

System Design and Signal Processing for Integrated Sensing and Communications

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Abstract

In addition to progressively scarce spectral resources and convergent hardware development, the increasing demand for higher data rates in communications systems and the need for coordinating non-collocated users in modern radar systems have driven efforts towards the development of integrated sensing and communications (ISAC) systems. In this context, an increasing number of research efforts towards enabling the use of a single waveform for simultaneously performing radar sensing and communications has been reported in the literature. While a significant number of studies addresses the issues related to waveform design, countermeasures to hardware impairments, and distributed sensing, a systematic analysis of ISAC systems both exposing and discussing details related to the mentioned challenges, besides proposing solutions and demonstrating them via meaningful simulations and verification measurements has yet not been reported. This gap is filled by the present dissertation.

In this context system design considerations and signal processing techniques to enable future ISAC systems are proposed, discussed, and verified. More specifically, a comprehensive description of processing chains for various modulation schemes, including phase-modulated continuous wave (PMCW), orthogonal frequency-division multiplexing (OFDM), and orthogonal chirp-division multiplexing (OCDM), introducing new concepts and signal processing techniques. A comparative performance analysis based on simulation and measurement results is presented, highlighting the suitability of PMCW and sector-modulated OCDM (SM-OCDM) for radar-centric applications where radar sensing performance is prioritized, and OFDM and conventional OCDM (C-OCDM) for communications-centric systems in which communications performance has priority over radar sensing.

Focusing on communications-centric ISAC based on OFDM, which is expected to become a reality in cellular networks in their sixth generation (6G) and beyond, this dissertation introduces a bistatic ISAC system concept, where radar signal processing strategies and a robust synchronization procedure to

accurately estimate and correct offsets, therefore enabling coherent radar signal processing, are included for the first time. Furthermore, a detailed discussion on hardware impairments, their effects on ISAC performance, and potential countermeasures via signal processing or during system design is also presented. Finally, an extension of the proposed bistatic OFDM-based ISAC system to multiple-input multiple-output (MIMO) configurations is proposed, and a signal processing chain as well as a distributed synchronization concept to ensure coherence among receive channels are introduced for the first time. Validated by simulations and verification measurements, the capabilities of the proposed single-input single-output (SISO) and MIMO bistatic OFDM-based ISAC system concepts and signal processing techniques are demonstrated under realistic conditions, highlighting their potential to push the development next-generation cellular-infrastructure-based sensing forward.

In conclusion, this dissertation addresses a multitude of relevant theoretical and practical challenges, contributing with system design considerations and novel signal processing concepts and algorithms to enabling future ISAC systems.

Zusammenfassung

Neben zunehmend knappen spektralen Ressourcen und der konvergenten Hardwareentwicklung haben die steigende Nachfrage nach höheren Datenraten in Kommunikationssystemen sowie die Notwendigkeit der Koordination nicht-kollozierter Nutzer in modernen Radarsystemen die Entwicklung integrierter *Sensing*- und Kommunikationssysteme (*Integrated Sensing and Communications*, ISAC) vorangetrieben. In diesem Zusammenhang wurde in der Literatur eine wachsende Anzahl von Forschungsarbeiten veröffentlicht, die darauf abzielen, die Nutzung einer einzigen Wellenform zur gleichzeitigen Durchführung von *Sensing* und Kommunikation zu ermöglichen. Während sich zahlreiche Studien mit Fragestellungen zur Wellenformentwicklung, mit Gegenmaßnahmen gegen Hardwarebeeinträchtigungen sowie mit verteiltem *Sensing* befassen, fehlt bislang eine systematische Analyse von ISAC-Systemen, die die genannten Herausforderungen sowohl offenlegt als auch detailliert diskutiert und darüber hinaus Lösungen vorschlägt sowie diese mittels aussagekräftiger Simulationen und Verifikationsmessungen demonstriert. Diese Lücke wird durch die vorliegende Dissertation geschlossen.

In diesem Kontext werden Systementwurfsaspekte und Signalverarbeitungstechniken zur Realisierung zukünftiger ISAC-Systeme vorgeschlagen, diskutiert und verifiziert. Insbesondere wird eine umfassende Beschreibung von Verarbeitungsketten für verschiedene Modulationsverfahren präsentiert, darunter *Phase-Modulated Continuous Wave* (PMCW), *Orthogonal Frequency-Division Multiplexing* (OFDM) und *Orthogonal Chirp-Division Multiplexing* (OCDM), wobei neue Konzepte und Signalverarbeitungstechniken eingeführt werden. Eine vergleichende Performance-Analyse auf Basis von Simulations- und Messergebnissen wird vorgestellt, die die Eignung von PMCW und *Sector-Modulated OCDM* (SM-OCDM) für radarzentrierte Anwendungen hervorhebt, bei denen die Radar-Performance priorisiert wird, sowie von OFDM und *Conventional OCDM* (C-OCDM) für kommunikation-

szentrierte Systeme, in denen die Kommunikation-Performance Vorrang vor *Sensing* hat.

Mit Fokus auf kommunikationszentrierte, OFDM-basierte ISAC-Systeme, deren Realisierung in zellulären Netzen der sechsten Generation (6G) und darüber hinaus erwartet wird, führt diese Dissertation erstmals ein bistatisches ISAC-Systemkonzept ein. Dabei werden Radar-Signalverarbeitungsstrategien sowie ein robustes Synchronisationsverfahren zur präzisen Schätzung und Korrektur von Offsets integriert, wodurch eine kohärente Radar-Signalverarbeitung ermöglicht wird. Darüber hinaus wird eine detaillierte Diskussion von Hardwarebeeinträchtigungen, deren Auswirkungen auf die *ISAC-Performance* sowie potenzieller Gegenmaßnahmen durch Signalverarbeitung oder bereits im Systementwurf präsentiert. Schließlich wird eine Erweiterung des vorgeschlagenen bistatischen OFDM-basierten ISAC-Systems auf *Multiple-Input-Multiple-Output*-(MIMO)-Konfigurationen vorgeschlagen, und erstmals werden eine entsprechende Signalverarbeitungskette sowie ein verteiltes Synchronisationskonzept zur Sicherstellung der Kohärenz zwischen den Empfangskanälen eingeführt. Validiert durch Simulationen und Verifikationsmessungen werden die Fähigkeiten der vorgeschlagenen *Single-Input-Single-Output*-(SISO)- und MIMO-bistatischen OFDM-basierten ISAC-Systemkonzepte sowie der Signalverarbeitungstechniken unter realistischen Bedingungen demonstriert und deren Potenzial zur Weiterentwicklung infrastrukturbasiertes *Sensings* der nächsten Generation in zellulären Netzen hervorgehoben.

Zusammenfassend adressiert diese Dissertation eine Vielzahl relevanter theoretischer und praktischer Herausforderungen und leistet mit Systementwurfsspekten sowie neuartigen Signalverarbeitungskonzepten und -algorithmen einen Beitrag zur Realisierung zukünftiger ISAC-Systeme.

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Declaration

The author declares that the findings reported in this work are the result of his original research. The manuscript's clarity and language have been enhanced with the assistance of AI-based tools, including generative AI.

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Acronyms and Symbols

Acronyms

2D	two-dimensional
3GPP	3rd Generation Partnership Project
5G	fifth generation
6G	sixth generation
A/D	analog-to-digital
ABE	analog back-end
ACDC	all-cell Doppler correction
ADC	analog-to-digital converter
AFDM	affine frequency-division multiplexing
AFE	analog front-end
AGV	automatic guided vehicle
AM-AM	amplitude-to-amplitude modulation
AM-PM	amplitude-to-phase modulation
AWGN	additive white Gaussian noise
BB	baseband
BER	bit error ratio

BP	bandpass
BPSK	binary phase-shift keying
C-OCDM	conventional OCDM
CCDF	complementary cumulative distribution function
CDM	code-division multiplexing
CFO	carrier frequency offset
CFR	channel frequency response
CIR	channel impulse response
CP	cyclic prefix
CPE	common phase error
CP-DSSS	CP direct-sequence spread spectrum
CRLB	Cramér–Rao lower bound
CS	chirp sequence
CSI	channel state information
CW	continuous wave
CZT	chirp Z-transform
D/A	digital-to-analog
DAC	digital-to-analog converter
DC	direct current
DDC	digital down-conversion
DFnT	discrete Fresnel transform
DFRC	dual-function(al) radar-communication

DFT	discrete Fourier transform
DL	downlink
DMRS	demodulation reference signal
DoA	direction of arrival
DoD	direction-of-departure
DPD	digital pre-distortion
DUC	digital up-conversion
ETSI	European Telecommunications Standards Institute
EVM	error vector magnitude
FBMC	filter bank multicarrier
FDE	frequency-domain equalization
FDM	frequency-division multiplexing
FDZP	frequency-domain zero padding
FFT	fast Fourier transform
FMCW	frequency-modulated continuous wave
FPGA	field programmable gate array
FR2	Frequency Range 2
FrDM	Fresnel-division multiplexing
FSP	frequency-shift precoding
GFDM	generalized frequency-division multiplexing
gNB	gNodeB
HAD	highly automated driving

HPBW	half-power beamwidth
I/Q	in-phase/quadrature
IBFD	in-band full duplex
IBO	input back-off
IC	interference cancellation
ICHl	interchirp interference
ICI	intercarrier interference
IDFT	inverse discrete Fourier transform
IDFnT	inverse discrete Fresnel transform
IF	intermediate frequency
IFFT	inverse fast Fourier transform
IHE	Institute of Radio Frequency Engineering and Electronics
IIP3	input-referred third-order intercept point
IM	index modulation
IP1dB	input-referred 1-dB compression point
ISAC	integrated sensing and communications
ISI	intersymbol interference
ISLR	integrated sidelobe level ratio
JCAS	joint communication and sensing
KIT	Karlsruhe Institute of Technology
LDPC	low-density parity-check
LFSR	linear-feedback shift register

LNA	low-noise amplifier
LO	local oscillator
LoS	line-of-sight
LP	low-pass
LS	least squares
LUT	lookup table
MIMO	multiple-input multiple-output
MLE	maximum likelihood estimator
MLS	maximum-length sequence
MMSE	minimum mean square error
mmWave	millimeter wave
MRC	maximal-ratio combining
MU	multi-user
MUSIC	multiple signal classification
NBI	narrowband interference
NLoS	non-line-of-sight
NR	New Radio
OCDM	orthogonal chirp-division multiplexing
OFDM	orthogonal frequency-division multiplexing
OIP3	output-referred third-order intercept point
OOB	out-of-band
OP1dB	output-referred 1 dB compression point

OTA	over-the-air
OTFS	orthogonal time-frequency space
P/S	parallel-to-serial
PA	power amplifier
PACF	periodic autocorrelation function
PAPR	peak-to-average power ratio
PCCF	periodic cross-correlation function
PDF	probability density function
PLL	phase-locked loop
PLS	physical layer security
PMCW	phase-modulated continuous wave
PMEPR	peak-to-mean envelope power ratio
PMN	perceptive mobile network
PN	oscillator phase noise
PoC	proof-of-concept
PPLR	peak power loss ratio
PRBS	pseudorandom binary sequence
PRS	positioning reference signal
PSD	power spectral density
PSLR	peak sidelobe level ratio
PSS	primary synchronization signal
PT-RS	phase tracking reference signal

QAM	quadrature amplitude modulation
QPSK	quadrature phase-shift keying
RadCom	radar-communication
RCS	radar cross section
RF	radio-frequency
RFT	Radon-Fourier transform
RIS	reflective intelligent surface
RMS	root mean square
RMSE	root mean squared error
RTS	radar target simulator
RU	radio unit
S/P	serial-to-parallel
SARA	sampling and reconstructing angular domains
S&C	Schmidl & Cox
SER	symbol error ratio
SFO	sampling frequency offset
S&H	sample and hold
SI	self-interference
SIC	self-interference cancellation
SINR	signal-to-interference-plus-noise ratio
SISO	single-input single-output
SIR	signal-to-interference ratio

SJ	sampling jitter
SM-OCDM	sector-modulated OCDM
SNR	signal-to-noise ratio
SoC	system-on-a-chip
SQNR	signal-to-quantization-noise ratio
SSB	synchronization signal block
STO	symbol time offset
TDD	time-division duplexing
TDM	time-division multiplexing
TITO	tilt inference of time offset
ToF	time of flight
TTD	true-time delay
UAV	unmanned aerial vehicle
UE	user equipment
UL	uplink
ULA	uniform linear array
V2I	vehicle-to-infrastructure
V2V	vehicle-to-vehicle
ZCZ	zero correlation zone
ZF	zero forcing
ZP	zero padding

Constants

c_0	Speed of light in vacuum: $299\,792\,458\,\text{m s}^{-1}$
e	Euler's number: $2.71828\dots$
j	Imaginary unit: $\sqrt{-1}$
k_B	Boltzmann constant: $1.380\,649 \times 10^{-23}\,\text{J K}^{-1}$
π	Pi: $3.14159\dots$

Latin Symbols and Variables

Symbols marked with [†] also appear with the superscripts listed under Subscripts and Superscripts.

Lowercase Letters

$\mathbf{b}_{\text{Rx}}$	Receive bit stream
$\mathbf{b}_{\text{Tx},p}$	Transmit steering vector for the p th steering direction
$\mathbf{b}_{\text{Rx},n_{\text{Rx}}^{\text{S}}}$	Receive steering vector for the n_{Rx}^{S} th evaluated DoA
$\mathbf{d}_{\text{Tx}}$	Transmit modulation symbol stream [†]
$\mathbf{d}_{\text{Rx}}$	Receive modulation symbol stream [†]
$\overline{\mathbf{d}_{\text{Tx}}}$	Normalized energy of the transmit modulation alphabet
f	Frequency
f_c	Carrier frequency
f_{corner}	Flicker corner frequency of the PN PSD

f_D	Doppler shift
$f_{D,p}, f_{D,\text{target},h}$	Doppler shift of the p th path / of the h th target
$f_{D,\text{path},p}$	Doppler shift of the p th communications path
f_{IF}	Digital intermediate frequency
f_k^{SFO}	SFO-induced frequency shift at the k th subcarrier
f_Δ	Carrier frequency offset between transmitter and receiver
f_Δ^{frac}	Fractional part of the carrier frequency offset
f_Δ^{int}	Integer part of the carrier frequency offset
h	Radar target index
h_n	n th element of the discrete-time CIR (Appendix)
k	Subcarrier index
$k_{\Delta,h}$	Normalized Doppler shift of the h th target
l	Auxiliary subcarrier index
m	OFDM symbol index
m_{pil}	Pilot OFDM symbol index
n	Discrete-time sample index within an OFDM symbol
n_A	Antenna element index
$n_{\text{Tx}}, n_{\text{Rx}}$	Transmit and receive channel indices
n_{Rx}^{S}	Index of the evaluated DoA
$n_{\Delta,p}, n_{\Delta,h}$	Delay of the p th path / range term of the h th target in samples
p	Propagation path index
$\mathbf{r}^{\text{OCDM}}$	Auxiliary matrix in the OCDM frequency-shift analysis (Appendix)
s	Discrete-time sample index
$\mathbf{s}^{\text{PMCW}}$	PRBS in the discrete-time domain
t	Time
v	Relative radial velocity [†]

v_p	Velocity magnitude of the target associated with the p th path
$v_{\max}$	Maximum tolerable relative radial velocity
$v_{\max,ua}$	Maximum unambiguous relative radial velocity [†]
$v_{\max,ICI}$	Maximum ICI-free relative radial velocity [†]
$v_{\max,FSP}$	Maximum tolerable relative radial velocity with FSP-based FDM (C-OCDM) [†]
x, x_{RF}	Baseband and RF transmit signals [†]
$\mathbf{x}$	Discrete-time domain transmit frame [†]
y, y_{RF}	Baseband and RF receive signals [†]
y^{corr}	CFO-corrected receive sequence
$\mathbf{y}$	Discrete-time domain receive frame [†]
z	Signal transmitted by ISAC #2

Uppercase Letters

A^{PMCW}	Number of PRBS repetitions per block
$AM(\cdot), PM(\cdot)$	AM-AM and AM-PM characteristics of the amplifier
B	Bandwidth [†]
B_{PLL}	PLL loop bandwidth
$\mathbf{C}^{\text{OCDM}}$	Matrix of modulation symbols
$\mathbf{D}$	Delay–slow-time profile matrix from pilot subcarriers
$\mathbf{D}^{n_{\text{Rx}}}$	Radar CFR matrix of the n_{Rx} th receive channel
D_k^{pb}	Ratio of the transmit preamble symbols at the k th subcarrier
$\mathcal{D}_{\text{image}}$	Required dynamic range of the radar image
E_b	Energy per bit

$\mathcal{E}^{\text{OCDM,rad}}$	Energy of the radar subchirp
$\mathcal{E}^{\text{OCDM,comm}}$	Energy of a communications subchirp
$\mathcal{E}^{\text{OCDM,ISAC}}$	Energy of the ISAC symbol
EVM	Error vector magnitude
F_s	Sampling frequency [†]
G_p	Radar processing gain [†]
$G_{\text{Tx}}, G_{\text{Rx}}$	Transmit and receive antenna gains
G	Zero-forcing frequency-domain equalization matrix
$\mathbf{G}_m$	Zero-forcing equalization vector of the m th block [†]
H	Number of radar targets
H	CFR matrix [†]
$H_{\text{AFE}}^{n_{\text{Tx}}}(f), H_{\text{ABE}}^{n_{\text{Rx}}}(f)$	CFR of the AFE of the n_{Tx} th transmit / ABE of the n_{Rx} th receive channel
$H_{\text{CFR}}^{n_{\text{Rx}}}$	CFR of the n_{Rx} th receive channel
I	Range–Doppler radar image [†]
$\mathcal{I}$	Three-dimensional radar image (range, Doppler shift, DoA)
$\mathcal{J}_{\text{HW}}$	Set of considered hardware impairments
IBO	Input back-off
IP1dB	Input-referred 1-dB compression point
L_0	In-band PN level
L_{floor}	PN floor
$L^{n_{\text{A}}}$	Range from the n_{A} th antenna to the planar wavefront
L^{origin}	Range from the origin to the planar wavefront
M	Number of OFDM symbols per frame [†]
M_{pil}	Number of pilot OFDM symbols
$M_{\text{pil}}^{\text{TITO}}$	Number of pilot OFDM symbols used for TITO SFO estimation
$\mathcal{M}$	Modulation constellation size

N	Number of subcarriers [†]
N_A	Number of antenna elements
N_b^M	Number of bits per modulation symbol
$N_{\text{bit}}^{\text{DAC}}, N_{\text{bit}}^{\text{ADC}}$	Number of DAC and ADC bits
N_{CP}	Cyclic prefix length in samples [†]
N_h	Length of the discrete-time CIR (Appendix)
N_{mod}	Number of modulation symbols per block
N_{pil}	Number of pilot subcarriers
$N_{\text{PRBS}}^{\text{PMCW}}$	PRBS length
N_{Rx}^S	Number of evaluated DoAs
$N_{\text{subc}}, N_{\text{subch}}$	Number of OFDM subcarriers / OCDM subchirps [†]
$N_{\text{Tx}}, N_{\text{Rx}}$	Number of transmit and receive channels
N_0	Noise power spectral density
NF	Noise figure
P	Number of propagation paths
P_{Tx}	Transmit power
PAPR	Peak-to-average power ratio
R_h	Range of the h th target
R_0, R_p	Length of the reference path / of the p th path
$R_{\text{Tx} \rightarrow \text{T}, p}, R_{\text{T} \rightarrow \text{Rx}, p}$	Transmitter–target and target–receiver ranges of the p th path
R_{max}	Maximum tolerable range
$R_{\text{max}, \text{ua}}$	Maximum unambiguous range [†]
$R_{\text{max}, \text{ISI}}$	Maximum ISI-free range [†]
$R_{\text{max}, \text{CP}}^{\text{SM-OCDM}}$	Maximum CP-limited range
R_{min}	Minimum target range
$\mathcal{R}$	Data rate [†]
S^{PMCW}	DFT of the PRBS

$S_{\theta^{\text{PN}}}(f)$	Double-sided PSD of the PN
SFO	Sampling frequency offset
SIR	Signal-to-interference ratio
SNR	Signal-to-noise ratio
SQNR	Signal-to-quantization-noise ratio
T_s	Critical sampling period [†]
T_{therm}	Standard room temperature
$T_{\text{block}}, T_{\text{frame}}$	Block and frame durations [†]
$T_{\text{symb}}, T_{\text{CP}}$	Symbol and cyclic prefix durations [†]
$T_{\text{PRBS}}^{\text{PMCW}}$	PRBS duration
W	Windowing matrix
X	Discrete-frequency domain transmit frame [†]
$\mathbf{\tilde{X}}^{\text{OCDM}}, \mathbf{\tilde{Y}}^{\text{OCDM}}$	Discrete-Fresnel domain transmit / receive frame
$X_k^{\text{pb},1}, X_k^{\text{pb},2}$	Transmit content of the k th subcarrier of the 1st / 2nd preamble symbol
Y	Discrete-frequency domain receive frame [†]
$Y_k^{\text{pb},1}, Y_k^{\text{pb},2}$	Receive content of the k th subcarrier of the 1st / 2nd preamble symbol
$\mathbf{Y}^{\text{MRC}}$	Receive frame after maximum ratio combining

Greek Symbols and Variables

α_0	Attenuation factor of the reference path
α_p	Attenuation factor of the p th path
α_h^{rad}	Attenuation factor of the h th target (monostatic radar)
α_p^{comm}	Attenuation factor of the p th communications path

α_k^{SFO}	SFO-induced amplitude modulation at the k th subcarrier
$\alpha_{\Delta}^{\text{I/Q-imbal}}$	I/Q amplitude imbalance
β_p	Bistatic angle of the p th path
γ	PN level scaling factor
γ_p	Aspect angle of the p th path w.r.t. the bistatic bisector
$\gamma^{\text{CFO}}[\kappa]$	Correlation metric for integer CFO estimation
$\gamma^{\text{STO}}[s]$	Timing metric for STO estimation
$\mathbf{\Gamma}$	Chirp phase vector in the OCDM receive processing
δ	Normalized SFO
$\delta_{\max}$	Maximum absolute normalized SFO (TITO)
$\delta_{\max}^{\text{ICI}}, \delta_{\max}^{\text{ISI}}$	Maximum normalized SFO without relevant ICI / ISI
$\delta_{\text{DAC}}^{\text{SJ}}, \delta_{\text{ADC}}^{\text{SJ}}$	SJ of the DAC / ADC sampling clock
$\delta_{\text{RMS}}^{\text{SJ}}$	RMS SJ
Δf	Subcarrier spacing [†]
Δf_{D}	Doppler shift resolution
$\Delta L^{n_{\text{A}}}$	Propagation length difference at the n_{A} th antenna
$\Delta L_p^{n_{\text{T}_x}}$	Propagation length difference at the n_{T_x} th transmit antenna for the p th path
$\Delta M_{\text{pil}}, \Delta N_{\text{pil}}$	Pilot spacing in OFDM symbols / subcarriers
ΔR	Range resolution [†]
Δv	Relative radial velocity resolution [†]
$\Delta \phi_{\text{Rx}}$	Rx azimuth resolution
$\Delta \tau_m^{\text{SFO}}$	SFO-induced delay migration at the m th OFDM symbol
ε	Tolerance factor of the TITO SFO bound

η	FDZP factor (signal oversampling in the SJ analysis; delay-axis interpolation in SFO estimation)
$\theta_{\text{Tx}}^{\text{PN}}, \theta_{\text{Rx}}^{\text{PN}}$	Transmitter / receiver PN
$\theta_{\text{DAC}}^{\text{PN}}$	PN of the DAC sampling clock oscillator
Θ_1, Θ_2	Chirp phase vectors of the DFnT (OCDM)
ι	Auxiliary summation index
κ	Auxiliary subcarrier index
λ_0	Wavelength at the carrier frequency
ν	Auxiliary sample index (OCDM)
$\xi^{\text{PN}}, \xi_{k,m}^{\text{PN}}$	PN-induced additive distortion
$\xi_{k,m}^{\text{PN,CPE}}, \xi_{k,m}^{\text{PN,ICI}}$	CPE / ICI part of the PN-induced distortion
$\xi_{k,m}^{\text{SFO,ICI}}$	SFO-induced ICI
ρ_p	Reflection coefficient of the p th path
ρ^{n_A}	Position vector of the n_A th antenna element
$\rho_{\text{Tx},x}^{n_{\text{Tx}}}, \rho_{\text{Rx},x}^{n_{\text{Rx}}}$	x coordinate of the n_{Tx} th transmit / n_{Rx} th receive antenna
σ_{RCS}	RCS
$\sigma_{\text{RCS},p}, \sigma_{\text{RCS},h}$	RCS of the target of the p th path / of the h th target
$\sigma_{\text{RCS,max}}$	Maximum expected target RCS
$\sigma_\tau, \sigma_\theta$	Delay / angular standard deviation among receive channels
$\sigma_{\hat{\delta}}, \sigma_{\Delta\tau}^{\text{SFO}}$	Standard deviation of the SFO / delay migration estimate
τ_0	Delay of the reference path
τ_p	Delay of the p th path
$\tau_{\text{Tx} \rightarrow \text{T},p}, \tau_{\text{T} \rightarrow \text{Rx},p}$	Transmitter–target / target–receiver delays of the p th path
$\tau_{\Delta,p}^{n_{\text{Tx}}}$	Delay difference at the n_{Tx} th transmit antenna for the p th path
$\tau_{\text{target},h}$	Round-trip delay of the h th target (monostatic radar)

$\tau_{\text{path},p}$	Delay of the p th communications path
τ_{Δ}	STO
τ_m^{SFO}	SFO-induced delay at the m th OFDM symbol
$\tau_{\text{max}}^{\text{CP}}$	Maximum delay covered by the CP
$\tau_{\text{ABE}}^{n_{\text{Rx}}}$	ABE CIR delay of the n_{Rx} th receive channel
φ^{CFO}	Phase for fractional CFO estimation
$\varphi_{\Delta}^{\text{I/Q-imbal}}$	I/Q phase imbalance
$\varphi_m, \varphi_{k,m}$	Phase of the modulation symbol [†]
$\varphi(s, s')$	Auxiliary function in the SJ derivation
$\phi_m, \phi_{m,h}$	Doppler-induced phase at the m th symbol (OCDM)
$\phi_{\text{Tx} \rightarrow \text{T},p}$	DoD of the p th path
$\phi_{\text{T} \rightarrow \text{Rx},p}$	DoA of the p th path
$\phi_{\text{Rx,max,ua}}$	Maximum unambiguous Rx azimuth
$\Phi_m^{\text{PN,CPE}}$	PN-induced CPE at the m th OFDM symbol
ψ_{Δ}	CFO-induced phase offset
$\psi_{\text{Tx}}, \psi_{\text{Rx}}$	Transmitter / receiver carrier phase offsets
ψ_p	Combined phase term of the p th path
$\psi_{k,m}^{\text{SFO}}$	SFO-induced phase rotation
ω	Auxiliary angular frequency in the SJ derivation

Operators and Mathematical Symbols

$\mathbb{C}, \mathbb{C}^{A \times B}$	Set of complex numbers / of complex $A \times B$ matrices
$\mathbb{N}$	Set of natural numbers
$\mathbb{N}_{\geq 0}, \mathbb{N}_{>0}$	Set of non-negative / positive integers
$\mathbb{R}, \mathbb{R}^{A \times B}$	Set of real numbers / of real $A \times B$ matrices

$\mathbb{R}_{\geq 0}, \mathbb{R}_{> 0}$	Set of non-negative / positive real numbers
$\mathbb{Z}$	Set of integers
$(\cdot)^*$	Complex conjugate
$(\cdot)^T$	Transpose
$\widehat{(\cdot)}$	Estimate
$(\tilde{\cdot})$	Impairment-free (ideal) version
$(\dot{\cdot})$	Discrete-Fresnel domain quantity (OCDM)
$ \cdot $	Absolute value
$\lfloor \cdot \rfloor$	Floor function
$\langle \cdot \rangle_N$	Modulo- N operation
$\langle \cdot \rangle$	Time average
$*$	Linear convolution
$\otimes_N$	N -point circular convolution
$\arg(\cdot)$	Argument (phase) of a complex number
$\arg \max_x(\cdot)$	Value of x that maximizes the input expression
$\mathbb{E}\{\cdot\}$	Expected value
$\Re\{\cdot\}, \Im\{\cdot\}$	Real / imaginary part
$\delta[\cdot]$	Kronecker delta
$\delta(\cdot)$	Dirac delta
$\text{rect}(\cdot)$	Rectangular function
$\triangleq$	Equal by definition
$\forall$	For all
$a \mid b$	a such that b
$\times$	Multiplication; matrix dimensions

Subscripts and Superscripts

$(\cdot)^{\text{OFDM}}$	OFDM
$(\cdot)^{\text{PMCW}}$	PMCW
$(\cdot)^{\text{OCDM}}$	OCDM
$(\cdot)^{\text{C-OCDM}}, (\cdot)^{\text{SM-OCDM}}$	C-OCDM / SM-OCDM
$(\cdot)_{\text{Tx}}, (\cdot)_{\text{Rx}}$	Transmitter / receiver
$(\cdot)_{\text{DAC}}, (\cdot)_{\text{ADC}}$	DAC / ADC
$(\cdot)_{\text{AFE}}, (\cdot)_{\text{ABE}}$	Analog front-end / back-end
$(\cdot)_{\text{RF}}$	Radio frequency
$(\cdot)^{\text{amp}}, (\cdot)_{\text{out}}$	Amplifier / output
$(\cdot)^{n_{\text{Tx}}}, (\cdot)^{n_{\text{Rx}}}, (\cdot)^{n_{\text{A}}}$	n_{Tx} th transmit / n_{Rx} th receive channel / n_{A} th antenna
$(\cdot)_p, (\cdot)_h$	p th path / h th target
$(\cdot)_{k,m}, (\cdot)_{n,m}$	Element at the k th / n th row and m th column
$(\cdot)^{\text{PN}}, (\cdot)^{\text{SJ}}$	PN / SJ
$(\cdot)^{\text{CFO}}, (\cdot)^{\text{SFO}}, (\cdot)^{\text{STO}}$	CFO / SFO / STO
$(\cdot)^{\text{I/Q-imbal}}$	I/Q imbalance
$(\cdot)^{\text{HW}}$	Hardware impairments
$(\cdot)^{\text{CPE}}, (\cdot)^{\text{ICI}}$	CPE / ICI component
$(\cdot)^{\text{rad}}, (\cdot)^{\text{comm}}, (\cdot)^{\text{ISAC}}$	Radar / communications / joint ISAC
$(\cdot)^{\text{monost}}, (\cdot)^{\text{bist}}$	Monostatic / bistatic
$(\cdot)^{\text{MIMO}}$	Combined over all MIMO receive channels
$(\cdot)_{\text{max}}, (\cdot)_{\text{min}}$	Maximum / minimum
$(\cdot)_{\text{ua}}$	Unambiguous
$(\cdot)_{\text{ISI}}, (\cdot)_{\text{ICI}}$	ISI-free / ICI-free limit
$(\cdot)_{\text{CP}}$	CP-limited
$(\cdot)_{\text{FSP}}$	FSP-based multiplexing

$(\cdot)_{\text{RMS}}$	Root mean square
$(\cdot)_{\text{eff}}$	Effective
$(\cdot)_{\text{pil}}$	Pilot-based
$(\cdot)_{\text{full}}$	Full-frame
$(\cdot)_{\text{pb}}$	Preamble
$(\cdot)_{\text{w/o CP}}$	Without CP
$(\cdot)_{\text{ZP}}$	Zero-padded (Appendix)
$(\cdot)^{\text{corr}}, (\cdot)^{\text{corr}'}, (\cdot)^{\text{corr}''}$	After first / second / third correction step
$(\cdot)^{\text{coarse}}, (\cdot)^{\text{res}}$	Coarse / residual estimate
$(\cdot)^{\text{frac}}, (\cdot)^{\text{int}}$	Fractional / integer part
$(\cdot)_{\text{MRC}}$	After maximum ratio combining
$(\cdot)_{\text{TITO}}$	TITO SFO estimation
$(\cdot)_{\text{image}}$	Radar image
$(\cdot)_{\text{target}}, (\cdot)_{\text{path}}$	Monostatic target / communications path (Ch. 2)
$(\cdot)_{\Delta}$	Offset between transmitter and receiver

1 Introduction

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This dissertation proposes, discusses, and verifies system design considerations and signal processing techniques to enable future integrated sensing and communications (ISAC) systems via numerical analyses and measurements. Before detailing technical aspects in the next chapters, Section 1.1 presents the motivation and background for the carried out research, and Section 1.2 offers an overview of the state of the art. Finally, Section 1.3 outlines the structure and the objectives of this dissertation.

