“Blind” Space-Frequency Beamforming for FFH & OFDM

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July 13, 2010

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Algorithmic Capabilities Offered

The proposed beamformer

- substantially improves the signal-to-interference-and-noise ratio (SINR):
  - could reject co-channel interference and noise,
  - could constructively summing the SOI.
- applies multi-dimensionally across receiving antennas & across hop-frequencies / subcarriers.
- operates "blindly", in the sense of requiring no prior information on
  - the SOI's angle-of-arrival,
  - the SOI's space-frequency propagation-channel impulse response,
    - any statistically decorrelated or independent fading across the antennas (i.e., the channel fading may be uncorrelated from antenna to antenna).
    - any spectral coherence of the channel's fading (i.e, the channel fading may be uncorrelated from frequency-hop bin to frequency-hop bin).
  - the other co-channel FFH-CDMA users' hop-sequences,
  - the other co-channel FFH-CDMA users' angles-of-arrival, nor
  - the receiving sensor-array's nominal or actual array manifold — the sensors may have unknown and arbitrary gain/phase/polarization responses and mutual coupling.

Algorithmic Philosophy

- To form two parallel sets of space-frequency data:
  1. The “S+I+N” data set contains the (signal-of-interest (SOI) plus any interference and noise.
  2. The “I+N” data set contains only the interference and noise.
- The “I+N” data set allows (implicit) estimation of the interference & noise in the “S+I+N” data set.

Key Question

How to obtain

- “S+I+N” data set vs. “I+N” data set ?
- for FFH-SS ? for OFDM?

Part I

“Blind” Beamformer for Fast Frequency-Hop (FFH) Spread Spectrum (SS) Signaling

FFH-SS Signal Model

For the kth user's sth symbol:

$$\sqrt{P_k}\sum_{h=1}^{H}\exp\left\{j\left[2\pi f_k^{(b_{k,s},h)}+\varphi_k^{(b_{k,s},h)}\right]\right\}u\left(t-(s-1)T_S-(h-1)T_H\right)$$

where

$b_{k,s}$ = information-symbol at the sth symbol for the kth user

$f_k^{(b_{k,s},h)}$ = hop-frequency at the hth hop-period for $b_{k,s}$

$\varphi_k^{(b_{k,s},h)}$ = initial phase at the hth hop-period for $b_{k,s}$

T_S = symbol-period = $H\,T_H$

T_H = hop-period

$u(\cdot)$ = symbol-pulse

P_k = power of the kth user

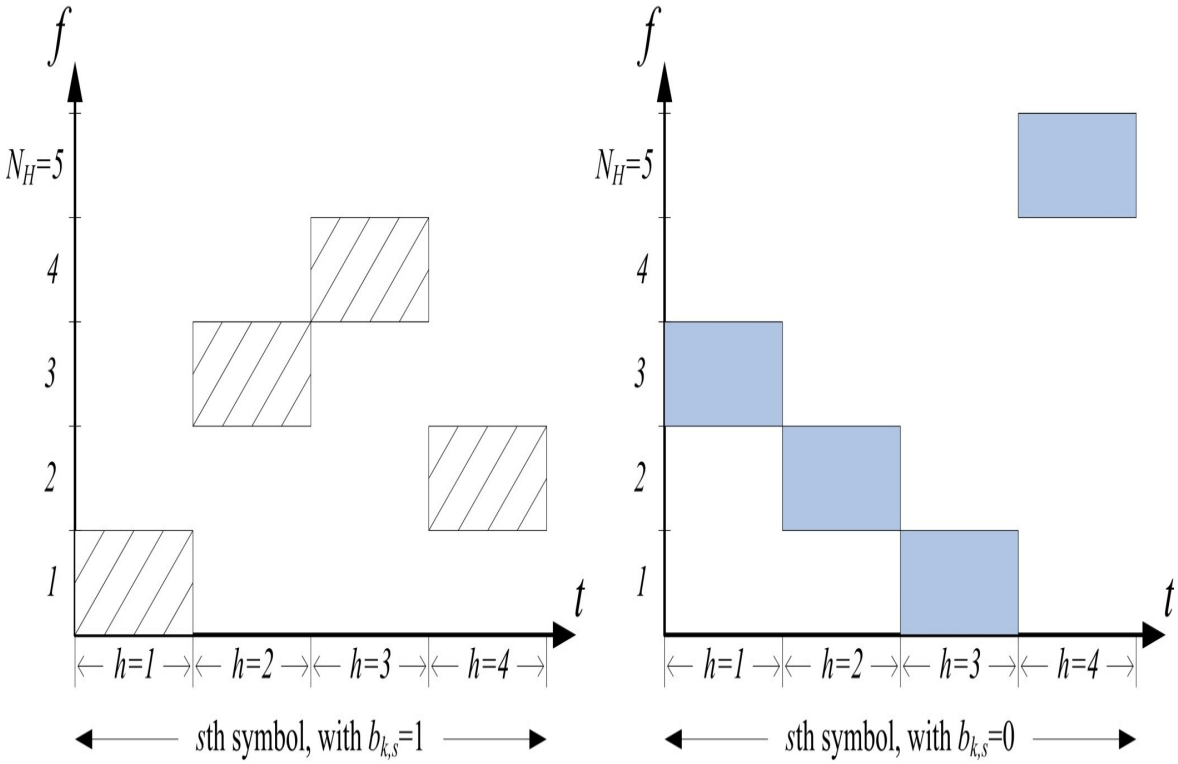


Figure 1:
Two sample hop-sequences for a hop-set of size N_H=5, and H=4 frequency-hops per symbol. For a binary constellation of M=2, the left diagram could correspond to a logical-*1* symbol-value, with the right diagram for logical-*0*.

