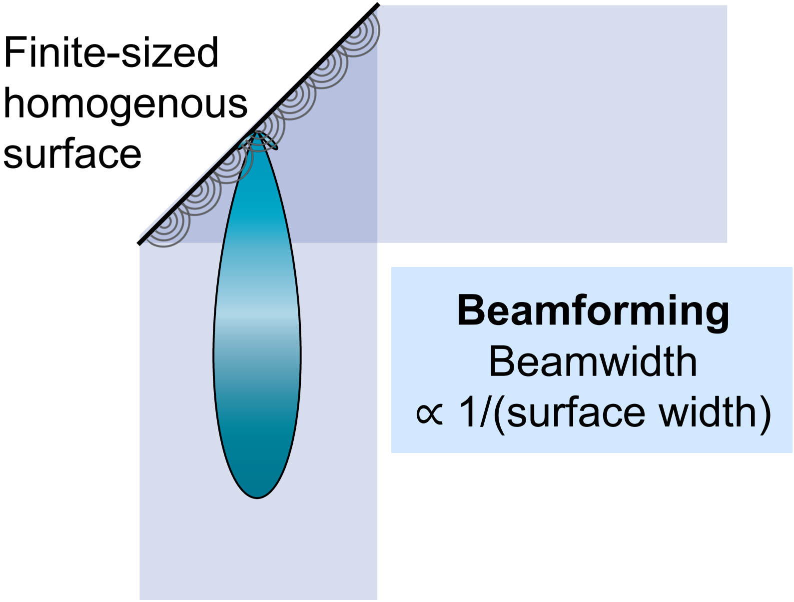
Reconfigurable intelligent surface (RIS)

Beamforming: Directivity by Constructive Interference



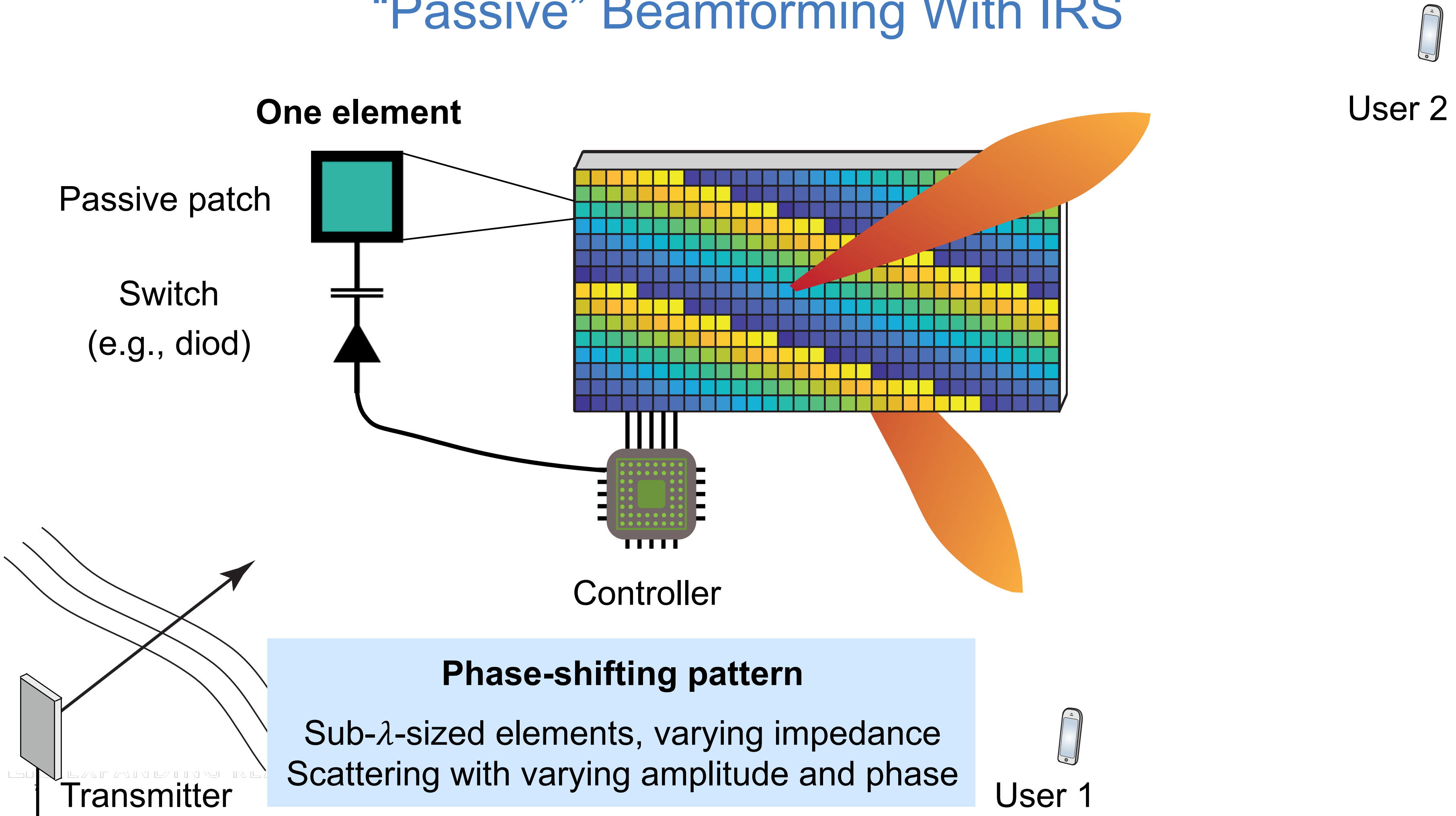
Interpreting Reflection via the Huygens-Fresnel Principle



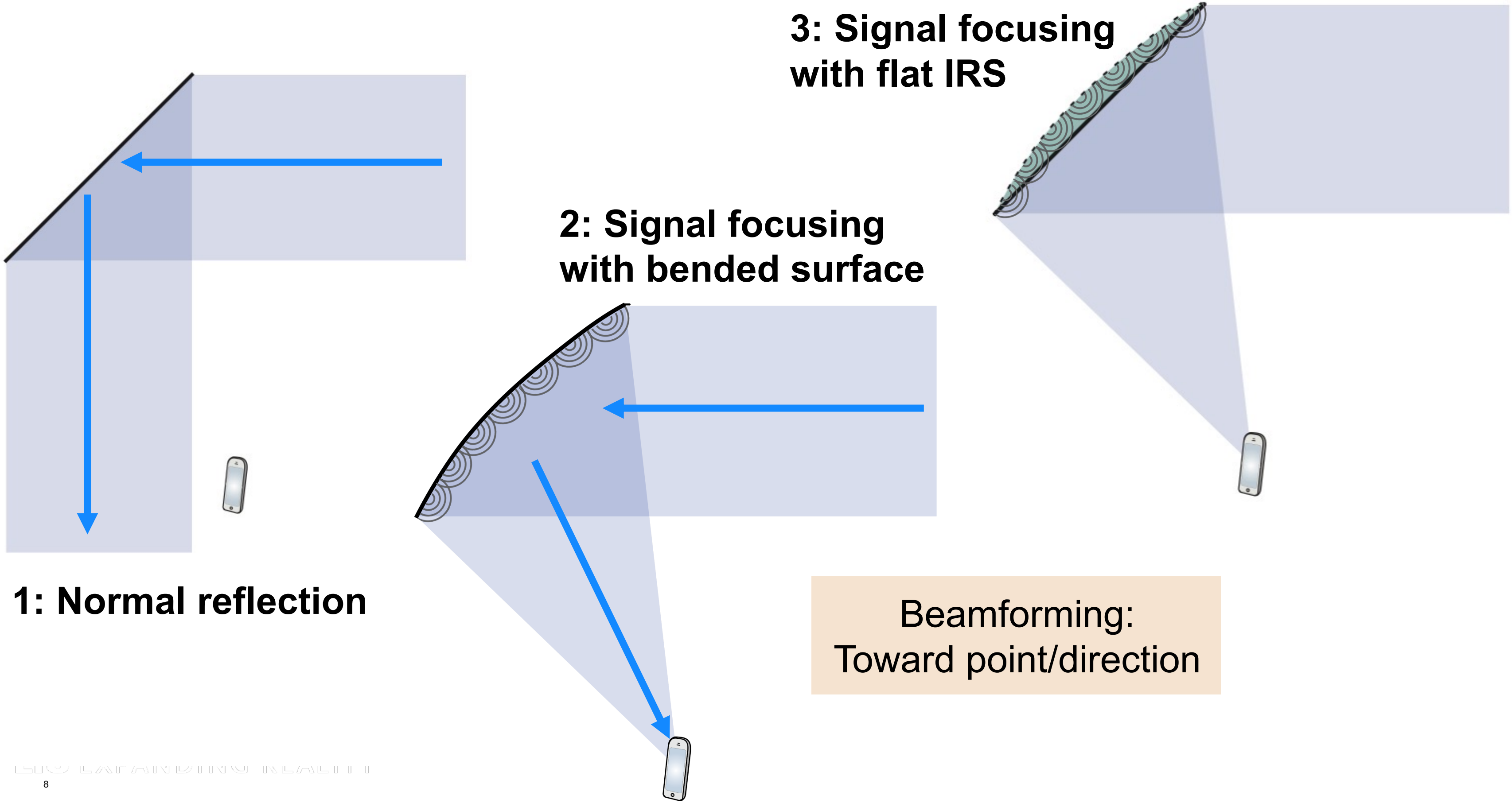


Every point “scatters” a spherical wave:
Constructive interference determines direction

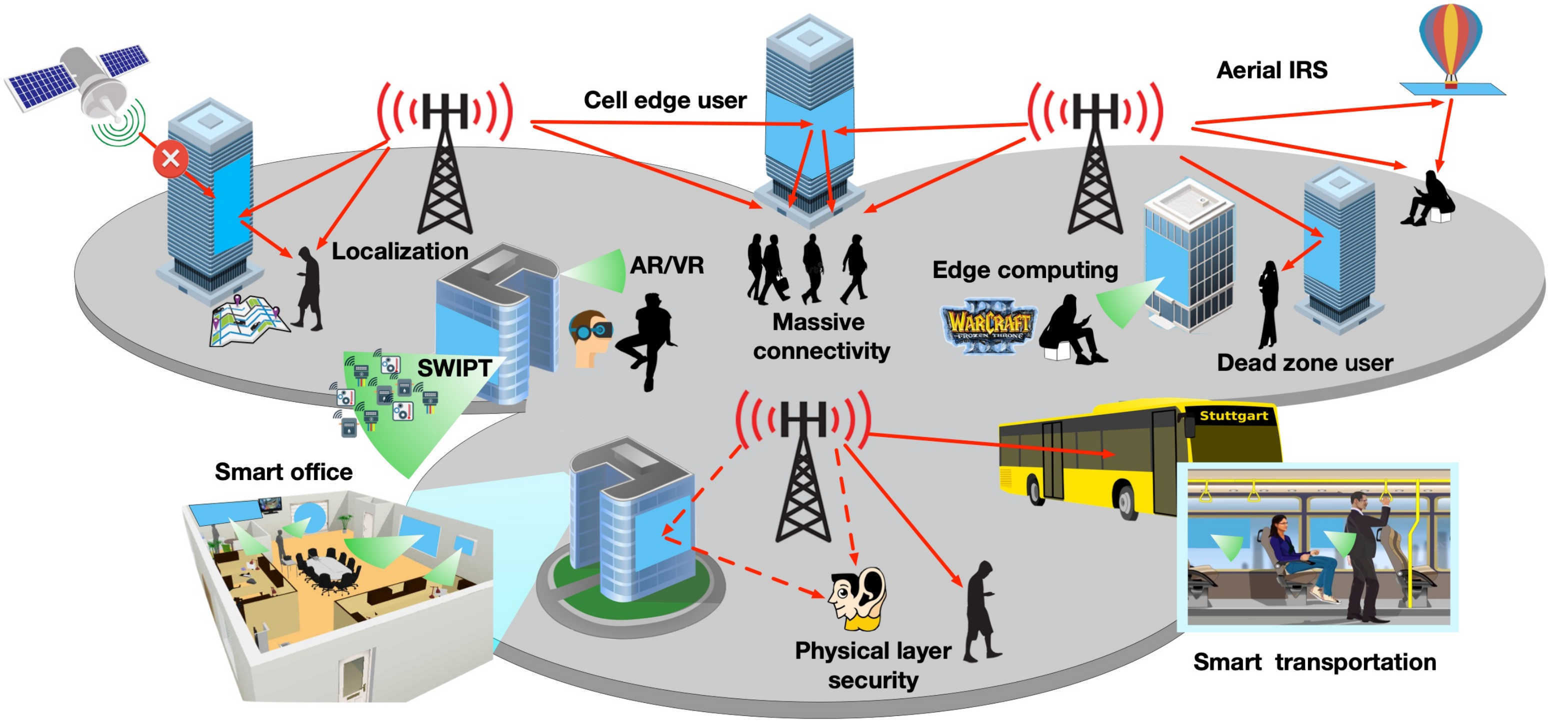
“Passive” Beamforming With IRS



Geometrical Interpretation at the Global Level



A Wealth of Prospective Use Cases

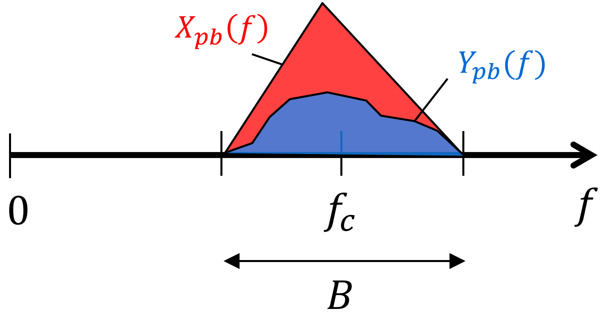


Q. Wu, S. Zhang, B. Zheng, C. You and R. Zhang, “Intelligent Reflecting Surface-Aided Wireless Communications: A Tutorial,” IEEE Transactions on Communications, 2021.