1.1 Motivation and context

Due to the growing number of applications that occupy frequency spectra at steadily increasing carrier frequencies as well as their need for connectivity and sensing capabilities [RXK⁺19], there has been a growing interest in the convergence of radar sensing and communications systems [DBB⁺21, QQGZ14]. Not only the coexistence of both systems has been investigated, but also their radar-centric, communications-centric, or multi-objective co-design [CPB17, ZLEW19, WBV21, ZLM⁺21]. In the context of co-design, both applications share the same hardware platform and therefore have common transmit and/or receive signals, while distinct processing strategies are adopted to obtain communications data and estimate parameters of radar targets. The concept of co-existing and co-designed systems supporting both radar sensing and communications is, for example, known as ISAC, also named joint communication and sensing (JCAS) [ZRW⁺21, WBV21, KGBR21]. In the particular case where the same signal is used for both applications, some early studies in the literature named the joint system dual-function(al) radar-communication (DFRC) [HAAH19, LMP⁺20, MSH⁺20] or radar-communication (RadCom) system [SW11, GCSM18], before converging to JCAS and then ISAC. Since

the proposal of what is known as the world's first ISAC system in 1963 [Mea63], through the introduction of an ISAC system based on orthogonal frequency-division multiplexing (OFDM) in 2009 [SZW09], until efforts towards the convergence of radar sensing and communications in the context of beyond fifth-generation (5G) and sixth-generation (6G) networks [DBB⁺21], significant advances have been made in the ISAC field as thoroughly described in [LMP⁺20].

In a radar-centric perspective, a possible application of the ISAC concept is in the context of highly automated driving (HAD) [WHM21], where vehicle-to-vehicle (V2V) and vehicle-to-infrastructure (V2I) communications can be achieved by embedding communications symbols into the radar signal and used, e.g., for traffic coordination and interference avoidance among radar sensors [PTWA17, DBSF19, DZB⁺20]. Such tasks can be performed by approaches such as coordination on physical and link layer levels among devices serving as a shared platform in a protocol-based approach to alternate radar and communications operation [AKG⁺21, AKW21], or by an ISAC approach consisting of the transmission of radar signals with embedded communications symbols [HY19, RBK⁺19, AKC⁺20, EHW⁺21]. A further radar-centric application of the ISAC concept is in the context of radar networks [Gen20], where hybrid active-passive radars can use the same signal to perform bistatic measurements as well as to extract communications data which aims to coordinate and synchronize radar nodes within the network.

Due to its potential of turning cellular networks into perceptive mobile networks (PMNs) [ZRH⁺21], the application of the ISAC concept in a communications-centric approach enables a wide range of possibilities for beyond 5G and 6G networks. As illustrated in Figure 1.1, those include monitoring of areas with campus networks in intralogistics scenarios with humans and automatic guided vehicles (AGVs) as well as environment sensing for critical infrastructure or safe urban mobility [KZPH23, SNS⁺24]. An additional application is the accurate detection and tracking of unmanned aerial vehicles (UAVs) based on echoes of communications signals emitted by base stations [NSF⁺17, WBV21]. In the context of communications-centric ISAC, efforts in reusing 5G New Radio (NR) reference signals for radar sensing have been, e.g., reported in [KGBR21] and [WWM⁺23], where positioning reference signals (PRSs) and demodulation reference signals (DMRSs) are reused for radar sensing, respectively.



Figure 1.1: Examples of ISAC use cases. Adapted from [24], © 2026 IEEE.

In both radar- and communications-centric ISAC approaches, efficient modulation schemes must be adopted, in which appropriate communications performance can be achieved and the influence of transmitted data on ultimately obtained radar images can be minimized or removed. On the one hand, radar-centric applications can be supported by ISAC systems based on, e.g., chirp sequence (CS) [KR14, RBK⁺19] or the digital phase-modulated continuous wave (PMCW) [GG⁺14, WGBC19], which jointly enable radar sensing and reliable, low-latency communications at low data rates. In the radar literature, the first modulation scheme is also known as fast-chirp frequency-modulated continuous wave (FMCW) [HY19], while the latter is alternatively named code-modulated, spread spectrum, or pseudo-noise radar [RBK⁺19, WHM21]. On the other hand, digital multicarrier modulation schemes such as OFDM [SBM⁺04, HYW⁺09] and orthogonal chirp-division multiplexing (OCDM) [OZ16a, Ouy17] are preferred in communications-centric applications where higher data rate and spectral efficiency are required.

Given this context, a detailed literature review on state-of-the-art research on enablement of ISAC systems is presented in Section 1.2.

1.2 State of the art

For an appropriate operation of ISAC systems, a number of aspects must be considered. In this context, some of the main defining factors that have been

investigated in the technical literature are the choice of a modulation scheme, hardware challenges, and signal processing techniques, which are addressed in Subsections 1.2.1 and 1.2.2, respectively.

1.2.1 Modulation schemes

To enable efficient communications and radar sensing in an ISAC system, a suitable modulation scheme must be adopted. Straightforward approaches to enable ISAC operation include embedding communications data onto any type of radar signal, e.g., via beam-pattern modulation. However, the choice of more elaborate modulation schemes [HY19, RBK⁺19] that are both compatible with existing radar and communications technologies and allow efficient system parameterization to yield desired performance in both tasks of ISAC systems is preferred. In this sense, the previously mentioned PMCW, OFDM, and OCDM modulation schemes have been perceived as attractive solutions in the ISAC literature.

In the binary PMCW context, a considerably low number of ISAC approaches has been reported in the literature. A possible approach is the binary phase-modulation of pseudorandom binary sequences (PRBSs) adopting binary phase-shift keying (BPSK) or its differential variant, which attains low data rates but sufficiently preserves the radar sensing performance [DHS⁺17, DSN⁺17, DMMO19, MBK⁺19]. The low data rate yielded by the use of BPSK could be compensated, e.g., by using information bits to be transmitted to generate the PRBS [DHS⁺17].

Aiming to keep the hardware platform relatively simple, the PMCW modulation scheme is usually adopted in radar-centric applications. In contrast, digital multicarrier modulation schemes such as OFDM [SZW09, LLX⁺17, SWSZ18, BRT⁺19, NDS⁺20, LBRV20, LBRV21] and OCDM [47, 49] are preferred in both radar-centric ISAC applications where more flexibility for, e.g., multiple-input multiple-output (MIMO) operation and in communications-centric ISAC applications where high spectral efficiency and data rates are targeted. The first modulation scheme is based on the transmission of orthogonal subcarriers and offers the possibility of achieving quasi-unbiased radar estimation via a zero forcing (ZF)-like radar channel estimation to remove the influence of transmit modulation symbols [SZW09, Bra14, Sit17, SNZ18, BRT⁺19]. The latter

modulation scheme, in turn, uses orthogonal subcarriers to perform transmission of data belonging to digital modulation constellations, and uses a correlation-based receiver that yields range sidelobes depending on the transmit modulation symbols [47]. Further processing approaches for OCDM-based systems that yield unbiased channel estimation for sensing purposes have already been proposed, e.g., in [GdPR21], but no extension to the ISAC concept has been reported in the literature yet. In terms of communications performance, however, OCDM-based systems tend to achieve better results compared to their OFDM counterpart [OM21] if no further measures such as frequency-selective bit and power allocation [GRHV18] are taken, which OCDM cannot straightforwardly profit from due to its spread spectrum nature. This is mainly due to their greater robustness against a number of impairments, including multipath propagation, Doppler shifts, intersymbol interference (ISI) even under reduced cyclic prefix (CP) lengths [BCF19], and narrowband interference (NBI) [OM20a]. The latter is associated with their spread spectrum characteristic and can be explained, e.g., by the discrete Fresnel transform (DFnT) circular convolution theorem [OAG⁺15].

Alternative digital multicarrier modulation schemes that have already been investigated in the ISAC context comprise generalizations of OFDM and OCDM such as orthogonal time-frequency space (OTFS) [GKCC20], affine frequency-division multiplexing (AFDM) [NWYH22, RRS⁺25], filter bank multicarrier (FBMC) [SGCM18], and generalized frequency-division multiplexing (GFDM) [SCGM19], which tend to achieve more flexible trade-offs between radar sensing and communications performances as the two aforementioned modulation schemes at the cost of increased computational complexity. An important similarity among the aforementioned OFDM- and OCDM-based modulation schemes in the ISAC context is the possibility of applying the index modulation (IM) principle [HAAH19] to jointly transmit modulated data and limit the occupancy of frequency tones within the considered bandwidth [HDOF21]. While the application of IM may not be straightforward in all OFDM and OCDM variations, the concept has been increasingly gaining attention in recent studies in the literature as an approach to avoid interference in highly populated scenarios.

1.2.2 Hardware challenges

In this section, reported challenges in the literature related to hardware development and hardware impairments related to future ISAC systems are discussed in Subsections 1.2.2.1 and 1.2.2.2, respectively.

1.2.2.1 Hardware development for ISAC

In the context of hardware development for ISAC-capable systems, recent advances show the possibility of modularizing the hardware design with independent front- and back-end blocks [SGR⁺21]. An example is the scalable MIMO platform presented in [NKAK17], which supports the implementation of PMCW-based systems. In the aforementioned design, the presence of BPSK modulators enables the transmission of PRBSs carrying binary data symbols for simultaneous radar and communications operation.

For flexible fully-digital ISAC systems, however, general-purpose front-ends can be adopted. Consequently, the necessary system adjustments for enabling the implementation of ISAC systems based on any digital modulation scheme rely on adjustments of the baseband processing, e.g., on a field programmable gate array (FPGA) back-end. This is, e.g., the case of the broadband massive MIMO testbed described in [NKN⁺24]. It has been developed to support ISAC with up to 128 transmit and 128 receive channels for an analog baseband (BB) bandwidth of 2 GHz or 256 transmit and 256 receive channels for a bandwidth of 1 GHz. In terms of operating frequency, the main analog front-end (AFE) and analog back-end (ABE) modules are planned to cover the range from 24 GHz to 30 GHz, but other frequencies can also be supported with dedicated AFE and ABE modules. The testbed has a modular architecture, and full synchronization of both BB and radio-frequency (RF) front- and back-ends is supported [KNN⁺24a, KNN⁺24b, SWR⁺24]. To alleviate requirements in front-ends and efficiently deal with the high data rates in such systems, features such as flexible digital up- and downconverters from Xilinx RFSoCs [FME18] were used.

In the context of communications-centric ISAC, [WGM⁺23] presented a ISAC proof-of-concept (PoC) based on commercial 5G communications hardware for the Frequency Range 2 (FR2) defined by the 3rd Generation Partnership Project

(3GPP). The developed PoC operates at an adjustable carrier frequency of around 27.6 GHz and is constituted by a communications radio unit (RU) acting as an ISAC transmitter and a dedicated RU that is placed near the transmitter in a quasi-monostatic configuration to serve as a sensing receiver, called *sniffer*. To control the sniffer, a server called sensing processing unit (SPU) is employed, which supports real-time OFDM radar processing. While the functionalities of the ISAC PoC are still being expanded, reported results so far show that is capable of enabling several of the intended cellular infrastructure-based ISAC applications such as human and AGV sensing in indoor scenarios [HMGt24, MAG⁺25].

1.2.2.2 Hardware impairments

Among the main hardware constraints that are common to both radar- and communications-centric ISAC systems [BSBF21] are (i) linearity and efficiency of power amplifiers (PAs), (ii) in-band full duplex (IBFD) operation of transmit and receive channels with enough isolation for monostatic sensing, where transmitter and receiver are collocated, (iii) low noise figure, high gain and compression point in low-noise amplifiers (LNAs), as well as (iv) digital-to-analog converters (DACs) and analog-to-digital converters (ADCs) with high resolutions in all-digital ISAC systems. Those constraints are mainly imposed by radar sensing requirements, and are described as follows.

Regarding (i), both radar and communications applications demand increasing bandwidths, which makes the PA design considerably more challenging. Furthermore, typical modulation schemes for radar-centric ISAC systems such as PMCW usually impose low PA linearity requirements [GGS⁺14]. However, if one adopts modulation schemes that yield significantly higher peak-to-average power ratios (PAPRs) such as OFDM [HYW⁺09] and OCDM [OZ16a], the PA efficiency suffers a fair reduction since a back-off from the PA compression point is necessary. While this has been extensively studied for communications systems, especially OFDM-based ones, a clear definition of the required back-off for sensing operation has not been reported in the literature.

The required isolation for appropriate IBFD operation in the context of monostatic sensing mentioned in (ii) is a long-known problem in digital radar systems. Usually, isolation of several tens of decibels are mandatory for extending the

maximum detectable range [BSBF21], since the self-interference (SI) power from transmit to receive channels defines the LNA compression point mentioned in (iii) may limit the ISAC system sensitivity to targets at far ranges. In communications-centric ISAC systems, however, this is a new challenge to be faced either via self-interference cancellation (SIC) and decoupling techniques [BLH⁺21] or by exploiting the inherently distributed nature of cellular networks and performing bistatic sensing, where the transmitter and receiver are not collocated [KGBR21].

Besides the aforementioned issues related to (ii) and (iii), the maximum detectable range in an ISAC system is also limited by the quantization noise [WGBC19]. To overcome this, the need for high-resolution ADCs and DACs mentioned in (iv) becomes a critical issue in ISAC systems based on digital modulation schemes such as PMCW, OFDM, and OCDM. In this context, a typically required sampling rate of 1 GSa/s at a bit resolution of 14 bits yields a data rate of 28 Gbit/s per transmit or receive channel with one in-phase/quadrature (I/Q) channel each to be handled by a field programmable gate array (FPGA), e.g., in a simple single-input single-output (SISO) configuration of an ISAC system [SGR⁺21]. The aforementioned high data rates in radar and ISAC systems based on digital modulation schemes result in the challenging task of handling the many digital signal processing operations in a short time interval, as well as in the requirement for storages ranging from about 35 to 70 MB for typical receive data frame durations between 10 ms and 20 ms. In MIMO configurations, this issue becomes even more critical due to the higher digital channel count. To overcome those issues, bandwidth enlargement and/or sampling rate reduction techniques can be adopted for radar-centric, OFDM-based ISAC systems. Due to the compatibility between OFDM and OCDM modulation schemes [OZ16a], such techniques can in principle also be applied to OCDM-based ISAC systems. Bandwidth enlargement methods comprise techniques such as the use of frequency combs [NMMZ20, 14], linear frequency-modulated carriers [SSK⁺18a], and stepped carrier frequencies [PFS15, SKSW18, SSK⁺19]. As a sampling rate reduction technique, compressed sensing can also be employed at the cost of increased computational complexity for radar processing [KSS⁺18]. While effective in sampling rate reduction, such techniques significantly degrade communications data rate and may therefore only be suitable for radar-centric ISAC.

While the previously discussed hardware constraints are usually considered during the design of radar, communications, and ISAC systems such as the ones discussed in Subsection 1.2.2.1, there are several other hardware impairments. If neither considered during system design nor countermeasures against them are taken afterwards, they may significantly affect the performance of ISAC systems. These impairments include, e.g., antenna and array effects such as beam squint and element coupling, mixer frequency-selective I/Q imbalance and oscillator phase noise (PN), DAC and ADC clipping, aliasing and sampling jitter (SJ). While most of these impairments have already been widely studied for OFDM-based communications systems, their influence on radar sensing performance and therefore on ISAC, and proposed compensation strategies have not always been given due attention. For most of them, conclusions can be drawn with little adaptation of existing knowledge for communications systems. However, impairments such as PN and SJ, which arise at the mixing and sampling stages respectively, require more detailed analysis, as each introduces multiple types of degradation that affect radar sensing and communications performance in different ways. Some of the discussed impairments in the ISAC context are addressed in [WTW⁺25], which, however, focuses on beamforming, array architectures and full duplex operation, and presents only brief discussions on PA nonlinear distortion and PN.

1.2.3 Signal processing techniques and challenges

Another key topic explored in the state-of-the-art literature on ISAC systems involves signal processing techniques. They are required, e.g., to enable simultaneous radar sensing and communications under several constraints, as well as to alleviate hardware limitations and mitigating the impairments discussed in Subsection 1.2.2.

To enable coexistence of multiple ISAC nodes, a number of techniques has been proposed in the ISAC literature. Those include, e.g., interference avoidance approaches such as the use of stepped-carrier OFDM [AKC⁺20, KAW20] mentioned in Subsection 1.2.2 to prevent multiple ISAC nodes from simultaneously transmitting at the same frequencies. The aforementioned techniques also encompass interference cancellation (IC) methods [Sit17, SNZ18], which allow compensating the influence of received signals from other nodes on the

radar sensing performance of the evaluated receiving node, therefore enabling simultaneous communications as well as bi- and monostatic sensing.

Further signal processing efforts have been reported in the context of communications-centric ISAC systems. For instance, techniques based on compressed sensing for extracting radar target parameters from ISAC signals have been proposed [RZH⁺20] to deal with discontinuities in the available bandwidth that commonly happen in practice due to communications resource allocation. Similarly, detection and compensation approaches of artifacts due to the time-division duplexing (TDD) between downlink (DL) and uplink (UL) modes have been proposed in [38].

In the context of array signal processing for ISAC, techniques such as the approach for sampling and reconstructing angular domains (SARA) from [MHD23] were proposed. SARA enables effective sensing of larger angular sectors in spite of the use of beamformed transmission. More specifically, sensing is performed for a limited set of points in the angular domain, and the full angular response is reconstructed via trigonometric interpolation. Attained results show that SARA is effective even in non-sparse scenarios, being also suitable for applications such as imaging. In addition, beam codebooks that enable SI suppression in monostatic deployments with moderate impact on communications performance were proposed in [HFCS25].

Shifting the focus to bistatic sensing in ISAC systems, recent studies in the literature have investigated bistatic radar sensing with 5G networks. Examples include [SAP⁺22, AKP⁺23, KPA⁺23], which consider bistatic sensing with a gNodeB (gNB) acting as an illuminator of opportunity and a receiver having two separate antennas and RF/BB channels for reference and sensing. During bistatic radar signal processing, the extracted signal from the sensing channel is correlated with the reference signal [TAD⁺19, SAP⁺22, EKDS25], enabling the measurement of relative range and Doppler shifts. In contrast to the aforementioned studies, no detailed considerations on bistatic ISAC approaches with either a dedicated reference path or full communications processing to estimate the transmit data to enable classical OFDM radar processing [SZW09] were reported in the literature. Complementary to these open issues, the solution to the angular sampling problem initially addressed in [MHD23], which becomes even more critical in the bistatic case than in the monostatic case, was extended to bistatic deployments in [FMH⁺25].

A further critical issue in bistatic ISAC is the required synchronization accuracy to enable coherent processing of long sample streams. While communications systems, e.g., based on OFDM, may rely on techniques such as the Schmidl & Cox (S&C) algorithm [SC97] for time and frequency synchronization and the Tsai algorithm [TKC05] for sampling frequency offset (SFO) estimation, much stricter synchronization requirements must be met to enable unbiased bistatic sensing. Although studies in the literature focus on addressing time and frequency synchronization offsets, synchronization approaches that consider both SFO and its interplay with the aforementioned offsets, along with the fine accuracy required, have not been reported.

1.3 Structure and objectives of the dissertation

In spite of the significant efforts reported in the literature towards enabling both radar-centric and communications-centric ISAC, a number of issues remain insufficiently addressed as discussed in previous sections. Regarding digital modulation schemes, these include, for example, a clear definition of the communications processing in PMCW-based ISAC, as well as a feasibility study of OCDM-based ISAC. Focusing on OFDM [HYW⁺09], which is already employed in current 5G cellular systems and is also expected to be the choice for 6G scenarios where ISAC could be deployed, a number of challenges also remain open. These include the definition of a signal processing chain and a synchronization concept for bistatic ISAC, as well as a thorough analysis of hardware impairments and possible countermeasures.

In this context, the objectives and the structure of this dissertation are outlined as follows:

- To provide a comprehensive description of the processing chains of the PMCW, OFDM, and OCDM modulation schemes for ISAC systems, introducing new concepts and signal processing techniques whenever applicable. To support the discussion, a comparative performance analysis based on simulation and measurement results is to be presented. These objectives shall be achieved in Chapter 2.

- Focusing on OFDM, to introduce a bistatic ISAC system concept based on a static reference path to enable synchronization, ensure sufficient signal-to-noise ratio (SNR) for communications, and provide a benchmark for bistatic radar sensing. In addition, bistatic signal processing techniques with or without prior communications processing to estimate the transmit data are to be presented, and a synchronization procedure to accurately estimate and correct not only time and frequency offsets, but also SFO and ensure unbiased sensing is to be proposed. Finally, the effects of impairments introduced by both residual synchronization offsets and hardware impairments are to be discussed, and possible countermeasures outlined. These objectives shall be achieved in Chapter 3.
- To validate the proposed bistatic OFDM-based ISAC system concept, processing chains, and synchronization procedure via simulation and measurement results. In addition, to quantify the performance degradation imposed by impairments resulting from synchronization offsets and hardware non-idealities. These objectives shall be achieved in Chapter 4.
- To extend the proposed bistatic OFDM-based ISAC system, signal processing, and synchronization concepts to MIMO. In addition, to verify the developed bistatic MIMO-OFDM-based ISAC system and analyze its performance via simulation and measurement results. These objectives shall be achieved in Chapter 5.

After achieving the above-described goals, the dissertation concludes with the remarks presented in Chapter 6.

2 Digital waveforms for ISAC

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This chapter describes and compares the system design and processing chain for ISAC systems based on the digital modulation schemes PMCW, OFDM, and OCDM. To this end, an ISAC system model is first outlined in Section 2.1. Next, ISAC systems utilizing each of these schemes are presented and discussed in Sections 2.2, 2.3, and 2.4, respectively. A comparative performance analysis, based on simulation and measurement results, follows in Section 2.5. Finally, conclusions for the chapter are provided in Section 2.6.

2.1 ISAC system model

Let a SISO ISAC system be based on an IBFD RF device that transmits signals jointly used for radar sensing and carrying modulated data for communicating with other ISAC or communications systems. Moreover, the considered ISAC system is also capable of performing both radar parameter estimation and extracting modulated data coming from other ISAC or communications systems from its receive signal. It is further assumed that this ISAC system performs transmission based on frames of time duration T_{frame} , which are comprised of $M \in \mathbb{N}_{>0}$ consecutive blocks of time duration T_{block} , resulting in $T_{\text{frame}} = MT_{\text{block}}$. A serial representation of the ISAC frame structure is depicted in Figure 2.1, where the blocks within the frame have index $m \in \{0, 1, \dots, M - 1\}$, fast time is the continuous time axis within the duration of a block and slow time is the discrete time axis with time step equal to T_{block} . At the transmitter of the ISAC system, each block is individually modulated with data belonging to a modulation constellation. Similarly, the blocks are individually evaluated at the receiver of the ISAC system. While the communications processing at the ISAC receiver consists of extracting modulated data

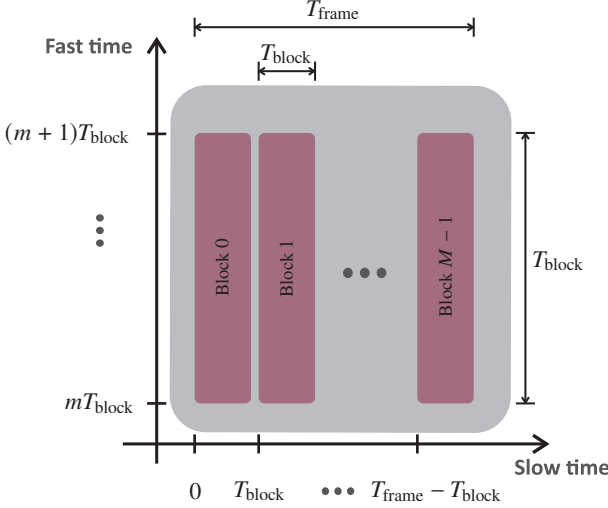


Figure 2.1: ISAC frame structure. The fast time axis comprises the time interval in which every m th block is transmitted, having a total duration T_{block} . For the sake of simplicity, no time granularity is defined for the fast time axis since it depends on the sampling period that changes with the adopted modulation scheme. In its turn, the slow time axis marks the start of each m th block, having therefore time granularity equal to T_{block} and total duration T_{frame} . © 2022 IEEE. Reprinted, with permission, from [18].

from each block individually, the radar processing consists of estimating one range profile per block, based on which a range-velocity radar image is finally computed.

In this context, the system model depicted in Figure 2.2 comprises the considered ISAC system, represented as ISAC #1, radar targets, and another ISAC system, which is represented as ISAC #2. It is assumed that the signals $x(t) \in \mathbb{R}$ and $z(t) \in \mathbb{R}$, where $t \in [0, T_{\text{frame}})$, carry frames transmitted by ISAC #1 and ISAC #2, respectively, besides occupying the frequency band $f \in [f_c - B/2, f_c + B/2]$. Consequently, both ISAC #1 and #2 occupy an RF bandwidth B centered at the carrier frequency $f_c \gg B$.

The analog signal $x(t)$ containing a transmit frame of the ISAC #1 in the form of a serial burst of blocks is transmitted with the power $P_{\text{Tx}} \in \mathbb{R}_{>0}$ by the transmit antenna of ISAC #1, propagates through the medium and: (i) travels

through a total of $P \in \mathbb{N}_{>0}$ paths, labeled as $p \in \{0, 1, \dots, P-1\}$, and is captured by the receive antenna of ISAC #2; and (ii) is reflected off a total of $H \in \mathbb{N}_{>0}$ radar point targets, labeled as $h \in \{0, 1, \dots, H-1\}$, and is captured by the receive antenna of ISAC #1 itself. For the sake of simplicity, it is assumed that ISAC #1 and ISAC #2 have quasi-colocated transmit and receive antennas and identical hardware so that the wireless communications channel between them is reciprocal, and that multipath-free propagation of reflections off radar point targets takes place. Although those assumptions neglect some aspects of the dominant scattering behavior in radar scenarios, they result in a simplified system model that allows fair assessment of the influence of adopted modulation schemes and processing strategies on the radar performance and have been also made in other studies in the literature [SW11, NSF⁺17, SNZ18]. Furthermore, it is considered that changes in the positions and velocities of scatterers are negligible, so that the only changes in the experienced radar and communications channels within the measurement time are Doppler-shift induced phase rotations. Based on those assumptions, the captured signal without noise $y(t) \in \mathbb{R}$ by the receive antenna of ISAC #1 can be expressed as

$$y(t) \approx \sum_{h=0}^{H-1} \alpha_h^{\text{rad}} x(t - \tau_{\text{target},h}) e^{j2\pi f_{\text{D},\text{target},h} t} + \sum_{p=0}^{P-1} \alpha_p^{\text{comm}} z(t - \tau_{\text{path},p} - \tau_{\Delta}) e^{j2\pi f_{\text{D},\text{path},p} t} e^{j2\pi f_{\Delta} t}. \quad (2.1)$$

The first term in (2.1) represents the quasi-monostatic radar reflections suffered by $x(t)$ off H radar point targets and captured by the receive antenna of ISAC #1. It accounts for the attenuation factor α_h^{rad} , the delay $\tau_{\text{target},h}$, and the Doppler shift $f_{\text{D},\text{target},h}$. In this term, the attenuation factor α_h^{rad} results from the propagation at the speed of light in vacuum c_0 through the overall traveled path of length $2R_h$ by $x(t)$ to the h th target and back to ISAC #1, being $R_h = c_0 \tau_{\text{target},h}/2$ the range of the h th target w.r.t. ISAC #1, the reflection off the h th target and other factors, and can be expressed as [Sit17]

$$\alpha_h^{\text{rad}} = \sqrt{\frac{G_{\text{Tx}} G_{\text{Rx}} \lambda_0^2 \sigma_{\text{RCS},h}}{(4\pi)^3 R_h^4}}. \quad (2.2)$$

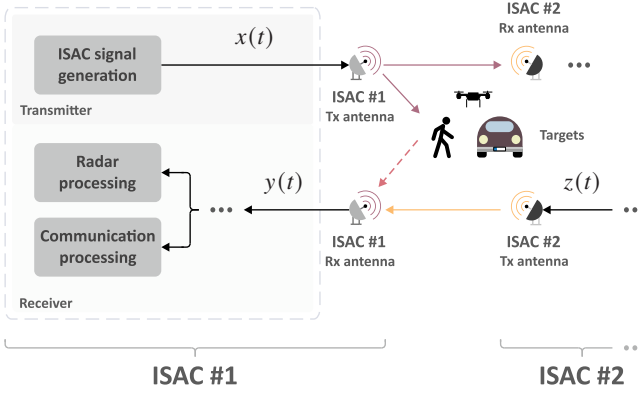


Figure 2.2: Simplified ISAC system representation. Adapted from [18], © 2022 IEEE.