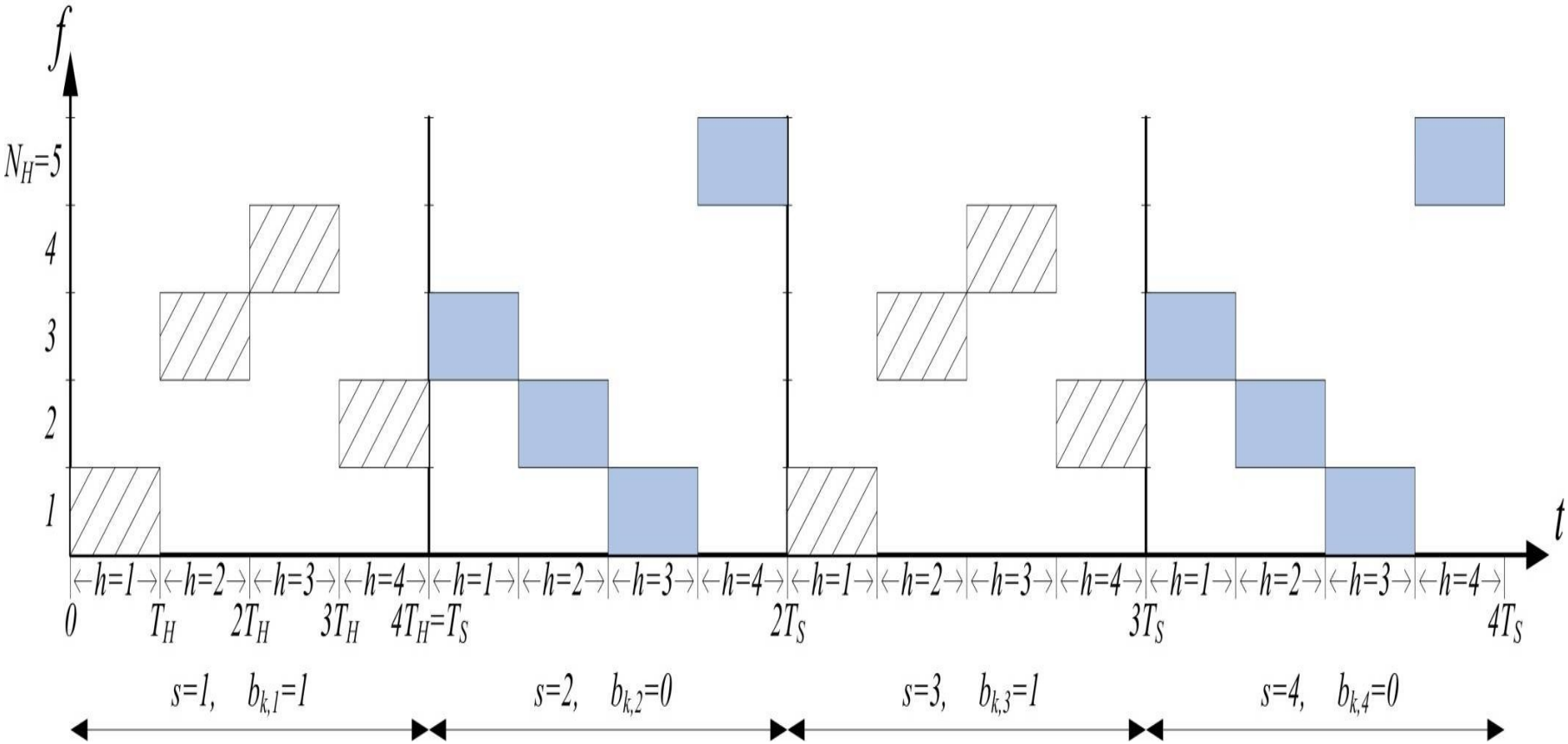


Figure 2:
$s_k(t)$ for the symbol-sequence of {$b_{k,s=1}$ =1 , $b_{k,s=2}$ =0, $b_{k,s=3}$ =1, $b_{k,s=4}$ =0}, using the hop-sequences of Figure 1.

Assumptions

- The transmitter's MFSK-FFH hopping-pattern is a priori known to the intended receiver.
- If the channel's coherence-time $\geq D$ symbols, the channel is statistically correlated over any $S < D$ symbols.
- The strongest co-channel interferers would persist for at least the next few information-symbols.

The Proposed "Blanking" Scheme

- Recall: The signal-of-interest (SOI) has H hop-periods / symbol.
- Proposed "Blanking":
  - "Blank" a different hop-period at each different symbol. I.e., during each symbol (= H hop-periods), transmits NO energy (i.e., to "blank") at 1 hop-period.
  - Over H symbol-periods, a different hop-period (out of the H hop-periods per symbol) is to be "blanked".
  - All H hop-periods would thus be "blanked" once, over the course of H consecutive information-symbols.

Figure 3:

The sample hop-sequences of Figure 1, after

“blanking” the *h*=*s* th hop-period at the *s*th symbol.

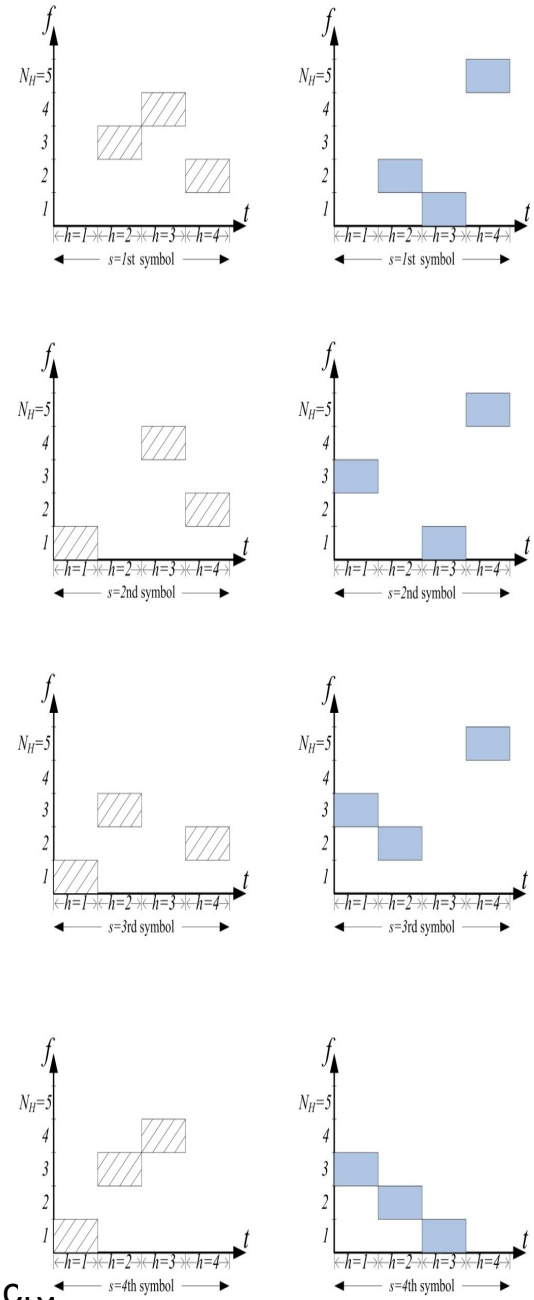
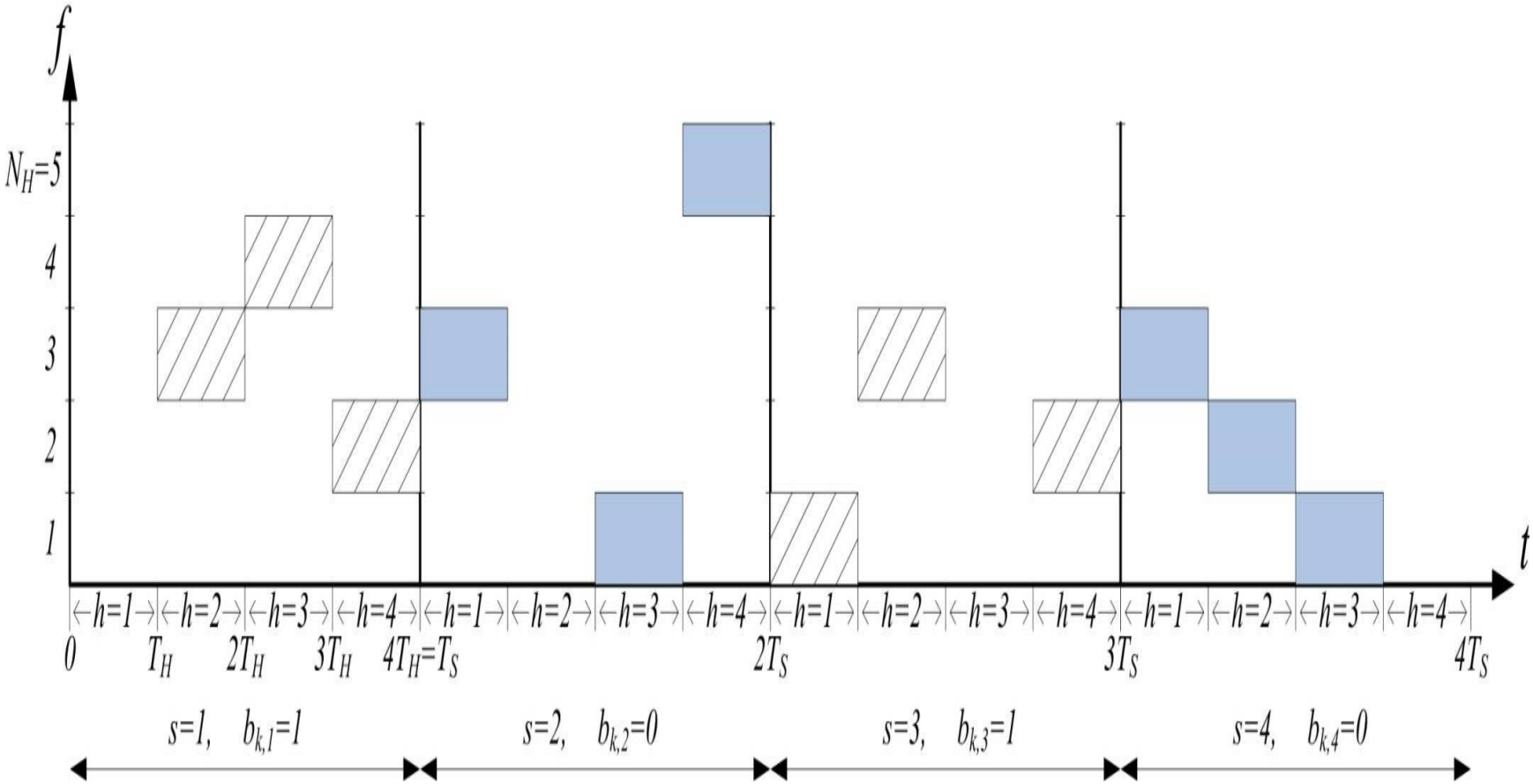