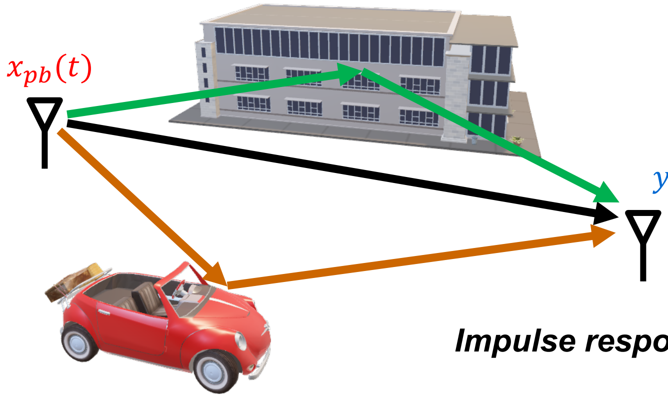
SYSTEM MODELING AND IRS OPTIMIZATION

Introduction to Signals and Systems

$x_{pb}(t)$ → **Linear-time invariant:** *Impulse response* $h_{pb}(t)$ → $y_{pb}(t) = (h_{pb} * x_{pb})(t) = \int_{\infty}^{-\infty} h_{pb}(u)x_{pb}(t-u)du$



- Communication channels are systems/filters:

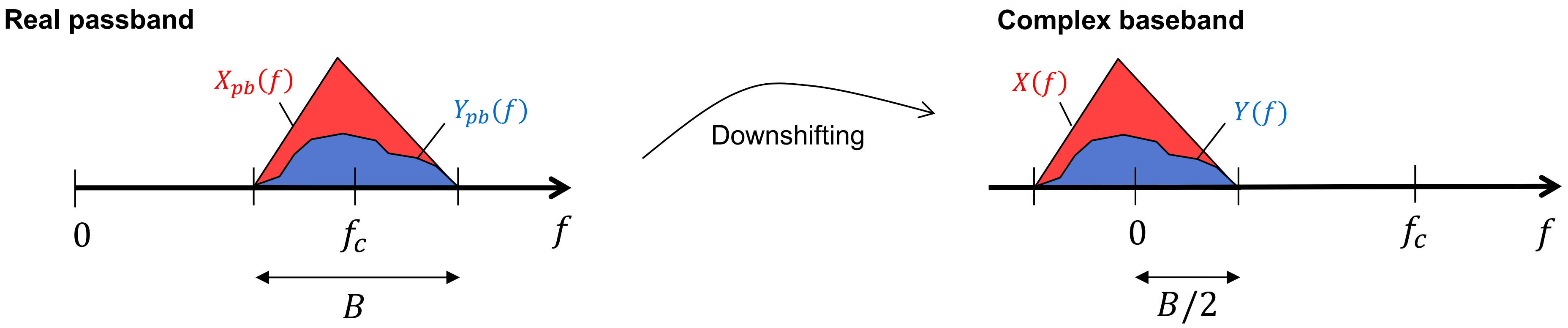


$$y_{pb}(t) = \sqrt{\rho_1}x_{pb}(t-\tau_1) + \sqrt{\rho_2}x_{pb}(t-\tau_2) + \sqrt{\rho_3}x_{pb}(t-\tau_3)$$

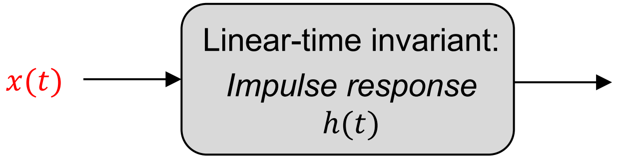
Impulse response: $h_{pb}(t) = \sqrt{\rho_1}\delta(t-\tau_1) + \sqrt{\rho_2}\delta(t-\tau_2) + \sqrt{\rho_3}\delta(t-\tau_3)$

Complex Baseband Representation

- Communication theory is developed for the baseband



- Connection: $X_{pb}(f) = \frac{X(f-f_c)+X^*(-f-f_c)}{\sqrt{2}}$, $Y_{pb}(f) = \frac{Y(f-f_c)+Y^*(-f-f_c)}{\sqrt{2}}$

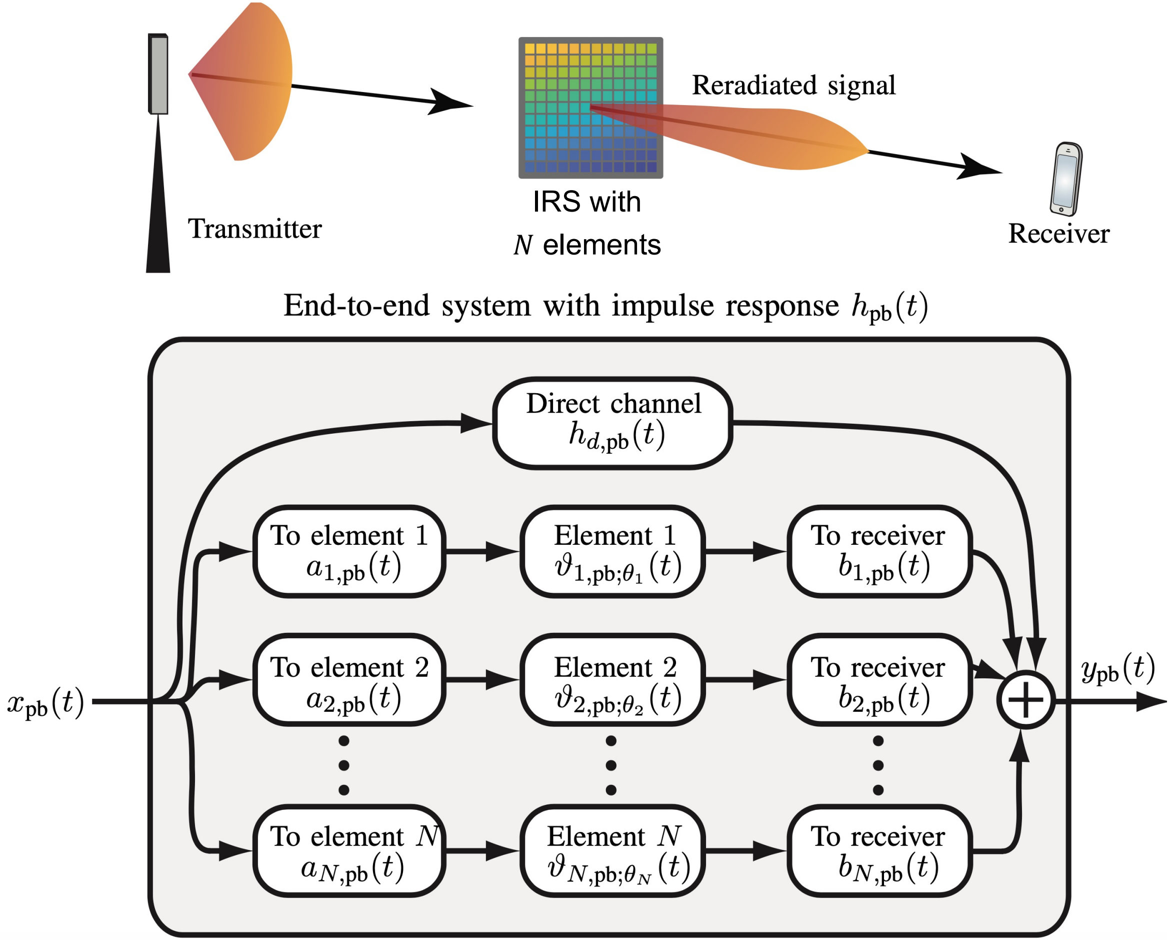


$y(t) = \int_{\infty}^{-\infty} h(u)x(t-u)du$

Downshifted channel:

$h(t) = h_{pb}(t)e^{-j2\pi f_c t}$

Analyzing System With Intelligent Reflecting Surface



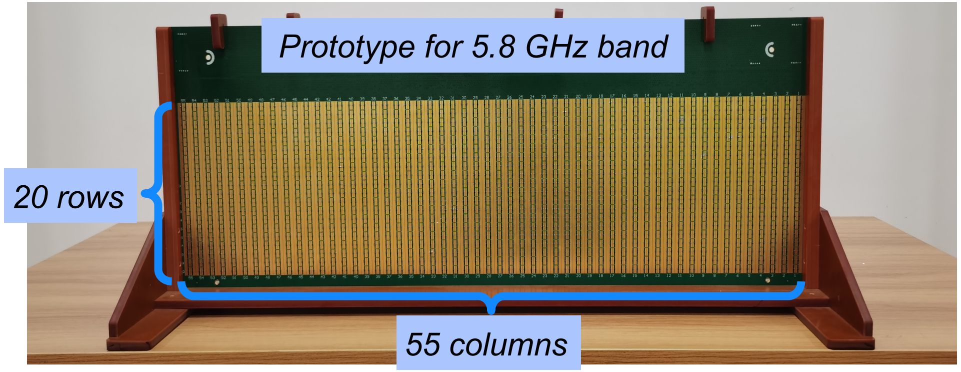
IRS control variables $\theta_1, \ldots, \theta_N$ in $\vartheta_{n,pb;\theta_n}$

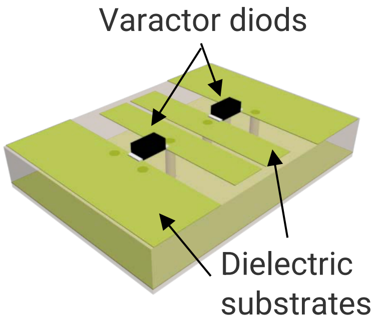
End-to-end impulse response:

$$h_{pb}(t) = h_{d,pb}(t) + \sum_{n=1}^{N} \left(b_{n,pb} * \vartheta_{n,pb;\theta_n} * a_{n,pb}\right)(t)$$

Conventional channel models for $a_{n,pb}, b_{n,pb}, h_{d,pb}$

How Will an IRS Element Filter the Signal?