In (2.2), G_{Tx} and G_{Rx} are the transmit and receive antenna gains, respectively, that are equal for both ISAC #1 and ISAC #2, $\lambda_0 = c_0/f_c$ is the wavelength associated with the carrier frequency f_c , and $\sigma_{\text{RCS},h}$ is the radar cross section (RCS) of the h th target. Additionally, the delay $\tau_{\text{target},h}$ from the first term in (2.1) is associated with the overall traveled path of length $2R_h$ by $x(t)$, while the Doppler shift $f_{\text{D,target},h} = -2v_h/\lambda_0$ results from the relative radial velocity v_h of the h th target w.r.t. ISAC #1. It is worth highlighting that carrier phases due to reflection off radar targets are neglected for simplicity since they are only relevant for direction of arrival (DoA) estimation, which would only be possible in a MIMO ISAC system.

In its turn, the second term in (2.1) represents the P versions of the transmit signal $z(t)$ of ISAC #2 after travelling through P paths and being captured by the receive antenna of ISAC #1. This term accounts for: (i) the attenuation factor α_p^{comm} , (ii) the delay $\tau_{\text{path},p}$, and (iii) the Doppler shift $f_{\text{D,path},p}$ suffered by $z(t)$ due to its propagation and reflection off scatterers through the p th path between the transmit antenna of ISAC #2 and the receive antenna of ISAC #1, as well as (iv) the time offset τ_{Δ} and (v) the frequency offset f_{Δ} between ISAC #1 and #2. It is worth highlighting that both τ_{Δ} and f_{Δ} , as well as resulting phase rotations, are affected by relative motion between ISAC #1 and #2, which ultimately leads to the need of more frequent channel state information (CSI) estimation at the receiving ISAC system for both communications and IC purposes and can be achieved by similar strategies to, e.g., the one introduced in [SC97]. Assuming

for simplicity that $z(t)$ suffers a single reflection while traveling through the the p th path, the attenuation factor α_p^{comm} can be expressed as [AKW21]

$$\alpha_p^{\text{comm}} = \sqrt{\frac{G_{\text{Tx}} G_{\text{Rx}} \lambda_0^2 \rho_p^2}{(4\pi)^2 R_{\text{path},p}^2}}, \quad (2.3)$$

where ρ_p is the reflection coefficient associated with the p th path, and $R_{\text{path},p} = c_0 \tau_{\text{path},p}/2$ is the overall path length traveled by $z(t)$ from ISAC #2 to ISAC #1.

After the reception by ISAC #1, the signal $y(t)$ undergoes analog conditioning, down-conversion to the baseband, and digital-to-analog (D/A) conversion. Next, two distinct processing strategies are performed on $y(t)$ to estimate radar parameters of the H targets and extract communications symbols transmitted by ISAC #2. Assuming that ideal separation of the radar and communications contributions of $y(t)$ is achieved and that all target reflections are clearly separated from each other according to the range resolution ΔR and relative radial velocity resolution Δv , besides being constrained to the maximum tolerable range R_{max} and relative radial velocity v_{max} , the radar processing ultimately yields a range-velocity radar image, in which the h th point target appears as a peak with SNR w.r.t. the resulting additive white Gaussian noise (AWGN) after signal processing given by

$$\text{SNR}_{\text{image},h} = \frac{P_{\text{Tx}} G_{\text{Tx}} G_{\text{Rx}} \sigma_{\text{RCS},h} \lambda_0^2 G_p}{(4\pi)^3 R_h^4 k_B B T_{\text{therm}} \text{NF}}, \quad (2.4)$$

where G_p is the modulation-scheme-dependent radar processing gain resulting from operations to form the range and relative radial velocity profiles, k_B is the Boltzmann constant, T_{therm} is the standard room temperature in Kelvin, and NF is the overall receiver noise figure in linear scale. In the denominator of (2.4), the term $k_B B T_{\text{therm}} \text{NF}$ accounts for the power of the sampled version of the AWGN $n(t)$. Besides influencing the ultimately experienced SNR, the considered parameters in (2.4) also play a role in defining the required ADC dynamic range and the corresponding number of ADC bits $N_{\text{bit}}^{\text{ADC}} \in \mathbb{N}_{>0}$. Assuming that full analog compensation of coupling between transmit and receive antennas takes place, which is, e.g., the case in analog CS-based systems, the required *effective* ADC dynamic range after radar signal processing $\mathcal{D}_{\text{image}}$ so that the

quantization noise and the thermal noise are at the same level in the ultimately obtained range-velocity radar image is

$$\mathcal{D}_{\text{image}} = \frac{P_{\text{Tx}} G_{\text{Tx}} G_{\text{Rx}} \sigma_{\text{RCS},\text{max}} \lambda_0^2 G_{\text{p}}}{(4\pi)^3 R_{\text{min}}^4 k_{\text{B}} B T_{\text{therm}} \text{NF}}, \quad (2.5)$$

where $\sigma_{\text{RCS},\text{max}}$ is the maximum expected target RCS and R_{min} is the minimum expected target range. In fully-digital radars such as PMCW, OFDM, OCDM, the aforementioned analog compensation of the coupling is not always possible, which can lead to a high coupling level between the transmit and receive antennas. Assuming that the coupled signal has much higher power than any received target reflection and defining a power coupling factor $C_{\text{Tx},\text{Rx}}$, the required *effective* ADC dynamic range becomes

$$\mathcal{D}_{\text{image}} = \frac{P_{\text{Tx}} C_{\text{Tx},\text{Rx}} G_{\text{p}}}{k_{\text{B}} B T_{\text{therm}} \text{NF}}. \quad (2.6)$$

It is worth highlighting that the processing gain G_{p} in (2.5) and (2.6) increases the power of the target peak in the radar image w.r.t. to both thermal noise and quantization noise and it is only experienced after radar signal processing. Considering this as well as the PAPR of the coupled signal, which will define the ADC full scale, the required physical ADC dynamic range is equal to $\mathcal{D}_{\text{image}} \cdot \text{PAPR}/G_{\text{p}}$. Consequently, the required number of ADC bits $N_{\text{bit}}^{\text{ADC}} \in \mathbb{N}_{>0}$ can finally be calculated based on the approximate relationship assumed in [VGG⁺14, Nus21], i.e.,

$$N_{\text{bit}}^{\text{ADC}} = (10 \log_{10}(\mathcal{D}_{\text{image}} \cdot \text{PAPR}/G_{\text{p}}) - 4.77) / 6.02. \quad (2.7)$$

For the sake of conciseness, it is henceforth assumed that the sufficient number of ADC bits is adopted so that the aforementioned dynamic range requirement are satisfied.

For communications purposes, the SNR measure of interest is the E_{b}/N_0 , which denotes energy per bit to noise power spectral density ratio that is jointly influenced by all P paths and depends on both the adopted modulation scheme and modulation constellation. It is worth highlighting that IC must be performed before further communications processing in case ISAC #1 receives signals from other ISAC or communications systems besides ISAC #2. As the power of received radar target reflections is typically much lower than that

of communications signals, IC for the radar contribution to the receive signal $y(t)$ is henceforth not considered. Furthermore, no additional constraints on the calculated ADC dynamic ranges according to (2.5) or (2.6) apply since radar operation of the considered ISAC system typically demands much higher number of bits $N_{\text{bit}}^{\text{ADC}}$ than required for communications purposes.

A more detailed analysis of both radar sensing and communications performances depends significantly on particular aspects of the adopted ISAC modulation scheme. In this sense, Sections 2.2 to 2.4 present an overview of ISAC systems based on PMCW, OFDM, and OCDM modulation schemes, providing further details on their respective block-frame structures, processing steps, and performance metrics for radar sensing and communications.

2.2 Phase-modulated continuous wave

In this section, the working principle of the PMCW scheme is explained in Section 2.2.1. Next, the proposed processing chain for PMCW-based ISAC is described in Section 2.2.2.

2.2.1 PMCW principle

The basic transmit signal in the discrete-time domain in a binary PMCW scheme is a PRBS with $N_{\text{PRBS}}^{\text{PMCW}} \in \mathbb{N}_{>0}$ chips or samples as depicted in Figure 2.3(a). Since a binary sequence is adopted, the samples of this sequence have constant amplitude and can either assume 0 or π phase shifts.

Assuming that the aforementioned chips are updated with a chip rate F_s^{PMCW} during D/A conversion, a PRBS time duration $T_{\text{PRBS}}^{\text{PMCW}} = N_{\text{PRBS}}^{\text{PMCW}} / F_s^{\text{PMCW}}$ is obtained. Before transmission, the resulting continuous-time domain sequence modulates the phase of the local oscillator (LO) signal with carrier frequency f_c , which is a continuous wave (CW) signal. Consequently, a phase-modulated CW signal, hence PMCW, is obtained as illustrated in Figure 2.3(b).

The processing at the receiver side of the binary PMCW scheme for both radar sensing and communications applications is based on pulse compression, which consists of correlating the receive sequence with the originally transmitted

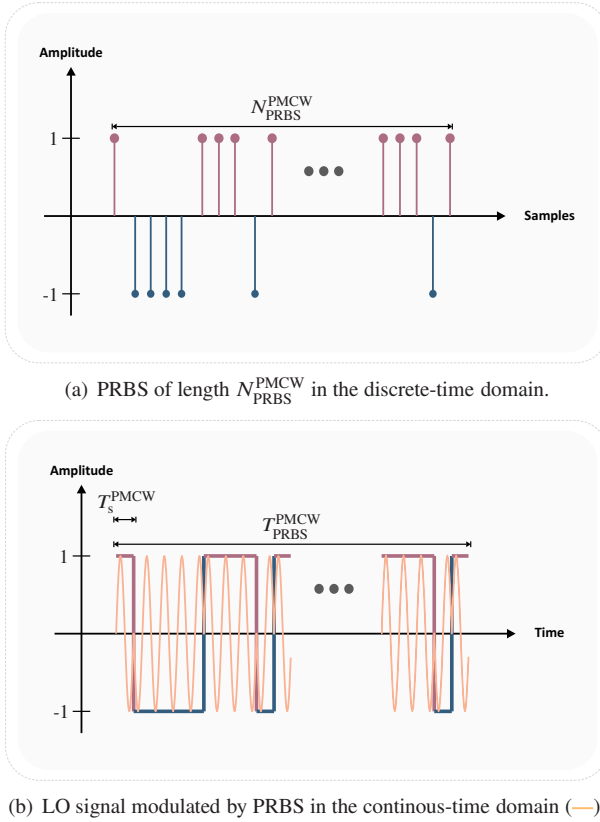


Figure 2.3: Binary PMCW principle: the RF transmit signal consists of an LO signal, which is a CW, modulated by a PRBS.

PRBS. Since the receive sequence is the result of the convolution of the transmit PRBS with the channel impulse response (CIR), the result of the pulse compression is a convolution between the PRBS autocorrelation function and the CIR. While for radar this can already be used to estimate the range of targets, PMCW-based communications require additional channel equalization, therefore resulting in a matched-filter communications receiver.

As both radar and communications performances in PMCW-based systems depend on the autocorrelation function of the PRBS, the adopted PRBS must be chosen in order to ensure low sidelobes and robustness to Doppler shifts. Based on the analysis in [19], m-sequences are an appropriate choice among typically adopted PRBSs for PMCW. Besides their robustness, using m-sequences as the reference PRBS in PMCW-based systems results in a simplified transmitter architecture. The reasons for that are the binary nature of m-sequences, which does not require I/Q modulation at the transmitter, and the fact that m-sequences can be generated using linear-feedback shift registers (LFSRs) instead of DACs.

Expanding on the brief explanation provided for the PMCW principle, Section 2.2.2 provides a detailed description of the processing steps for both radar and communications functions in a PMCW-based ISAC system.

2.2.2 PMCW-based ISAC processing chain

The transmit frame of the considered SISO PMCW-based ISAC system contains $M^{\text{PMCW}} \in \mathbb{N}_{>0}$ blocks of time duration $T_{\text{block}}^{\text{PMCW}}$, each of them containing $A^{\text{PMCW}} \in \mathbb{N}_{>0}$ repetitions of a same BPSK-modulated version of a PRBS. The aforementioned PRBS has $N_{\text{PRBS}}^{\text{PMCW}} \in \mathbb{N}_{>0}$ chips that are updated with a chip rate F_s^{PMCW} , which results in a PRBS time duration $T_{\text{PRBS}}^{\text{PMCW}} = N_{\text{PRBS}}^{\text{PMCW}} / F_s^{\text{PMCW}}$. Alternatively, the chips and the chip rate can be respectively interpreted as samples and sampling rate. Consequently, the frame duration is $T_{\text{frame}}^{\text{PMCW}} = M^{\text{PMCW}} T_{\text{block}}^{\text{PMCW}} = M^{\text{PMCW}} (A^{\text{PMCW}} T_{\text{PRBS}}^{\text{PMCW}})$.

To choose an appropriate PRBS, aspects such as achievable sidelobe level and robustness against Doppler shifts must be investigated [OJUY20], besides requirements for sequence generation. A possible PRBS generation strategy is the use of a LFSR, which allows simpler and more cost-effective generation of, e.g., maximum-length sequences (MLSs), also known as m-sequences, than employing a DAC with the drawback of not supporting interpolation at the D/A conversion and therefore causing out-of-band (OOB) emission due to the sinc-shaped repetitions of the PRBS spectral content at the first Nyquist zone. This ultimately results in a wider occupied RF bandwidth, which according to [GGs⁺14] can be defined as $B^{\text{PMCW}} = 2F_s^{\text{PMCW}}$ as shown in Figure 2.4 for a single PRBS repetition assuming the use of an MLS. To suppress OOB emission, one can adopt strategies such as the spectral sidelobe suppression

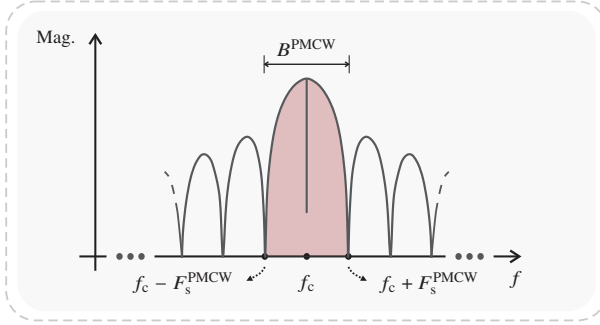


Figure 2.4: Evaluated RF spectrum without LP or BP filtering for a single PRBS repetition, which is an MLS in the considered example. © 2022 IEEE. Reprinted, with permission, from [18].

proposed in [GG⁺14], which is based on 3rd harmonic rejection and low-pass (LP) filtering before modulation of the RF carrier.

Although PMCW-based systems can adopt polyphase sequences, the use of PRBSs is preferred to avoid the need for an I/Q transmitter and to have higher energy efficiency, since the resulting transmit signal of a PRBS-based PMCW system is a CW, which allows operating the PA in compression [GG⁺14]. As for the choice for BPSK modulation, it is intended to keep the resulting modulated copies of the PRBS binary and consequently profit from the typical characteristics from binary PMCW radar systems.

The processing chain of the considered PMCW-based ISAC system with the previously described transmit signal characteristics is depicted in Figure 2.5. At the transmitter side, the m th symbol, $m \in \{0, 1, \dots, M^{\text{PMCW}} - 1\}$, of the modulated data stream $\mathbf{d}_{\text{Tx}}^{\text{PMCW}} \in \mathbb{C}^{M^{\text{PMCW}} \times 1}$, which is denoted as $d_{\text{Tx},m}^{\text{PMCW}} \in \mathbb{C} | d_{\text{Tx},m}^{\text{PMCW}} = e^{j\varphi_m^{\text{PMCW}}}$ and is assumed to belong to a BPSK constellation, i.e., $\varphi_m^{\text{PMCW}} \in \{0, \pi\}$, undergoes modulation in the slow-time with the adopted PRBS [DHS⁺17, DSN⁺17, DMMO19, MBK⁺19]. The resulting BPSK-modulated PRBS is then repeated A^{PMCW} times to allow accumulation at the receive side of the PMCW-based ISAC system [GG⁺14, BAG⁺16, GSD⁺17], which is denoted in Figure 2.5 as $A^{\text{PMCW}} \times \text{REP}$. It is worth highlighting that the A^{PMCW} -time repetition at the transmit side and the accumulation of each A^{PMCW} repetitions of the BPSK-modulated PRBS are complementary

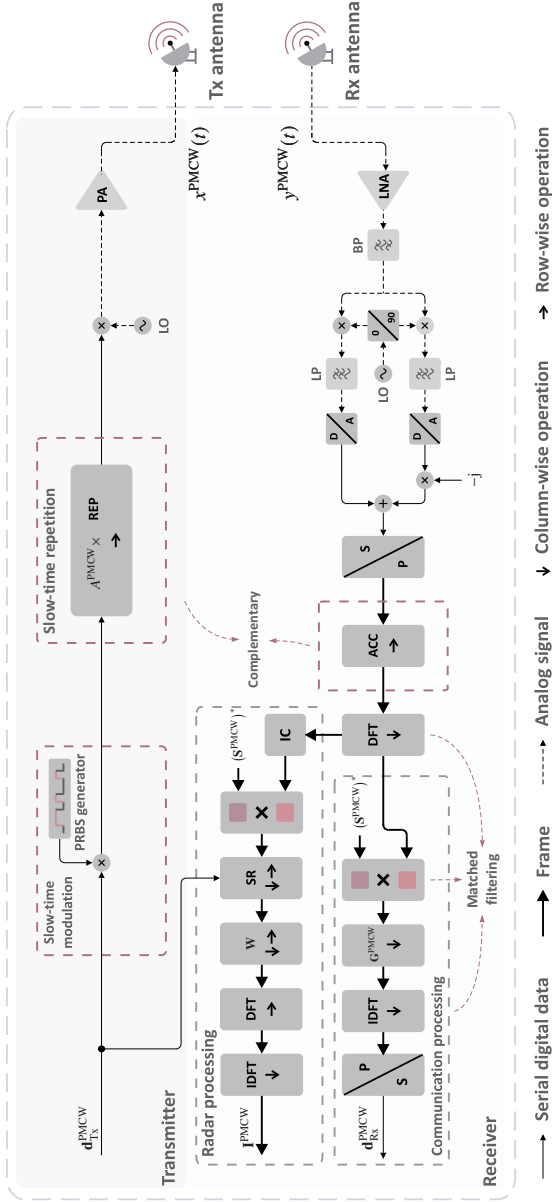


Figure 2.5: PMCW-based ISAC system representation. Adapted from [18], © 2022 IEEE.

operations, which are optional in PMCW-based radar systems where no modulation of PRBSs take place. In the considered ISAC context for MIMO based on outer coding, however, these operations are necessary. Next, the modulated frame consisting of M^{PMCW} blocks carrying a total of M^{PMCW} BPSK symbols modulates the RF carrier of frequency f_c , which results in a CW of frequency f_c that is phase-modulated by both PRBS chips and BPSK symbols, hence the name PMCW. The aforementioned signal then undergoes amplification by a PA, which can operate in compression since its input signal is a CW, and the resulting signal $x^{\text{PMCW}}(t) \in \mathbb{R}$ is transmitted by the transmit antenna. If power factors are omitted for the sake of simplicity, then the continuous-time domain signal containing the m th block transmitted within the transmit frame of the PMCW-based ISAC system can be expressed as

$$x_m^{\text{PMCW}}(t) = \left[\sum_{a=0}^{A^{\text{PMCW}}-1} d_{\text{Tx},m}^{\text{PMCW}} s^{\text{PMCW}} \left(t - aT_{\text{PRBS}}^{\text{PMCW}} - mT_{\text{block}}^{\text{PMCW}} \right) \right. \\ \left. \text{rect} \left(\frac{t - aT_{\text{PRBS}}^{\text{PMCW}} - mT_{\text{block}}^{\text{PMCW}}}{T_{\text{PRBS}}^{\text{PMCW}}} \right) \right] \cos(2\pi f_c t), \\ t \in [mT_{\text{block}}^{\text{PMCW}}, (m+1)T_{\text{block}}^{\text{PMCW}}]. \quad (2.8)$$

In this equation, $s^{\text{PMCW}}(t) \in \{-1, 1\}$ is the adopted PRBS defined for $t \in [0, T_{\text{PRBS}}^{\text{PMCW}})$. The continuous-time domain transmit signal $x^{\text{PMCW}}(t)$ containing the frame comprising M^{PMCW} blocks transmitted by the PMCW-based ISAC system can be therefore expressed as

$$x^{\text{PMCW}}(t) = \sum_{m=0}^{M^{\text{PMCW}}-1} x_m^{\text{PMCW}}(t) \text{rect} \left(\frac{t - mT_{\text{block}}^{\text{PMCW}}}{T_{\text{block}}^{\text{PMCW}}} \right). \quad (2.9)$$

Based on the presented concepts, Figure 2.6 depicts a serial representation of the frame in the considered PMCW-based ISAC system.

At the receiver of the PMCW-based ISAC system, the receive signal $y^{\text{PMCW}}(t) \in \mathbb{R}$ containing the output version of $x^{\text{PMCW}}(t)$ by the radar channel, the output version of the transmit signal $z^{\text{PMCW}}(t) \in \mathbb{R}$ of another ISAC system by the communications channel, and additive noise according to (2.1) is captured by the receive antenna. $y^{\text{PMCW}}(t)$ then undergoes analog conditioning

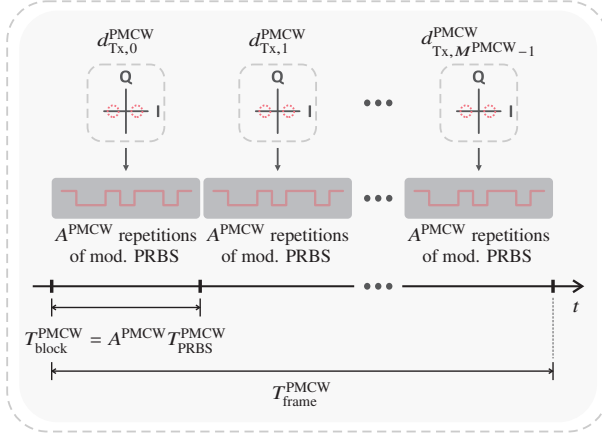


Figure 2.6: PMCW transmit signal: frame serial representation. Adapted from [18], © 2022 IEEE.

in an I/Q receiver and analog-to-digital (A/D) conversion with sampling rate F_s^{PMCW} and corresponding sampling period $T_s^{\text{PMCW}} = 1/F_s^{\text{PMCW}}$, being the samples from both I and Q channels into the real and imaginary parts of a serial vector. After serial-to-parallel (S/P) conversion on the aforementioned vector, accumulation of A^{PMCW} PRBS repetitions, each modulated with the same BPSK symbol, is optionally performed in the slow-time direction. Since there may be phase breaks between consecutive blocks modulated with different BPSK symbols S_m^{PMCW} , the first out of the A^{PMCW} PRBS repetitions of each block must be discarded before accumulation. The aforementioned processing is similar to what happens in outer-code-based MIMO-PMCW radar systems [BAG⁺16], and intends to avoid an effect that can be compared to the well-known ISI in communications systems [HYW⁺09]. Furthermore, A^{PMCW} must in principle assume the highest value possible to avoid processing gain loss due to the discarded sequence repetition at the beginning of each block. However, very high A^{PMCW} values may lead to incoherence of the accumulation due to Doppler-shift-induced phase rotations, which also leads to processing gain reduction and data rate reduction. This is described in detail in [17], where an example for a MIMO-PMCW radar is given.

After slow-time accumulation, discrete Fourier transforms (DFTs) are finally performed in the fast-time direction, i.e., along the chips of each PMCW block

and different processing strategies are adopted for radar sensing and communications. These strategies, as well as, typical multiplexing schemes for MIMO or multiuser operation of the considered PMCW-based ISAC system are described as follows.

2.2.2.1 Radar processing

After IC, the remaining radar processing steps in the considered PMCW-based ISAC system comprises an element-wise multiplication of each block by the vector $(\mathbf{S}^{\text{PMCW}})^* \in \mathbb{C}^{N_{\text{PRBS}}^{\text{PMCW}} \times 1}$, which is the complex-conjugate version of the DFT of the vector $\mathbf{s}^{\text{PMCW}} \in \{-1, 1\}^{N_{\text{PRBS}}^{\text{PMCW}} \times 1}$ that represents the adopted PRBS in the discrete-time domain. The resulting frame undergoes symbol removal by multiplying each of its columns with the respective BPSK symbols used to modulate the transmit blocks, which is denoted in Figure 2.5 as SR. Next, range and velocity windowing are performed in fast- and slow-time directions, respectively, with the aim of suppressing sidelobes in the radar image to be formed. The output matrix from the aforementioned windowing subsequently undergoes row-wise DFTs, i.e., DFTs in the slow-time direction, to convert Doppler-shift-induced phase rotations into Doppler shift estimates, which are translated into relative radial velocities. Finally, column-wise inverse discrete Fourier transforms (IDFTs), i.e., IDFTs in the fast-time direction, are performed to estimate time delays, which are translated into range. Together with the previously element-wise multiplication by $(\mathbf{S}^{\text{PMCW}})^*$ in the discrete-frequency domain, the IDFT operation yields a matched filtering, which is more computationally efficient and yields the same result as the periodic cross-correlation function (PCCF) of the receive blocks with the adopted PRBS [OJUY20].

The ultimately obtained range-velocity radar image $\mathbf{I}^{\text{PMCW}} \in \mathbb{C}^{N_{\text{PRBS}}^{\text{PMCW}} \times M^{\text{PMCW}}}$ experiences a processing gain G_p^{PMCW} and has range resolution ΔR^{PMCW} and a maximum unambiguous range $R_{\text{max,ua}}^{\text{PMCW}}$, as well as a relative radial velocity resolution Δv^{PMCW} and a maximum unambiguous relative radial velocity $v_{\text{max,ua}}^{\text{PMCW}}$ as shown in Table 2.2 in Section 2.5 based on [GGs⁺14, WGBC19, OJUY20]. While no further restrictions on maximum tolerable range apply in the considered PMCW-based ISAC system, the definition of a maximum tolerable relative radial velocity depends on the Doppler shift tolerance of the adopted PRBS [OJUY20] and is therefore not defined in Table 2.2.

2.2.2.2 Communications processing

Before starting the communications processing of the considered PMCW-based ISAC system, it is first assumed that the time, frequency and phase offsets from the second term in (2.1) are removed via synchronization based on a preamble comprised of unmodulated PRBSs. The remaining communications processing steps comprise a joint equalization and matched filtering in the discrete-frequency domain, which shares some similarities with the previously described radar processing. In this sense, every m th column of the resulting frame from discrete-frequency domain matched filtering undergoes element-wise multiplication by the corresponding vector $\mathbf{G}_m^{\text{PMCW}} \in \mathbb{C}^{N_{\text{PRBS}}^{\text{PMCW}} \times 1}$. Assuming that $\mathbf{G}_m^{\text{PMCW}}$ contains the multiplicative inverse of the communications channel frequency response (CFR) estimate $\mathbf{H}_m^{\text{PMCW}} \in \mathbb{C}^{N_{\text{PRBS}}^{\text{PMCW}} \times 1}$ for the m th block, the aforementioned processing yields a ZF frequency-domain equalization (FDE). To obtain the aforementioned CFR estimates, unmodulated blocks of PRBSs are used similarly to what is done for synchronization as previously mentioned. For conciseness, no further details on synchronization and channel estimation in the considered PMCW-based ISAC system are given and the reader is referred to [51]. To compensate phase rotation introduced by Doppler shifts indicated in (2.1), the CFR estimate must be periodically updated for increasing m via unmodulated PRBSs. The resulting frame from the joint equalization and matched filtering finally undergoes parallel-to-serial (P/S) conversion, which consists of extracting the receive BPSK symbols from the peak at the first samples of the frame columns and forming the receive modulated data stream $\mathbf{d}_{\text{Rx}}^{\text{PMCW}} \in \mathbb{C}^{M^{\text{PMCW}} \times 1}$. The described communications processing steps in the considered PMCW-based ISAC system yield data rate denoted by $\mathcal{R}_{\text{BPSK}}^{\text{PMCW}}$, as well as energy per bit to noise power spectral density ratio $(E_b/N_0)_{\text{BPSK},m}^{\text{PMCW}}$ and bit error ratio (BER) denoted by $\text{BER}_{\text{BPSK},m}^{\text{PMCW}}$ assuming BPSK modulation as presented in Table 2.2 in Section 2.5, where $H_{k,m}^{\text{PMCW}} \in \mathbb{C}$ is the k th element, $k \in \{0, 1, \dots, N_{\text{PRBS}}^{\text{PMCW}} - 1\}$, of $\mathbf{H}_m^{\text{PMCW}}$.