Figure 4: The counterpart to $s_k(t)$ of Figure 2, but after the "blanking" of Figure 1.



For a sequence of S number of M-ary symbols, there are MS different symbol-sequences $\{\mathbf{m}_q, q = 1, 2, \cdots, M\}$ as estimation-candidates of $\mathbf{b}_k$.

The Proposed "Blanking" Scheme

To form the "**I+N**" data set, $\mathbf{X}_q^{(I+N)}$:

Time-sample at each "blanked" hop-period, at only the hop-frequency of that particular candidate symbol-sequence, if there were no "blanking".

To form the "**S+I+N**" data set, $\mathbf{X}_q^{(S+I+N)}$:

Time-sample at each hop-period not "blanked", at where the hop-frequency would be for that particular candidate symbol-sequence.

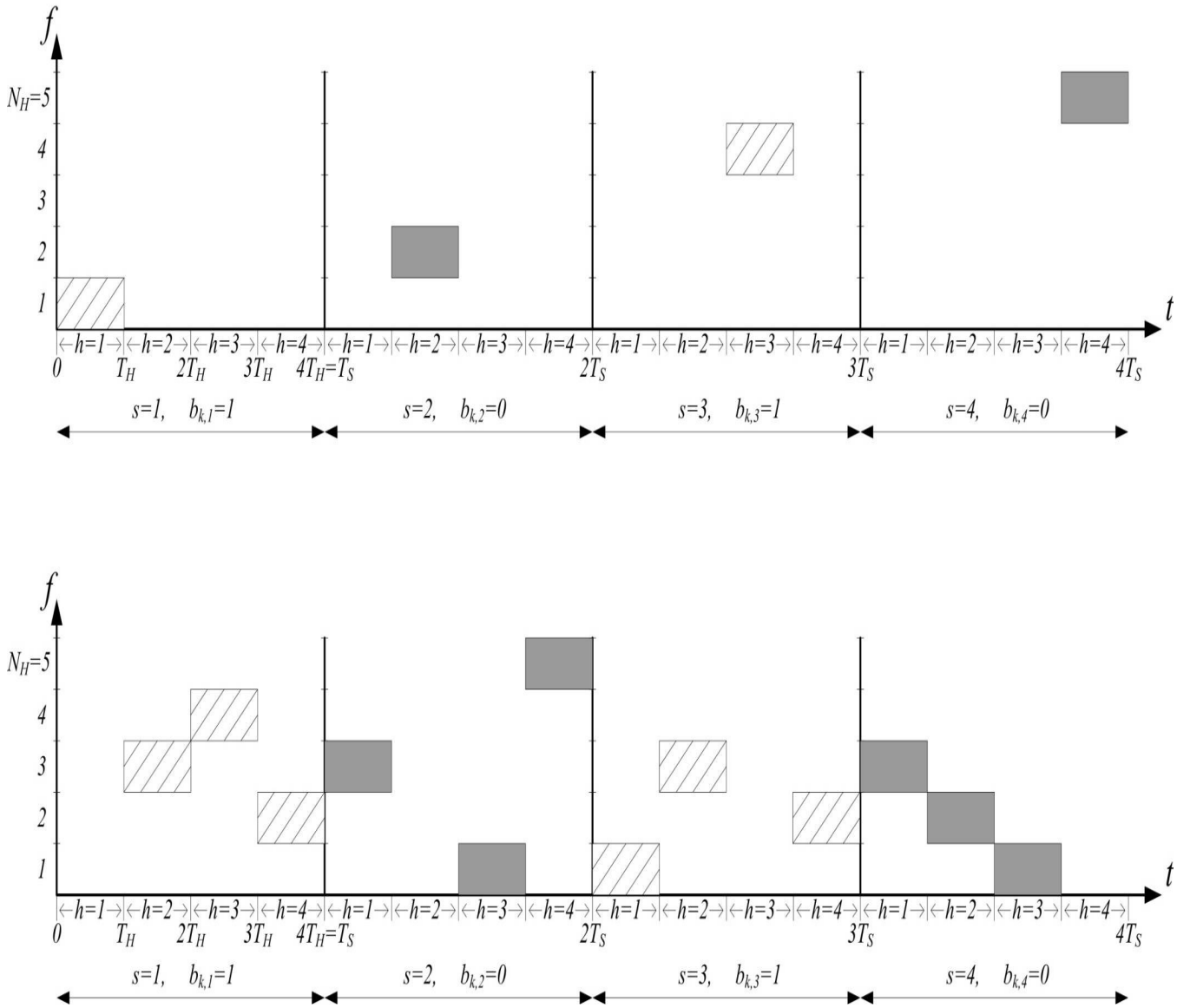


Figure 5: For candidate symbol-sequence of {$b_{k,s=1}$ =1, $b_{k,s=2}$ =0, $b_{k,s=3}$ =1, $b_{k,s=4}$ =0}, top diagram = “I+N” data-set, bottom diagram = “S+I+N” data-set.