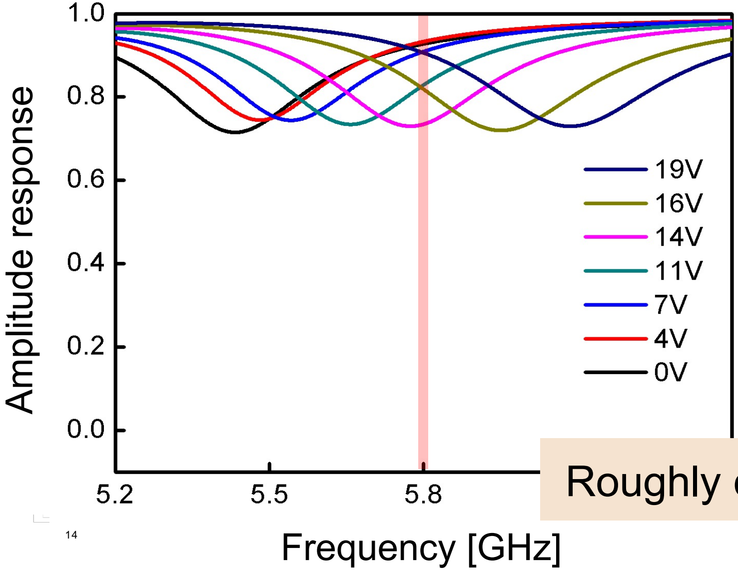


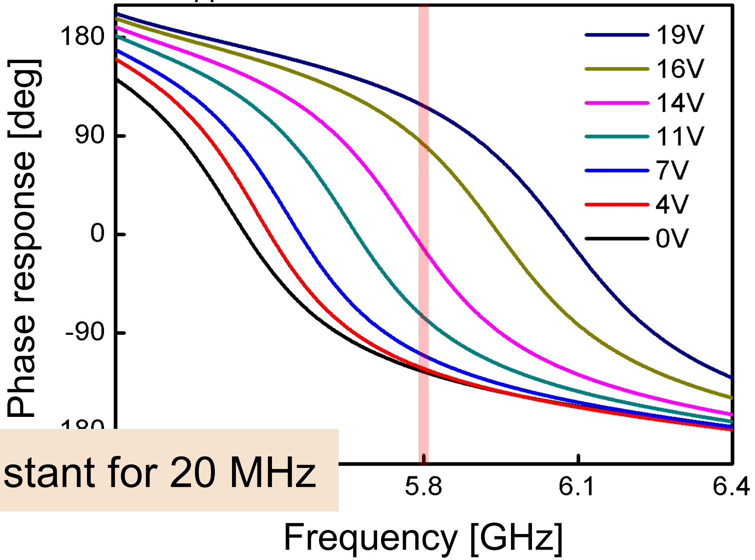
Example: Patch with bias voltage V

Reflection coefficient:

$$\frac{Z_n(V) - Z_0}{Z_n(V) + Z_0}$$

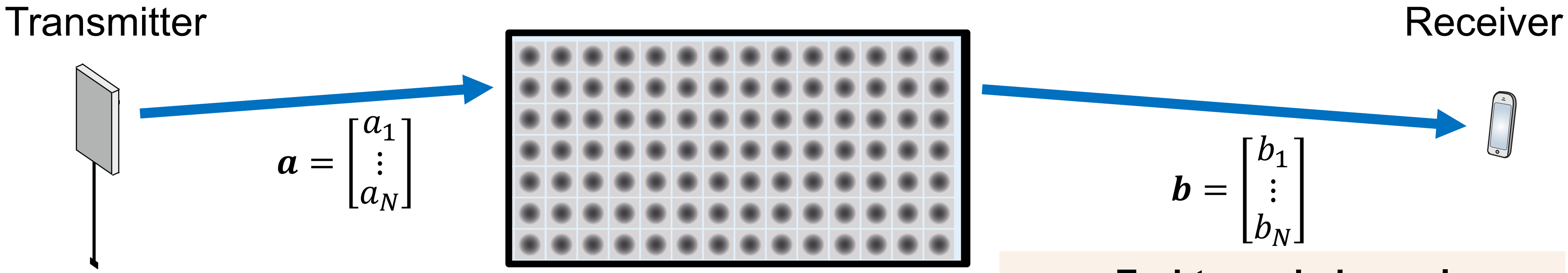
Reference: X. Pei, H. Yin, L. Tan, L. Cao, Z. Li, K. Wang, K. Zhang, E. Björnson, "RIS-Aided Wireless Communications: Prototyping, Adaptive Beamforming, and Indoor/Outdoor Field Trials," IEEE Trans. Commun, to appear.





Roughly constant for 20 MHz

Narrowband System Modelling: N IRS elements



$$\boldsymbol{a} = \begin{bmatrix} a_1 \\ \vdots \\ a_N \end{bmatrix}$$

$$\boldsymbol{b} = \begin{bmatrix} b_1 \\ \vdots \\ b_N \end{bmatrix}$$

Phase-shifts: $\boldsymbol{\omega}_\theta = \begin{bmatrix} e^{-j\theta_1} \\ \vdots \\ e^{-j\theta_N} \end{bmatrix}$

End-to-end-channel

$$\gamma = \sum_{n=1}^{N} a_n e^{-j\theta_n} b_n = (\boldsymbol{a} \odot \boldsymbol{b})^T \boldsymbol{\omega}_\theta$$

Element-wise product

Configure IRS to maximize $|\gamma|^2$:

Opposite phases in $\boldsymbol{\omega}_\theta$ and $\boldsymbol{a} \odot \boldsymbol{b}$:

$$\theta_n = \arg(a_n b_n)$$

$$|\gamma|^2 = \left| \sum_{n=1}^{N} a_n e^{-j\theta_n} b_n \right|^2 \leq \left| \sum_{n=1}^{N} |a_n| \cdot |b_n| \right|^2 = \|\boldsymbol{a} \odot \boldsymbol{b}\|_1^2$$

Adding a "Direct" Path

$$|h_d + \gamma|^2 = \left| h_d + \sum_{n=1}^{N} a_n e^{-j\theta_n} b_n \right|^2 \leq \left| |h_d| + \sum_{n=1}^{N} |a_n|\,|b_n| \right|^2$$

Match phases to direct path: $\theta_n = \arg(a_n b_n) - \arg(h_d)$

Power scaling behavior

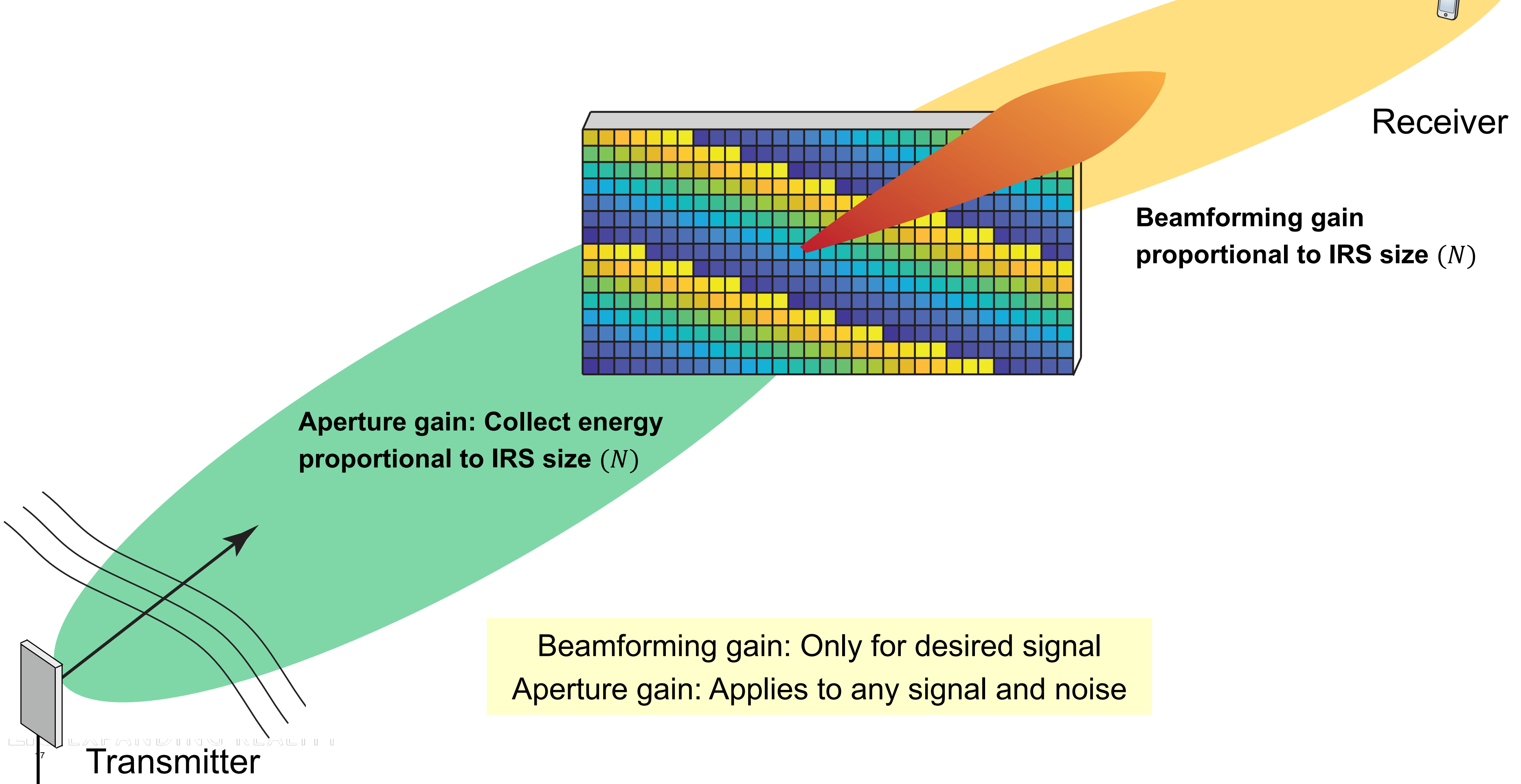
- Assume $|a_n| = \sqrt{\alpha}, |b_n| = \sqrt{\beta}, |h_d| = \sqrt{\rho}$
- SNR proportional to

$$\left|\sqrt{\rho} + N\sqrt{\alpha\beta}\right|^2 = N^2\alpha\beta + \rho + 2N\sqrt{\alpha\beta\rho}$$

$$\approx \begin{cases} N^2\alpha\beta, & \text{if } N\sqrt{\alpha\beta} \gg \sqrt{\rho} \\ \rho, & \text{if } N\sqrt{\alpha\beta} \ll \sqrt{\rho} \end{cases}$$

"Squaring law"

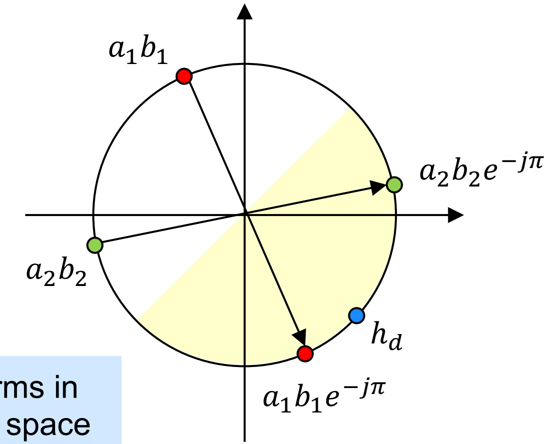
Why is SNR Proportional to N^2?