2.2.2.3 Multiplexing for MIMO and multiuser scenarios

For enabling MIMO or multiuser operation in PMCW-based ISAC systems, a common practice is to adopt code-division multiplexing (CDM) aiming for

interference-free operation of N_{Tx} transmitters. The reason why time-division multiplexing (TDM)-based approaches are usually not adopted is the greater flexibility in the design of coded PMCW signals with the possibility of not experiencing the processing gain loss caused by reduced number of transmitted PRBSs per transmitter that would occur in the TDM case. As for frequency-division multiplexing (FDM), it is usually not adopted since the design of PRBSs or general pseudorandom sequences is usually not flexible in the discrete-frequency domain. Approaches such as the commonly used interleaved FDM in OFDM systems [SSBZ13] are therefore not suitable for PMCW-based ISAC systems.

In the CDM context, a possible multiplexing approach is based on the use of different PRBSs that are orthogonal to each other by each of the N_{Tx} transmitters [WGBC19, OJUY20]. While this approach results in no changes in the radar and communications performance parameters listed in Table 2.2, it requires a total of N_{Tx} different correlators at the receiver side to support the N_{Tx} orthogonal PRBSs. A possible issue with this approach is the relatively high sidelobe level of the PCCFs of orthogonal PRBSs, which tends to limit the radar dynamic range to about 30 dB for the commonly adopted PRBSs [WGBC19].

Alternatively, outer codes [BAG⁺16, GSD⁺17, WGBC19] can be integrated into the signal structure of the SISO case described earlier in this section. A possible outer code choice is a binary Hadamard code with N_{Tx} codewords of length N_{Tx} [BAG⁺16, GSD⁺17, WGBC19]. In this approach, each of the N_{Tx} codewords is assigned to a different transmitter, which transmits N_{Tx} copies of each block modulated with the corresponding binary element of its codeword. The simplest example would be for $N_{\text{Tx}} = 2$, in which the first transmitter transmits two copies of the same block, both with 0° phase shifts since the first transmitter is associated with the Hadamard codeword $[1, 1]$. Similarly, the second transmitter transmits one copy of the block with 0° phase shift and another one with 180° phase shift following its Hadamard codeword $[1, -1]$. At the receiver side, the receive blocks are decoded via phase-shifting by corresponding element of the codeword of the evaluated transmitter and summed. Ideally, this should result in the block transmitted by the evaluated transmitter with a processing gain of $10 \log_{10}(N_{\text{Tx}})$ dB due to the averaging-like effect of the decoding. The described CDM multiplexing approach increases $T_{\text{block}}^{\text{PMCW}}$ by a factor of N_{Tx} , becoming $N_{\text{Tx}} T_{\text{block}}^{\text{PMCW}}$. Consequently, $v_{\text{max,ua}}^{\text{PMCW}}$ is decreased by a factor of N_{Tx} to $v_{\text{max,ua}}^{\text{PMCW}} = F_s^{\text{PMCW}} c_0 / (4 f_c N_{\text{Tx}} N_{\text{PRBS}}^{\text{PMCW}} A^{\text{PMCW}})$. Additionally, the processing

gain is changed to $G_p^{\text{PMCW}} = N_{\text{Tx}} N_{\text{PRBS}}^{\text{PMCW}} A^{\text{PMCW}} M^{\text{PMCW}}$, although it may assume the same value as in the SISO case due to the necessary reduction of M^{PMCW} by a factor of N_{Tx} to keep $T_{\text{frame}}^{\text{PMCW}}$ constant. In terms of communications performance, the only change w.r.t. to the listed parameters in Table 2.2 is the reduction of the achieved data rate assuming BPSK modulation by a factor of N_{Tx} to $\mathcal{R}_{\text{BPSK}}^{\text{PMCW}} = F_s^{\text{PMCW}} / [N_{\text{Tx}} (N_{\text{PRBS}}^{\text{PMCW}} A^{\text{PMCW}})]$.

2.3 Orthogonal frequency-division multiplexing

In this section, the OFDM principle is discussed in Section 2.3.1. The processing chain for OFDM-based ISAC is then described in Section 2.3.2.

2.3.1 OFDM principle

At the transmitter side of the OFDM scheme, a vector representing an OFDM symbol in the discrete-frequency domain contains $N_{\text{subc}}^{\text{OFDM}} \in \mathbb{N}_{>0}$ modulation symbols that are to be transmitted through $N_{\text{subc}}^{\text{OFDM}}$ orthogonal subcarriers. The aforementioned vector undergoes an IDFT, which can be implemented as an inverse fast Fourier transforms (IFFTs) for reduced computational complexity. The result is a vector representing the discrete-time domain OFDM symbol, whose n th element $x_n^{\text{OFDM}} \in \mathbb{C}$, $n \in \{0, 1, \dots, N_{\text{subc}}^{\text{OFDM}} - 1\}$, is expressed as

$$x_n^{\text{OFDM}} = \frac{1}{N_{\text{subc}}^{\text{OFDM}}} \sum_{k=0}^{N_{\text{subc}}^{\text{OFDM}}-1} X_k^{\text{OFDM}} e^{j \frac{2\pi n k}{N_{\text{subc}}^{\text{OFDM}}}}. \quad (2.10)$$

$X_k^{\text{OFDM}} \in \mathbb{C}$, $k \in \{0, 1, \dots, N_{\text{subc}}^{\text{OFDM}} - 1\}$, in this equation represents the k th element of the discrete-frequency domain OFDM symbol. Consequently, the OFDM symbol after transformation into the discrete-time domain symbol consists of the sum of $N_{\text{subc}}^{\text{OFDM}}$ orthogonal, modulated subcarriers² as illustrated in Figure 2.7. In OFDM-based systems, transmission is usually based on frames

² Hence the subscript “subc” in $N_{\text{subc}}^{\text{OFDM}}$.

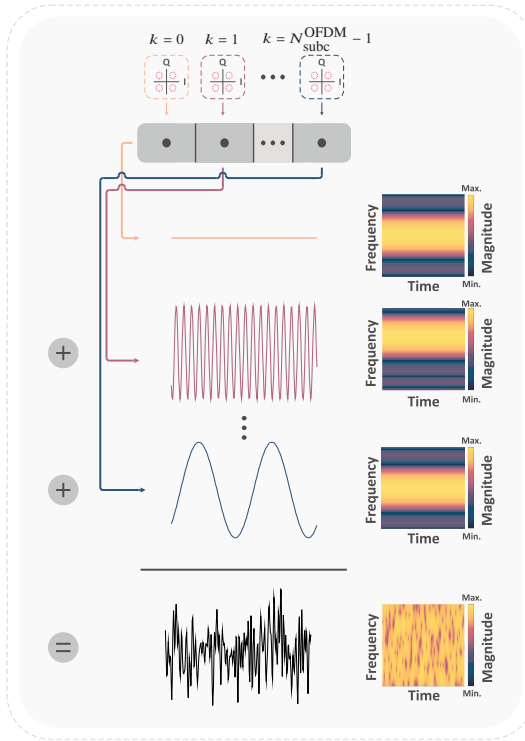


Figure 2.7: OFDM principle: sum of orthogonal, modulated subcarriers with modulation symbols belonging to a digital modulation constellation.

that are constituted by multiple OFDM symbols. To avoid interference among consecutively transmitted OFDM symbols, which is called ISI, and simultaneously turn the linear convolution of the OFDM symbols with the CIR into a circular convolution at the receiver side, a CP is prepended to each OFDM symbol in the discrete-time domain.

At the receiver side, the CP must be removed from each OFDM symbol so that they can be transformed into the discrete-frequency domain via DFT and therefore allow recovering the originally transmitted modulation symbols. Neglecting the effects of the propagation through a channel and AWGN for simplicity, the received samples in the discrete-time and frequency domains will be

equal to x_n^{OFDM} and X_k^{OFDM} , respectively. The aforementioned DFT operation, which can be implemented as a fast Fourier transform (FFT) for computational efficiency, can be therefore expressed as

$$X_k^{\text{OFDM}} = \sum_{n=0}^{N_{\text{subc}}^{\text{OFDM}}-1} x_n^{\text{OFDM}} e^{-j \frac{2\pi n k}{N_{\text{subc}}^{\text{OFDM}}}}. \quad (2.11)$$

If a CIR were considered, the DFT convolution theorems states that the experienced circular convolution with it in the discrete-time would yield a multiplication of the OFDM symbol with the CFR in the discrete-frequency domain provided that sufficient CP is used.

The preceding discussion summarizes OFDM processing for communications in a simplified scenario. Additional considerations are, however, required to enable ISAC functionality. In this sense, Section 2.3.2 provides a detailed examination of the OFDM-based processing chain tailored for ISAC systems.

2.3.2 OFDM-based ISAC processing chain

The considered SISO OFDM-based ISAC system transmits a frame that comprises of $M^{\text{OFDM}} \in \mathbb{N}_{>0}$ blocks, each containing one OFDM symbol that carries $N_{\text{subc}}^{\text{OFDM}} \in \mathbb{N}_{>0}$ quadrature phase-shift keying (QPSK)-modulated subcarriers and a CP to prevent ISI when propagating through both radar and communications channels [HYW⁺09]. The OFDM frame carries therefore data from a modulated data stream $\mathbf{d}_{\text{Tx}}^{\text{OFDM}} \in \mathbb{C}^{N_{\text{subc}}^{\text{OFDM}} M^{\text{OFDM}} \times 1}$ containing a total of $N_{\text{subc}}^{\text{OFDM}} M^{\text{OFDM}}$ QPSK symbols, which ultimately leads to a considerably higher data rate as in the PMCW-based ISAC system from Section 2.2. Although higher-order digital modulations could be adopted for enabling higher data rates, the choice for QPSK is justified by the fact that it is less sensitive to phase rotations and results in the same allocated power at OFDM subcarriers, which ultimately yields more reliable radar measurements.

Following [SW11, Bra14, Sit17, BRT⁺19], the processing chain of the considered OFDM-based ISAC system with the previously described transmit signal characteristics is depicted in Figure 2.8. The modulated data stream $\mathbf{d}_{\text{Tx}}^{\text{OFDM}} \in \mathbb{C}^{N_{\text{subc}}^{\text{OFDM}} M^{\text{OFDM}} \times 1}$ first undergoes S/P conversion, generating the

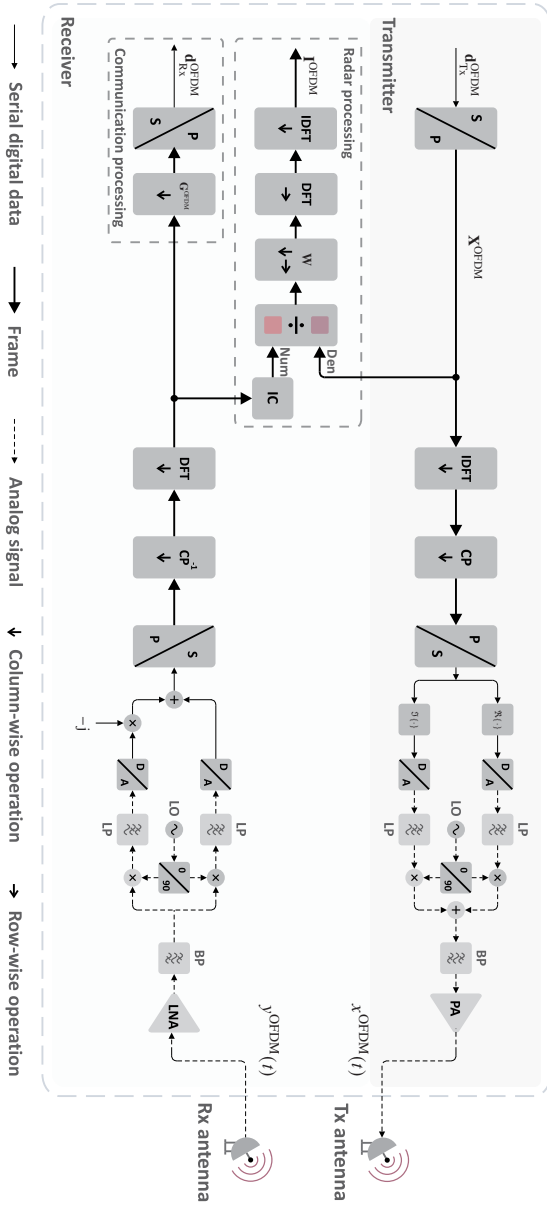


Figure 2.8: OFDM-based ISAC system representation. Adapted from [18], © 2022 IEEE.

discrete-frequency domain OFDM frame $\mathbf{X}^{\text{OFDM}} \in \mathbb{C}^{N_{\text{sub}}^{\text{OFDM}} \times M^{\text{OFDM}}}$. The frame $\mathbf{X}^{\text{OFDM}}$ then undergoes column-wise IDFTs along the $N_{\text{sub}}^{\text{OFDM}}$ QPSK-modulated subcarriers of each of its M^{OFDM} OFDM symbols, and the resulting discrete-time domain OFDM frame has CPs of length $N_{\text{CP}}^{\text{OFDM}} \in \mathbb{N}_{>0}$ prepended to each of its M^{OFDM} OFDM symbols. Next, the discrete-time domain OFDM frame with prepended CPs, which has size $(N_{\text{sub}}^{\text{OFDM}} + N_{\text{CP}}^{\text{OFDM}}) \times M^{\text{OFDM}}$, undergoes P/S conversion and the real and imaginary parts of the resulting vector are fed to DACs of sampling rate $F_{\text{s,DAC}}^{\text{OFDM}}$. The output analog signals by the DACs undergo analog conditioning and I/Q modulation to the carrier of frequency f_c , producing the analog signal $x^{\text{OFDM}}(t) \in \mathbb{R}$ that is transmitted by the transmit antenna of the OFDM-based ISAC system. It is worth highlighting that, in contrast to the PMCW-based ISAC system from Section 2.2, an input power back-off from the PA compression is required in the OFDM-based ISAC system due to the high PAPR experienced in OFDM signals [HYW⁺09], which can be higher than 10 dB depending on the adopted modulation scheme and number of subcarriers $N_{\text{sub}}^{\text{OFDM}}$.

Omitting power factors for simplicity and assuming that no CP is used, the continuous-time domain signal $x_m^{\text{OFDM, w/o CP}}(t, \tau) \in \mathbb{C}$ carrying the m th OFDM symbol, $m \in \{0, 1, \dots, M^{\text{OFDM}} - 1\}$, can be expressed as

$$\begin{aligned} x_m^{\text{OFDM, w/o CP}}(t, \tau) = & \Re \left\{ \frac{1}{N_{\text{sub}}^{\text{OFDM}}} \sum_{k=0}^{N_{\text{sub}}^{\text{OFDM}}-1} X_{\langle k \rangle_{N/2}, m}^{\text{OFDM}} e^{j2\pi(k-N/2)\Delta f^{\text{OFDM}} t} \right\} \cos(2\pi f_c \tau) \\ & + \Im \left\{ \frac{1}{N_{\text{sub}}^{\text{OFDM}}} \sum_{k=0}^{N_{\text{sub}}^{\text{OFDM}}-1} X_{\langle k \rangle_{N/2}, m}^{\text{OFDM}} e^{j2\pi(k-N/2)\Delta f^{\text{OFDM}} t} \right\} \sin(2\pi f_c \tau), \\ & t \in [0, T_{\text{symb}}^{\text{OFDM}}), \tau \in [mT_{\text{block}}^{\text{OFDM}}, (m+1)T_{\text{block}}^{\text{OFDM}}). \end{aligned} \quad (2.12)$$

In this equation, $X_{k,m}^{\text{OFDM}} \in \mathbb{C} | X_{k,m}^{\text{OFDM}} = e^{j\varphi_{k,m}^{\text{OFDM}}}$ is the QPSK symbol with phase $\varphi_{k,m}^{\text{OFDM}} \in \{\pi/4, 3\pi/4, 5\pi/4, 7\pi/4\}$ that is modulated onto the k th subcarrier, $k \in \{0, 1, \dots, N_{\text{sub}}^{\text{OFDM}} - 1\}$, of the m th OFDM symbol. Furthermore, $\Delta f^{\text{OFDM}} = 1/T_{\text{symb}}^{\text{OFDM}}$ is the subcarrier frequency spacing defined by the OFDM symbol time duration disregarding CP $T_{\text{symb}}^{\text{OFDM}} = N_{\text{sub}}^{\text{OFDM}}/B^{\text{OFDM}}$, which in

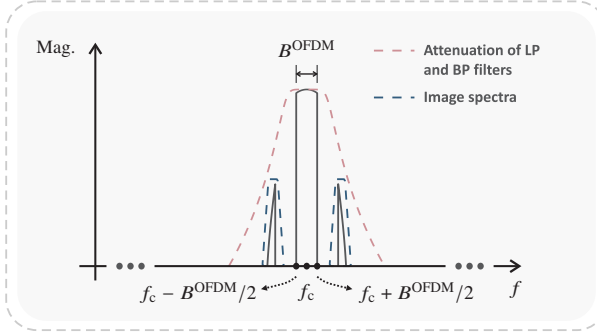


Figure 2.9: Evaluated spectrum for one OFDM symbol. © 2022 IEEE. Reprinted, with permission, from [18].

turns depends on the RF bandwidth B^{OFDM} occupied by the OFDM signal. It is worth highlighting that B^{OFDM} is defined assuming that the adopted sampling rate $F_{s,\text{DAC}}^{\text{OFDM}}$ is higher than the Nyquist critical sampling rate and that the LP filtering right after D/A conversion is capable to suppress the repetitions of the OFDM signal spectral content outside the first Nyquist zone. Those repetitions are produced by the sample and hold (S&H) circuit in the DACs, which also yields a sinc-shaped attenuation that can be digitally compensated by a inverted sinc filter. The aforementioned interpolation and upconversion of the transmit OFDM signal ultimately result in an RF spectrum for a single OFDM symbol as depicted in Figure 2.9 and can be achieved by using extra, inactive subcarriers at the edges of the OFDM symbol in the discrete-frequency domain. Consequently, the continuous-time domain signal $x_m^{\text{OFDM}}(t) \in \mathbb{R}$ carrying the m th OFDM symbol with $N_{\text{CP}}^{\text{OFDM}}$ -length CP can be expressed as

$$x_m^{\text{OFDM}}(t) = \begin{cases} x_m^{\text{OFDM, w/o CP}} \left(t + T_{\text{symb}}^{\text{OFDM}} - T_{\text{CP}}^{\text{OFDM}} - mT_{\text{block}}^{\text{OFDM}}, t \right), \\ t \in \left[mT_{\text{block}}^{\text{OFDM}}, mT_{\text{block}}^{\text{OFDM}} + T_{\text{CP}}^{\text{OFDM}} \right) \\ \\ x_m^{\text{OFDM, w/o CP}} \left(t - T_{\text{CP}}^{\text{OFDM}} - mT_{\text{block}}^{\text{OFDM}}, t \right), \\ t \in \left[mT_{\text{block}}^{\text{OFDM}} + T_{\text{CP}}^{\text{OFDM}}, (m+1)T_{\text{block}}^{\text{OFDM}} \right) \end{cases} . \quad (2.13)$$

In this equation, $T_{\text{block}}^{\text{OFDM}} = T_{\text{symbol}}^{\text{OFDM}} + T_{\text{CP}}^{\text{OFDM}}$ is the duration of each OFDM block consisting of an OFDM symbol and its corresponding CP of time duration $T_{\text{CP}}^{\text{OFDM}} = N_{\text{CP}}^{\text{OFDM}}/B^{\text{OFDM}}$. The continuous-time domain signal $x^{\text{OFDM}}(t) \in \mathbb{R}$ containing the frame comprising M^{OFDM} blocks transmitted by the OFDM-based ISAC system can be finally expressed as

$$x^{\text{OFDM}}(t) = \sum_{m=0}^{M^{\text{OFDM}}-1} x_m^{\text{OFDM}}(t) \text{rect}\left(\frac{t - mT_{\text{block}}^{\text{OFDM}}}{T_{\text{block}}^{\text{OFDM}}}\right). \quad (2.14)$$

Based on the presented concepts, Figures 2.10(a) and 2.10(b) depict serial representations of the frame and a block in the considered OFDM-based ISAC system, respectively.

At the receiver side of the OFDM-based ISAC system, the output version of $x^{\text{OFDM}}(t)$ by the radar channel, the output version of the transmit signal $z^{\text{OFDM}}(t) \in \mathbb{R}$ of another ISAC system by the communications channel, and additive noise according to (2.1) are combined into the continuous-time domain signal $y^{\text{OFDM}}(t) \in \mathbb{R}$ that is captured by the receive antenna. After the same analog conditioning, A/D conversion with sampling rate $F_s^{\text{OFDM}} = B^{\text{OFDM}}$ and corresponding sampling period $T_s^{\text{OFDM}} = 1/F_s^{\text{OFDM}}$ is performed for both I and Q channels as described for the PMCW case in Section 2.2. The resulting complex-valued vector undergoes S/P conversion, forming a discrete-time domain OFDM frame containing M^{OFDM} receive symbols with CP. Next CP removal and DFTs are performed along the M^{OFDM} columns of the aforementioned frame, which produces a discrete-frequency domain receive OFDM frame of size $N_{\text{subc}}^{\text{OFDM}} \times M^{\text{OFDM}}$, on which distinct further processing steps are performed for radar sensing and communications. Both aforementioned steps as well as a multiplexing strategy to enable MIMO or multiuser operation of the considered OFDM-based ISAC system are discussed as follows.

2.3.2.1 Radar processing

The remaining radar processing steps in the considered OFDM-based ISAC system start with IC, which consists of removing the influence of communications signals originally transmitted by other ISAC or communications systems from the discrete-frequency domain receive OFDM frame. Afterwards, the resulting frame is divided by the frame $\mathbf{X}^{\text{OFDM}}$ that contains the transmit QPSK

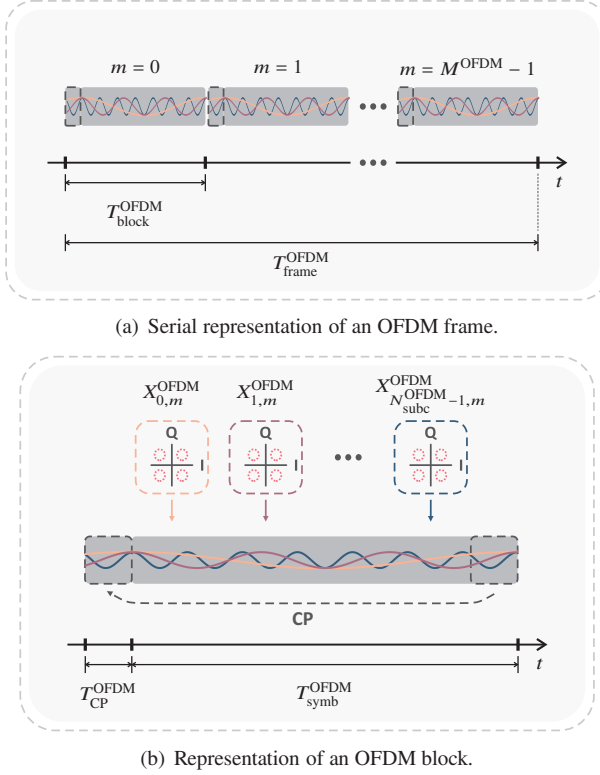


Figure 2.10: OFDM transmit signal structure. Adapted from [18], © 2022 IEEE.

modulation symbols, which yields one ZF estimate of the radar CFR for each of the M^{OFDM} symbols. The remaining processing steps are the same from the PMCW case in Subsection 2.2.2.1, i.e., range and velocity windowing, row-wise DFTs for Doppler shift estimation, and column-wise IDFTs for time delay estimation, being the estimated Doppler shifts and time delays translated into relative radial velocity and range estimates, respectively.

Following [FNG06, SZW09, NMZ18], the aforementioned processing steps in the OFDM-based ISAC system ultimately yield the range-velocity radar image $\mathbf{I}^{\text{OFDM}} \in \mathbb{C}^{N^{\text{OFDM}}_{\text{subc}} \times M^{\text{OFDM}}}$, which experiences a processing gain $G^{\text{OFDM}}_{\text{p}}$ and is as-

sociated with range resolution ΔR^{OFDM} , maximum unambiguous range $R_{\text{max,ua}}^{\text{OFDM}}$, and maximum ISI-free range $R_{\text{max,ISI}}^{\text{OFDM}}$, as well as relative radial velocity resolution, Δv^{OFDM} , maximum unambiguous relative radial velocity $v_{\text{max,ua}}^{\text{OFDM}}$, and maximum intercarrier interference (ICI)-free relative radial velocity $v_{\text{max,ICI}}^{\text{OFDM}}$ values as shown in Table 2.2 in Section 2.5.

2.3.2.2 Communications processing

Considering that time, frequency and phase offsets from the second term in (2.1) have already been compensated via synchronization based on a preamble defined by, e.g., the Schmidl and Cox algorithm [SC97], the remaining communications processing in the considered OFDM-based ISAC system comprises of a ZF FDE via fast-time, element-wise multiplication of every m th block by the corresponding vector $\mathbf{G}_m^{\text{OFDM}} \in \mathbb{C}^{N_{\text{sub}}^{\text{OFDM}} \times 1}$ that contains the multiplicative inverse of the estimate of the experienced communications CFR $\mathbf{H}_m^{\text{OFDM}} \in \mathbb{C}^{N_{\text{sub}}^{\text{OFDM}} \times 1}$. As in the PMCW case from Section 2.2, the CFR estimate must be updated periodically with increasing m to compensate for phase rotations introduced by Doppler shifts of scatterers indicated in (2.1), which is achieved with the use of pilot subcarriers in OFDM-based systems [SBM⁺04, HYW⁺09, SNZ18]. For an example of a full communications processing chain in OFDM-based ISAC, the reader is referred to [50]. The resulting frame from the FDE finally undergoes P/S conversion, generating the receive modulated data stream $\mathbf{d}_{\text{Rx}}^{\text{OFDM}} \in \mathbb{C}^{(N_{\text{sub}}^{\text{OFDM}} M^{\text{OFDM}}) \times 1}$.

The communications processing steps in the considered OFDM-based ISAC system result in the overall data rate denoted by $\mathcal{R}_{\text{QPSK}}^{\text{OFDM}}$, as well as energy per bit to noise power spectral density ratio $(E_b/N_0)_{\text{QPSK},k,m}^{\text{OFDM}}$ and BER denoted by $\text{BER}_{\text{QPSK},k,m}^{\text{OFDM}}$ for the k th QPSK-modulated subcarrier in the m th OFDM symbol as presented in Table 2.2 in Section 2.5, where $H_{k,m}^{\text{OFDM}} \in \mathbb{C}$ is the k th element of $\mathbf{H}_m^{\text{OFDM}}$.

2.3.2.3 Multiplexing for MIMO and multiuser scenarios

For MIMO and multiuser operation of OFDM-based ISAC systems, the widely adopted multiplexing approach to achieve orthogonality of a total of

N_{Tx} transmit signals is FDM based on an interleaved subcarrier assignment scheme [SSBZ13]. This scheme is adopted to ensure that signals from different transmit channels are orthogonal in the frequency domain as long as tolerable Doppler shifts are experienced, being those assumed to be equal to or less than the same upper bound of $0.1\Delta f^{\text{OFDM}}$ associated with the tolerable ICI in OFDM systems [HYW⁺09, SSBZ13, NSF⁺17, SNZ18]. The simplest example of this approach would be for $N_{\text{Tx}} = 2$ with even $N_{\text{subc}}^{\text{OFDM}}$, where every subcarrier of even index, i.e., $k \in \{0, 2, 4, \dots, N_{\text{subc}}^{\text{OFDM}} - 2\}$, would be assigned to the first transmitter, while every subcarrier of odd index, i.e., $k \in \{1, 3, 5, \dots, N_{\text{subc}}^{\text{OFDM}} - 1\}$, would be assigned to the second transmitter. After performing the processing steps until the DFT at the receiver side, the $N_{\text{subc}}^{\text{OFDM}}/N_{\text{Tx}}$ subcarriers assigned to the evaluated transmitter are selected and further processed. Consequently, the processing gain listed in Table 2.2 is changed to $G_{\text{p}}^{\text{OFDM}} = (N_{\text{subc}}^{\text{OFDM}}/N_{\text{Tx}})M^{\text{OFDM}}$, which tends to be $10 \log_{10}(N_{\text{Tx}})$ dB lower than in the SISO case. The last changed radar performance parameter w.r.t. the listed ones in Table 2.2 is the maximum unambiguous range, which is reduced to $R_{\text{max,ua}}^{\text{OFDM}} = (N_{\text{subc}}^{\text{OFDM}}/N_{\text{Tx}})c_0/(2B^{\text{OFDM}})$. Finally, it is worth highlighting that no communications performance parameters are changed in this multiplexing approach in the MIMO case if each transmit channel transmits a different payload, whereas the data rate assuming QPSK modulation is reduced by a factor of N_{Tx} to $\mathcal{R}_{\text{QPSK}}^{\text{OFDM}} = 2N_{\text{subc}}^{\text{OFDM}}B^{\text{OFDM}}/[N_{\text{Tx}}(N_{\text{subc}}^{\text{OFDM}} + N_{\text{CP}}^{\text{OFDM}})]$ in the multiuser case.

Straightforward alternative MIMO strategies for OFDM-based ISAC systems include TDM [SBM⁺04], which is usually not considered for the same reasons as in the PMCW case from Section 2.2. A further possibility is CDM based on either the use of uncorrelated OFDM symbols [NMZ18], yielding considerably poor isolation among different transmitters, or outer codes [HYW⁺09], which are same as discussed for the PMCW case in Section 2.2. Since the aforementioned approaches tend to be outperformed by the previously discussed interleaved FDM, they are not discussed in further detail. Although optimized CDM strategies have been recently investigated in the literature [KES⁺21], only widely adopted interleaved FDM will be henceforth considered for the sake of simplicity.

2.4 Orthogonal chirp-division multiplexing

This section starts with a discussion of the OCDM principle in Section 2.4.1. Two distinct applications of this principle for enabling OCDM-based ISAC, along with their associated processing chains, are then discussed in Sections 2.4.2 and 2.4.3.