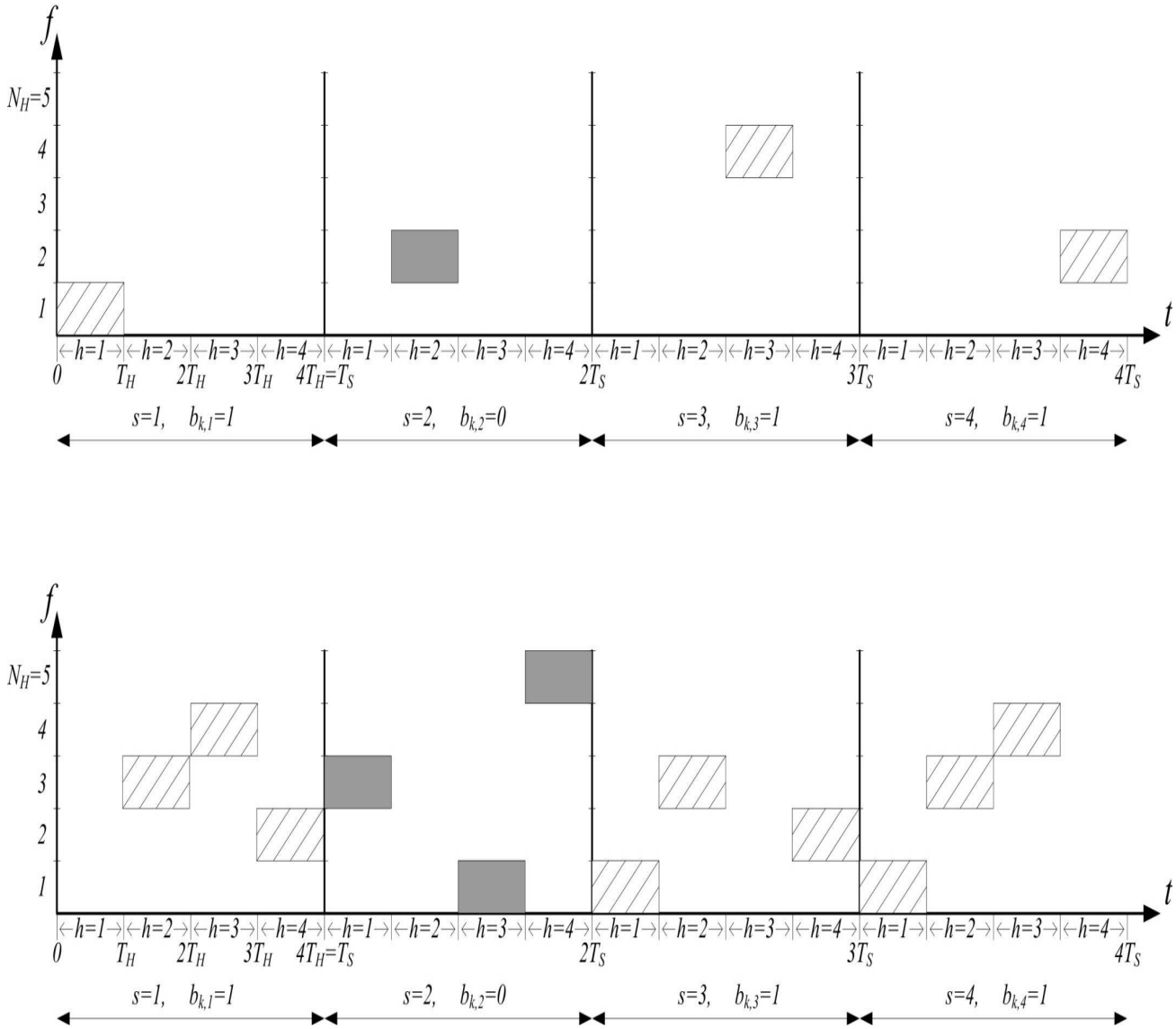


Figure 6: For candidate symbol-sequence of {$b_{k,s=1}$ =1, $b_{k,s=2}$ =0, $b_{k,s=3}$ =1, $b_{k,s=4}$ =1}, top diagram = “I+N” data-set, bottom diagram = “S+I+N” data-set.

The Proposed "Blanking" Scheme

Next, for the q th candidate symbol-sequence, do the following:

- Form the spatial covariance matrices

$$\mathbf{R}_q^{(\mathrm{S+I+N})} = \mathbf{X}_q^{(\mathrm{S+I+N})} \left[\mathbf{X}_q^{(\mathrm{S+I+N})}\right]^H$$

$$\mathbf{R}_q^{(\mathrm{I+N})} = \mathbf{X}_q^{(\mathrm{I+N})} \left[\mathbf{X}_q^{(\mathrm{I+N})}\right]^H$$

  (Each row / column = an antenna-frequency pairing.)

- Compute the principal generalized eigenvector $\mathbf{w}_q$ for the matrix-pencil pair { $\mathbf{R}_q^{(\mathrm{S+I+N})}$, $\mathbf{R}_q^{(\mathrm{I+N})}$ }. (Using this $\mathbf{w}_q$ as the beamformer weight vector, the SINR_q is maximized at the beam output.)

Lastly, the estimate of $\mathbf{b}_k$ is $\mathbf{m}_{q_{\max}}$, where $q_{\max}$ is the q that gives the largest SINR_q