Beamforming gain: Only for desired signal
Aperture gain: Applies to any signal and noise

Finite-Resolution Phase Shift Configurations

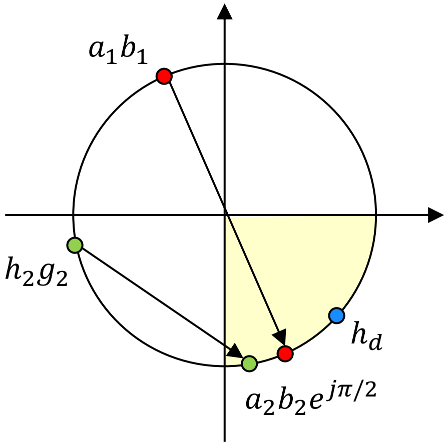
Two configurations: $\theta_n \in \{0, \pi\}$



Gather all terms in the same half space

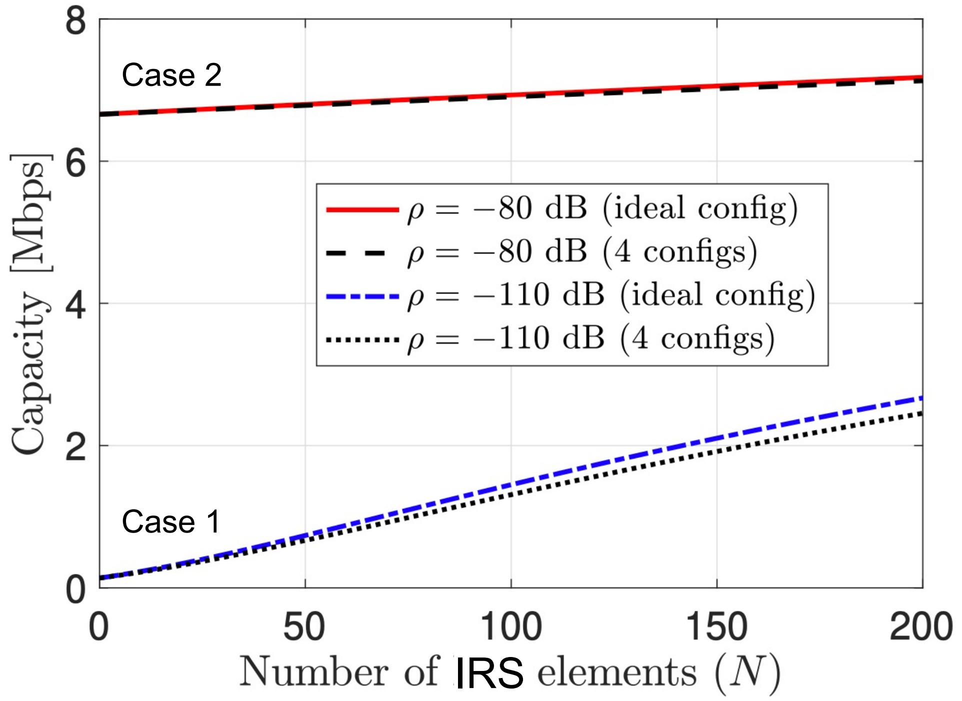
$$\left|h_d + a_1 b_1 e^{-j\theta_1} + a_2 b_2 e^{-j\theta_2}\right|^2$$

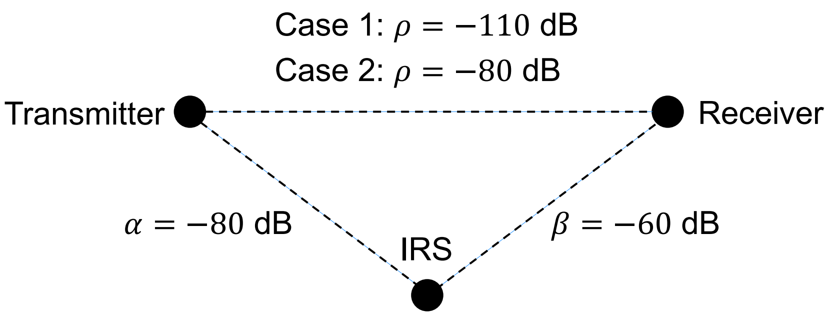
Four configurations: $\theta_n \in \left\{-\frac{\pi}{2}, 0, \frac{\pi}{2}, \pi\right\}$



Gather all terms in the same quadrant

Basic Performance Benefit





Bandwidth: $B = 1$ MHz

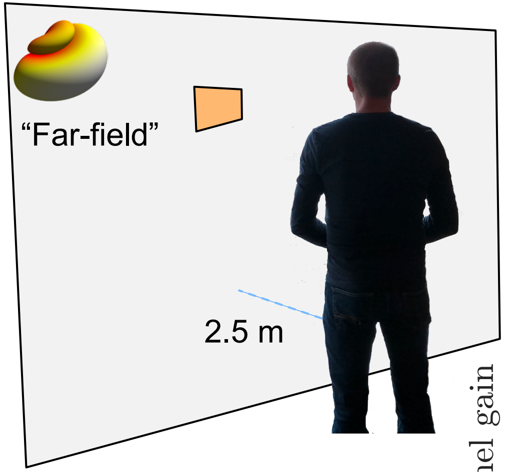
Transmit SNR: 100 dB

IRS is particularly helpful

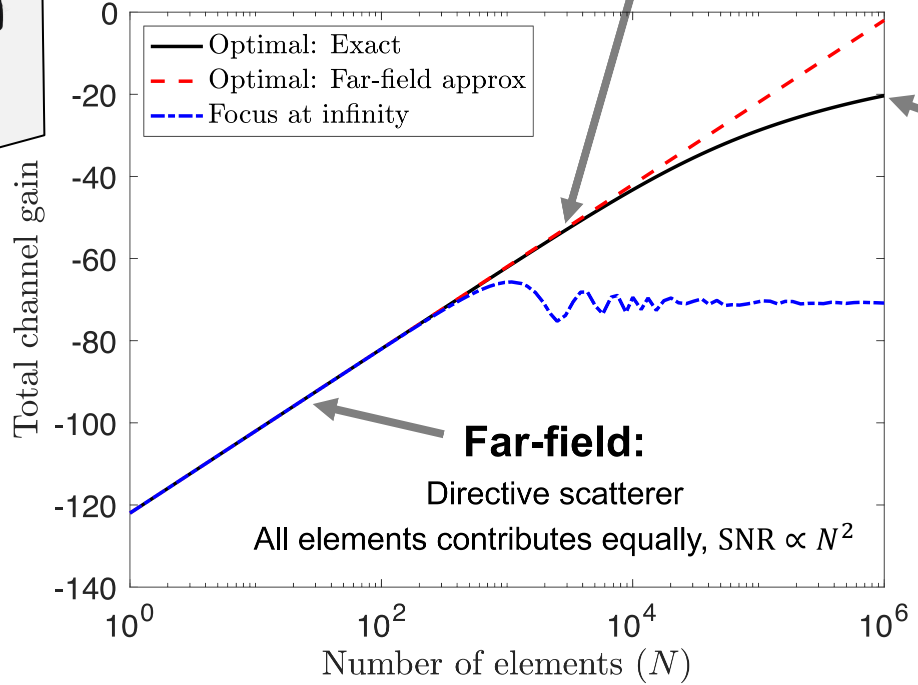
when direct path is relatively weak

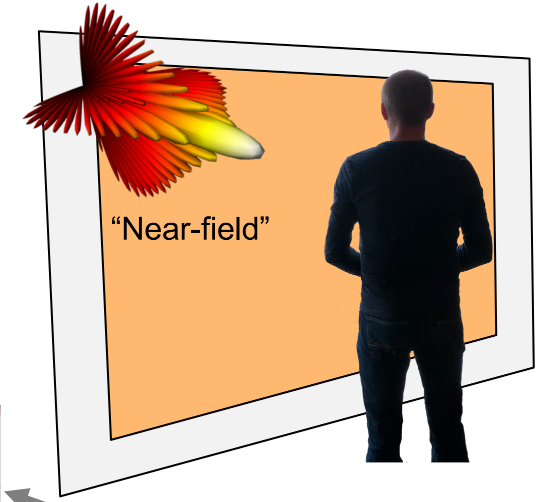
Reference: *E. Björnson, H. Wymeersch, B. Matthiesen, P. Popovski, L. Sanguinetti, E. de Carvalho, "A Signal Processing Perspective on Reconfigurable Intelligent Surfaces With Wireless Applications", Available on arXiv:2102.00742.*

What Happens Asymptotically?



Near-field:
Focus at a point
(not in a direction)





Asymptotic limit
Receive almost all
transmit power

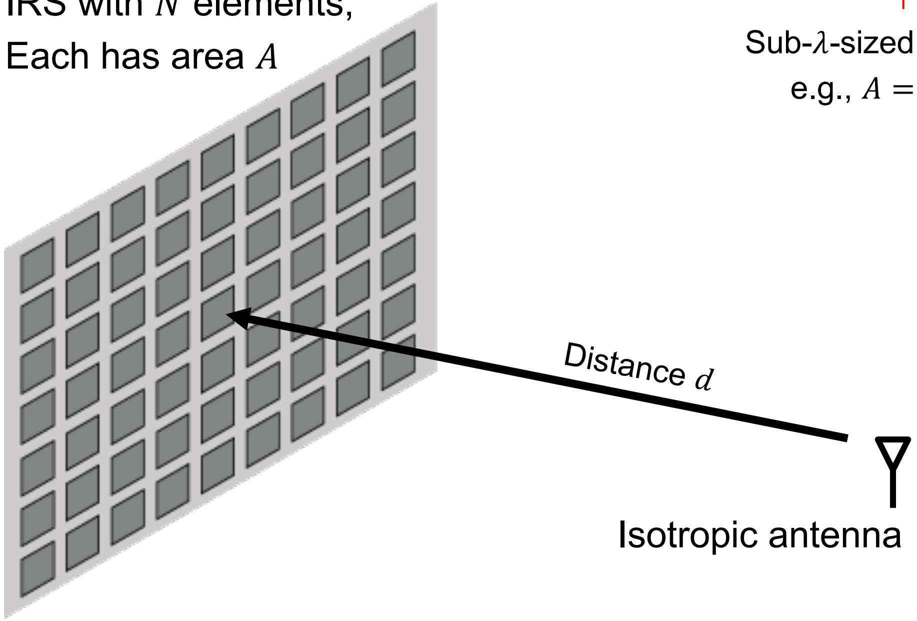
Assumptions:

25 m from transmitter
2.5 m to receiver
Isotropic antennas (3 GHz)

NARROWBAND CHANNEL MODELING

Basic Pathloss Modeling

IRS with N elements,
Each has area A



Pathloss: $\alpha = \frac{\text{Effective area}}{4\pi d^2} \approx \frac{A}{4\pi d^2}$

Sub-λ-sized elements,
e.g., $A = (\lambda/5)^2$

Compare to isotropic receive antenna:

$$\alpha = \alpha_{\text{iso}} \frac{A}{\lambda^2/4\pi}$$

Smaller than 1

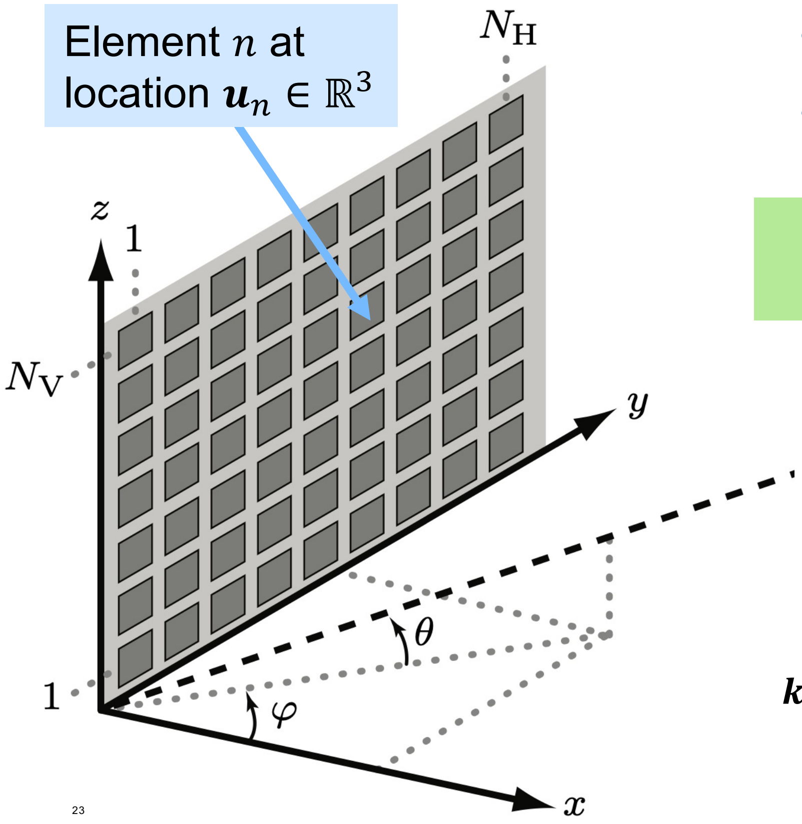
Example: Use 3GPP Urban Micro Model

$$\alpha_{\text{iso}}\ [\text{dB}] = -37.5 - 22\log_{10}(d/1\ \text{m})$$

$$\alpha\ [\text{dB}] = \alpha_{\text{iso}}\ [\text{dB}] + 10\log_{10}\left(\frac{A}{\lambda^2/4\pi}\right)$$

Extra reduction

Channel Modeling Using Array Response Vector



Far-field: Incoming/outgoing plane wave determined by

- Azimuth angle $\varphi \in [-\pi, \pi]$
- Elevation angle $\theta \in \left[-\frac{\pi}{2}, \frac{\pi}{2}\right]$

Channels $\boldsymbol{a}$ or $\boldsymbol{b}$ for wave in direction (φ, θ)
$\text{constant} \cdot \boldsymbol{r}(\varphi, \theta)$