2.4.1 OCDM principle

In the OCDM scheme, a discrete-Fresnel domain vector carrying $N_{\text{subch}}^{\text{OCDM}} \in \mathbb{N}_{>0}$ modulation symbols is transformed into a discrete-time domain vector at the transmitter side via the inverse discrete Fresnel transform (IDFnT) as [OZ16a]

$$x_n^{\text{OCDM}} = \frac{1}{N_{\text{subch}}^{\text{OCDM}}} e^{j\frac{\pi}{4}} \sum_{k=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_k^{\text{OCDM}} e^{-j\frac{\pi(n-k)^2}{N_{\text{subch}}^{\text{OCDM}}}}. \quad (2.15)$$

In this equation, $x_n^{\text{OCDM}} \in \mathbb{C}$ and $\dot{X}_k^{\text{OCDM}} \in \mathbb{C}$, where $n \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} - 1\}$ and $k \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} - 1\}$, represent the elements of the discrete-time domain and discrete-Fresnel domain OCDM symbols, respectively. The resulting discrete-time domain OCDM symbol consists of the sum of $N_{\text{subch}}^{\text{OCDM}}$ orthogonal, modulated subchirps² as depicted in Figure 2.11. As in the OFDM case as described in Section 2.3.1, a frame containing multiple OCDM symbols is usually transmitted and a CP is also prepended to each of the discrete-time domain OCDM symbols in the discrete-time domain to avoid ISI.

After CP removal at the receiver side, a DFNT is used to yield a discrete-Fresnel domain OCDM symbol and recover the transmitted modulation symbols as

$$\dot{X}_k^{\text{OCDM}} = e^{-j\frac{\pi}{4}} \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} x_n^{\text{OCDM}} e^{j\frac{\pi(n-k)^2}{N_{\text{subch}}^{\text{OCDM}}}}. \quad (2.16)$$

For the sake of simplicity, AWGN is neither considered in (2.16) nor in the following expressions throughout this chapter since the DFNT operation does

² Hence the subscript “subch” in $N_{\text{subch}}^{\text{OCDM}}$.

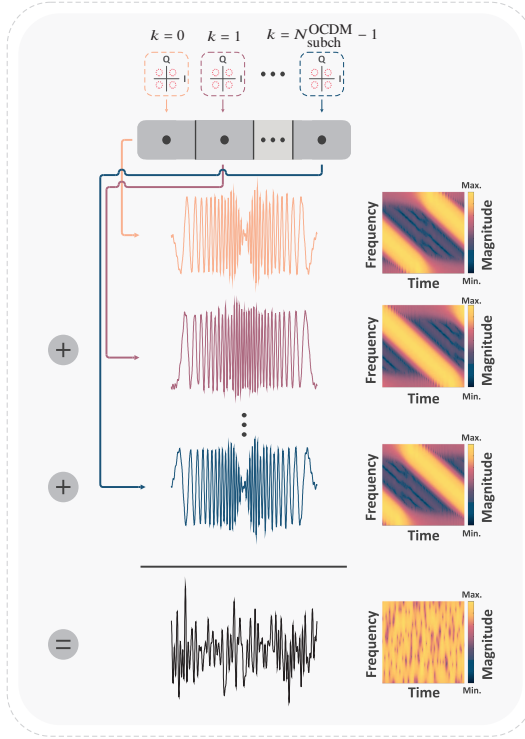


Figure 2.11. OFDM principle: sum of orthogonal, modulated subcarriers with modulation symbols belonging to a digital modulation constellation.

not change its statistics. This is supported by the claims in [OZ16a] and further discussion on this topic can be found in [OM21].

For computational efficiency, both DFNT and its inverse can be performed based on FFTs and/or IFFT and multiplications by phase terms. The aforementioned alternative DFNT implementation also allows to compensate for the propagation through a channel via FDE [OZ16a]. Alternatively, the DFNT can be interpreted as a circular convolution between the discrete-time domain OFDM symbol and the root Zadoff-Chu sequence $e^{j\pi n^2 / N_{\text{subch}}^{\text{OFDM}}}$ multiplied by $e^{-j\frac{\pi}{4}}$. Similarly, the IDFNT can be interpreted as a circular convolution between the vector carrying modulated data and the complex-conjugate version of the root Zadoff-Chu

sequence, i.e., $e^{-j\pi k^2/N_{\text{subch}}^{\text{OCDM}}}$, multiplied by $e^{j\frac{\pi}{4}/N_{\text{subch}}^{\text{OCDM}}}$. OCDM can therefore be classified as a CP direct-sequence spread spectrum (CP-DSSS) scheme in which a unit spreading gain is observed. The fact that OCDM is a special case of CP-DSSS based on circular convolutions with the root Zadoff-Chu sequence and its complex-conjugate version also explains the convolution theorem of the DF_nT. Unlike the DFT convolution theorem in OFDM-based systems, which states that a circular convolution in the discrete-time yields a multiplication in the discrete-frequency domain, the DF_nT convolution theorem states that a circular convolution in the discrete-time also yields a circular convolution in the discrete-Fresnel domain.

While the previous discussion summarizes the OCDM processing for communications purposes, further aspects must be considered to enable ISAC operation. In this context, two approaches for enabling radar ranging in OCDM-based systems are proposed next. Both are based on pulse compression and aim to reduce or eliminate the influence of transmitted modulation symbols onto the radar channel estimate, which can be interpreted as a range spectrum. The first is based on a circular-matched filtering of the receive OCDM symbol with the originally transmitted one in the discrete-time domain [18]. This compresses the transmit OCDM symbol into a narrow pulse and delivers a radar channel estimate that equals the convolution of the actual CIR with the aforementioned pulse. For the sake of clarity, the OCDM-based ISAC system adopting this first radar range processing approach will be henceforth referred to as conventional OCDM (C-OCDM). In the second radar range processing approach, the sector-modulated OCDM (SM-OCDM)-based ISAC system, a strategic subchirp allocation at the transmitter side is performed [20]. At the receiver side, the performed pulse compression consists of performing a DF_nT on the receive discrete-time domain OCDM symbol. In either OCDM-based ISAC system variants, the estimation of Doppler shifts, which can be translated into relative radial velocities, can be performed after the influence of transmit modulation symbols is suppressed during range processing. For that purpose, DFTs along consecutive range spectra are usually performed to generate a range-velocity radar image. Further details on the processing chains of ISAC systems based on the C-OCDM and SM-OCDM schemes are discussed in Sections 2.4.2 and 2.4.3, respectively.

2.4.2 C-OCDM-based ISAC processing chain

OCDM-based systems share many similarities with their OFDM counterparts. Some of their few differences are, e.g., data modulation onto orthogonal subchirps [ODGZ17] instead of subcarriers and the use of the DF_nT and its inverse instead of the DFT-IDFT pair [OZ16a]. Similarly to its OFDM counterpart from Section 2.3, the considered SISO C-OCDM-based ISAC system transmits a frame that comprises of $M^{\text{OCDM}} \in \mathbb{N}_{>0}$ blocks, each containing one OCDM symbol that carries $N_{\text{subch}}^{\text{OCDM}}$ QPSK-modulated subchirps and a CP. The OCDM frame carries data from a modulated data stream $\mathbf{d}_{\text{Tx}}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}} \times 1}$ containing a total of $N_{\text{subch}}^{\text{OCDM}} M^{\text{OCDM}}$ QPSK symbols, achieving the same data rate as an equally parameterized OFDM-based ISAC system.

Based on [47], the processing chain of the considered C-OCDM-based ISAC system is depicted in Figure 2.12. In this figure, the modulated data stream $\mathbf{d}_{\text{Tx}}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}} \times 1}$ first undergoes S/P conversion, generating the discrete-Fresnel domain OCDM frame $\hat{\mathbf{X}}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$. As opposed to the column-wise IDFTs in the OFDM case described in Section 2.3, the frame $\hat{\mathbf{X}}^{\text{OCDM}}$ undergoes column-wise IDF_nTs along the $N_{\text{subch}}^{\text{OCDM}}$ QPSK-modulated subchirps of each of its M^{OCDM} OCDM symbols. According to [OZ16a, Ouy17, 47, 49], the aforementioned IDF_nTs can be accomplished via element-wise pre-multiplication of every $N_{\text{subch}}^{\text{OCDM}}$ -length column of the OCDM frame by the complex conjugate of vector $\Theta_2 \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$, followed by column-wise IDFTs on the resulting frame and element-wise multiplication of each of its columns by the complex conjugate of vector $\Theta_1 \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$. Assuming even $N_{\text{subch}}^{\text{OCDM}}$, the k th element of Θ_2 is expressed as [OZ16a]

$$\Theta_{2,k} = e^{j\pi k^2 / N_{\text{subch}}^{\text{OCDM}}}, \quad (2.17)$$

while the n th element of Θ_1 is expressed as

$$\Theta_{1,n} = e^{-j\frac{\pi}{4}} e^{j\pi n^2 / N_{\text{subch}}^{\text{OCDM}}}. \quad (2.18)$$

The choice of distinct indexes k and n for $\Theta_{2,k}$ and $\Theta_{1,n}$ is due to the fact that they are multiplied by the OCDM symbols before and after the aforementioned IDFTs, respectively. Next, CPs of length $N_{\text{CP}}^{\text{OCDM}} \in \mathbb{N}_{>0}$ are prepended to each of the M^{OCDM} OCDM symbols of the output discrete-time domain OCDM

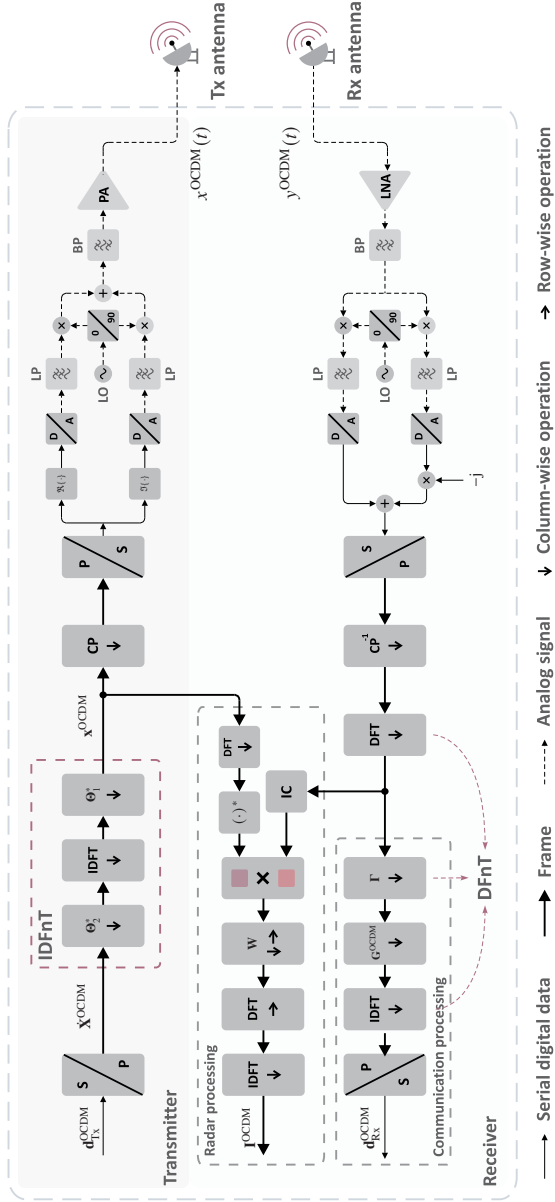


Figure 2.12: C-OCDM-based ISAC system representation. Adapted from [18], © 2022 IEEE.

frame $\mathbf{x}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ from the aforementioned IDFnT operation. For communications purposes, shorter CP lengths can be used in OCDM-based systems compared to their OFDM counterparts [OZ16a, BCF20, OM21], which results from the fact that ISI is spread over all subchirps and can be explained by the DFnT convolution theorem [OAG⁺15]. However, the use of CPs with shorter length than the radar CIR may bias radar measurements. The following operations are the same as the ones described in Section 2.3 for the OFDM case, and the resulting transmit analog signal $x^{\text{OCDM}}(t) \in \mathbb{R}$ also has a similar PAPR than its OFDM counterpart due to many similarities between the OCDM and OFDM schemes [OZ16a].

Based on the aforementioned processing steps and omitting the power factors for simplicity, the continuous-time domain signal $x_m^{\text{OCDM, w/oCP}}(t, \tau) \in \mathbb{R}$ carrying the m th OCDM symbol, $m \in \{0, 1, \dots, M^{\text{OCDM}} - 1\}$, without CP can be expressed as

$$\begin{aligned}
 x_m^{\text{OCDM, w/oCP}}(t, \tau) = & \Re \left\{ \mathcal{A}_{B^{\text{OCDM}}} \left\{ \frac{1}{N_{\text{subch}}^{\text{OCDM}}} e^{j\pi/4} \sum_{k=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{k,m}^{\text{OCDM}} e^{-j\pi \frac{N_{\text{subch}}^{\text{OCDM}}}{T_{\text{symb}}^{\text{OCDM}}{}^2} \left[k \frac{T_{\text{symb}}^{\text{OCDM}}}{N_{\text{subch}}^{\text{OCDM}}} - t \right]^2} \right\} \right\} \\
 & \times \cos(2\pi f_c \tau) \\
 & + \Im \left\{ \mathcal{A}_{B^{\text{OCDM}}} \left\{ \frac{1}{N_{\text{subch}}^{\text{OCDM}}} e^{j\pi/4} \sum_{k=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{k,m}^{\text{OCDM}} e^{-j\pi \frac{N_{\text{subch}}^{\text{OCDM}}}{T_{\text{symb}}^{\text{OCDM}}{}^2} \left[k \frac{T_{\text{symb}}^{\text{OCDM}}}{N_{\text{subch}}^{\text{OCDM}}} - t \right]^2} \right\} \right\} \\
 & \times \sin(2\pi f_c \tau), \\
 t \in [0, T_{\text{symb}}^{\text{OCDM}}], \tau \in [mT_{\text{block}}^{\text{OCDM}}, (m+1)T_{\text{block}}^{\text{OCDM}}]. \tag{2.19}
 \end{aligned}$$

In this equation, $\Re\{\cdot\}$ and $\Im\{\cdot\}$ denote the real and imaginary parts of a complex number. The operator $\mathcal{A}_{B^{\text{OCDM}}} \{\cdot\}$ aliases back high-frequency components of a signal into the $[-B^{\text{OCDM}}/2, B^{\text{OCDM}}/2]$ frequency range. $\dot{X}_{k,m}^{\text{OCDM}} \in \mathbb{C} | \dot{X}_{k,m}^{\text{OCDM}} = e^{j\varphi_{k,m}^{\text{OCDM}}}$ is the QPSK symbol with phase $\varphi_{k,m}^{\text{OCDM}} \in \{\pi/4, 3\pi/4, 5\pi/4, 7\pi/4\}$ that is modulated onto the k th subchirp of the m th OCDM symbol. Additionally, $T_{\text{symb}}^{\text{OCDM}} = N_{\text{subch}}^{\text{OCDM}}/B^{\text{OCDM}}$ is time

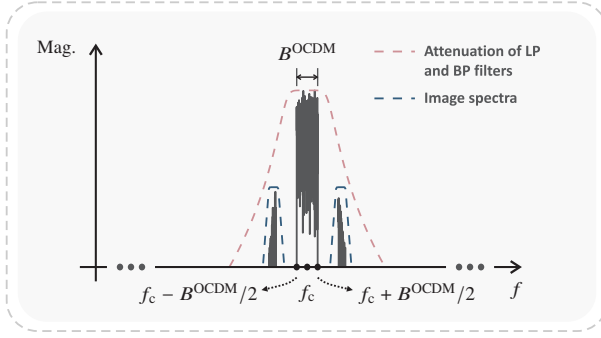


Figure 2.13: Evaluated spectrum for one OCDM symbol. © 2022 IEEE. Reprinted, with permission, from [18].

duration of an OCDM symbol disregarding CP, which depends on the RF bandwidth B^{OCDM} occupied by the OCDM signal. Unlike the subcarriers in the OFDM case from Section 2.3, all subchirps occupy the whole discrete-frequency domain spectrum of the OCDM signal. However, interpolation of the transmit OCDM signal can still be achieved via frequency-domain zero padding (FDZP) as described in [OM20b], which consists of appending empty tones to the edges of the OCDM symbol in the discrete-frequency domain. This ultimately results in an RF spectrum for a single OCDM symbol as depicted in Figure 2.13, where $F_{s,\text{DAC}}^{\text{OCDM}}$ is the sampling frequency adopted at the DACs. The continuous-time domain signal $x_m^{\text{OCDM}}(t) \in \mathbb{R}$ carrying the m th OCDM symbol with $N_{\text{CP}}^{\text{OCDM}}$ -length CP can therefore be expressed as

$$x_m^{\text{OCDM}}(t) = \begin{cases} x_m^{\text{OCDM,w/oCP}} \left(t + T_{\text{symp}}^{\text{OCDM}} - T_{\text{CP}}^{\text{OCDM}} - mT_{\text{block}}^{\text{OCDM}}, t \right), \\ t \in [mT_{\text{block}}^{\text{OCDM}}, mT_{\text{block}}^{\text{OCDM}} + T_{\text{CP}}^{\text{OCDM}}) \\ x_m^{\text{OCDM,w/oCP}} (t - T_{\text{CP}}^{\text{OCDM}} - mT_{\text{block}}^{\text{OCDM}}, t), \\ t \in [mT_{\text{block}}^{\text{OCDM}} + T_{\text{CP}}^{\text{OCDM}}, (m+1)T_{\text{block}}^{\text{OCDM}}) \end{cases} \quad (2.20)$$

In this equation, $T_{\text{block}}^{\text{OCDM}} = T_{\text{symp}}^{\text{OCDM}} + T_{\text{CP}}^{\text{OCDM}}$ is the duration of each OCDM block comprised of an OCDM symbol and a CP of time duration $T_{\text{CP}}^{\text{OCDM}} = N_{\text{CP}}^{\text{OCDM}} / B^{\text{OCDM}}$. Consequently, the continuous-time domain

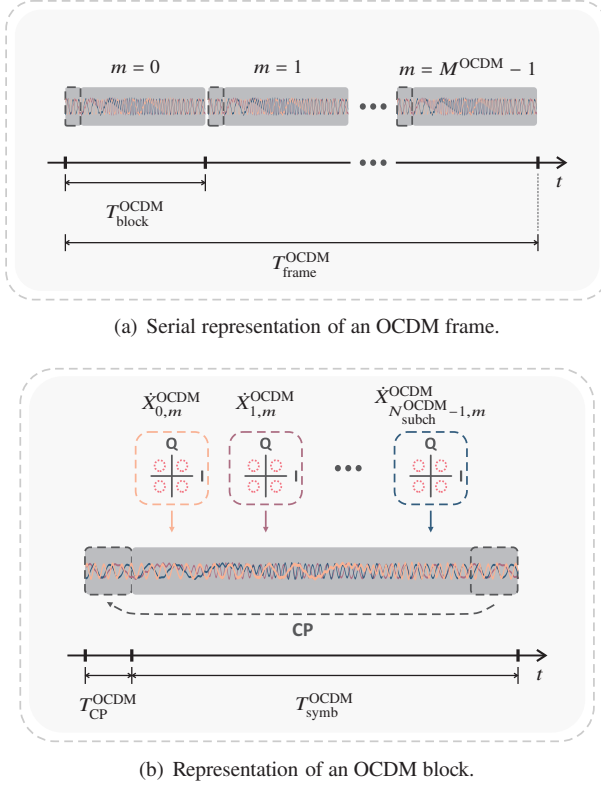


Figure 2.14: OCDM transmit signal structure. Adapted from [18], © 2022 IEEE.

transmit signal $x^{\text{OCDM}}(t) \in \mathbb{R}$ carrying the frame comprised of M^{OCDM} blocks that is transmitted by the C-OCDM-based ISAC system can be expressed as

$$x^{\text{OCDM}}(t) = \sum_{m=0}^{M^{\text{OCDM}}-1} x_m^{\text{OCDM}}(t) \text{rect}\left(\frac{t - mT^{\text{OCDM}}_{\text{block}}}{T^{\text{OCDM}}_{\text{block}}}\right). \quad (2.21)$$

Based on the presented concepts, Figure 2.14(a) depicts a serial representation of the frame in the considered C-OCDM-based ISAC system, while Figure 2.14(b) shows the detailed structure of a single block.

At the receiver side of the C-OCDM-based ISAC system, the receive antenna captures the continuous-time domain signal $y^{\text{OCDM}}(t) \in \mathbb{R}$, which comprises of the output version of $x^{\text{OCDM}}(t)$ by the radar channel, the output version of the transmit signal $z^{\text{OCDM}}(t) \in \mathbb{R}$ of another ISAC system by the communications channel, and additive noise according to (2.1). The initial processing steps at the receiver side are similar to the ones from the OFDM case described in Section 2.3, i.e., analog conditioning, A/D conversion with sampling rate $F_s^{\text{OCDM}} = B^{\text{OCDM}}$ and corresponding sampling period $T_s^{\text{OCDM}} = 1/F_s^{\text{OCDM}}$ is performed for both I and Q channels. After S/P conversion on the sampled complex-valued vector, CP removal, and DFTs are performed along the M^{OCDM} columns of the output frame. The resulting discrete-frequency domain receive frame of size $N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}$ finally undergoes distinct processing steps for radar sensing and communications purposes. Both these steps as well as the application of multiplexing strategies to the considered C-OCDM-based ISAC system to enable MIMO or multiuser operation are discussed as follows.

2.4.2.1 Radar processing

As in the previously presented modulation schemes, the radar processing in the considered C-OCDM-based ISAC system starts with IC. Next, the resulting frame undergoes element-wise multiplication by the complex conjugate of the output frame from column-wise DFTs on $\mathbf{x}^{\text{OCDM}}$. Similarly to the radar processing in the PMCW case described in Section 2.2, the aforementioned operation yields a matched filtering in the discrete-frequency domain, which, after further processing, will ultimately generate radar range profiles whose sidelobes will be defined by the transmit QPSK symbols in the considered C-OCDM-based ISAC system. Compared to [47], the use of simple DFTs ultimately yields a PCCF between the transmit and receive OCDM symbols, which has less associated computational complexity than the use of zero-padded DFTs in [47] used for performing a linear cross-correlation and introduces no degradation of the range sidelobe level. Furthermore, the reason why a simple element-wise division in the discrete-frequency domain is not performed as in the OFDM case from Section 2.3 is that OCDM symbols do not present flat spectral content as their OFDM counterparts and can even have null frequency tones, which prevents estimation of radar CFR [47]. In fact, although alternative radar processing strategies that explore properties of the DF_{FT} could be used, no such approaches have been reported in the literature so far.

The remaining processing steps after matched-filtering in the discrete-frequency domain are the same from the PMCW and OFDM case in Sections 2.2 and 2.3, namely range and velocity windowing, row-wise DFTs for Doppler shift estimation, and column-wise IDFTs for time delay estimation. After the estimated Doppler shifts and time delays are translated into relative radial velocity and range estimates, respectively, a range-velocity radar image $\mathbf{I}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ is ultimately obtained. According to [47, 49], the aforementioned radar image experiences a processing gain $G_p^{\text{C-OCDM}}$ and is associated with range resolution $\Delta R^{\text{C-OCDM}}$, maximum unambiguous range $R_{\text{max,ua}}^{\text{C-OCDM}}$, and maximum ISI-free range $R_{\text{max,ISI}}^{\text{C-OCDM}}$, as well as relative radial velocity resolution, $\Delta v^{\text{C-OCDM}}$ and maximum unambiguous relative radial velocity $v_{\text{max,ua}}^{\text{C-OCDM}}$ values as shown in Table 2.2 based on [47, 49]. Regarding a restriction on maximum tolerable relative radial velocity for the considered C-OCDM-based ISAC system, results from [18] showed that a similar effect to the ICI in the OFDM-based ISAC system from Section 2.3 is experienced in C-OCDM. Consequently, the $v_{\text{max,ICI}}^{\text{C-OCDM}}$ is defined as a maximum tolerable relative radial velocity based on the same criterion as for $v_{\text{max,ICI}}^{\text{OFDM}}$ and also listed in Table 2.2 in Section 2.5.

2.4.2.2 Communications processing

First, it is assumed that time, frequency and phase offsets from the second term in (2.1) are compensated via synchronization based on a preamble that can be defined by the Schmidl and Cox algorithm [SC97] as in the OFDM case from Section 2.3 [Ouy17]. A specific preamble design using OCDM symbols is presented in [7]. The remaining communications processing in the considered OCDM-based ISAC system starts with an element-wise multiplication of each block by the vector $\mathbf{\Gamma} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$ [OZ16a], whose k th element is expressed for even $N_{\text{subch}}^{\text{OCDM}}$ as

$$\Gamma_k = e^{-j\pi k^2 / N_{\text{subch}}^{\text{OCDM}}}. \quad (2.22)$$

Next, a fast-time, element-wise multiplication of every m th block by the corresponding vector $\mathbf{G}_m^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$ containing the multiplicative inverse of the estimate of the communications CFR $\mathbf{H}_m^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$ is performed. To obtain the aforementioned CFR estimates, pilots can be inserted via several multiplexing strategies, e.g., in the discrete-frequency domain as described in

detail in [OM22]. Column-wise IDFTs are then performed to complete the communications processing. Together with the previously performed column-wise DFTs before the specific radar and communications processing steps, the multiplication by $\mathbf{\Gamma} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$ and the IDFTs comprise a DFnT. The output discrete-Fresnel domain OCDM frame from the aforementioned processing steps undergoes P/S conversion, generating the modulated data stream $\mathbf{d}_{\text{Rx}}^{\text{OCDM}} \in \mathbb{C}^{(N_{\text{subch}}^{\text{OCDM}} M^{\text{OCDM}}) \times 1}$.

The presented communications processing steps in the considered C-OCDM-based ISAC system ultimately yield the overall data rate denoted by $\mathcal{R}_{\text{QPSK}}^{\text{C-OCDM}}$, as well as energy per bit to noise power spectral density ratio $(E_b/N_0)_{\text{QPSK},k,m}^{\text{C-OCDM}}$ and BER denoted by $\text{BER}_{\text{QPSK},k,m}^{\text{C-OCDM}}$ for the k th QPSK-modulated subchirp in the m th OCDM symbol as presented in Table 2.2 in Section 2.5, where $H_{k,m}^{\text{OCDM}} \in \mathbb{C}$ is the k th element of $\mathbf{H}_m^{\text{OCDM}}$.

2.4.2.3 Multiplexing for MIMO and multiuser scenarios

To enable MIMO or multiuser operation of N_{Tx} transmitters based on the transmission of orthogonal signals in OCDM-based ISAC systems, FDM and CDM approaches are usually adopted, being TDM disregarded for the same reasons as in the PMCW and OFDM cases from Subsections 2.2 and 2.3, respectively.

A typical FDM-based approach is the so-called frequency-shift precoding (FSP) [OM19, 49]. The use of FSP consists of allocating a set of $N_{\text{subch}}^{\text{OCDM}}/N_{\text{Tx}}$ interleaved frequency tones to each of the N_{Tx} transmitters via a precoding, which is undone at the receiver side by selecting the associated frequency tones with the evaluated transmitter. This results in similar changes in both radar and communications performances as in the OFDM case from Section 2.3, namely a reduction of the listed processing gain in Table 2.2 to $G_{\text{p}}^{\text{C-OCDM}} = (N_{\text{subch}}^{\text{OCDM}}/N_{\text{Tx}})M^{\text{OCDM}}$ and a reduction of the maximum unambiguous range to $R_{\text{max,ua}}^{\text{C-OCDM}} = (N_{\text{subch}}^{\text{OCDM}}/N_{\text{Tx}})c_0/(2B^{\text{OCDM}})$. To avoid isolation loss among transmitters due to a similar effect to ICI in OFDM systems, an additional constraint on the maximum tolerable relative radial velocity for OCDM-based ISAC systems adopting FSP-based FDM as a multiplexing strategy is defined as $v_{\text{max,FSP}}^{\text{C-OCDM}} = B^{\text{OCDM}}c_0/(20N_{\text{subch}}^{\text{OCDM}}f_c)$. As in the OFDM case,

the communications performance is only affected in the multiuser case, where the achieved data rate assuming QPSK modulation is reduced by a factor of N_{Tx} to $\mathcal{R}_{\text{QPSK}}^{\text{C-OCDM}} = 2N_{\text{subch}}^{\text{OCDM}}B^{\text{OCDM}}/[N_{\text{Tx}}(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}})]$.

A CDM-based approach for enabling MIMO or multiuser operation of OCDM-based ISAC systems is the use of uncorrelated OCDM symbols, which yields poor isolation among transmitters as described for the OFDM case Section 2.3 [NMZ18]. Alternatively, the same outer coding described for PMCW and OFDM in Subsections 2.2 and 2.3, respectively, can also be performed in the OCDM context [49], changing the required duration to transmit a block to $N_{\text{Tx}}T_{\text{block}}^{\text{OCDM}}$ and consequently the maximum unambiguous relative radial velocity to $v_{\text{max,ua}}^{\text{OCDM}} = B^{\text{OCDM}}c_0/[4f_cN_{\text{Tx}}(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}})]$. Additionally, the outer-coding based CDM changes the listed processing gain in Table 2.2 to $G_{\text{p}}^{\text{OCDM}} = N_{\text{Tx}}N_{\text{subch}}^{\text{OCDM}}M^{\text{PMCW}}$, although the effectively experienced $G_{\text{p}}^{\text{OCDM}}$ tends to be the same as in the SISO case due to the necessary reduction of M^{PMCW} to keep $T_{\text{frame}}^{\text{OCDM}}$ constant. Finally, it is worth highlighting that, as in the previously described FSP-based FDM approach, the only experienced change in the communications performance is the reduction of the data rate by a factor of N_{Tx} in the multiuser case.