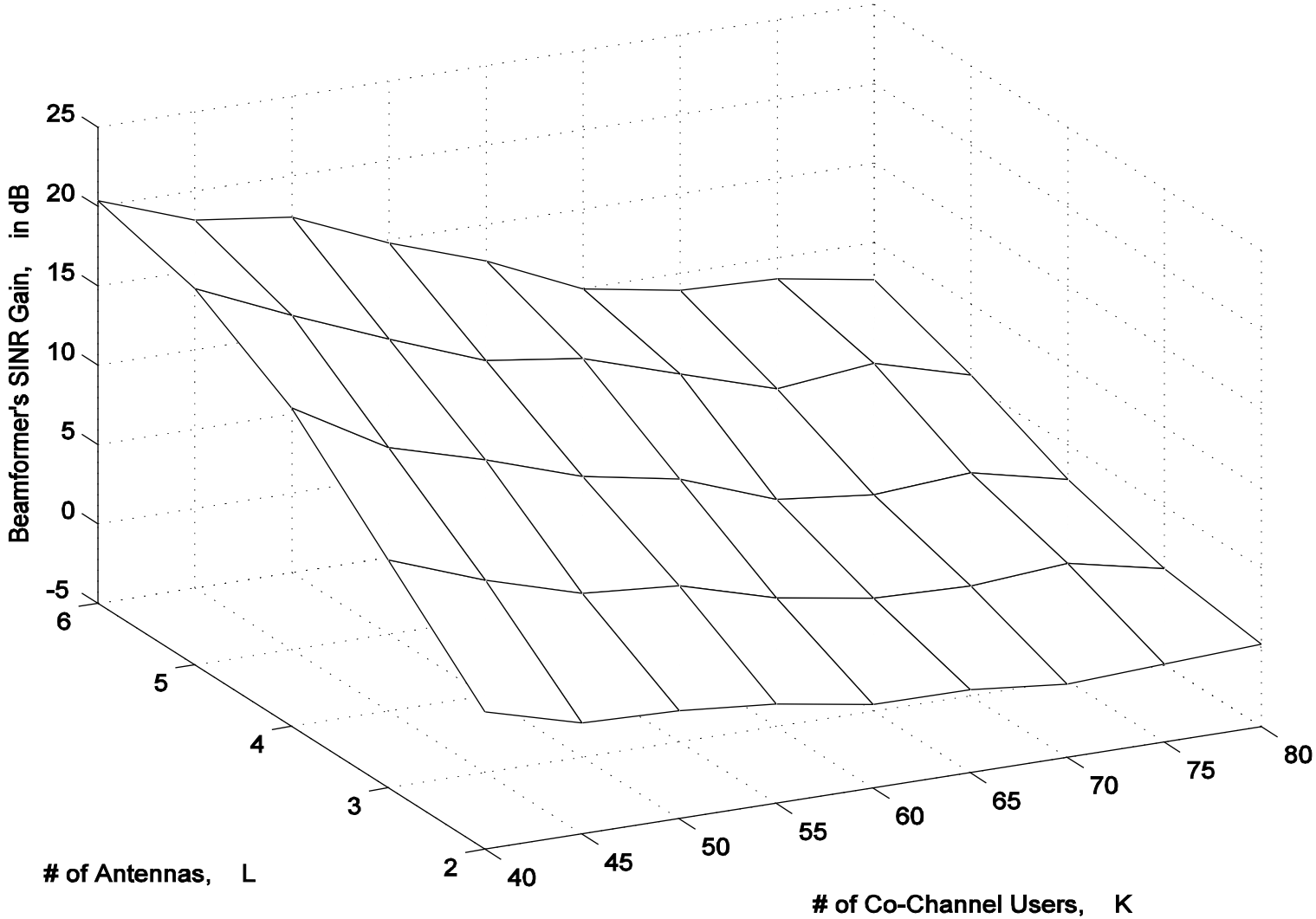


Figure 7:
The proposed “blind” space-frequency beamformer's SINR-gain (= output-SINR / input-SINR)

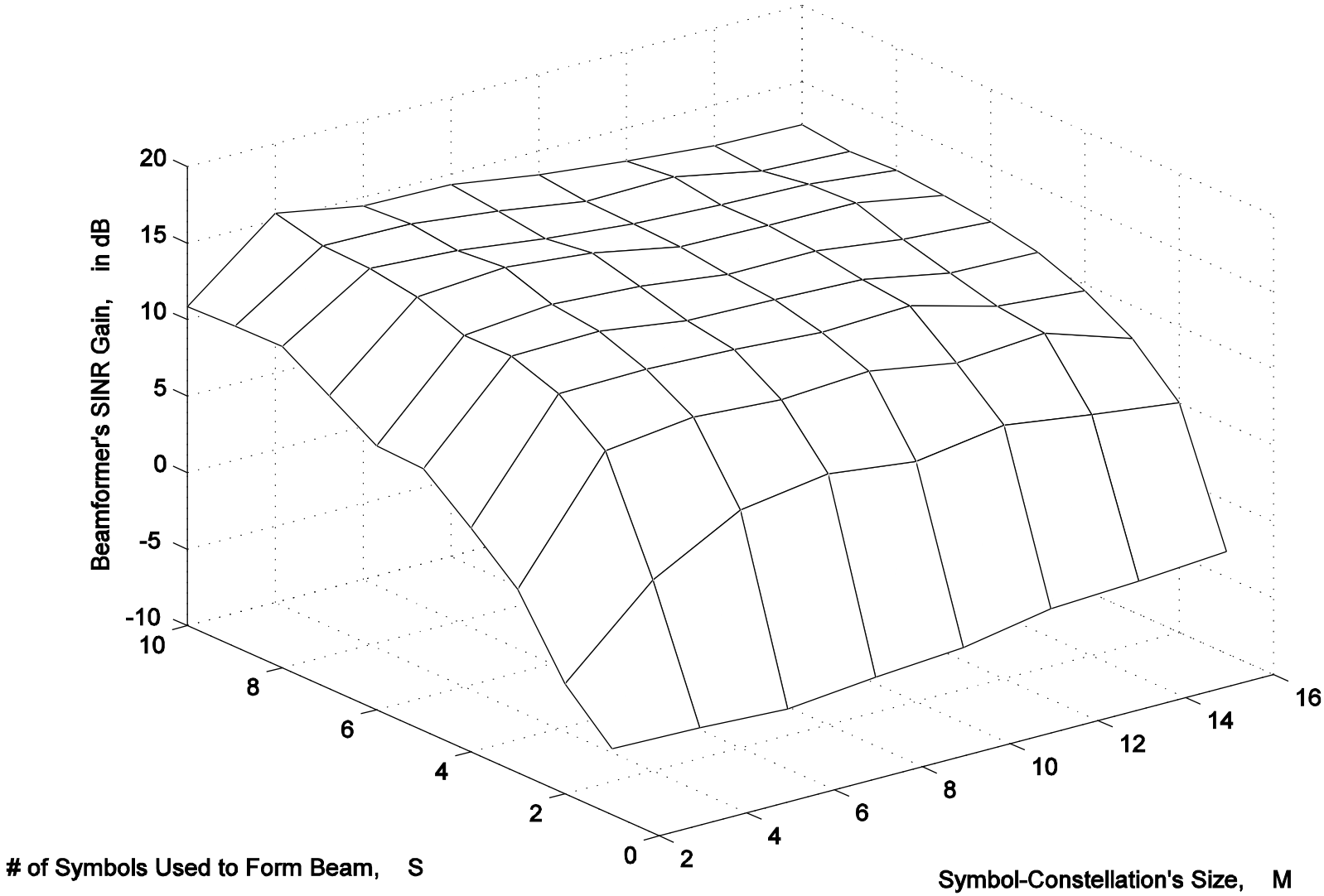


Figure 8:
The proposed “blind” space-frequency beamformer's SINR-gain (= output-SINR / input-SINR)