Array response vector:

$$\boldsymbol{r}(\varphi, \theta) = \begin{bmatrix} e^{j\boldsymbol{k}(\varphi,\theta)^T \boldsymbol{u}_1} \\ \vdots \\ e^{j\boldsymbol{k}(\varphi,\theta)^T \boldsymbol{u}_N} \end{bmatrix}$$

$$\boldsymbol{k}(\varphi, \theta) = \frac{2\pi}{\lambda} \begin{bmatrix} \cos(\theta)\cos(\varphi) \\ \cos(\theta)\sin(\varphi) \\ \sin(\theta) \end{bmatrix}$$

Basic Multipath Channels

Sparse channels

L is small or one path is much stronger than all other

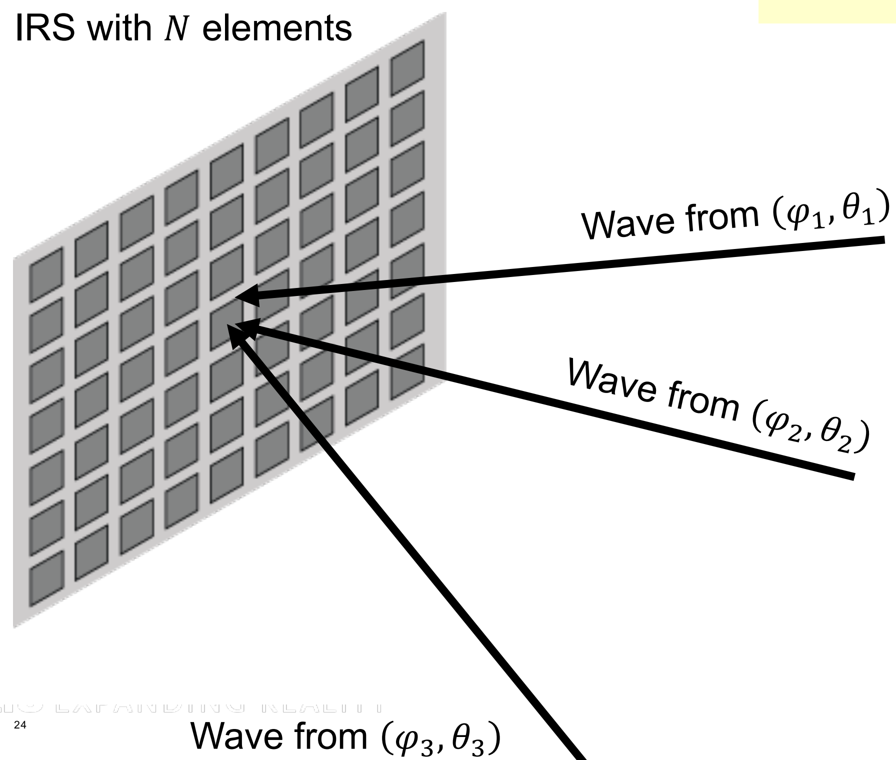
Example: Line-of-sight channels

Channel vector for L plane wave

$$\boldsymbol{a} = \sum_{l=1}^{L} \frac{c_l}{\sqrt{L}} \cdot \boldsymbol{r}(\varphi_l, \theta_l)$$

N complex parameters in $\boldsymbol{a}$

or $4L$ real parameters

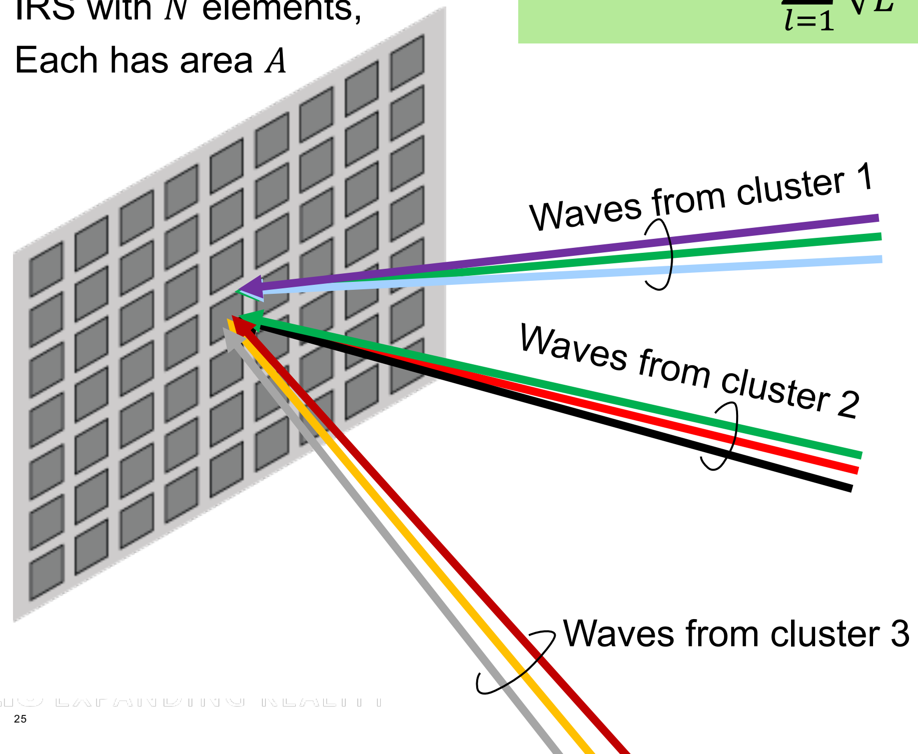


Rich Multipath Channels

Channel vector for *L* plane wave

$$\boldsymbol{a} = \sum_{l=1}^{L} \frac{c_l}{\sqrt{L}} \cdot r(\varphi_l, \theta_l)$$

IRS with N elements,
Each has area A



Rayleigh fading: $\boldsymbol{a} \sim CN(\boldsymbol{0}, \boldsymbol{C})$

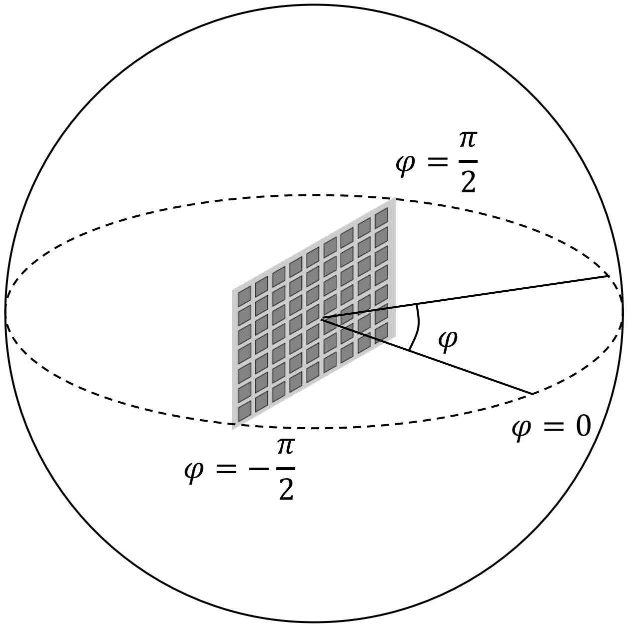
Achieved as $L \to \infty$ if

c_l i.i.d. with zero mean, variance $A\mu$

(φ_l, θ_l) i.i.d. with PDF $f(\varphi, \theta)$

$$[\boldsymbol{C}]_{nm} = A\mu \, E\left\{ e^{j\boldsymbol{k}(\varphi,\theta)^T(\boldsymbol{u}_n - \boldsymbol{u}_m)} \right\}$$

Isotropic Scattering



Uniform distribution in one half-space

(φ_l, θ_l) i.i.d. with PDF

$$f(\varphi, \theta) = \begin{cases} \cos(\theta)/2\pi, & \varphi \in \left[-\frac{\pi}{2}, \frac{\pi}{2}\right] \\ 0, & \text{otherwise} \end{cases}$$

Spatial correlation matrix:

$$[\boldsymbol{C}]_{nm} = A\mu \, E\left\{e^{j\boldsymbol{k}(\varphi,\theta)^T(\boldsymbol{u}_n - \boldsymbol{u}_m)}\right\}$$
$$= A\mu \cdot \mathrm{sinc}\left(\frac{2}{\lambda}\|\boldsymbol{u}_n - \boldsymbol{u}_m\|\right)$$

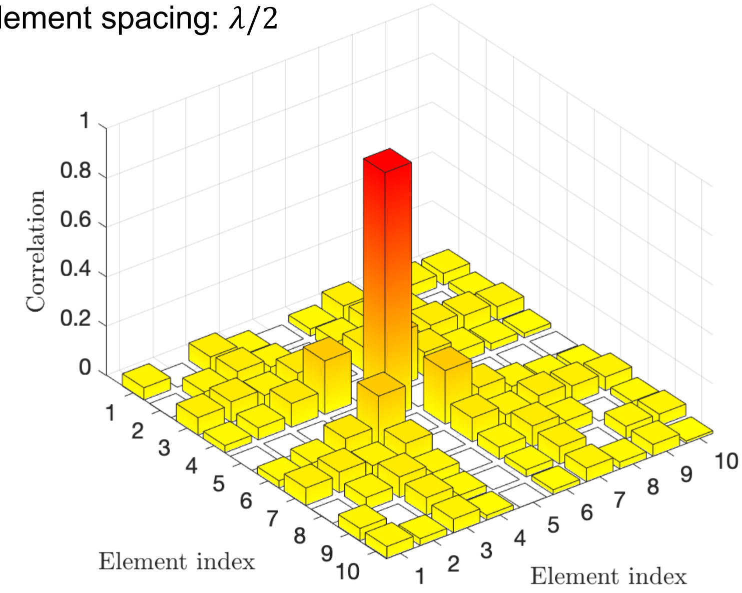
Reference: E. Björnson, L. Sanguinetti, “Rayleigh Fading Modeling and Channel Hardening for Reconfigurable Intelligent Surfaces,” IEEE Wireless Communications Letters, 2021.

Misconception: “We can assume i.i.d. Rayleigh fading”

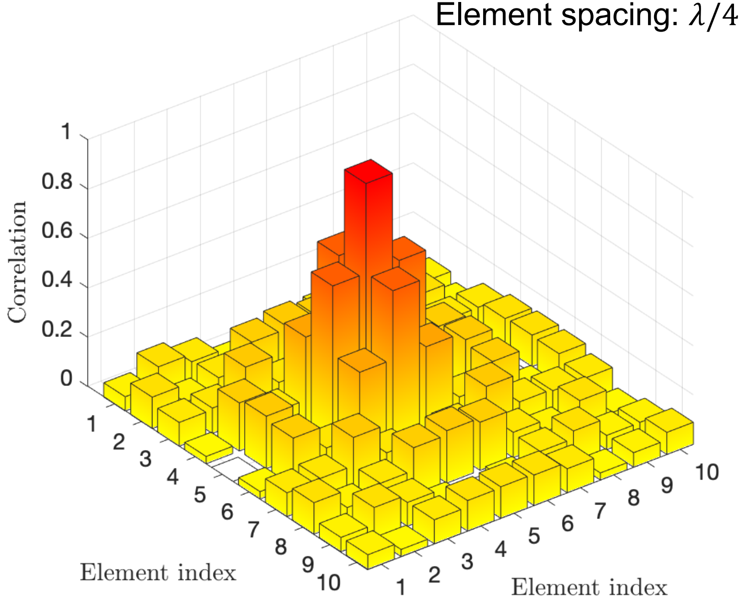
Spatial correlation matrix in isotropic scattering:

$$[\boldsymbol{C}]_{nm} = A\mu \cdot \mathrm{sinc}\left(\frac{2}{\lambda}\|\boldsymbol{u}_n - \boldsymbol{u}_m\|\right)$$

Element spacing: $\lambda/2$



Element spacing: $\lambda/4$



Reference: E. Björnson, L. Sanguinetti, “Rayleigh Fading Modeling and Channel Hardening for Reconfigurable Intelligent Surfaces,” IEEE Wireless Communications Letters, 2021.