2.4.3 SM-OCDM-based ISAC processing chain

If a DFnT is performed on the receive OCDM symbols in the discrete-time domain for radar processing instead of the matched-filtering approach described for C-OCDM in Section 2.4.2, the effects of time and frequency shifts in the discrete-Fresnel domain must be considered to describe the effects of the range and velocity of targets in the OCDM signal propagation. In this sense, a thorough description of both aforementioned effects on the originally transmitted discrete-Fresnel domain OCDM symbols represented by the columns of $\mathbf{X}^{\text{OCDM}}$, which result in the discrete-Fresnel domain frame $\mathbf{Y}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ after propagation through a channel, is presented in the Appendix A.1. Based on the achieved results therein, it is possible to predict the relationship between the channel estimates in OCDM-based systems and the actual channel CIR. Given the fact that this principle was first discussed in [20] and is not widely addressed in the literature, a more thorough mathematical formulation than presented for PMCW, OFDM and C-OCDM in the previous sections will be henceforth provided.

To enable radar processing based on the DF_nT, the discrete-Fresnel domain pilot design strategy introduced in [ODZ17] for coherent optical OFDM systems, which was also investigated for reflectometric sensing of power lines in baseband OCDM systems as reported in [GdPR21] and further improved in [OATT18], is considered. The original pilot design from [ODZ17] performs unbiased channel estimation under the assumption that no frequency shifts take place. This is done by exploiting the convolution theorem of the DF_nT [OAG⁺15] and having a Kronecker comb with $S \in \mathbb{N}_{>0}$ active subchirps in the discrete-Fresnel domain at the transmitter side, and averaging the resulting S sections of length $N_{\text{subch}}^{\text{OCDM}}/S$ in the discrete-Fresnel domain at the receiver side to estimate the CIR. Conversely, the improvement proposed in [OATT18] achieves optimal CIR estimates in the minimum mean square error (MMSE) sense by activating only the first subchirp in the discrete-Fresnel domain at the transmitter side, and discarding noisy samples beyond the maximum expected CIR length in the obtained discrete-Fresnel domain receive OCDM symbol to obtain a CIR estimate. Since dividing the transmission power among S active subchirps and averaging their corresponding CIR estimates yields the same SNR as in the case where the full transmission power is allocated to a single subchirp and only one CIR is obtained, the superiority of the improved pilot design approach is solely due to the noise-rejection windowing performed in the discrete-Fresnel domain at the receiver side.

Considering the characteristics of the discussed pilot design strategies, namely the one from [ODZ17] and the one from [OATT18], the latter one is henceforth adopted. Consequently, the element at the k th row and m th column of the matrix $\dot{\mathbf{X}}^{\text{OCDM}}$ that represents a frame containing M^{OCDM} discrete-Fresnel domain pilot OCDM symbols in its columns is given by

$$\dot{X}_{k,m}^{\text{OCDM}} = \delta[k], \quad (2.23)$$

where $\delta[k]$ is the Kronecker delta function. As indicated by the discussions in [FMM20] and [OZ16a], this results in the fact that the OCDM symbols in the discrete-time domain are equal to a weighted complex-conjugate root Zadoff-Chu root sequence. As all M^{OCDM} OCDM symbols within the frame are equal, no CP is required. The implications of the pilot OCDM symbol design on the performance of radar channel estimation, as well as extensions of the discussed concept to multi-user (MU) or MIMO scenarios as well as to ISAC applications with the previously mentioned SM-OCDM scheme are discussed as follows.

2.4.3.1 Discrete-Fresnel domain radar processing

For the sake of simplicity, it is assumed in this subsection that the OCDM-based system acts as a monostatic radar, in which transmit and receive antennas are virtually collocated. In practice, however, only a quasi-monostatic radar is possible, in which the distance between transmit and receive antennas is negligible in comparison to the range of expected target distances with respect to the radar. To estimate the range and relative radial velocity of multiple point targets with respect to the radar, consecutive radar CIRs are estimated. Each CIR is composed of multiple taps at delays corresponding to the time of flight (ToF) taken by the OCDM signal to be sent out by the transmit antenna, reflected off targets, and captured by the receive antenna. If no effects such as range migration occur within the measurement time, all of the estimated radar CIRs allow estimating the range of resolved point targets present in the monitored scenario. The observed difference among consecutive CIRs, which is the phase progression of its taps, ultimately allows obtaining a radar image that has not only range information, but also estimates of the experienced Doppler shifts related to the relative radial velocities of the point targets. To enable the estimation of range and Doppler shifts of H point targets, the matrix $\tilde{\mathbf{y}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ that would ideally represent the discrete-time domain receive OCDM frame has elements at its n th row and m th column expressed as

$$\begin{aligned} \tilde{y}_{n,m}^{\text{OCDM,rad}} = & \sum_{h=0}^{H-1} \left[\sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} x_{\nu,m}^{\text{OCDM}} \left(\frac{e^{j2\pi(n_{\Delta,h}-n+\nu)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(n_{\Delta,h}-n+\nu)} - 1} \right) \right] \\ & \times e^{j2\pi k_{\Delta,h} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_{m,h}}. \end{aligned} \quad (2.24)$$

In this equation, $x_{\nu,m}^{\text{OCDM}} \in \mathbb{C}$ is the element at the ν th row, with $\nu \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} - 1\}$, and m th column of the discrete-time domain OCDM frame $\mathbf{x}^{\text{OCDM}}$, $n_{\Delta,h} \in \mathbb{R}_{\geq 0} | n_{\Delta,h} = 2R_h B^{\text{OCDM}} / c_0$ is a discrete normalized range term, and R_h is the range of the h th target as defined in Section 2.1. Additionally, the fraction containing complex exponential terms is to account for normalized range terms $n_{\Delta,h}$ that yield fractional delays in the

discrete-time domain and can be derived based on the property from (A.11). The terms $k_{\Delta,h} \in \mathbb{R} | k_{\Delta,h} = f_{D,\text{target},h}/\Delta f^{\text{OCDM}}$, where

$$\Delta f^{\text{OCDM}} \triangleq B^{\text{OCDM}}/N_{\text{subch}}^{\text{OCDM}}, \quad (2.25)$$

and $\phi_{m,h} \in \mathbb{R} | \phi_{m,h} = 2\pi k_{\Delta,h} \left[m \left(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}} \right) + N_{\text{CP}}^{\text{OCDM}} \right] / N_{\text{subch}}^{\text{OCDM}}$ are in turn the discrete normalized Doppler shift and the Doppler phase associated to the h th point target. The actual Doppler shift $f_{D,\text{target},h} = -2v_h/\lambda_0$, is caused by the fact that the h th point target moves with relative radial velocity v_h with respect to the radar as defined in Section 2.1. In the specific case where $n_{\Delta,h} \in \mathbb{Z}$, (2.24) can be rewritten as

$$\tilde{y}_{n,m}^{\text{OCDM,rad}} = \sum_{h=0}^{H-1} x_{n,m}^{\text{OCDM}} \delta[n - n_{\Delta,h}] e^{j2\pi k_{\Delta,h} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_{m,h}}. \quad (2.26)$$

With the OCDM pilot symbols contained in the columns of $\mathbf{\tilde{X}}^{\text{OCDM}}$ having their k th elements defined as in (2.23), the resulting ideal discrete-Fresnel domain receive OCDM frame matrix representation $\tilde{\mathbf{Y}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ can be alternatively defined as the matrix $\tilde{\mathbf{h}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ that contains M^{OCDM} consecutive radar CIR estimates of length $N_{\text{subch}}^{\text{OCDM}}$. The element at the n th row and m th column of $\tilde{\mathbf{h}}^{\text{OCDM,rad}}$ can be obtained by substituting (2.23) into (2.24) to yield

$$\tilde{h}_{n,m}^{\text{OCDM,rad}} = \sum_{h=0}^{H-1} \left(\frac{e^{j2\pi(n_{\Delta,h}-n)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(n_{\Delta,h}-n)} - 1} \right) e^{j2\pi k_{\Delta,h} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_{m,h}} \quad (2.27)$$

or, for $n_{\Delta,h} \in \mathbb{Z}$, by substituting (2.23) into (2.26) to yield

$$\tilde{h}_{n,m}^{\text{OCDM,rad}} = \sum_{h=0}^{H-1} \delta[n - n_{\Delta,h}] e^{j2\pi k_{\Delta,h} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_{m,h}}. \quad (2.28)$$

In reality, however, a matrix $\mathbf{\tilde{Y}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ $|\mathbf{\tilde{Y}}^{\text{OCDM,rad}} \neq \tilde{\mathbf{Y}}^{\text{OCDM,rad}}$ is obtained at the receiver side of the OCDM-based system. Since pilot OCDM symbols are transmitted, $\mathbf{\tilde{Y}}^{\text{OCDM,rad}}$ can be alternatively defined as the matrix

$\hat{\mathbf{h}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}} | \hat{\mathbf{h}}^{\text{OCDM,rad}} \neq \tilde{\mathbf{h}}^{\text{OCDM,rad}}$ that contains the actual consecutive radar CIR estimates in its columns. Assuming a single target scenario and considering only the m th of the M^{OCDM} radar channel estimates, a corresponding continuous-frequency domain CFR $\hat{H}_m^{\text{OCDM,rad}}(f) \in \mathbb{C}$ of the aforementioned channel presents a linear phase progression along with the frequency in the RF band $f \in [f_c - B^{\text{OCDM}}/2, f_c + B^{\text{OCDM}}/2]$ as represented at the top of Figure 2.15(a), which is caused by the continuous-time delay associated with the target's range. In the discrete-frequency domain, the CIR contained in the m th column of $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ is represented by the vector $\hat{\mathbf{H}}_m^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1}$. Since the samples $k = 0$ to $k = N_{\text{subch}}^{\text{OCDM}}/2 - 1$ of $\hat{\mathbf{H}}_m^{\text{OCDM,rad}}$ can be used to reconstruct spectral information on the RF frequencies f_c to $f_c + (N_{\text{subch}}^{\text{OCDM}}/2 - 1)\Delta f^{\text{OCDM}} = f_c + B^{\text{OCDM}}/2 - \Delta f^{\text{OCDM}}$, while the samples $k = N_{\text{subch}}^{\text{OCDM}}/2 + 1$ to $k = N_{\text{subch}}^{\text{OCDM}}$ allow reconstructing spectral information on the frequencies $f_c + (-N_{\text{subch}}^{\text{OCDM}}/2)\Delta f^{\text{OCDM}} = f_c - B^{\text{OCDM}}/2$ to $f_c - \Delta f^{\text{OCDM}}$, the phase of $\hat{\mathbf{H}}_m^{\text{OCDM,rad}}$ is circularly shifted by $N_{\text{subch}}^{\text{OCDM}}/2$ samples with respect to an ideally linearly progressing phase [Nus21]. Considering a single-target scenario, an exemplary phase progression in the BB of $\hat{\mathbf{H}}_m^{\text{OCDM,rad}}$ can be seen in Figure 2.15(a).

In order for the ToFs, and consequently target ranges, to be appropriately estimated, the effectively obtained matrix that represents the discrete-time domain receive OCDM frame and is denoted by $\mathbf{y}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}} | \mathbf{y}^{\text{OCDM,rad}} \neq \tilde{\mathbf{y}}^{\text{OCDM,rad}}$ must consist of circularly-shifted versions of $\mathbf{x}^{\text{OCDM}}$. If the experienced spectrum shift is not compensated, the columns of the estimated radar CIR matrix $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ are equal to the corresponding columns of the ideally obtained matrix $\tilde{\mathbf{h}}^{\text{OCDM,rad}}$ with the addition of discrete-frequency domain shift by $N_{\text{subch}}^{\text{OCDM}}/2$ samples. In other words

$$\hat{h}_{n,m}^{\text{OCDM,rad}} = \tilde{h}_{n,m}^{\text{OCDM,rad}} e^{-j2\pi(N_{\text{subch}}^{\text{OCDM}}/2)n/N_{\text{subch}}^{\text{OCDM}}} = \tilde{h}_{n,m}^{\text{OCDM,rad}} e^{-j\pi n}, \quad (2.29)$$

where $\hat{h}_{n,m}^{\text{OCDM,rad}} \in \mathbb{C}$ denotes the element of $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ at its n th row and m th column. Although the amplitudes of the elements of $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ are the same as the corresponding ones of $\tilde{\mathbf{h}}^{\text{OCDM,rad}}$, the introduced phase rotations would ultimately prevent correctly estimating the ranges of targets when reconstructing the analog equivalent of $\hat{\mathbf{h}}^{\text{OCDM,rad}}$, e.g., via zero padding (ZP). Consequently,

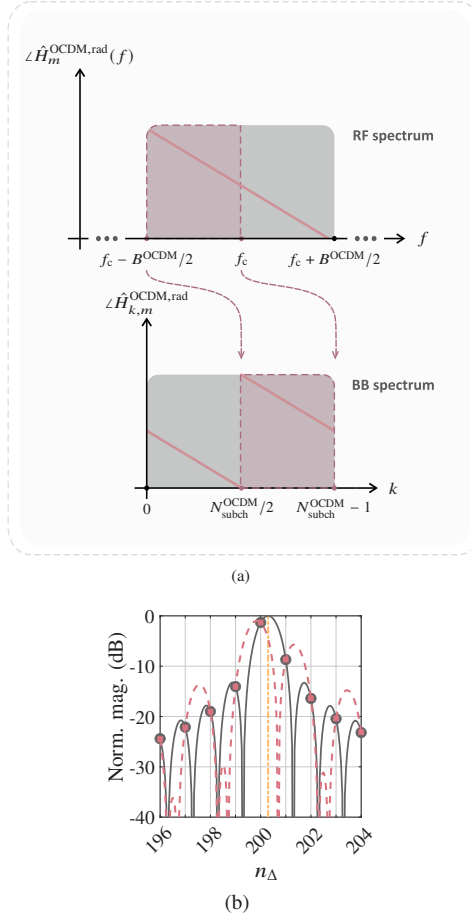


Figure 2.15: Spectrum shift effect. (a) CFR phase progression due to propagation time delay in RF and BB spectra [Nus21]. (b) Range profile obtained for $n_\Delta = 200.3$ (—). The same normalized magnitude at integer n_Δ values (i.e., multiples of the range resolution) are observed with (●) and without (■) spectrum shift correction. However, the sinc-interpolated range profile via FDZP without correction (---) deviates from the expected shape and true target range (—), which is only attained by its corrected counterpart (—). Adapted from [20], © 2023 IEEE.

the experienced spectrum shift must be corrected by performing a circular shift in the discrete-frequency domain by $N_{\text{subch}}^{\text{OCDM}}/2$ on the columns of $\hat{\mathbf{Y}}^{\text{OCDM,rad}}$ to produce $\hat{\mathbf{Y}}^{\text{OCDM,rad,corr}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$, which can be alternatively defined as the matrix $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ that contains an estimate of the desired radar CIR matrix $\hat{\mathbf{h}}^{\text{OCDM,rad}}$. The necessary spectrum shift compensation can be achieved in two manners. The first one is by multiplying every k th element of the columns of $\hat{\mathbf{Y}}^{\text{OCDM,rad}}$ by $e^{j\pi k}$. A second solution is based on the OCDM receiver structure #2 from [OZ16a], where the DFNT is calculated by means of a DFT, followed by an element-wise multiplication by the vector $\mathbf{\Gamma}$ whose elements are defined in (2.22) and an IDFT. In this approach, a circular shift in the discrete-frequency domain by $N_{\text{subch}}^{\text{OCDM}}/2$ samples is performed right after the element-wise multiplication by $\mathbf{\Gamma}$. Taking advantage of the fact that samples are in the discrete-frequency domain at this point, windowing for reduction of range and velocity sidelobes can be performed by element-wise multiplying the frame between the aforementioned multiplication by $\mathbf{\Gamma}$ and the IDFT that finalizes the DFNT implementation by a windowing matrix $\mathbf{W} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$. For simplicity, it is assumed that rectangular windowing is performed in both directions, which results in a matrix $\mathbf{W}$ whose all elements are equal to one and therefore allows omitting it from the following discussion. In addition, this DFNT implementation approach allows zero performing FDZP before its last IDFT in order to produce sinc-interpolated range profiles and enhance the accuracy of the estimated ranges by the OCDM-based radar system. If no FDZP is performed, the described spectrum shift does not alter the magnitude of the obtained range profiles as shown in Figure 2.15(b) and its correction would not be mandatory. However, if FDZP is used for improved accuracy in range estimation, the experienced spectrum shift will degrade the sinc-interpolation of the range profile samples if not compensated for, as in the example in Figure 2.15(b).

Assuming that no Doppler shifts take place, i.e., $k_{\Delta,h} = 0$ and $\phi_{m,h} = 0 \forall m, h$, both described spectrum shift correction approaches yield the CIR matrix estimate $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$, whose element at its n th row and m th column is given by

$$\hat{h}_{n,m}^{\text{OCDM,rad,corr}} = e^{j\pi n} \hat{h}_{n,m}^{\text{OCDM,rad}} = \tilde{h}_{n,m}^{\text{OCDM,rad}}. \quad (2.30)$$

If, however, a frequency shift also takes place, the discrete-time domain CIR matrix representation without spectrum shift correction $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ has the ele-

ment at its n th row and m th column expressed as in (2.31). For the specific case where $n_{\Delta,h} \in \mathbb{Z}$ and $k_{\Delta,h} \in \mathbb{Z} \forall h$, (2.31) becomes (2.32). After performing spectrum shift correction, $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}}$ is obtained such that the element in its n th row and m th column is expressed as in (2.33) or, for $n_{\Delta,h} \in \mathbb{Z}$ and $k_{\Delta,h} \in \mathbb{Z} \forall h$, as in (2.34).

The obtained expressions in (2.33) and (2.34) reveal that the delay-Doppler coupling mentioned in Section A.1.3 of the Appendix still takes place after spectrum shift correction. Additionally, it is observed for the h th target contribution to the aforementioned equations that the columns of $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}}$ are circularly-shifted versions of the columns of $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ by $k_{\Delta,h}$ samples. Compared to their counterparts in $\hat{\mathbf{h}}^{\text{OCDM,rad}}$, each n th element of these columns is also multiplied by complex exponentials with linearly increasing phase along with the index n .

Since each of the M^{OCDM} columns of $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}}$ contains information on the range of the H targets, performing row-wise DFTs on $\hat{\mathbf{h}}^{\text{OCDM,rad,corr}}$ converts the phases $\phi_{m,h}$ into Doppler shift information. This processing produces the matrix $\mathbf{I}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$, whose element $I_{n,m}^{\text{OCDM,rad}} \in \mathbb{C}$ at its n th row and m th column is given by

$$I_{n,m}^{\text{OCDM,rad}} = \sum_{m=0}^{M^{\text{OCDM}}-1} \hat{h}_{n,m}^{\text{OCDM,rad,corr}}. \quad (2.35)$$

From $\mathbf{I}^{\text{OCDM,rad}}$, one can clearly extract information on the range and relative radial velocity of all H targets if they are resolved in at least one of the radar image dimensions. For clearer visualization, the radar processing chain described in this subsection is depicted in Figure 2.16. Based on the analyses in [NMZ18, 18, 47, 49], the obtained radar image experiences a processing gain G_p^{OCDM} and is associated with range resolution ΔR^{OCDM} and maximum unambiguous range $R_{\text{max,ua}}^{\text{OCDM}}$, as well as relative radial velocity resolution Δv^{OCDM} and maximum unambiguous relative radial velocity $v_{\text{max,ua}}^{\text{OCDM}}$ values as shown in Table 2.1. Since all transmit OCDM symbols within the frame are equal and therefore no CP is required, no further restriction on the maximum range value such as in [18, 47], and [49] is considered. Furthermore, since a restriction on the maximum tolerable relative radial velocity for the considered OCDM-based

$$\begin{aligned}
\hat{h}_{n,m}^{\text{OCDM,rad}} &= \sum_{h=0}^{H-1} \frac{e^{j\phi_{m,h}}}{N_{\text{OCDM}}^{\text{subch}}} \sum_{\kappa=0}^{N_{\text{OCDM}}^{\text{subch}}-1} \left[\sum_{\nu=0}^{N_{\text{OCDM}}^{\text{subch}}-1} \left(e^{j2\pi(n_{\Delta,h}-\nu)} \frac{-1}{-1} \right) e^{-j\pi\nu} \delta \left[\langle \kappa - \nu \rangle_{N_{\text{OCDM}}^{\text{subch}}} \right] e^{j\frac{\pi}{N_{\text{OCDM}}^{\text{subch}}}(n^2 - \kappa^2)} \right] \\
&\quad \times \left(\frac{e^{j2\pi(k_{\Delta,h}-n+\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{OCDM}}^{\text{subch}}}(k_{\Delta,h}-n+\kappa)} - 1} \right) \\
&= \sum_{h=0}^{H-1} \frac{e^{j\phi_{m,h}}}{N_{\text{OCDM}}^{\text{subch}}} \sum_{\kappa=0}^{N_{\text{OCDM}}^{\text{subch}}-1} \left[\left(\frac{e^{j2\pi(n_{\Delta,h}-\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{OCDM}}^{\text{subch}}}(n_{\Delta,h}-\kappa)} - 1} \right) e^{-j\pi\kappa} e^{j\frac{\pi}{N_{\text{OCDM}}^{\text{subch}}}(n^2 - \kappa^2)} \right] \left(\frac{e^{j2\pi(k_{\Delta,h}-n+\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{OCDM}}^{\text{subch}}}(k_{\Delta,h}-n+\kappa)} - 1} \right) \quad (2.31)
\end{aligned}$$

$$\begin{aligned}
\hat{h}_{n,m}^{\text{OCDM,rad}} &= \sum_{h=0}^{H-1} e^{j\phi_{m,h}} \delta \left[\langle n - (n_{\Delta,h} + k_{\Delta,h}) \rangle_{N_{\text{OCDM}}^{\text{subch}}} \right] e^{-j\pi[n - (n_{\Delta,h} + k_{\Delta,h})]} e^{j\frac{\pi}{N_{\text{OCDM}}^{\text{subch}}}(n^2 - (n - k_{\Delta,h})^2)} \\
&= \sum_{h=0}^{H-1} e^{j\phi_{m,h}} \delta \left[\langle n - (n_{\Delta,h} + k_{\Delta,h}) \rangle_{N_{\text{OCDM}}^{\text{subch}}} \right] e^{-j\pi[n - (n_{\Delta,h} + k_{\Delta,h})]} e^{j\frac{\pi}{N_{\text{OCDM}}^{\text{subch}}}(2nk_{\Delta,h} - k_{\Delta,h}^2)} \quad (2.32)
\end{aligned}$$

$$\begin{aligned}
 \hat{h}_{n,m}^{\text{OCDM,rad,corr}} &= e^{j\pi n} \hat{h}_{n,m}^{\text{OCDM,rad}} = e^{j\pi n} \left\{ \sum_{h=0}^{H-1} \frac{e^{j\phi_{n,h}}}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \left[\left(\frac{e^{j2\pi(n_{\Delta,h}-\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(n_{\Delta,h}-\kappa)} - 1} \right) e^{-j\pi\kappa} \right] e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(n^2-\kappa^2)} \right\} \\
 &\quad \times \left(\frac{e^{j2\pi(k_{\Delta,h}-n+\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(k_{\Delta,h}-n+\kappa)} - 1} \right) \quad (2.33)
 \end{aligned}$$

$$\begin{aligned}
 \hat{h}_{n,m}^{\text{OCDM,rad,corr}} &= e^{j\pi n} \hat{h}_{n,m}^{\text{OCDM,rad}} = \sum_{h=0}^{H-1} e^{j\phi_{n,h}} \delta \left[\langle n - (n_{\Delta,h} + k_{\Delta,h}) \rangle N_{\text{subch}}^{\text{OCDM}} \right] e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} \left[2nk_{\Delta,h} - k_{\Delta,h}^2 + N_{\text{subch}}^{\text{OCDM}}(n_{\Delta,h} + k_{\Delta,h}) \right]} \quad (2.34)
 \end{aligned}$$

Table 2.1: Performance parameters in an OCDM-based radar system with discrete-Fresnel domain channel estimation. Adapted from [20], © 2023 IEEE.

Radar performance parameters	
Processing gain	$G_p^{\text{OCDM}} = N_{\text{subch}}^{\text{OCDM}} M^{\text{OCDM}}$
Range resolution	$\Delta R^{\text{OCDM}} = c_0 / (2B^{\text{OCDM}})$
Max. unamb. range	$R_{\text{max,ua}}^{\text{OCDM}} = N_{\text{subch}}^{\text{OCDM}} c_0 / (2B^{\text{OCDM}})$
Velocity resolution	$\Delta v^{\text{OCDM}} = B c_0 / \left 2f_c \left(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}} \right) M^{\text{OCDM}} \right $
Max. unamb. velocity	$v_{\text{max,ua}}^{\text{OCDM}} = B^{\text{OCDM}} c_0 / \left 4f_c \left(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}} \right) \right $

system depends on the tolerable delay-Doppler coupling as later investigated in Section 2.5.2, it is not listed in Table 2.1.

2.4.3.2 Multiplexing for MIMO and multiuser scenarios

If N_{Tx} synchronized OCDM transmitters labeled as $n_{\text{Tx}} \in \{0, 1, \dots, N_{\text{Tx}} - 1\}$ operate simultaneously in an MU or MIMO scenario, then the elements of the matrix $\dot{\mathbf{X}}^{\text{OCDM}, n_{\text{Tx}}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ that represents the discrete-Fresnel domain transmit frame of the n_{Tx} th transmitter are given by

$$\dot{X}_{k,m}^{\text{OCDM}, n_{\text{Tx}}} = \delta \left[\left\langle k - n_{\text{Tx}} N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}} \right\rangle_{N_{\text{subch}}^{\text{OCDM}}} \right]. \quad (2.36)$$

This equation indicates that the Fresnel-division multiplexing (FrDM) scheme [47], which was originally proposed in [GdPR21] for distributed reflectometric sensing of power lines, is adopted to enable orthogonal transmission of signals associated with each of the N_{Tx} transmitters as depicted in Figure 2.17.

Assuming that there are $N_{\text{Rx}} \in \mathbb{N}_{>0}$ synchronized OCDM radar receivers labeled as $n_{\text{Rx}} \in \{0, 1, \dots, N_{\text{Rx}} - 1\}$ and based on the formulation for the SISO

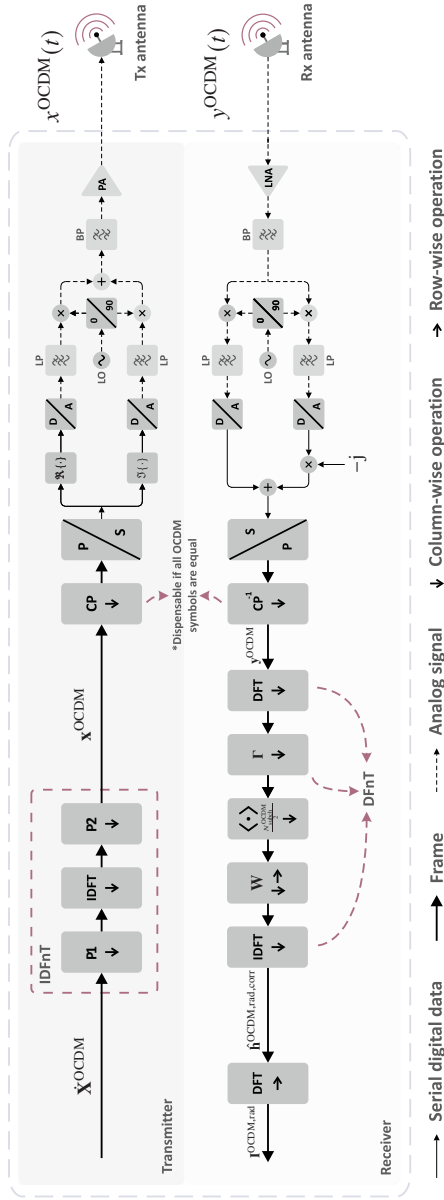


Figure 2.16: OCDM-based discrete-Fresnel domain radar processing chain. Adapted from [20], © 2023 IEEE.

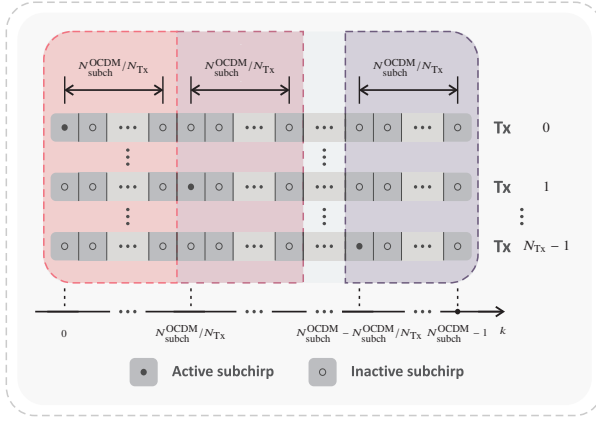


Figure 2.17: Subchirp allocation in the discrete-Fresnel domain among distinct transmit channels for MU or MIMO operation based on FrDM. Adapted from [20], © 2023 IEEE.

case, the elements of the matrix $\dot{\mathbf{Y}}^{\text{OCDM}, n_{\text{Rx}}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ representing the discrete-Fresnel domain receive frame of the n_{Rx} th receiver are given by

$$\dot{Y}_{k,m}^{\text{OCDM}, n_{\text{Rx}}} = \sum_{n_{\text{Tx}}=0}^{N_{\text{Tx}}-1} \hat{h}_{\langle k - n_{\text{Tx}} N_{\text{subch}}^{\text{OCDM}} / n_{\text{Tx}} \rangle_{N_{\text{subch}}^{\text{OCDM}}, m}}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}}, \quad (2.37)$$

where $\hat{h}_{k,m}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}} \in \mathbb{C}$ is the element at the k th row and m th column of the matrix $\hat{\mathbf{h}}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ that represents consecutive CIR estimates associated with the n_{Tx} th transmitter and n_{Rx} th receiver. If the CIRs in $\hat{\mathbf{h}}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}}$ are assumed to have length equal to or smaller than $N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}}$, i.e., $\hat{\mathbf{h}}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}}$ can be redefined as $\hat{\mathbf{h}}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}} \times M^{\text{OCDM}}}$, then its elements $\hat{h}_{k',m}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}} \in \mathbb{C}$, $k' \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}} - 1\}$, can be extracted from $\dot{\mathbf{Y}}^q$ as

$$\hat{h}_{k',m}^{\text{OCDM}, \text{rad}, \text{corr}, n_{\text{Tx}}, n_{\text{Rx}}} = \dot{Y}_{k' + n_{\text{Tx}} N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}}, m}^{\text{OCDM}, n_{\text{Rx}}}. \quad (2.38)$$

With the described FrDM approach for enabling MU or MIMO radar operation, most performance parameters from Table 2.1 are kept as all OCDM symbols in the transmit frame are equal and CPs are not needed.