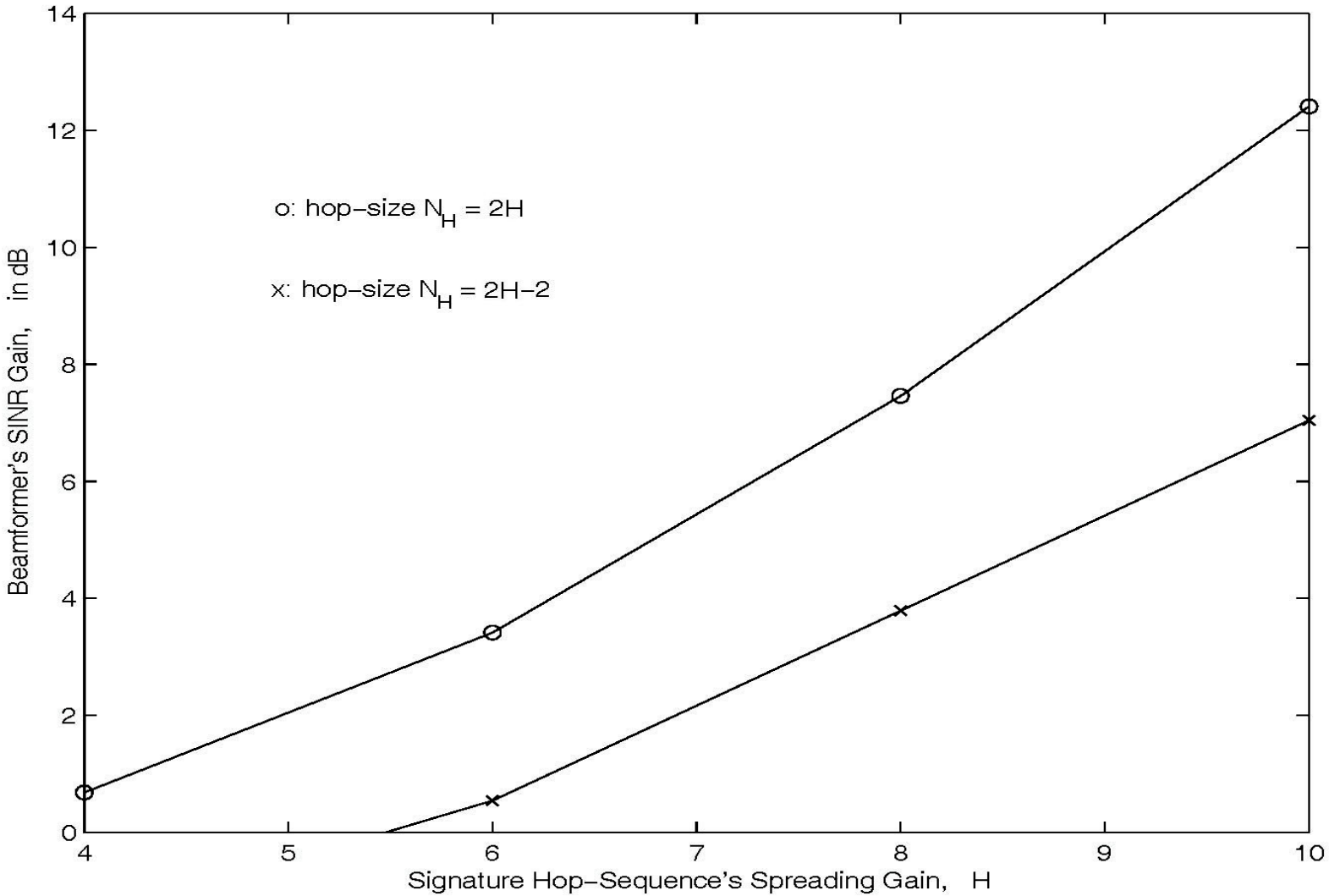


Figure 9:
The proposed “blind” space-frequency beamformer's SINR-gain (= output-SINR / input-SINR)

References

P. Tichavsky & **K. T. Wong**, "A Reduced-Complexity ‘Blind’ Space-Frequency Algorithm for Peer-to-Peer MFSK-FFH-CDMA Detection & Co-Channel Interference Suppression,” *IEEE Global Telecommunications Conference,* vol. 5, pp. 2437-2441, 2003.

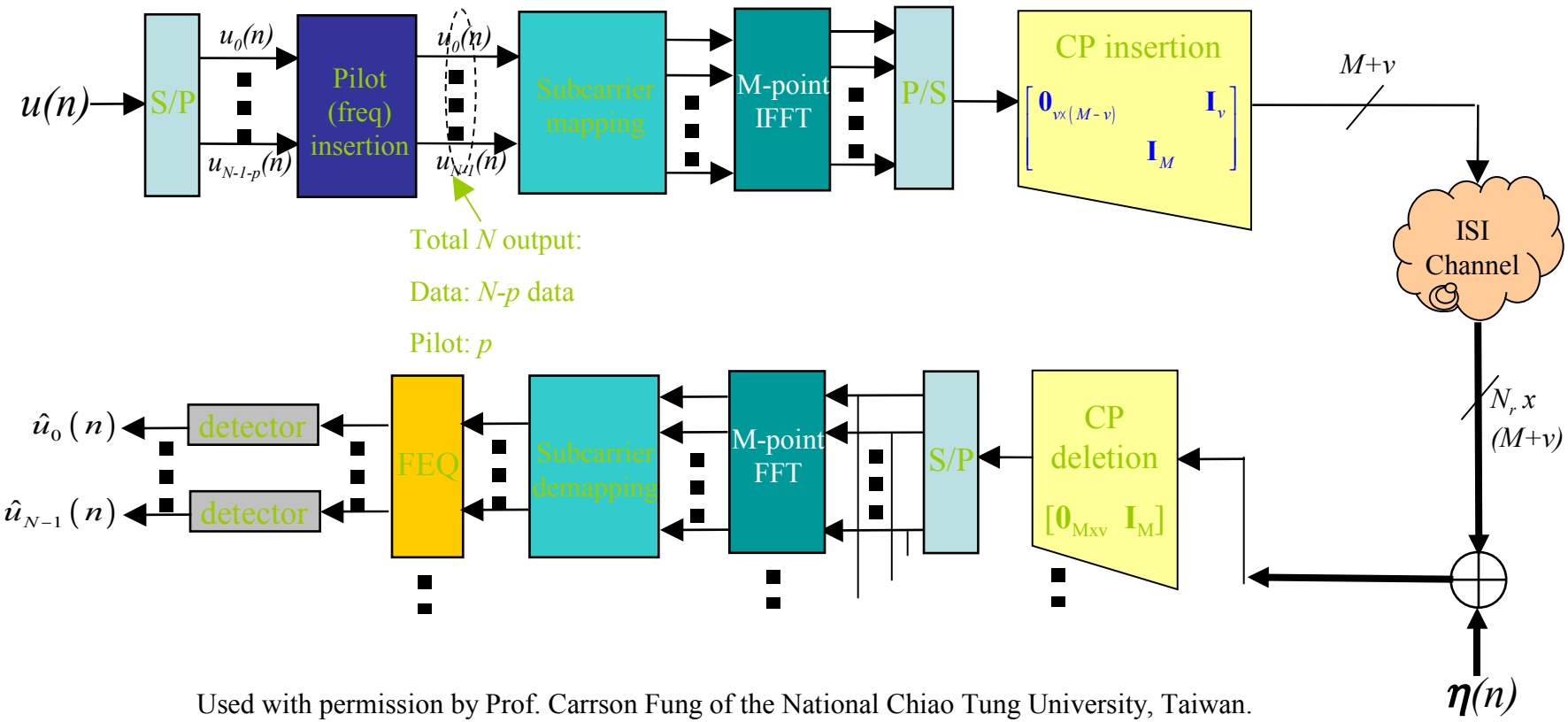
K. T. Wong, “A New Space-Frequency Beamspace ‘Blind’ Co-Channel Interference Suppression & Detection Algorithm for MFSK-FFH-CDMA Peer-to-Peer Mobile Communications,” *IEEE Global Telecommunications Conference,* vol. 2, pp. 1950-1954, 2002.

K. T. Wong, “A New Transmission/Reception Scheme for “Blind” Suppression of Co-Channel Interference in FH-CDMA Peer-to-Peer Mobile Communications,” *IEEE Military Communications Conference, vol. 2,* pp. 1229-1232, 2001.