Spatial Correlation Appears Naturally

Square array, isotropic scattering

$$\frac{\mathrm{rank}(\boldsymbol{C})}{\pi N A/\lambda^2} \to 1$$

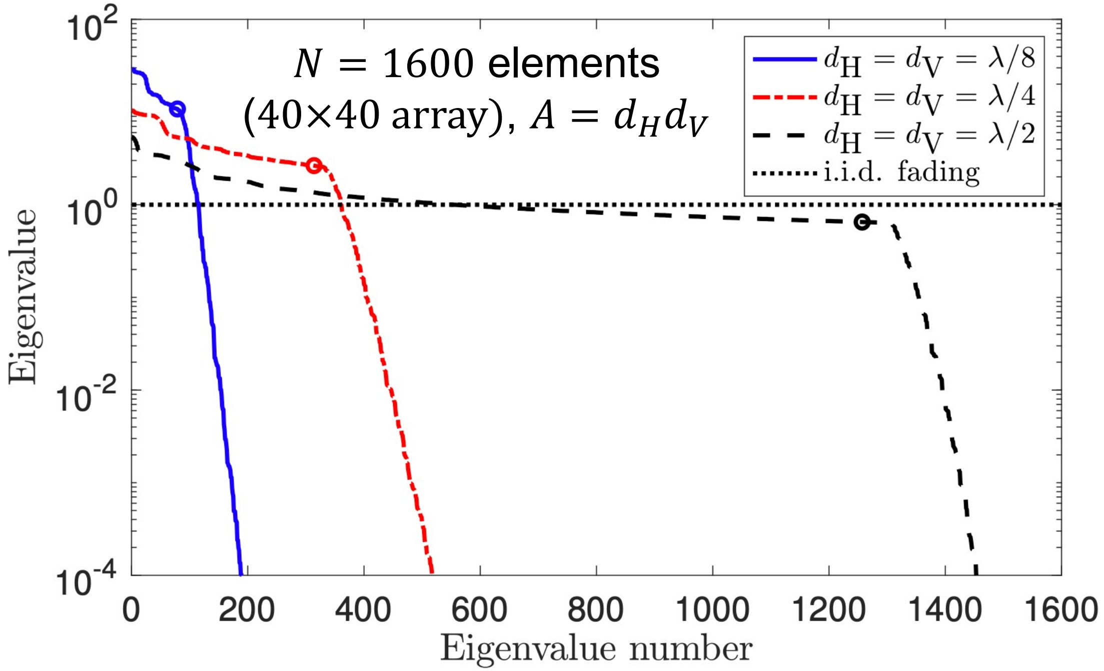
when $N \to \infty, A \to 0$ so that $NA \to \infty$

Channel sparsity

Even in isotropic fading!

Two contributing factors

1. Sub-wavelength antenna spacing
2. Two-dimensional surface



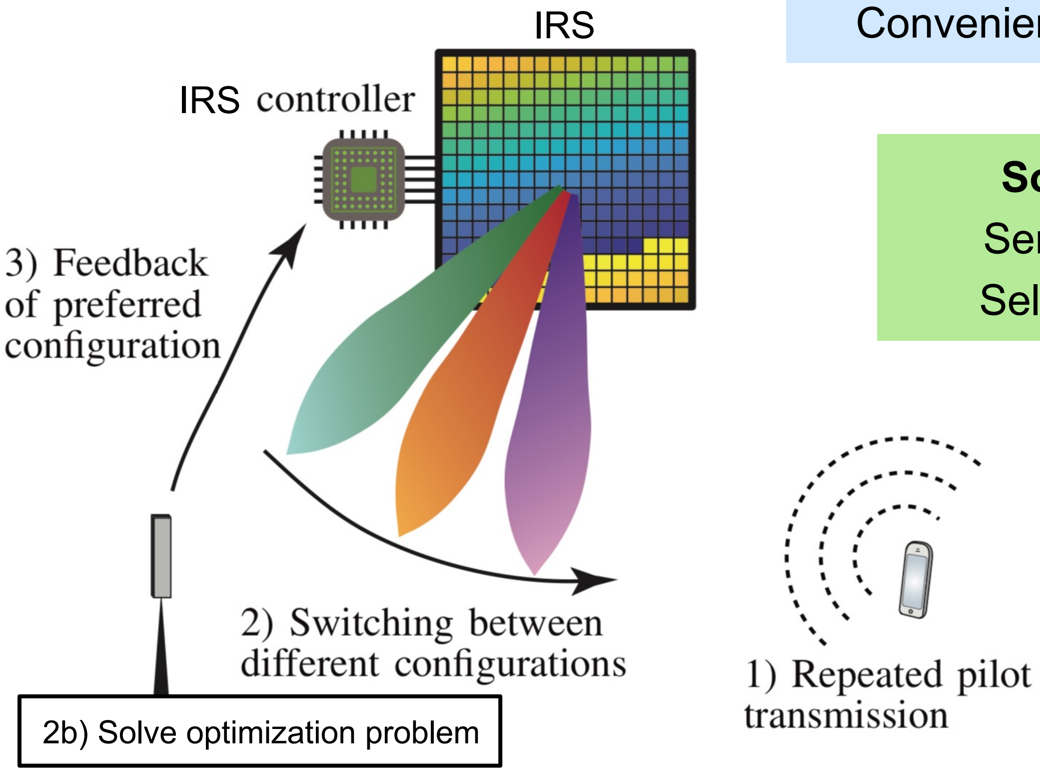
CHANNEL ESTIMATION

IRS Reconfigurability is Complicated

The IRS is blind!

Solution 1: Have a fraction of active IRS elements
Convenient, but removes the "low-cost" argument

Solution 2: Codebook approach
Send pilots and switch configuration
Select the best configuration in a set



Basic Codebook-Based Channel Estimation

Received signal: $z[n] = (h_d + (\boldsymbol{a} \odot \boldsymbol{b})^T \boldsymbol{\omega}_\theta)x[n] + w[n]$

Repeat a known pilot signal L times: $x[n] = s,\ n = 1, \dots, L$

- Call cascaded channel $\boldsymbol{c} = \boldsymbol{a} \odot \boldsymbol{b}$
- Use different configurations: $\boldsymbol{\omega}_{\theta_1}, \dots, \boldsymbol{\omega}_{\theta_L}$

- Received signal: $z[n] = \left(h_d + \boldsymbol{c}^T \boldsymbol{\omega}_{\theta_n}\right)x[n] + w[n]$

$$\underbrace{\left[z[1], \dots, z[L]\right]}_{= \boldsymbol{Z}} = h_d[1 \ \dots \ 1]s + \boldsymbol{c}^T \underbrace{\left[\boldsymbol{\omega}_{\theta_1}, \dots, \boldsymbol{\omega}_{\theta_L}\right]}_{= \boldsymbol{\Omega}} s + \underbrace{\left[w[1], \dots, w[L]\right]}_{= \boldsymbol{W}}$$

$$\boldsymbol{Z} = \begin{bmatrix} h_d \\ \boldsymbol{c} \end{bmatrix}^T \begin{bmatrix} 1 & \dots & 1 \\ & \boldsymbol{\Omega} & \end{bmatrix} s + \boldsymbol{W}$$

Received signal

If $\widetilde{\boldsymbol{\Omega}} = \begin{bmatrix} 1 & \dots & 1 \\ & \boldsymbol{\Omega} & \end{bmatrix}$ is invertible $\longrightarrow$

$$\boldsymbol{Z}\widetilde{\boldsymbol{\Omega}}^{-1}s^{-1} = \begin{bmatrix} h_d \\ \boldsymbol{c} \end{bmatrix}^T + \boldsymbol{W}\widetilde{\boldsymbol{\Omega}}^{-1}s^{-1}$$

Channel estimate

Requires $L = N + 1$

Example: $\widetilde{\boldsymbol{\Omega}}$ =DFT or Hadamard matrix

Exploiting Channel Sparsity

$$\boldsymbol{Z} = \begin{bmatrix} h_d \\ \boldsymbol{c} \end{bmatrix}^T \begin{bmatrix} 1 & \dots & 1 \\ & \boldsymbol{\Omega} & \end{bmatrix} s + \boldsymbol{W}$$

Received signal

Sparsity will exist in $\boldsymbol{c}$!

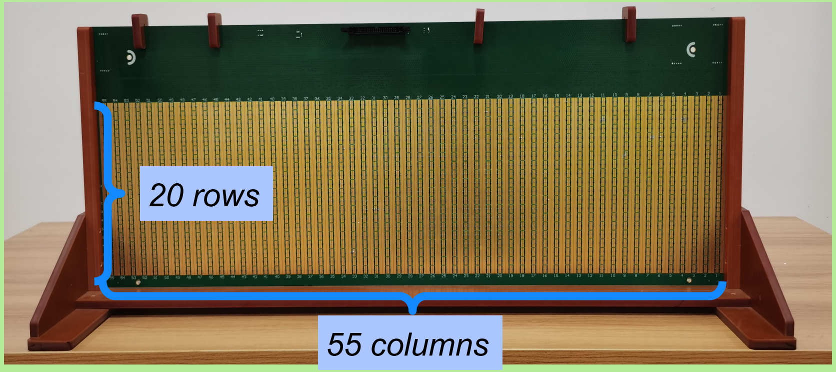
An open research challenge to utilize it most effectively

References: "*Reconfigurable Intelligent Surfaces: Three Myths and Two Critical Questions*", *2020*

Use sparse channel models

Group elements together

Basic configuration algorithm



Two states per element: ± 90 degrees

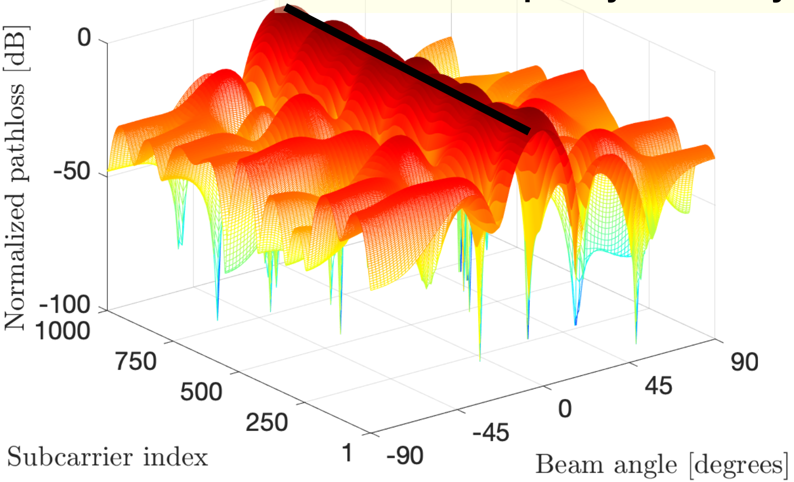
- Flip one row or column, pick best option
- Sweep horizontally, then vertically

Reference: X. Pei, H. Yin, L. Tan, L. Cao, Z. Li, K. Wang, K. Zhang, E. Björnson, "RIS-Aided Wireless Communications: Prototyping, Adaptive Beamforming, and Indoor/Outdoor Field Trials," to appear

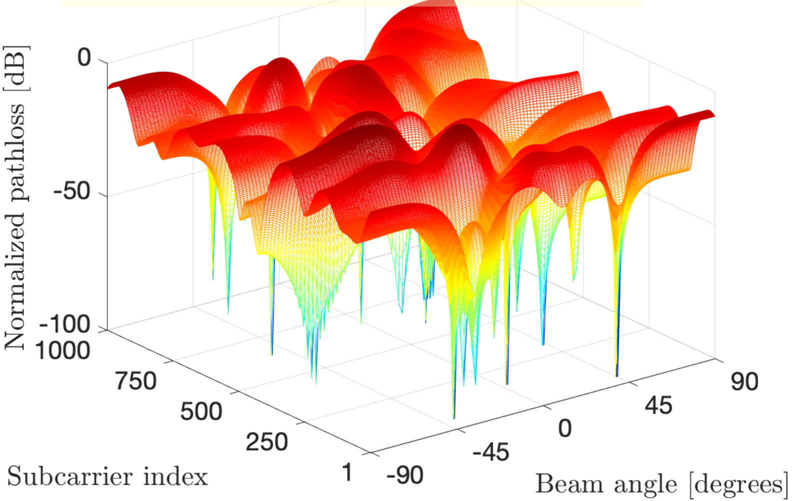
WIDEBAND MODELING

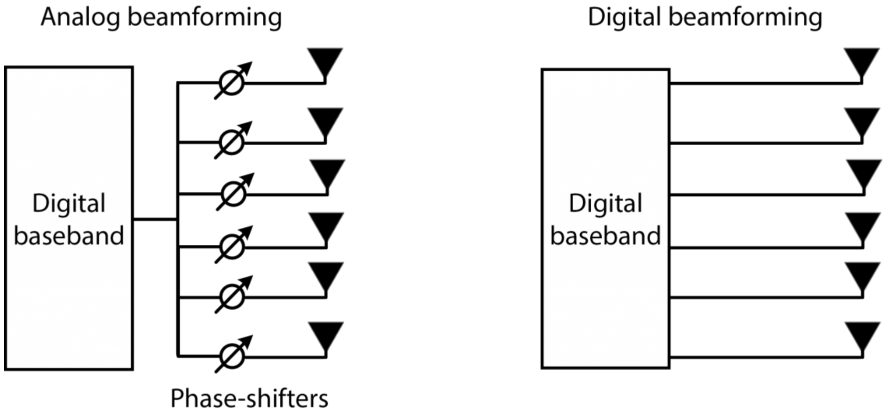
Frequency selective fading

LOS: Weak frequency-selectivity



NLOS: Strong frequency-selectivity



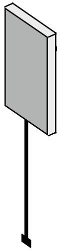


One *angular* beam	Sum of many beams
Same on all subcarriers	Different over subcarriers