The only exception is the maximum unambiguous range, which becomes $R_{\max, \text{ua}, \text{FrDM}}^{\text{OCDM}} = \left(N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}} \right) c_0 / (2B^{\text{OCDM}})$. It is worth highlighting that, although the ultimately obtained CIR estimates have the reduced length of $N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}}$, the processing gain remains proportional to $N_{\text{subch}}^{\text{OCDM}}$ since it is already experienced after the DFNT at the receiver, which performs a pulse-compression like operation on the $N_{\text{subch}}^{\text{OCDM}}$ discrete-time domain samples and takes place before the $N_{\text{subch}}^{\text{OCDM}} / N_{\text{Tx}}$ samples associated with each transmitter are selected at each receiver.

2.4.3.3 Extension to ISAC operation with SM-OCDM

To enable joint ISAC operation of an OCDM-based radar system with discrete-Fresnel domain estimation, the FrDM principle explained for MU and MIMO radar operation is applied. In this context, an OCDM symbol structure in the discrete-Fresnel domain based on the autocorrelation pattern of zero correlation zone (ZCZ) sequences used in PMCW radars and similar to the OCDM symbol structure from [ZWM22] is adopted. Unlike in the previously described radar-only case, each of the M^{OCDM} OCDM symbols in the transmit frame will be unique as they carry different modulated data. Consequently, a CP of length $N_{\text{CP}}^{\text{OCDM}} \in \mathbb{N}_{>0} | \{ 2N_{\text{CP}}^{\text{OCDM}} - 1 < N_{\text{subch}}^{\text{OCDM}} \}$ must be appended to each OCDM symbol in the discrete-time domain to avoid ISI. The resulting maximum tolerable range in the considered OCDM-based ISAC system is therefore equal to $N_{\text{CP}}^{\text{OCDM}} c_0 / (2B^{\text{OCDM}})$. For a given $N_{\text{CP}}^{\text{OCDM}}$, the FrDM-based design of the OCDM symbol depicted in Figure 2.18 is adopted and described as follows. First, the active, unmodulated subchirp of index $k = 0$ from (2.23) is kept and allocated and energy $\mathcal{E}^{\text{OCDM}, \text{rad}} \in \mathbb{R}_{>0}$, being followed by $N_{\text{CP}}^{\text{OCDM}} - 1$ inactive subchirps. At the receiver side, these first $N_{\text{CP}}^{\text{OCDM}}$ elements of the discrete-Fresnel domain OCDM symbol will ultimately contain the radar CIR estimate. Next, subchirps $k = N_{\text{CP}}^{\text{OCDM}}$ to $k = N_{\text{subch}}^{\text{OCDM}} - N_{\text{CP}}^{\text{OCDM}}$, which comprise a total of $N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1$ subchirps, are modulated with symbols belonging to a digital modulation constellation that are contained in the matrix $\mathbf{C}^{\text{OCDM}} \in \mathbb{C}^{(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1) \times M^{\text{OCDM}}}$, which results from S/P conversion on the modulated data stream that is redefined as $\mathbf{d}_{\text{Tx}}^{\text{OCDM}} \in \mathbb{C}^{(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1) M^{\text{OCDM}} \times 1}$. Finally, the last $N_{\text{CP}}^{\text{OCDM}} - 1$ are kept inactive, which, taking into account energy factors unlike in the previous

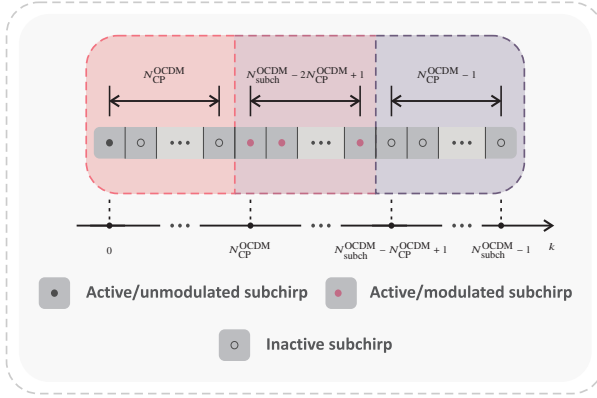


Figure 2.18: Subchirp allocation in the discrete-Fresnel domain for enabling ISAC operation. The first active, unmodulated subchirp and the next $N_{\text{CP}}^{\text{OCDM}} - 1$ inactive subchirps are used for radar sensing, while the $N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1$ active, modulated subchirps are used for communication. The last $N_{\text{CP}}^{\text{OCDM}} - 1$ inactive subchirps are used as a guard interval. Adapted from [20], © 2023 IEEE.

sections, yields a discrete-Fresnel domain transmit frame whose elements are given by

$$\begin{aligned} \dot{X}_{k,m}^{\text{OCDM}} = & \sqrt{\mathcal{E}^{\text{OCDM},\text{rad}}} \delta[k] \\ & + \sum_{k''=0}^{(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1) - 1} C_{k'',m}^{\text{OCDM}} \delta \left[\left\langle k - (N_{\text{CP}}^{\text{OCDM}} + k'') \right\rangle_{N_{\text{subch}}^{\text{OCDM}}} \right]. \end{aligned} \quad (2.39)$$

In this equation, $k'' \in \{0, 1, \dots, (N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1) - 1\}$ and $C_{k'',m}^{\text{OCDM}} \in \mathbb{C}$ denotes an element of $\mathbf{C}^{\text{OCDM}}$. The modulation symbols contained in $\mathbf{C}^{\text{OCDM}}$ have mean constellation energy that can be expressed as $\mathcal{E}^{\text{OCDM},\text{comm}} = \mathbb{E}\{|C_{k'',m}^{\text{OCDM}}|^2\}$ if $C_{k'',m}^{\text{OCDM}}$ is regarded as a random variable. The total energy allocated to the transmit OCDM symbol for joint ISAC operation is consequently $\mathcal{E}^{\text{OCDM},\text{ISAC}} = \mathcal{E}^{\text{OCDM},\text{rad}} + (N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1)\mathcal{E}^{\text{OCDM},\text{comm}}$. The last $N_{\text{CP}}^{\text{OCDM}} - 1$ inactive subchirps of each OCDM symbol are used as

a guard interval to avoid interference of the $N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1$ communications subchirps onto the $N_{\text{CP}}^{\text{OCDM}}$ radar subchirps, as the OCDM symbols experience a circular convolution with the radar CIRs. Due to the OCDM symbol structure from (2.39), the proposed OCDM-based ISAC system is named SM-OCDM-based ISAC system. In order for the communications subchirps to have the same allocated energy and therefore same communications performance as the subchirps in the C-OCDM scheme, $\mathcal{E}^{\text{OCDM,comm}} = \mathcal{E}^{\text{OCDM,ISAC}} / N_{\text{subch}}^{\text{OCDM}}$ is henceforth assumed. Consequently, the allocated energy to the active, unmodulated radar subchirp is $\mathcal{E}^{\text{OCDM,rad}} = \mathcal{E}^{\text{OCDM,ISAC}} - (N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1)\mathcal{E}^{\text{OCDM,comm}} = \mathcal{E}^{\text{OCDM,comm}} [(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1)\mathcal{E}^{\text{OCDM,ISAC}} / N_{\text{subch}}^{\text{OCDM}}]$.

After transmission, the OCDM signal will not only reflect off radar targets and be received by the same SM-OCDM ISAC system that originally transmitted it, but also propagate towards another OCDM communications or ISAC device. The radar case is first explained as follows. Assuming $n_{\Delta,h} \in \mathbb{Z}$ and $k_{\Delta,h} \in \mathbb{Z}$ for the sake of conciseness and using the result from (2.34), the elements of the matrix $\mathbf{Y}^{\text{OCDM,rad}}$ representing the radar receive frame after CP removal and transformation into the discrete-Fresnel domain are expressed as

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM,rad},q} = & \hat{h}_{k,m}^{\text{OCDM,rad,corr}} + \sum_{k''=0}^{(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1) - 1} C_{k'',m}^{\text{OCDM}} \sum_{h=0}^{H-1} e^{j\phi_{m,h}} \\ & \times \delta \left[\left\langle k - (n_{\Delta,h} + N_{\text{CP}}^{\text{OCDM}} + k_{\Delta,h}) \right\rangle_{N_{\text{subch}}^{\text{OCDM}}} \right] \\ & \times e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} \left[2nk_{\Delta,h} - k_{\Delta,h}^2 + N_{\text{subch}}^{\text{OCDM}} (n_{\Delta,h} k_{\Delta,h}) \right]}. \end{aligned} \quad (2.40)$$

Similarly, the result for $n_{\Delta,h} \in \mathbb{R}_{\geq 0}$ and $k_{\Delta,h} \in \mathbb{R}$ can be achieved based on the result from (2.33). Since the length of the CIRs in $\hat{\mathbf{h}}^{\text{OCDM,rad}}$ is equal to or smaller than $N_{\text{CP}}^{\text{OCDM}}$, the aforementioned CIR matrix can be redefined as $\hat{\mathbf{h}}^{\text{OCDM,rad}} \in \mathbb{C}^{N_{\text{CP}}^{\text{OCDM}} \times M^{\text{OCDM}}}$. Based on both (2.40) and the sufficient guard interval composed by null subchirps at the end of the discrete-Fresnel domain OCDM symbols in (2.39), the elements $\hat{h}_{k''',m}^{\text{OCDM,rad,corr}} \in \mathbb{C}$, $k''' \in \{0, 1, \dots, N_{\text{CP}}^{\text{OCDM}} - 1\}$, of the CIR matrix can be extracted from $\dot{\mathbf{Y}}^{\text{OCDM,rad}}$ as

$$\hat{h}_{k''',m}^{\text{OCDM,rad,corr}} \approx \dot{Y}_{k''',m}^{\text{OCDM,rad}}. \quad (2.41)$$

Apart from the discussed maximum tolerable range, all other radar performance parameters from Table 2.1 also hold for the considered SM-OCDM-based ISAC system with discrete-Fresnel domain channel estimation. For comparison with PMCW, OFDM, and C-OCDM, these parameters are listed as $G_p^{\text{SM-OCDM}}$, $\Delta R^{\text{SM-OCDM}}$, $R_{\text{max,ua}}^{\text{SM-OCDM}}$, $R_{\text{max,CP}}^{\text{SM-OCDM}}$, $\Delta v^{\text{SM-OCDM}}$, and $v_{\text{max,ua}}^{\text{SM-OCDM}}$ in Table 2.2 in Section 2.5.

As for the captured signal by the receiver of another OCDM communications or ISAC device, time and frequency synchronization can be performed in the same way as described for C-OCDM in Section 2.4.2, i.e., via techniques such as Schmidl & Cox's algorithm [SC97, Ouy17] and possibly using the preamble design from [7]. After CP removal and transformation into the discrete-Fresnel domain, unbiased CIR estimates can be obtained from the same unmodulated pilot subchirps that used for radar sensing as previously mentioned and, e.g., performed in [ZWM22]. After channel equalization, the modulation symbols contained in subchirps $k = N_{\text{CP}}^{\text{OCDM}}$ to $k = N_{\text{subch}}^{\text{OCDM}} - N_{\text{CP}}^{\text{OCDM}}$ can finally be extracted and demodulated following typical OCDM communications processing, e.g., as described for C-OCDM in Section 2.4.2. Consequently, a data rate denoted by $\mathcal{R}_{\text{QPSK}}^{\text{SM-OCDM}}$, as well as an energy per bit to noise power spectral density ratio $(E_b/N_0)_{\text{QPSK},k,m}^{\text{SM-OCDM}}$ and a BER denoted by $\text{BER}_{\text{QPSK},k,m}^{\text{SM-OCDM}}$ for the k th QPSK-modulated subchirp in the m th OCDM symbol as presented in Table 2.2 in Section 2.5.

To extend the SM-OCDM scheme for MIMO or multiuser operation, the same CDM and FDM approaches discussed for C-OCDM in Section 2.4.2 can be adopted. They are, however, associated with the loss of the inherent Doppler shift robustness in SM-OCDM. More specifically, if CDM based on outer codes is used, the non-coherent accumulation effect discussed for PMCW in [17] happens and reduces the ultimately experienced processing gain and SNR of targets in the radar image as defined in (2.4). As for FDM based on FSP, the fact that multiple transmit channels or users are separated in the frequency domain also reduces the Doppler shift tolerance of the SM-OCDM scheme. While in the SISO case Doppler shifts would only lead to range-velocity coupling, which tends to be tolerable for Doppler shifts experienced in practice, they lead to loss of orthogonality among originally separated channels or users if FSP is adopted as a multiplexing strategy. An alternative to enable MIMO or multiuser operation and keep the Doppler shift robustness of SM-OCDM is to use FrDM. However, empty subchirps would have to be kept as a guard

interval not only between the radar and communications subchirps of each users, but also between the subchirps allocated to each individual user. As a consequence, either the communications data rate $\mathcal{R}_{\text{QPSK}}^{\text{SM-OCDM}}$ or the maximum tolerable range, which is equal to $R_{\text{max,CP}}^{\text{SM-OCDM}} = N_{\text{CP}}^{\text{OCDM}} c_0 / (2B^{\text{OCDM}})$ in the SISO case, would be significantly reduced. Out of this reasons, multiplexing schemes will not further be analyzed for the proposed SM-OCDM-based ISAC system.

2.5 Comparative performance analysis

In this section, a comparative performance analysis of the ISAC modulation schemes presented in Sections 2.2 to 2.4 is performed. More specifically, Section 2.5.1 compares the performances of ISAC systems based on PMCW, OFDM, and C-OCDM. Next, Section 2.5.2 analyzes the performance of DFNT-based radar signal processing for OCDM-based radar systems. Due to its multicarrier nature, a comparative performance analysis of this scheme and its extension for ISAC applications, i.e., the SM-OCDM scheme, is presented having only OFDM and C-OCDM and no longer PMCW as alternative schemes. To assess the performance of the aforementioned modulation schemes for ISAC, the radar and communications performance parameters listed in Table 2.2 are considered alongside other performance metrics.

2.5.1 Comparison among PMCW, OFDM, and C-OCDM

In this section, the performance of the PMCW, OFDM, and C-OCDM modulation schemes for ISAC presented in Sections 2.2 to 2.4 are compared. Assuming an HAD context, the parameterization of ISAC systems based on all three considered modulation schemes is also presented and discussed. Next, the influence of a same-modulation-scheme interferer on the radar and communications performances of the ISAC systems considered in the previous sections are analyzed. Finally, the same scenario discussed in the previous sections is considered in a MIMO and multiuser context, and introduced performance changes in the considered ISAC systems by the adoption of multiplexing schemes are discussed.

Table 2.2: Radar and communications performance parameters of the ISAC systems based on PMCW, OFDM, C-OCDM, and SM-OCDM. Except for PMCW, which adopts BPSK modulation, all schemes are assumed to adopt QPSK modulation. Adapted from [18] and [20], © 2022, 2023 IEEE.

		Radar performance parameters		
	PMCW	OFDM	C-OCDM	SM-OCDM
Range resolution	$\Delta R^{\text{PMCW}} = \frac{c}{2f_c v_{\text{max}}}$	$\Delta R^{\text{OFDM}} = \frac{c}{2B_{\text{OFDM}}}$	$\Delta R^{\text{C-OCDM}} = \frac{c}{2B_{\text{C-OCDM}}}$	$\Delta R^{\text{SM-OCDM}} = \frac{c}{2B_{\text{SM-OCDM}}}$
Max. unamb. range	$R_{\text{max,unamb}}^{\text{PMCW}} = N_{\text{PRBS}} \frac{c}{2f_c v_{\text{max}}}$	$R_{\text{max,unamb}}^{\text{OFDM}} = N_{\text{OFDM}} \frac{c}{2B_{\text{OFDM}}}$	$R_{\text{max,unamb}}^{\text{C-OCDM}} = N_{\text{OCDM}} \frac{c}{2B_{\text{C-OCDM}}}$	$R_{\text{max,unamb}}^{\text{SM-OCDM}} = N_{\text{OCDM}} + N_{\text{CP}} \frac{c}{2B_{\text{C-OCDM}}}$
Max. tolerable range	No further restrictions	$R_{\text{max,ISI}}^{\text{OFDM}} = N_{\text{CP}} \frac{c}{2B_{\text{OFDM}}}$	$R_{\text{max,ISI}}^{\text{C-OCDM}} = N_{\text{CP}} \frac{c}{2B_{\text{C-OCDM}}}$	$R_{\text{max,ISI}}^{\text{SM-OCDM}} = N_{\text{CP}} \frac{c}{2B_{\text{C-OCDM}}}$
Relative radial velocity resolution	$\Delta v^{\text{PMCW}} = \frac{f_c (N_{\text{PRBS}} \Delta f_{\text{PRBS}} / \Delta f^{\text{PMCW}})}{2f_c (N_{\text{PRBS}} \Delta f_{\text{PRBS}} / \Delta f^{\text{PMCW}})}$	$\Delta v^{\text{OFDM}} = \frac{B_{\text{OFDM}}}{2f_c (N_{\text{OFDM}} + N_{\text{CP}}^2 v)} \Delta f^{\text{OFDM}}$	$\Delta v^{\text{C-OCDM}} = \frac{B_{\text{C-OCDM}}}{2f_c (N_{\text{OCDM}} + N_{\text{CP}}^2 v)} \Delta f^{\text{C-OCDM}}$	$\Delta v^{\text{SM-OCDM}} = \frac{B_{\text{SM-OCDM}}}{2f_c (N_{\text{OCDM}} + N_{\text{CP}}^2 v)} \Delta f^{\text{C-OCDM}}$
Max. unamb. relative radial velocity	$v_{\text{max,unamb}}^{\text{PMCW}} = \frac{f_c v_{\text{max}}}{\Delta f^{\text{PMCW}}}$	$v_{\text{max,unamb}}^{\text{OFDM}} = \frac{B_{\text{OFDM}}}{\Delta f^{\text{OFDM}}}$	$v_{\text{max,unamb}}^{\text{C-OCDM}} = \frac{B_{\text{C-OCDM}}}{\Delta f^{\text{C-OCDM}}}$	$v_{\text{max,unamb}}^{\text{SM-OCDM}} = \frac{B_{\text{SM-OCDM}}}{\Delta f^{\text{C-OCDM}}}$
Max. tolerable relative radial velocity	Depends on adopted PRBS	$v_{\text{max,ISI}}^{\text{OFDM}} = \frac{B_{\text{OFDM}}}{\Delta f^{\text{OFDM}}}$	$v_{\text{max,ISI}}^{\text{C-OCDM}} = \frac{B_{\text{C-OCDM}}}{\Delta f^{\text{C-OCDM}}}$	Depends on tolerable range-Doppler coupling
Processing gain	$G_P^{\text{PMCW}} = N_{\text{PRBS}} \Delta f^{\text{PMCW}} / \Delta f^{\text{PMCW}}$	$G_P^{\text{OFDM}} = N_{\text{OFDM}} \Delta f^{\text{OFDM}} / \Delta f^{\text{OFDM}}$	$G_P^{\text{C-OCDM}} = N_{\text{OCDM}} \Delta f^{\text{C-OCDM}} / \Delta f^{\text{C-OCDM}}$	$G_P^{\text{SM-OCDM}} = N_{\text{OCDM}} \Delta f^{\text{C-OCDM}} / \Delta f^{\text{C-OCDM}}$
Communications performance parameters				
	PMCW	OFDM	C-OCDM	SM-OCDM
Data rate	$R_{\text{data}}^{\text{PMCW}} = \frac{R_{\text{PRBS}}^{\text{PMCW}}}{N_{\text{PRBS}} \Delta f^{\text{PMCW}}}$	$R_{\text{data}}^{\text{OFDM}} = \frac{2 N_{\text{OFDM}} B_{\text{OFDM}}}{N_{\text{OFDM}} + N_{\text{CP}}}$	$R_{\text{data}}^{\text{C-OCDM}} = \frac{2 N_{\text{OCDM}} B_{\text{C-OCDM}}}{N_{\text{OCDM}} + N_{\text{CP}}}$	$R_{\text{data}}^{\text{SM-OCDM}} = \frac{2 N_{\text{OCDM}} B_{\text{C-OCDM}}}{N_{\text{OCDM}} + N_{\text{CP}}}$
E_b/N_0	$(E_b/N_0)_{\text{BPSK,unamb}}^{\text{PMCW}} = \frac{P_{\text{avg}} (G_{\text{PMCW}} - 1) N_{\text{PRBS}}^2}{N_{\text{PRBS}} \Delta f_{\text{PRBS}}^2}$	$(E_b/N_0)_{\text{QPSK,unamb}}^{\text{OFDM}} = \frac{P_{\text{avg}} B_{\text{OFDM}}^2}{N_0}$	$(E_b/N_0)_{\text{QPSK,unamb}}^{\text{C-OCDM}} = \frac{P_{\text{avg}}}{N_0 \Delta f_{\text{C-OCDM}}^2}$	$(E_b/N_0)_{\text{QPSK,unamb}}^{\text{SM-OCDM}} = \frac{P_{\text{avg}}}{N_0 \Delta f_{\text{C-OCDM}}^2}$
BER	$\text{BER}_{\text{BPSK,unamb}}^{\text{PMCW}} = Q \left(\sqrt{\frac{2(E_b/N_0)_{\text{BPSK,unamb}}^{\text{PMCW}}}{1}} \right)$	$\text{BER}_{\text{QPSK,unamb}}^{\text{OFDM}} = Q \left(\sqrt{2(E_b/N_0)_{\text{QPSK,unamb}}^{\text{OFDM}}} \right)$	$\text{BER}_{\text{QPSK,unamb}}^{\text{C-OCDM}} = Q \left(\sqrt{2(E_b/N_0)_{\text{QPSK,unamb}}^{\text{C-OCDM}}} \right)$	$\text{BER}_{\text{QPSK,unamb}}^{\text{SM-OCDM}} = Q \left(\sqrt{2(E_b/N_0)_{\text{QPSK,unamb}}^{\text{SM-OCDM}}} \right)$

2.5.1.1 Radar and communications parameter dependencies

To analyze the dependencies of radar and communications performance of ISAC systems based on PMCW, OFDM, and C-OCDM modulation schemes on key signal parameters, an HAD context is assumed. Consequently, the physical parameters listed in Table 2.3 are adopted, which results in a range resolution of 0.15 m for all considered modulation schemes. Additionally, $N_{\text{CP}}^{\text{OFDM}} = N_{\text{subc}}^{\text{OFDM}}/4$ and $N_{\text{CP}}^{\text{OCDM}} = N_{\text{subch}}^{\text{OCDM}}/4$ are adopted for the OFDM- and C-OCDM-based ISAC systems, respectively, which are typically considered values for the CP length in the literature. In order for the fraction of unused samples to be equal to 1/5 as in the OFDM and C-OCDM cases, $A^{\text{PMCW}} = 5$ is adopted for the PMCW-based ISAC system.

Based on the described assumptions and on the expressions in Table 2.2, Figure 2.19 shows the dependency of several parameters on the number of samples $N \in \mathbb{N}_{>0}$ of a single obtained range profile after radar signal processing, which corresponds to the PRBS length $N_{\text{PRBS}}^{\text{PMCW}}$ in the PMCW case, the number of subcarriers $N_{\text{subc}}^{\text{OFDM}}$ in the OFDM case, and the number of subchirps $N_{\text{subch}}^{\text{OCDM}}$ in the C-OCDM case. For all modulation schemes, a direct proportion to the block duration and maximum unambiguous and tolerable ranges is observed in Figures 2.19(a) and 2.19(b), respectively, while the maximum unambiguous and tolerable relative radial velocities are inversely proportional to N as shown in Figure 2.19(c).

In terms of communications performance, the considered PMCW ISAC systems experiences data rate degradation along with increasing N as observed in Figure 2.19(d). This is explained by the fact that its block duration becomes longer, while the number of transmitted BPSK symbols remains equal to one. Conversely, the communications data rate is kept constant for increasing N values in both OFDM and C-OCDM ISAC schemes, which also yield higher data rates than the two aforementioned modulation schemes due to their multicarrier nature. Although the number of subcarriers or subchirps and consequently the total number of transmitted QPSK symbols increases along with N , the block duration also does, which leads to a fixed communications data rate. Assuming an AWGN channel with no attenuation and a noise power defined as in the noise contribution in (2.4), i.e., $k_B B T_{\text{therm}} N F$, the obtained E_b/N_0 values are shown in Figure 2.19(e). In this figure, higher values are experienced for increasing N in the PMCW modulation scheme due to the fact that P_{Tx} is integrated into

Table 2.3: Adopted physical parameters for the considered SISO RadCom systems. Adapted from [18], © 2022 IEEE.

Physical parameters			
	PMCW	OFDM	C-OCDM
Transmission power	$P_{Tx} = 0 \text{ dBm}$		
Antenna gains	$G_{Tx} = G_{Rx} = 10 \text{ dB}$		
Noise figure	$NF = 10 \text{ dB}$		
Carrier frequency	$f_c = 79 \text{ GHz}$		
RF frequency bandwidth	$B_{PMCW} = 2 \text{ GHz}$	$B_{OFDM} = 1 \text{ GHz}$	$B_{OCDM} = 1 \text{ GHz}$
ADC sampling rate	$F_s^{PMCW} = 1 \text{ GHz}$	$F_s^{OFDM} = 1 \text{ GHz}$	$F_s^{OCDM} = 1 \text{ GHz}$

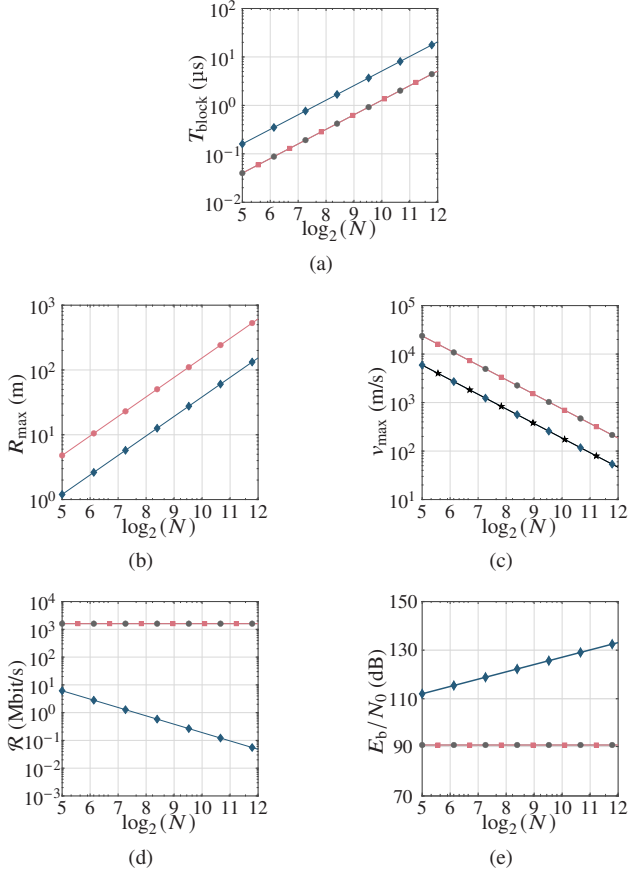


Figure 2.19. Calculated dependency of block duration and radar and communications performance parameters on the number of usable samples within a block $N \in \{32, 64, 128, 256, 512, 1024, 2040, 4096\}$, i.e., $N_{\text{PRBS}}^{\text{PMCW}}$, $N_{\text{subc}}^{\text{OFDM}}$, and $N_{\text{subch}}^{\text{OCDM}}$ for PMCW, OFDM, and C-OCDM, respectively. (a) shows the block duration for PMCW (◆), OFDM (●), and C-OCDM (■). (b) shows the maximum unambiguous range for all schemes (●) and the maximum ICI-free range for OFDM and C-OCDM (◆). (c) shows the maximum unambiguous relative radial velocity for all schemes and the maximum tolerable relative radial velocity for OFDM and C-OCDM due to ICI. (d) and (e) shows the achieved data rate and E_b/N_0 , respectively, both for PMCW (◆), OFDM (●), and C-OCDM (■). Adapted from [18], © 2022 IEEE.

the modulation symbol energy over a longer time. In the OFDM and C-OCDM ISAC systems, a lower, constant E_b/N_0 than in the PMCW case is experienced. This is due to the fact that the total transmission power P_{Tx} is divided over multiple subcarriers or subchirps. Furthermore, the fact that E_b/N_0 remains constant over increasing N values can be similarly explained as in the data rate case. While the power per subcarrier or subchirp is reduced by a certain factor with increasing N , the block duration is increased by the same factor, being therefore integrated into the same energy.

Considering the block parameter dependencies depicted in Figure 2.19 and aiming for typical parameterization of all ISAC modulation schemes in an HAD context, which must cover a maximum range of at least 70 m and a maximum relative radial velocity of at least 50 m/s, $N_{PRBS}^{PMCW} = 1023$, $N_{subc}^{OFDM} = 2048$, $N_{subch}^{OCDM} = 2048$ are fixed for the considered ISAC systems based on PMCW, OFDM, and C-OCDM, respectively. Based on the assumed parameters, the dependency of the frame duration and the remaining radar performance parameters on the number of blocks $M \in \mathbb{N}_{>0}$ is shown in Figure 2.20. In ISAC systems based on PMCW, OFDM, and C-OCDM, M is expressed by M^{PMCW} , M^{OFDM} , and M^{OCDM} , respectively. In this context, Figure 2.20(a) shows the resulting frame duration. For typical HAD applications, the frame duration should be below 20 ms, which yields appropriate relative radial velocity resolution and processing gain in all considered modulation schemes as shown in Figures 2.20(b) and 2.20(c).