Part II

“Blind” Beamformer for Offset-Frequency Division Multiplexing (OFDM)

Review of OFDM System



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Problem with Cyclic Prefix (CP)

- Function: IFFT & FFT need to avoid inter-symbol interference (ISI), i.e., need circular convolution (not linear convolution), made possible by a "cyclic prefix" (CP) that exceeds the channel's delay spread.
- Problem:
  - The CP displaces data, hence reduces the data-rate.
  - Without a CP,
    - There would be ISI and ICI.
    - Orthogonality would be lost among the subcarriers.
    - The simple one-tap FEQ no longer applicable.

PROPOSED SCHEME'S ADVANTAGES

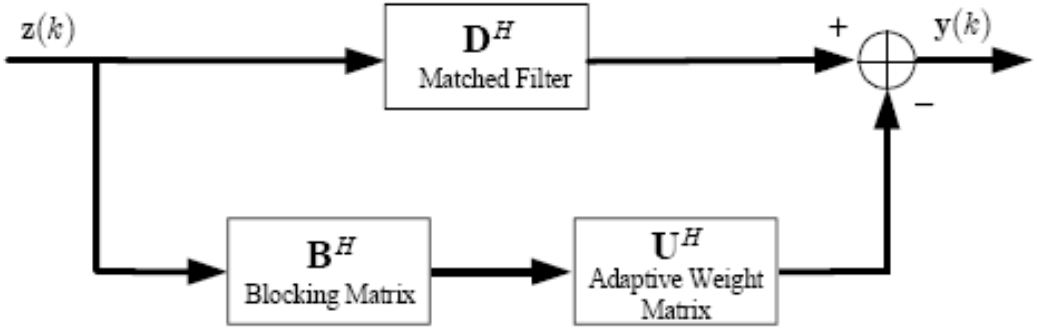
The proposed scheme can

- allow the "cyclic prefix" (CP) to be "*in*sufficiently" long for the channel's delay spread (to increase data-rate).

- blindly suppress many types of interference, e.g.,
  - Inter-carrier interference (ICI),
  - Inter-symbol interference (ISI),
  - co-channel interference,
  - Interference from overlaid systems.

But the proposed scheme needs perfect prior knowledge of channel's per-tap gain, for ***D****.*

“S+I+N” Dataset & “I+N” Dataset



(a)

$\mathbf{z}(k)$: $N \times 1$ data-vector collected at the N subcarriers at kth time-sample.

$\mathbf{D}$: $N \times N$ diagonal matrix, each element = one-tab equalizer, assumed prior known

$\mathbf{B}, \mathbf{U}$: to-be-designed.

Motivation:

- Upper leg: passes “ SOI + I + N ”, while equalizing SOI.
- Lower Leg: ideally to pass “ I + N ” only, by a suitable design of **B** and **U**.
- Hence, to subtract 2) from 1) to estimate SOI.

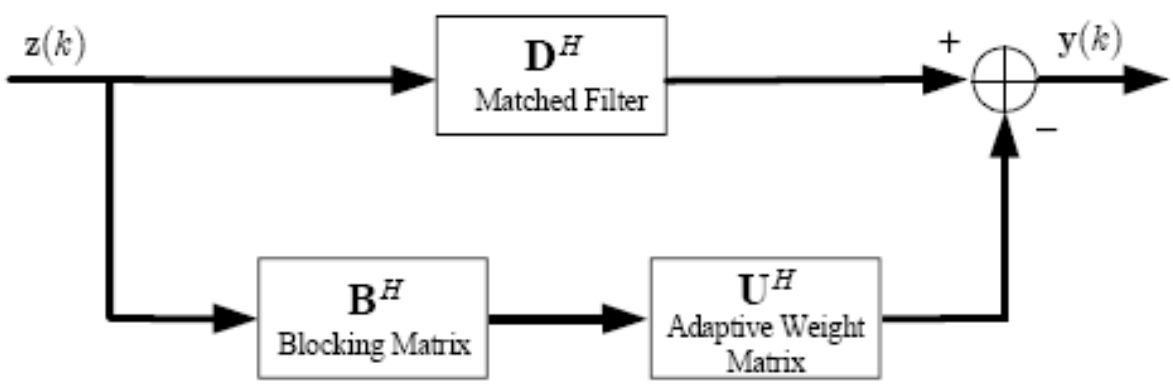
Lower Leg: Uses <u>null subcarriers</u> (a.k.a. virtual / unused / unmodulated subcarriers) , already mandated in most wireless OFDM standards.

B is a selection matrix,

- with 1's at the null subcarriers, to pass the " I + N ", but
- with 0's to suppress the data-subcarriers, to block the SOI.

$$\mathbf{U} = \arg\min_{\mathbf{U}} E\left\|\mathbf{i}(k) - \mathbf{U}^H\mathbf{B}^H\mathbf{x}(k)\right\|_2^2 = \left(\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{B}\right)^{-1}\mathbf{B}^H\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)}\mathbf{D}$$

$$\mathbf{R}_{\mathbf{i}(k)\mathbf{i}(k)} = \mathbf{H}_{ICI}\mathbf{R}_{\mathbf{s}(k)\mathbf{s}(k)}\mathbf{H}_{ICI}^H + \mathbf{H}_{ISI}\mathbf{R}_{\mathbf{s}(k-1)\mathbf{s}(k-1)}\mathbf{H}_{ISI}^H + \mathbf{W}_N\mathbf{R}_{\mathbf{n}(k)\mathbf{n}(k)}\mathbf{W}_N^H$$



(a)