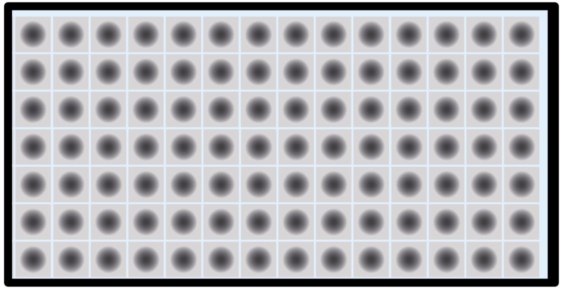
Same principle as for IRS

OFDM System Modeling

Transmitter



Distinct paths:
$\boldsymbol{a}_1, \dots, \boldsymbol{a}_{L_h}$



Phase-shifts: $\boldsymbol{\omega}_\theta = \begin{bmatrix} e^{-j\theta_1} \\ \vdots \\ e^{-j\theta_N} \end{bmatrix}$

Receiver

Distinct paths:
$\boldsymbol{b}_1, \dots, \boldsymbol{b}_{L_g}$

K subcarrier channels:

$$\boldsymbol{\gamma} = \begin{bmatrix} \gamma_1 \\ \vdots \\ \gamma_K \end{bmatrix} = \boldsymbol{F}\boldsymbol{V}^T\boldsymbol{\omega}_\theta$$

DFT matrix $(K \times M)$

Time-domain end-to-end channel matrix (M taps)

$$\boldsymbol{V} = \sum_{l=1}^{L_h}\sum_{i=1}^{L_g}(\boldsymbol{a}_l \odot \boldsymbol{b}_i)\left[\delta[\tau_{h,l} + \tau_{g,i}] \dots \delta[\tau_{h,l} + \tau_{g,i} - (M-1)]\right]$$

Integer delays

IRS Optimization for OFDM system

Sum rate expression:

$$R = \frac{B}{K+M-1}\sum_{k=1}^{K}\log_2\left(1+\frac{P}{\sigma^2}|\gamma_k|^2\right) = \frac{B}{K+M-1}\sum_{k=1}^{K}\log_2\left(1+\frac{P}{\sigma^2}|\boldsymbol{f}_k^H\boldsymbol{V}^T\boldsymbol{\omega}_\theta|^2\right)$$

Row k of $\boldsymbol{F}$

Upper bound: Optimize $\boldsymbol{\omega}_\theta$ for each subcarrier:

$$\left|\boldsymbol{f}_k^H\boldsymbol{V}^T\boldsymbol{\omega}_\theta\right|^2 \leq \|\boldsymbol{f}_k^H\boldsymbol{V}^T\|_1^2$$

Heuristic solution: Strongest-path maximization (STM),

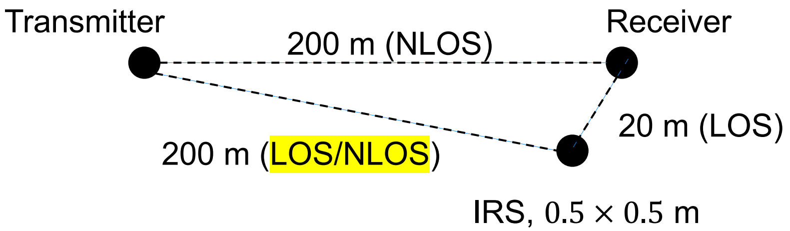
$$m = \arg\max_i \|\boldsymbol{v}_i\|_1^2 \ \text{ where } \boldsymbol{V} = [\boldsymbol{v}_1 \ \ldots \boldsymbol{v}_M]$$

$$\boldsymbol{\omega}_\theta = \arg\max_{\boldsymbol{\omega}} |\boldsymbol{v}_m^T\boldsymbol{\omega}|^2$$

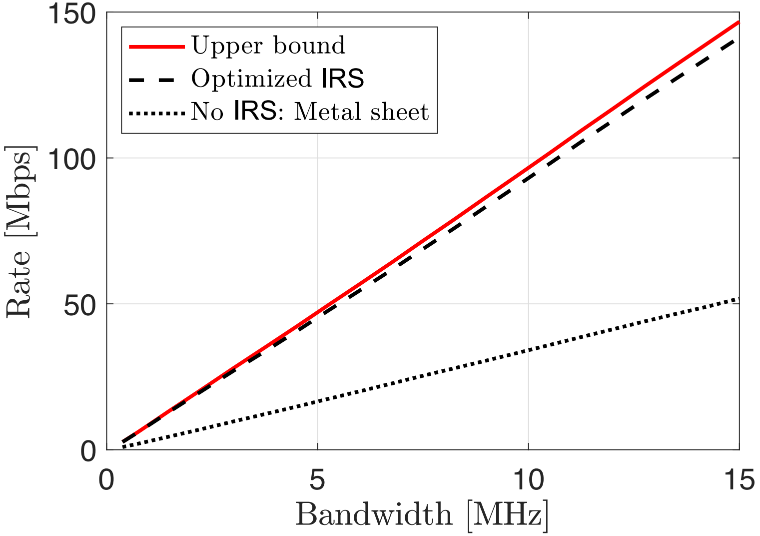
Reference: B. Zheng, R. Zhang, "Intelligent Reflecting Surface-Enhanced OFDM: Channel Estimation and Reflection Optimization," IEEE Wireless Communications Letters, 2020

IRS in Frequency Selective Channels

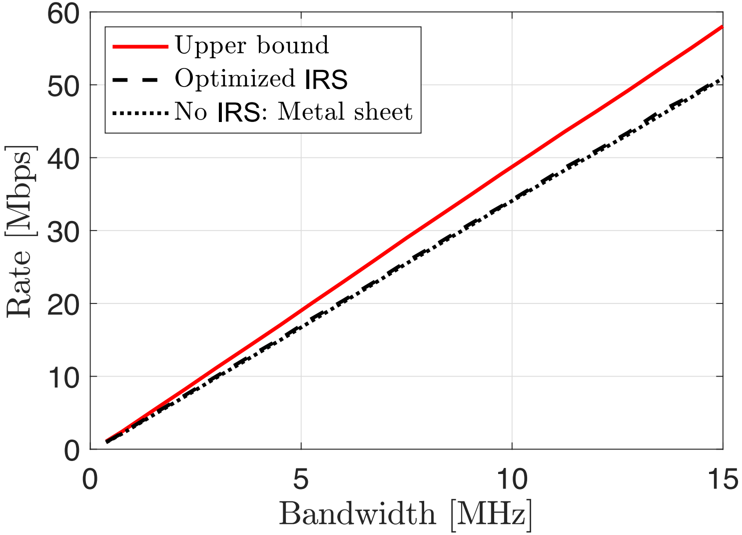
IRS mainly effective with LOS to transmitter and receiver!



LOS: Weakly frequency-selective

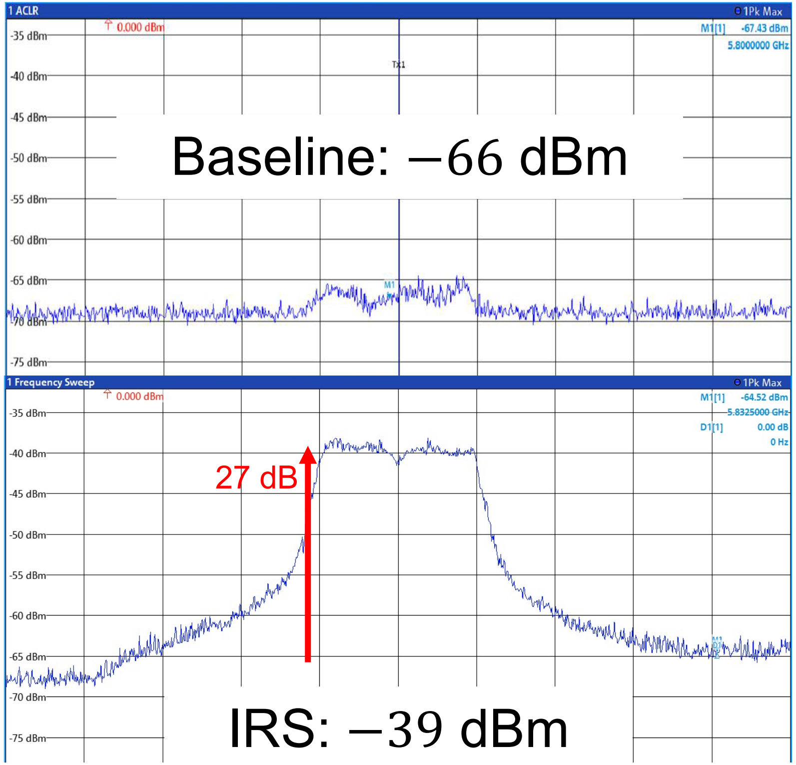


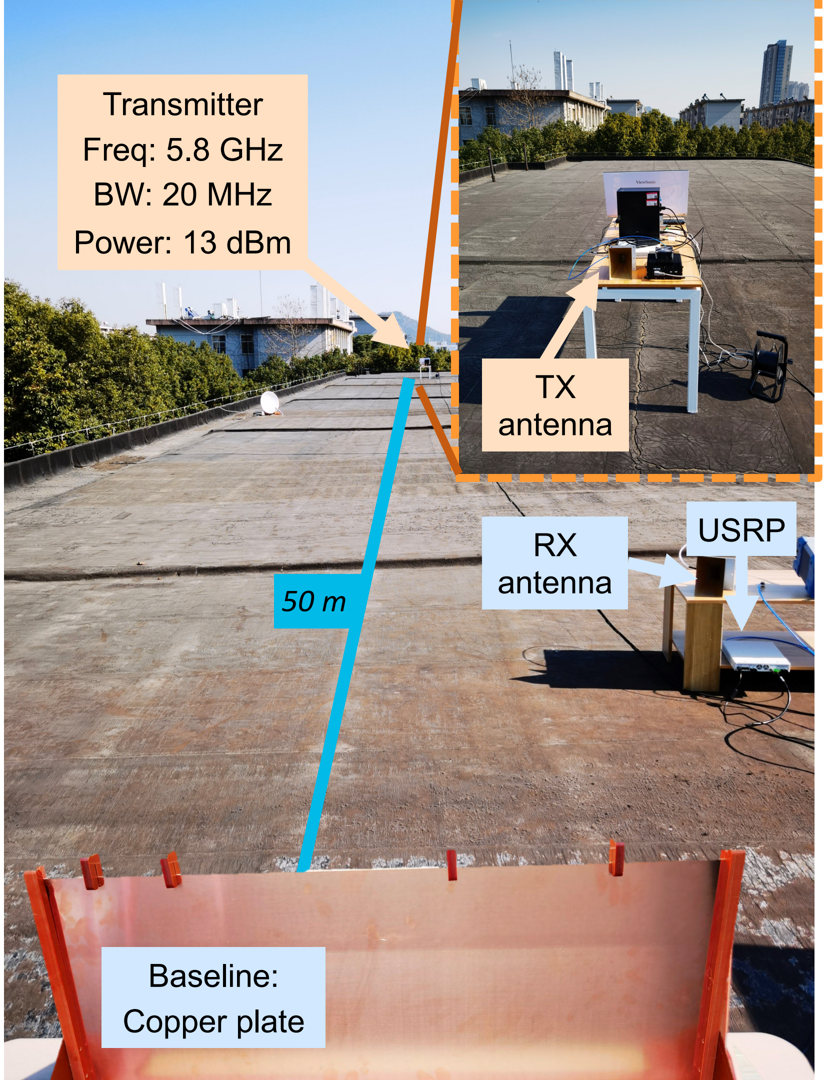
NLOS: Strongly frequency-selective



Reference: *E. Björnson, et al.*, "Reconfigurable Intelligent Surfaces: A Signal Processing Perspective With Wireless Applications", *Available on arXiv:2102.00742.*