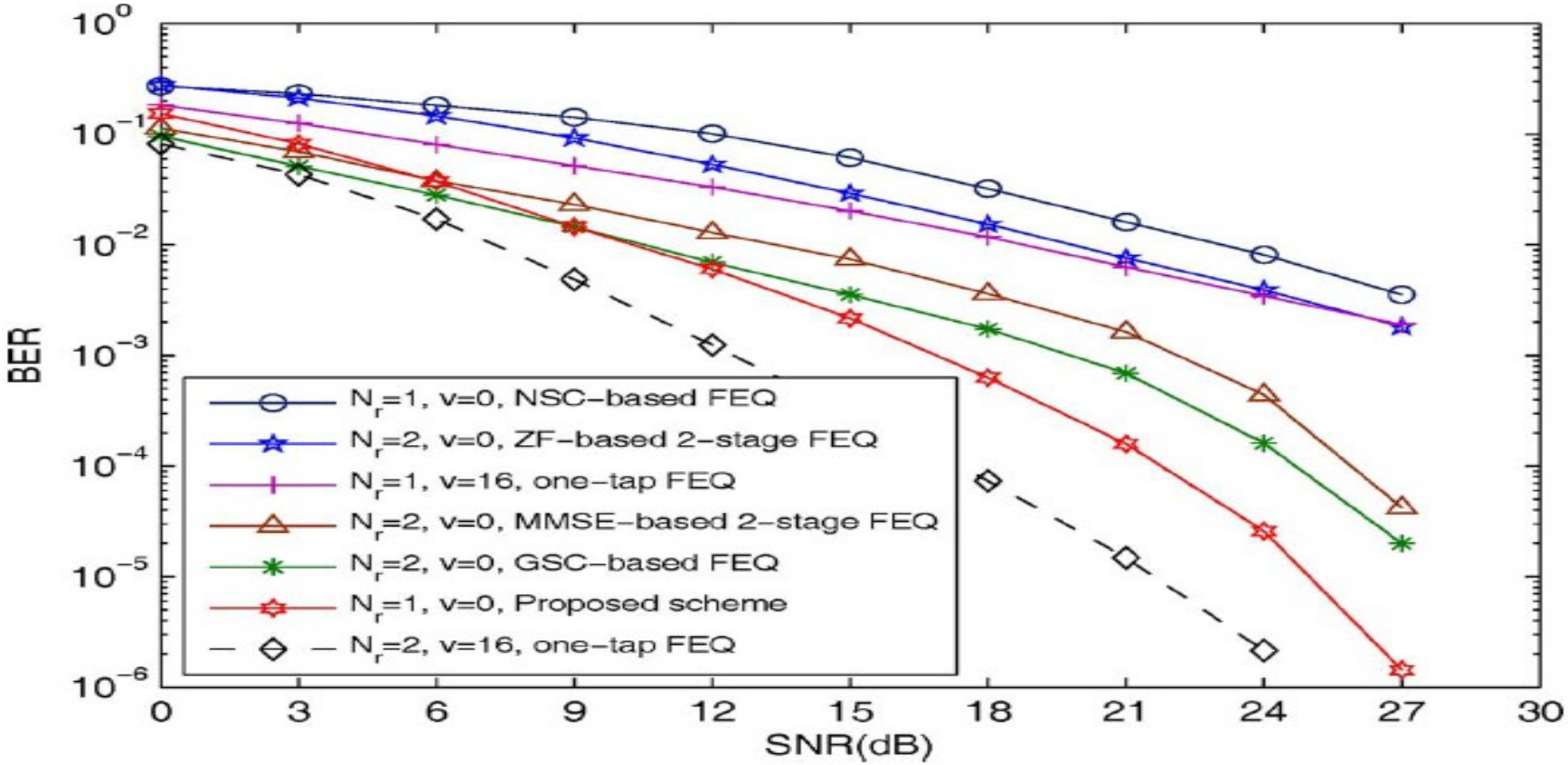


Fig. 1. BER versus SNR performance for competing equalizers: NSC-based FEQ [3], ZF-based 2-stage FEQ [10], MMSE-based 2-stage FEQ [9], and GSC-based FEQ [8].

SYMBOLS:

- v = CP length
- N_r = no. of receive-antennas
- p = 12 null subcarriers
- q = channel delay-spread = 10

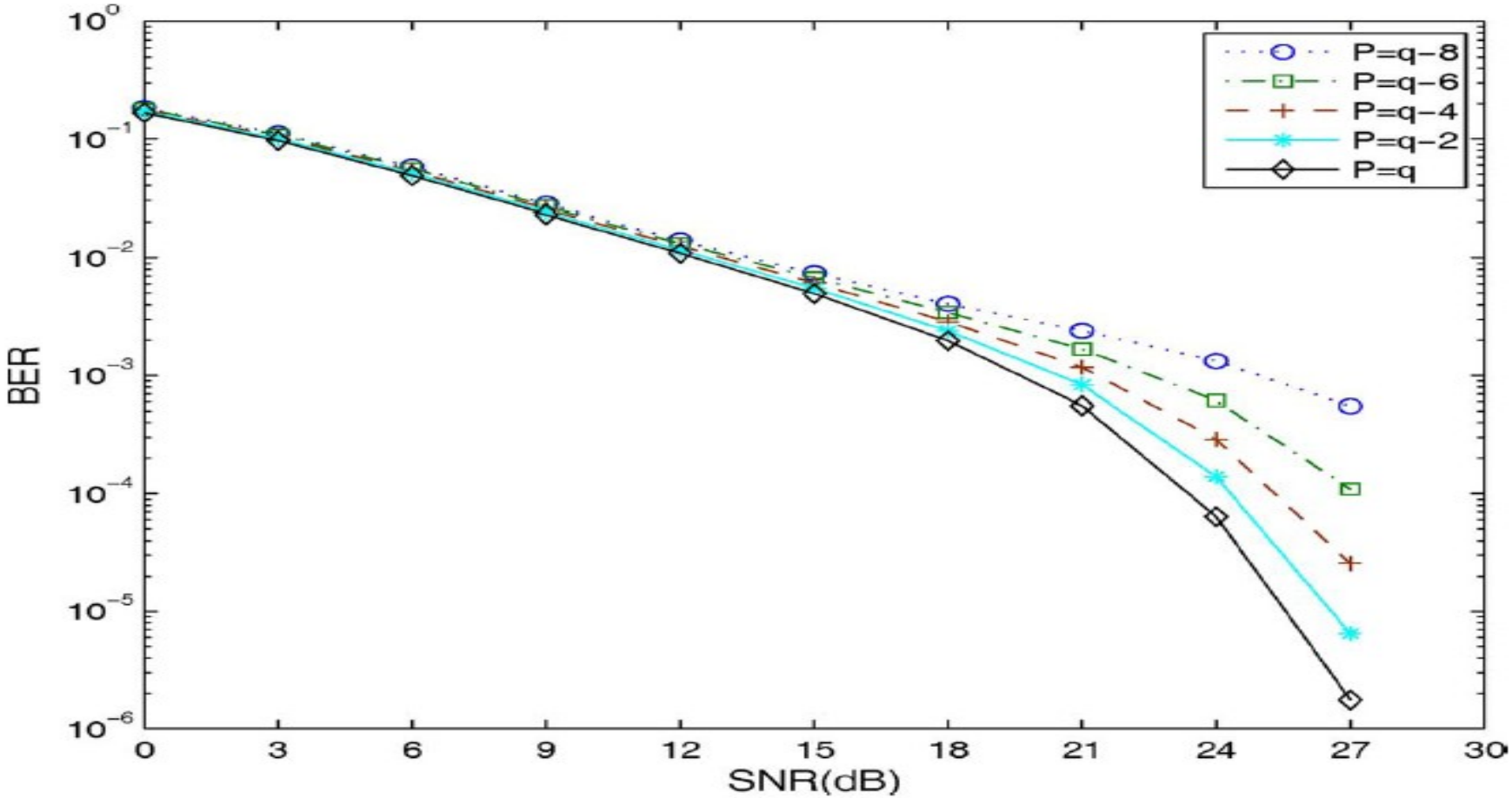


Fig. 2. BER vs. SNR for different number of NSC (P) for the proposed FEQ (no CP).

SYMBOLS:

- v = CP length = 0
- N_r = no. Of receive-antennas = 1
- p = no. of null subcarriers
- q = channel delay-spread = 10

References

Y.-R. Huang, C. C. Fung & **K. T. Wong**, “Interference Suppression for OFDM Systems With Insufficient Guard Interval Using Null Subcarriers,” *IEEE Signal Processing Letters*, vol. 16, no. 11, pp. 929-932, November 2009.

Y.-R. Huang, C. Fung & **K. T. Wong**, “Frequency Domain Equalization for OFDM Systems With Insufficient Guard Interval Using Null Subcarriers,” *European Signal Processing Conference*, Glasgow, Scotland, U.K., 24-28 August, 2009.

Thank you !