Experimental Validation

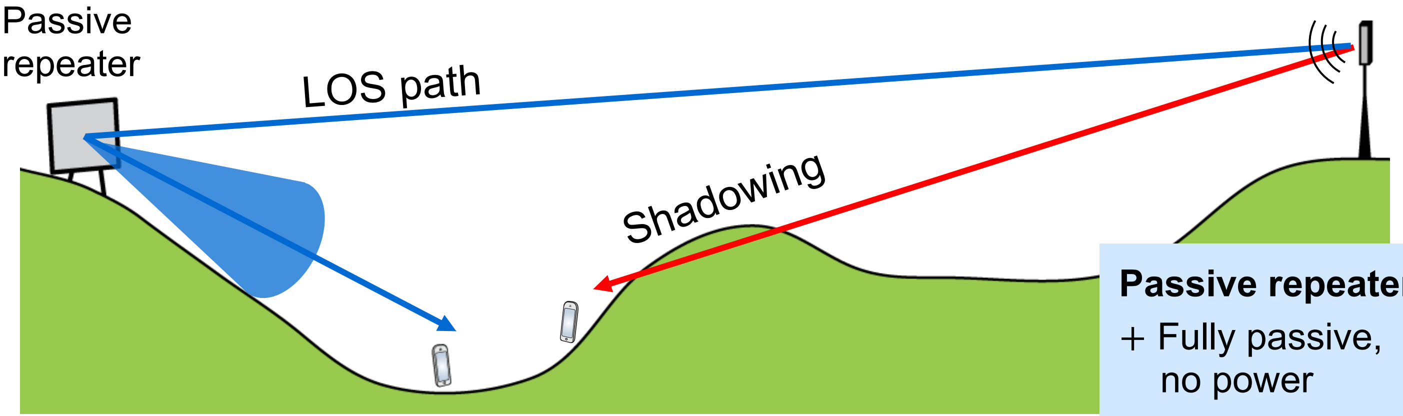




Reference: X. Pei, H. Yin, L. Tan, L. Cao, Z. Li, K. Wang, K. Zhang, E. Björnson, "RIS-Aided Wireless Communications: Prototyping, Adaptive Beamforming, and Indoor/Outdoor Field Trials," to appear

COMPETING TECHNOLOGIES AND POTENTIAL WEAKNESSES

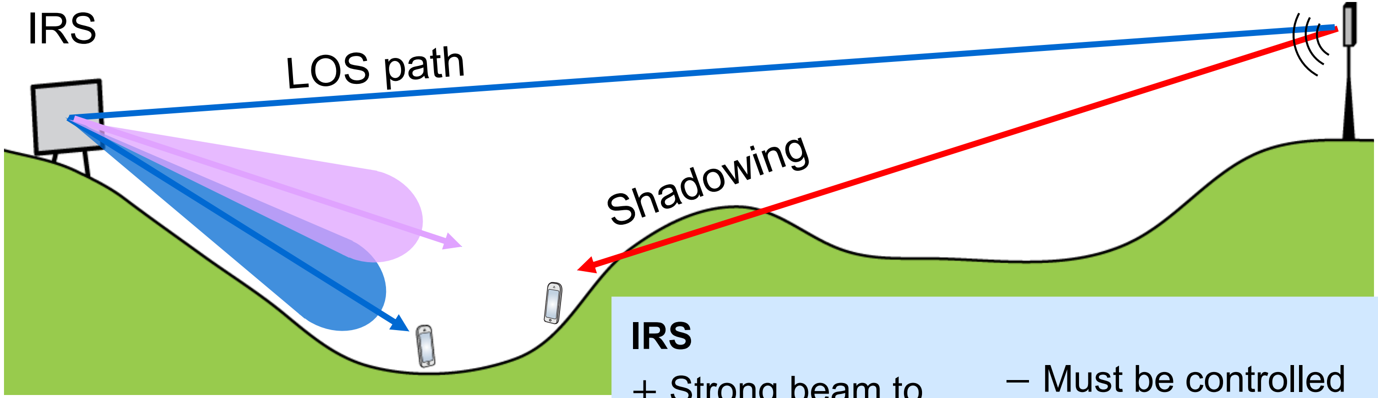
Passive Repeaters versus IRS



Passive repeater

+ Fully passive, no power

– Fixed beam pattern: wide or strong?



IRS

+ Strong beam to desired location

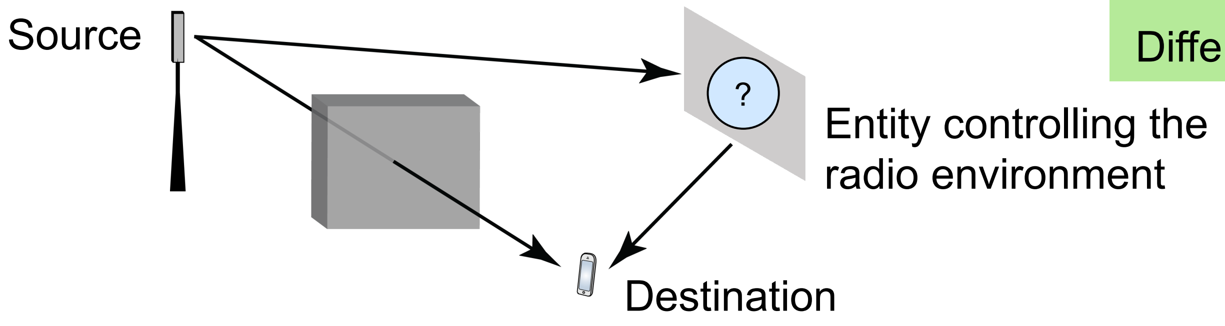
– Must be controlled

– Requires power

Taxonomy of “Cooperative Communications”

Not one or the other!

Different choice in different setups



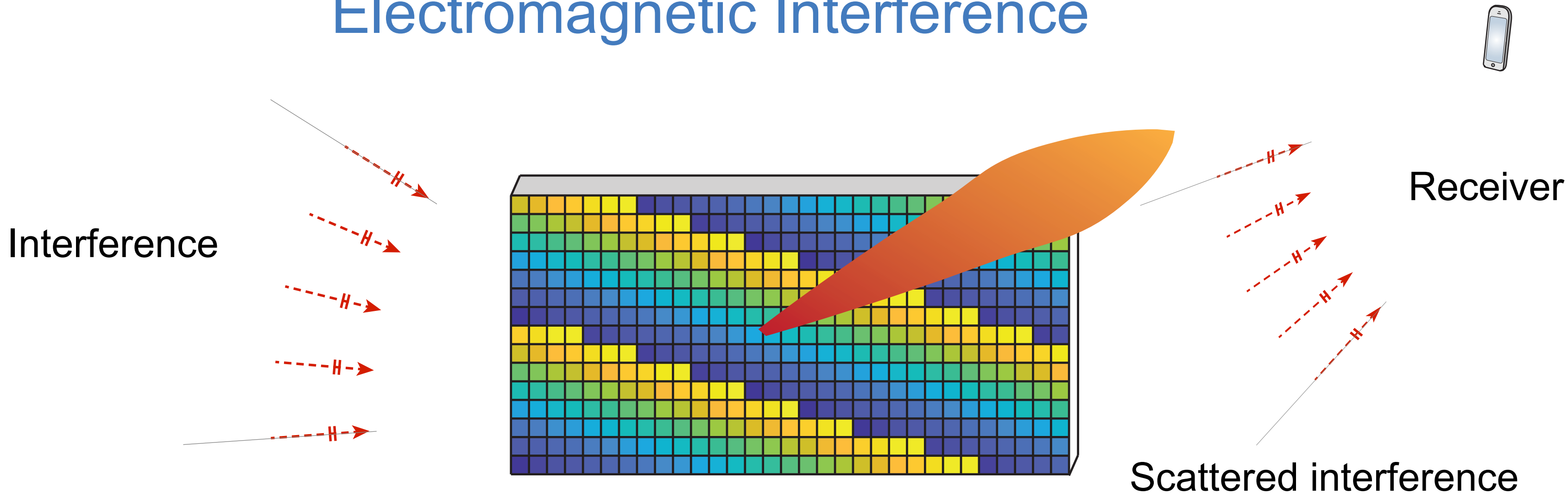
Used in cellular networks

Old-school analog repeater

	Regenerative	Transparent
Half duplex	Decode-and-forward (DF) relay	Amplify-and-forward (AF) relay
Full duplex	**AF and DF relay**	Passive repeater, **IRS**

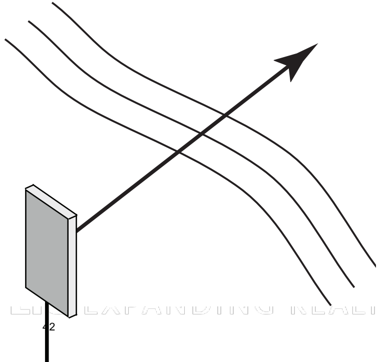
Emerging technologies

Electromagnetic Interference



Like painting a room with bright colors

Interference reflected by IRS proportional to N



IRS Might Reduce “SNR”

Simulation setup

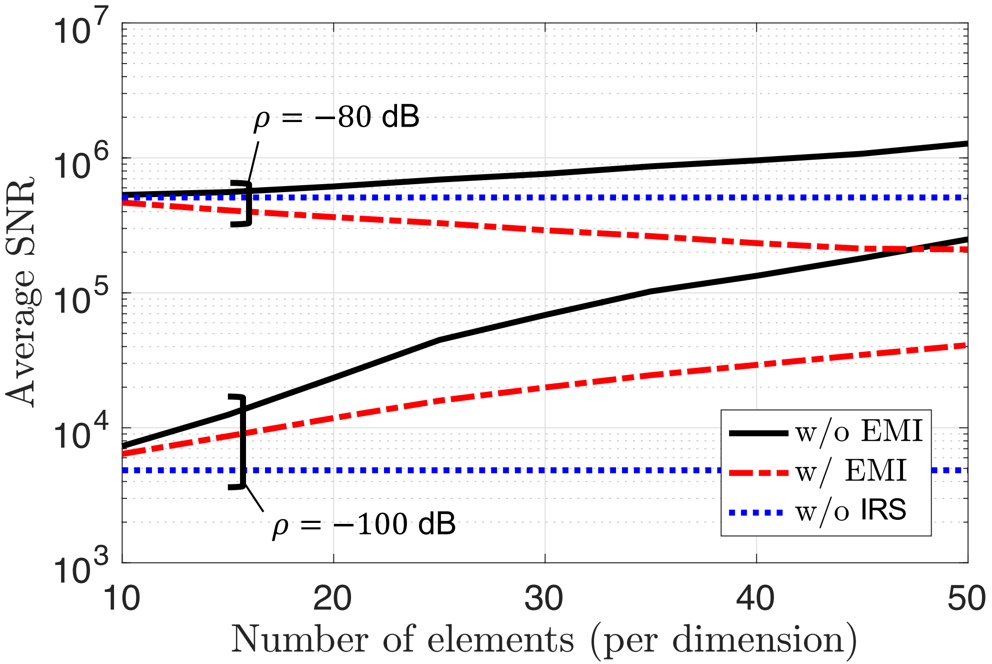
SNR at IRS elements: 20 dB

Isotropic electromagnetic interference (EMI)

IRS can sometimes reduce performance,
Asymptotic SNR proportional to N, not N^2

Careful deployment is necessary!

$$\mathrm{SNR} \approx \frac{P\left(\sqrt{\rho} + N\sqrt{\alpha\beta}\right)^2}{N \cdot \underbrace{\mathrm{constant}}_{\text{EMI}} + \sigma^2}$$



Reference: A. D. J. Torres, L. Sanguinetti, E. Björnson, “Electromagnetic Interference in RIS-Aided Communications,” IEEE Wireless Communications Letters, to appear.

The Most Promising Example

	NLOS to/from IRS	LOS to/from IRS
Strong direct path	IRS not effective in OFDM Difficult to estimate	IRS adds little to direct path
Weak direct path		Large gains in OFDM Few parameters to estimate

END OF PART 1