Furthermore, the aforementioned fixed values for the N parameters result in the complementary cumulative distribution functions (CCDFs) of the RF PAPR for all considered ISAC modulation schemes shown in Figure 2.21. The attained results were obtained by calculating the peak-to-mean envelope power ratio (PMEPR) [LMI16], i.e., the PAPR of the complex-baseband equivalent of the transmit signals expressed in (2.9), (2.14), and (2.21), assuming an oversampling factor of 20 to capture fast variations of signals. Since f_c is much higher than the bandwidth in all considered cases, the PAPRs of the RF signals are finally obtained by adding 3.0 dB to the results from the aforementioned PAPR calculation for oversampled complex-baseband equivalents of signals [Mon19]. PMCW yields a fixed PAPR of nearly 3.0 dB due to CW characteristic of its transmit waveforms, which does not change for other PRBS lengths N_{PRBS}^{PMCW} . The worst observed PAPR in Figure 2.21, with probability 0.99 of being higher than 11.5 dB, is attained by both OFDM and C-OCDM due to their similar

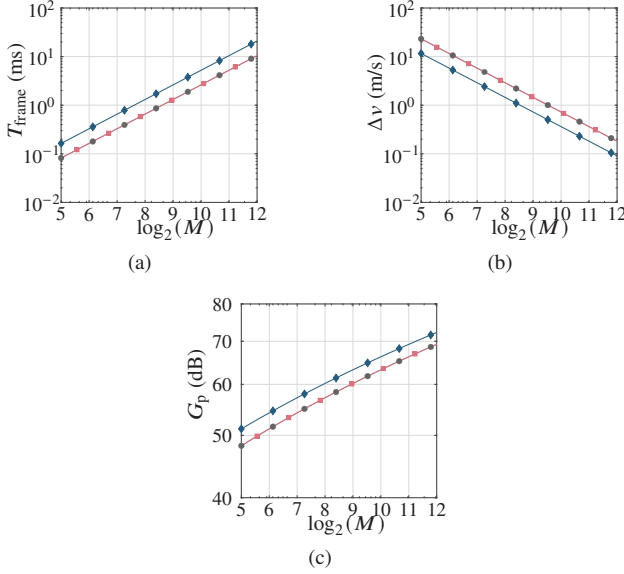


Figure 2.20. Calculated dependency of frame duration and radar performance parameters on the number of blocks $M \in \{32, 64, 128, 256, 512, 1024, 2040, 4096\}$, i.e., M^{PMCW} , M^{OFDM} , and M^{OCDM} for PMCW, OFDM, and C-OCDM, respectively. (a) shows the resulting frame duration, (b) shows the achieved relative radial velocity resolution, and (c) shows the processing gain. In all figures, results are shown for PMCW (◆), OFDM (●), and C-OCDM (■). Adapted from [18], © 2022 IEEE.

transmitter structure [OZ16a]. For higher number of subcarriers $N_{\text{subc}}^{\text{OFDM}}$ and subchirps $N_{\text{subch}}^{\text{OCDM}}$ in OFDM and C-OCDM systems, respectively, increasing PAPR values are expected. For the sake of conciseness, the reader is referred to [OZ16a, OZ16b, OM21] for an extended discussion on PAPR in OFDM- and OCDM-based systems.

2.5.1.2 ISAC system parameterization

In this section, a comparative performance analysis of the ISAC modulation schemes presented in Sections 2.2 to 2.4 is performed. Since PMCW-based ISAC systems are primarily used for radar sensing, offering low-data-rate com-

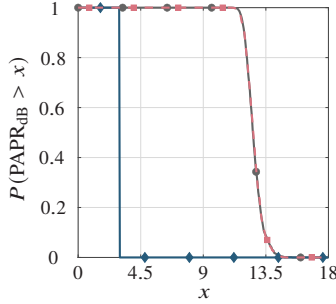


Figure 2.21. Simulated RF PAPR performance based on the carrier frequency $f_c = 79$ GHz from Table 2.3 and assuming $N_{\text{PRBS}}^{\text{PMCW}} = 1023$, $N_{\text{subc}}^{\text{OFDM}} = 2048$, and $N_{\text{subch}}^{\text{OCDM}} = 2048$ for PMCW (◆), OFDM (●), and C-OCDM (■), respectively. Adapted from [18], © 2022 IEEE.

munications capabilities simply as an additional feature, a radar-centric parameterization of ISAC systems is performed aiming to satisfy typical mid-range HAD requirements, which comprise a maximum range of 70 m with a resolution of up to 0.2 m, and a maximum relative radial velocity of 50 m/s with a resolution of up to 0.2 m/s. In this context, the physical parameters from Table 2.3 are further considered and additionally adopted waveform parameters for the considered ISAC modulation schemes are listed in Table 2.4.

The attained radar and communications performance parameters by the SISO ISAC systems with parameters from Table 2.4 are listed in Table 2.5. As a proof of concept, the system-on-a-chip (SoC) platform Zynq UltraScale+ RFSoC ZCU111 from Xilinx, Inc. was used to emulate SISO ISAC systems based on the considered modulation schemes and adopted parameters together with the radar target simulator (RTS) described in [DANZ21] and used in [DAN⁺21, DKK⁺21, DNPZ22], which receives transmitted signals by the ISAC system and modifies them according to adjustable range, relative radial velocity, and attenuation factors depending on, e.g., range and RCS. In the considered measurement setup, the transmitters and the receivers of both ISAC and RTS have I/Q channels. To avoid unwanted reflections of the surrounding, the DACs of the ISAC transmitter are directly connected to the corresponding ADCs of the RTS and the DACs of RTS transmitter are directly connected to the ADCs of the ISAC receiver. The region of interest of the obtained radar images from measurements with a single point target with RCS $\sigma_{\text{RCS}} = 20$ dBsm at an

Table 2.4: Adopted signal parameters for the considered SISO ISAC systems. Adapted from [18], © 2022 IEEE.

Signal parameters				
		PMCW	OFDM	C-OFDM
Block parameters	Samples	$N_{\text{PRBS}}^{\text{PMCW}} = 1023$ $A^{\text{PMCW}} = 5$	$N_{\text{subc}}^{\text{OFDM}} = 2048$ $N_{\text{CP}}^{\text{OFDM}} = 512$	$N_{\text{subch}}^{\text{OFDM}} = 2048$ $N_{\text{CP}}^{\text{OFDM}} = 512$
	Time duration	$T_{\text{PRBS}}^{\text{PMCW}} = 1.02 \mu\text{s}$ $T_{\text{block}}^{\text{PMCW}} = 5.11 \mu\text{s}$	$T_{\text{subc}}^{\text{OFDM}} = 2.05 \mu\text{s}$ $T_{\text{CP}}^{\text{OFDM}} = 0.51 \mu\text{s}$ $T_{\text{block}}^{\text{OFDM}} = 2.56 \mu\text{s}$	$T_{\text{subc}}^{\text{OFDM}} = 2.05 \mu\text{s}$ $T_{\text{CP}}^{\text{OFDM}} = 0.51 \mu\text{s}$ $T_{\text{block}}^{\text{OFDM}} = 2.56 \mu\text{s}$
Frame parameters	Samples	$M^{\text{PMCW}} = 2048$	$M^{\text{OFDM}} = 4096$	$M^{\text{OFDM}} = 4096$
	Time duration	$T_{\text{frame}}^{\text{PMCW}} = 10.48 \text{ ms}$	$T_{\text{frame}}^{\text{OFDM}} = 10.49 \text{ ms}$	$T_{\text{frame}}^{\text{OFDM}} = 10.49 \text{ ms}$

Table 2.5: Resulting radar and communications performance parameters for the considered SISO ISAC systems. Adapted from [18], © 2022 IEEE.

Radar performance parameters			
	PMCW	OFDM	C-OCDM
Range resolution	$\Delta R^{\text{PMCW}} = 0.15 \text{ m}$	$\Delta R^{\text{OFDM}} = 0.15 \text{ m}$	$\Delta R^{\text{C-OCDM}} = 0.15 \text{ m}$
Max. unamb. range	$R_{\text{max,ua}}^{\text{PMCW}} = 153.45 \text{ m}$	$R_{\text{max,ua}}^{\text{OFDM}} = 307.20 \text{ m}$	$R_{\text{max,ua}}^{\text{C-OCDM}} = 307.20 \text{ m}$
Max. tolerable range	—	$R_{\text{max,ISI}}^{\text{OFDM}} = 76.80 \text{ m}$	$R_{\text{max,ISI}}^{\text{C-OCDM}} = 76.80 \text{ m}$
Relative radial velocity resolution	$\Delta v^{\text{PMCW}} = 0.18 \text{ m/s}$	$\Delta v^{\text{OFDM}} = 0.18 \text{ m/s}$	$\Delta v^{\text{C-OCDM}} = 0.18 \text{ m/s}$
Max. unamb. relative radial velocity	$v_{\text{max,ua}}^{\text{PMCW}} = 185.60 \text{ m/s}$	$v_{\text{max,ua}}^{\text{OFDM}} = 370.85 \text{ m/s}$	$v_{\text{max,ua}}^{\text{C-OCDM}} = 370.85 \text{ m/s}$
Max. tolerable relative radial velocity	—	$v_{\text{max,ICI}}^{\text{OFDM}} = 92.71 \text{ m/s}$	$v_{\text{max,ICI}}^{\text{C-OCDM}} = 92.71 \text{ m/s}$
Processing gain	$G_p^{\text{PMCW}} = 69.23 \text{ dB}$	$G_p^{\text{OFDM}} = 69.24 \text{ dB}$	$G_p^{\text{C-OCDM}} = 69.24 \text{ dB}$
Communications performance parameters			
	PMCW	OFDM	C-OCDM
Data rate	$\mathcal{R}_{\text{BPSK}}^{\text{PMCW}} = 195.50 \text{ kbit/s}$	$\mathcal{R}_{\text{QPSK}}^{\text{OFDM}} = 1.60 \text{ Gbit/s}$	$\mathcal{R}_{\text{QPSK}}^{\text{C-OCDM}} = 1.60 \text{ Gbit/s}$

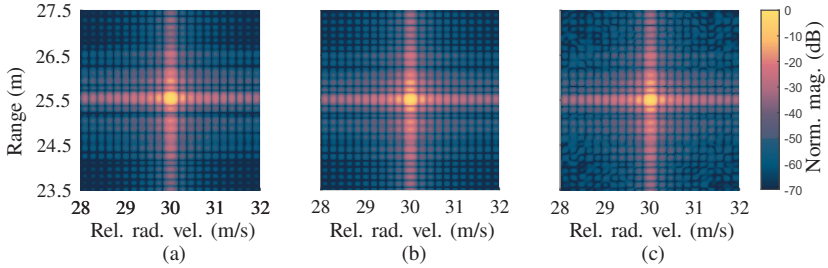


Figure 2.22. Obtained range-velocity radar images from measurements with a simplified baseband setup with ISAC systems based on: (a) PMCW, (b) OFDM, and (c) C-OCDM modulation schemes. Adapted from [18], © 2022 IEEE.

emulated range of 25.5 m and relative radial velocity of 30 m/s is shown in Figure 2.22. For those images, rectangular windowing in range and velocity directions was considered. A slightly higher sidelobe floor is observed for the C-OCDM-based ISAC system, while the remaining modulation schemes yield a similarly lower sidelobe level. This is explained by the influence of the modulated QPSK symbols onto the OCDM signal, which is not removed by the correlation processing at the receiver side and ultimately biases the range sidelobes. Since a DFT-based Doppler shift estimation is performed in all considered ISAC systems, the sidelobe level decreases at the same ratio for all modulation schemes when moving away from the estimated target velocity along the relative radial velocity direction.

Although only slight differences are observed in the radar images shown in Figure 2.22, where typical range and relative radial velocity values are assumed, an increase in the latter parameter may induce Doppler shifts that impose non-negligible degradation to the radar images that strongly depends on the adopted modulation scheme. Aiming to quantify the Doppler-shift robustness of the radar sensing performed by ISAC systems based on each of the three considered modulation schemes, peak power loss ratio (PPLR), peak sidelobe level ratio (PSLR), and integrated sidelobe level ratio (ISLR) of their range ambiguity functions were evaluated along Doppler shifts associated with relative radial velocities from 0 m/s to the maximum tolerable value of 92.71 m/s in the OFDM scheme. The simulation results, which assume an oversampling factor of 20 to account for the influence of sidelobes on point targets that do not exactly fall

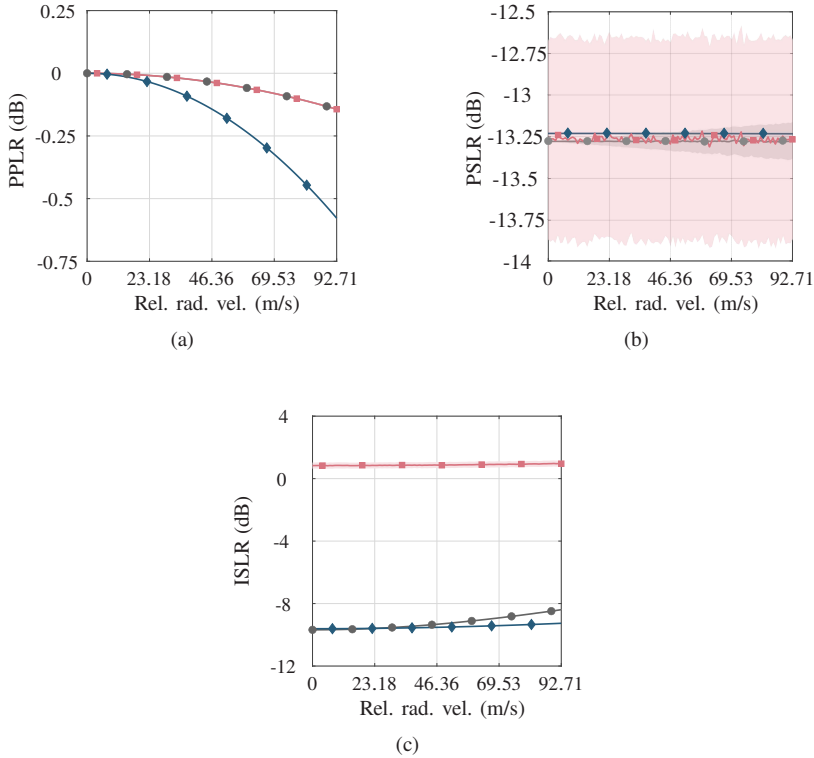


Figure 2.23. Simulated Doppler-shift tolerance: range (a) PPLR, (b) PSLR, and (c) ISLR for PMCW (◆), OFDM (●), and C-OCDF (■). The continuous lines represent the mean value of the calculated parameters and the shading in the background represents the standard deviation. Adapted from [18], © 2022 IEEE.

onto a single range bin, are shown in Figure 2.23. The PPLR results shown in Figure 2.23(a) serve as a measure of SNR loss, expressing the ratio between the mainlobe power under the evaluated Doppler shift and the one under no Doppler shift [OJUY20]. An analysis of the obtained results reveals that PMCW yields the highest PPLR degradation along with increasing relative radial velocities, which is caused by both the degradation of the MLS periodic autocorrelation function (PACF) and the incoherent accumulation effect under increasing

Doppler shifts, which is described in further detail in [17]. A maximum PPLR degradation of 0.14 dB is observed in both OFDM and C-OCDM cases. For all modulation schemes, the calculated PPLR values present negligible standard deviation. In their turn, the PSLR and ISLR express the ratio between the peak powers of the highest sidelobe and the mainlobe and the ratio between the integrated sidelobe and mainlobe powers, respectively [LMI16]. Both metrics serve as parameters to predict the radar sensing performance in multi-target scenarios, since high sidelobes associated with a strong target detection may not only mask weaker target reflections, but also be erroneously detected as targets. As shown in Figure 2.23(b), all modulation schemes present roughly the mean PSLR value along all evaluated relative radial velocities. In terms of standard deviation of the PSLR, PMCW presents negligible values, while increasing values are observed for both OFDM and C-OCDM. The high standard deviation values experiences in the C-OCDM-based ISAC system are explained by the dependency of the obtained range profiles on the transmit QPSK symbols. The Doppler-shift robustness analysis is finally completed with the ISLR results shown in Figure 2.23(c). ISLR degradations of 0.12 dB and 0.35 dB at 92.71 m/s are observed in the C-OCDM and PMCW cases, respectively, w.r.t. the values attained at 0 m/s. Although a small ISLR degradation is experienced in the C-OCDM case, high ISLR values are attained for all considered velocities due to the aforementioned absence of compensation for the influence of transmit QPSK symbols on the correlation-based C-OCDM radar processing. Lastly, the most severe ISLR degradation is observed in the OFDM case, where the experienced ICI at the highest considered velocity yields a 1.31 dB ISLR degradation w.r.t. the static case. As in the PPLR case, negligible standard deviation of the calculated ISLRs is observed. If other parameterizations of the SISO ISAC systems are adopted, attention must be drawn to the fact that increasing PRBS length $N_{\text{PRBS}}^{\text{PMCW}}$ and number of accumulations A^{PMCW} in the PMCW case, as well as increasing numbers of subcarriers $N_{\text{subc}}^{\text{OFDM}}$ and sub-chirps $N_{\text{subch}}^{\text{OCDM}}$ in the OFDM and C-OCDM cases, respectively, lead to more severe degradation of PPLR, PSLR, and ISLR values along increasing relative radial velocities if both the carrier frequency and the frequency bandwidth are kept constant.

The comparative performance analysis in this subsection is finally completed with a simulative BER analysis, whose results assuming a 4-path Rayleigh channel are shown in Figure 2.24. For the ISAC systems based on OFDM and C-OCDM, mean BER results were obtained for all $N_{\text{subc}}^{\text{OFDM}}$ subcarriers

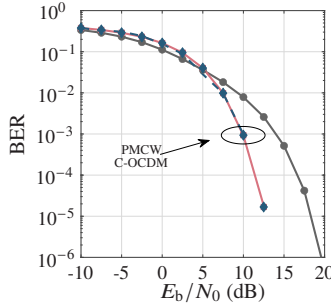


Figure 2.24. Simulated BER performance in a 4-path Rayleigh channel for PMCW (◆), OFDM (●), and C-OCDM (■). Adapted from [18], © 2022 IEEE.

and $N_{\text{subch}}^{\text{OCDM}}$ subchirps, respectively. The simulated BER performance results are represented in the form of curves in Figure 2.24 and agree with the calculated BER performance results based on the closed-form E_b/N_0 and BER expressions from Table 2.2, which in turn are represented in the form of markers. Due to their spread-spectrum nature, ISAC systems based on PMCW and C-OCDM schemes yield virtually the same BER performance, outperforming their OFDM counterpart for E_b/N_0 higher than 5 dB in the considered communications channel. If the same transmit power P_{Tx} is assumed for all considered ISAC systems, very high E_b/N_0 values will be attained by PMCW, since it only carries 1 bit per block. This results in the fact that symbols may be correctly demodulated in those modulation schemes even under negative SNR regime, since very high processing gains are yielded by the ISAC systems based on PMCW. In turn, OFDM and C-OCDM yield much higher data rates as listed in Table 2.5, while having considerably lower E_b/N_0 values since 2 bits are transmitted per subcarrier and subchirp, respectively. This indicates that C-OCDM yields the best communications performance among the three considered modulation schemes, achieving the same data rate and spectral efficiency as OFDM while also presenting sufficiently robust BER performance.

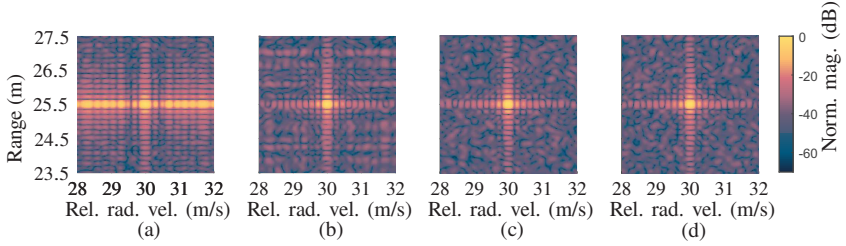


Figure 2.25. Simulated range-velocity radar images under influence of a same-modulation-scheme interferer: (a) PMCW with same PRBS for interferer and victim ISAC systems, (b) PMCW with orthogonal PRBSs for interferer and victim ISAC systems, (c) OFDM, and (d) C-OFDM. In all subfigures, same-modulation-scheme interferers and $\text{SIR} = -25$ dB before radar signal processing were assumed. Adapted from [18], © 2022 IEEE.

2.5.1.3 Influence of interference on radar performance

Still considering the parameters from Tables 2.4 and 2.3, as well as their resulting radar and communications performance parameters from Table 2.5, the influence of a ISAC interferer on both radar and communications performances of a victim ISAC system is analyzed assuming that both adopt the same modulation scheme, but transmit different payload data.

For the analysis of radar performance under interference, it is assumed that the interferer and victim ISAC systems are at the same position and that the same considered point target for the measurements with the RTS, which has $\text{RCS } \sigma_{\text{RCS}} = 20$ dBsm, range of 25.5 m and relative radial velocity of 30 m/s, is present in the scenario. Additionally, it is assumed that there is no coupling between the receiver of the victim ISAC system and the both the victim and interferer ISAC transmitters. Consequently, the victim ISAC receives both the reflection of its own signal and the one of the signal transmitted by the ISAC interferer off the point target, being also impaired by AWGN. Although this scenario is not realistic, it allows evaluating the sole influence of the adopted modulation scheme on the experienced interference in a worst-case scenario and has already been assumed in the literature [SSBZ13].

Given the assumed scenario, the simulated radar images keeping the aforementioned parameters and adding an interfering signal with power such that the victim ISAC experiences a signal-to-interference ratio (SIR) before signal pro-

cessing of -25 dB are shown in Figure 2.25. In this image, both the use of the same MLS for the ISAC interferer and the ISAC victim and the use of orthogonal MLSs as described in Section 2.2 are considered for PMCW, which are henceforth occasionally referred to as the worst and best cases in the considered scenario, respectively. While only a higher noise floor is observed for PMCW with orthogonal MLSs, OFDM and C-OCDM, a stripe along the relative radial velocity axis at the point target range and its corresponding range sidelobes is observed in the radar images obtained by PMCW with same MLS for the ISAC interferer and the ISAC victim. This is explained by the fact that the correlation with an orthogonal MLS in the best PMCW case, symbol division in the OFDM case, and the PCCF in the C-OCDM case turn the interfering signal into a noise-like signal, therefore experiencing the full range processing gain that is proportional to $N_{\text{subc}}^{\text{OFDM}}$ and $N_{\text{subch}}^{\text{OCDM}}$, respectively. Conversely, the correlation with the same MLS in the worst PMCW case does not contribute to the experienced processing gain w.r.t. the interfering signal. In fact, only the DFTs for relative radial velocity estimation are able to transform the interfering signal contribution to the already estimated range cells into a noise-like signal, being the effective processing gains w.r.t. the interfering signal in the worst PMCW case around $10 \log_{10} A^{\text{PMCW}} N_{\text{PRBS}}^{\text{PMCW}} = 37.10$ dB lower than the listed values in Table 2.22. This would result in expected dynamic range reductions for the considered point target w.r.t. the interfering signal and noise of around 37.10 dB compared to OFDM and C-OCDM. The aforementioned values, however, only hold with the simplified assumption that the resulting noise-like signal from the interfering signal after radar signal processing is Gaussian.

Considering SIR values before radar signal processing that range from -90 dB to 30 dB, Figure 2.26 shows the achieved dynamic range for the considered point target with the adopted signal and system parameters. The calculated dynamic range values can be described as the ratio between the point target contribution and the peak of the joint contribution of interference and noise in the radar image, and the ultimately presented results are the average dynamic ranges over multiple simulations. At the low SIR regime in Figure 2.26, the achieved dynamic range by both OFDM and C-OCDM ISAC systems is around 32.70 dB and 12.26 dB higher than in the worst and best PMCW cases, respectively. The aforementioned numbers indicate that the experienced dynamic ranges in the worst PMCW case is around 4 dB higher than originally predicted, which is explained by the fact that simply performing a DFT along the same range cells transforms the BPSK modulation symbols of the inter-

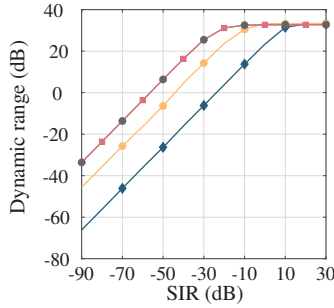


Figure 2.26. Simulated dynamic range as a function of SIR w.r.t. same-modulation-scheme interfering signal before radar signal processing for ISAC systems based on all considered modulation schemes and parameterization. In this figure, PMCW worst (♦) represents the case where the interferer and victim ISAC systems use the same PRBS, whereas PMCW best (●) represents the case where orthogonal PRBSs are adopted. In addition, results are also shown for OFDM (●) and C-OCDM (■). Adapted from [18], © 2022 IEEE.

fering signal into a pseudonoise that follows a non-Gaussian distribution. In its turn, the reduced dynamic range in the best PMCW is explained by the considerably high sidelobes resulting from the PCCF between the orthogonal PRBSs [WGBC19]. Furthermore, the presented results in Figure 2.26 show that negligible dynamic range losses are observed in the best PMCW, OFDM and C-OCDM cases for SIR values higher than around -20 dB, while SIRs of 20 dB or higher are demanded in the worst PMCW case. For low SIR regime, SIC can be performed by iteratively decoding signals from present interferers in the scenario, reconstructing them and subtracting their contributions from the overall received signal similarly to what is described in [SNZ18].

2.5.1.4 Influence of interference on communications performance

For evaluating the influence of an interfering signal of the same modulation scheme on the communications performance of a victim ISAC receiver, the signal parameters from Table 2.4 are further considered, yielding the listed data rates in Table 2.5. Similarly to the analysis of interference influence on radar performance, it is assumed that the ISAC transmitter transmitting the signal of interest and the ISAC interferer adopt the same modulation scheme

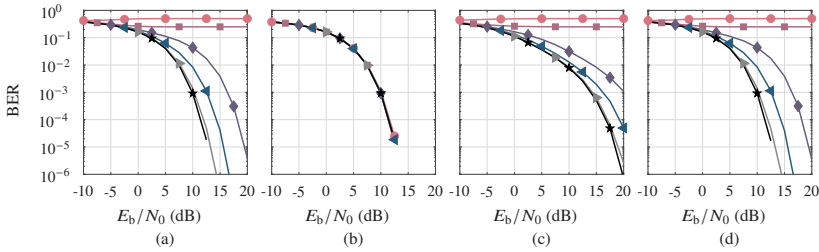


Figure 2.27. Simulated BER performance under interference with ISAC systems based on: (a) PMCW with same PRBS, (b) PMCW with orthogonal PRBSs, (c) OFDM, and (d) C-OCDM modulation schemes. In all subfigures, same-modulation-scheme interferers and different SIR levels before communications signal processing were assumed, namely -10 dB (●), 0 dB (■), 5 dB (◆), 10 dB (◀), 20 dB (▶), and no interference (★). Adapted from [18], © 2022 IEEE.

and are at the same position, therefore experiencing the same channel CIR which is assumed to be the same of the adopted Rayleigh channel for the results in Figure 2.24. Furthermore, both the use of same or orthogonal MLSs are considered for PMCW.

The simulated BER performances of all considered ISAC modulation schemes in the described scenario are presented in Figure 2.27, where BER versus E_b/N_0 curves for SIRs ranging from -10 dB to 20 dB are considered alongside with the interference-free case. The achieved results show that an SIR of 20 dB results in nearly the same BER performance as in the interference-free case for all three modulation schemes, except in the PMCW case where orthogonal sequences are assigned to the interferer and victim ISAC systems. As for the analysis of the influence of interference on radar sensing, a better performance is achieved in the aforementioned case w.r.t. its counterpart with same MLSs for the ISAC interferer and the ISAC victim. This is explained by the yielded isolation of the ISAC victim's signal w.r.t. the interfering one with the use of orthogonal MLSs, which already assumes a value of 24.81 dB at the lowest considered SIR value of -10 dB and therefore makes the influence of the interfering signal at all considered SIR values negligible. Additionally, it can be observed that PMCW assuming the use of the same MLS for the interferer and victim ISAC systems, and C-OCDM ISAC systems still tend to outperform their counterparts based on PMCW with orthogonal sequences and OFDM counterparts for reasonable SIR values, while all modulation schemes present

equally poor performance for lower SIR values than 5 dB. This also includes a lower communications performance degradation due to interference compared to the ISAC systems based on PMCW with orthogonal sequences and OFDM in the considered scenario. To compensate for SIR values that may yield non-negligible performance degradation, which can either be caused by interference of an external ISAC system as so far considered or by reflections off radar targets of the signal transmitted by the same ISAC device whose receiver is evaluated, iterative SIC can also be performed following, e.g., a similar approach as the radar-oriented one in [SNZ18].

2.5.1.5 Analysis of multiplexing schemes for MIMO operation

As discussed in Sections 2.2, 2.3, and 2.4 for ISAC systems based on the PMCW, OFDM, and C-OCDFM modulation schemes, respectively, the use of multiplexing schemes for enabling MIMO or multiuser operation may change some of the signal parameters, as well as some of the radar and communications performance parameters. Assuming $N_{Tx} = 4$ transmitters and adjusting the necessary parameters to keep the same frame durations from Table 2.4, the effects of each multiplexing scheme on the aforementioned parameters is discussed as follows.

Based on the discussion on multiplexing strategies for PMCW-based ISAC systems in Section 2.2, the use of orthogonal PRBSs as a CDM strategy produces neither changes in the signal parameters nor in the radar and communications ones, being the only difference the reduced dynamic range even under absence of Doppler shifts. If, however, CDM based on outer coding is adopted, then the required duration to transmit a block becomes $T_{\text{block}}^{\text{PMCW}} = 20.44 \mu\text{s}$ and the number of transmitted blocks becomes $M^{\text{PMCW}} = 512$ so that $T_{\text{frame}}^{\text{PMCW}}$ is kept the same as listed in Table 2.4. Consequently, the maximum unambiguous relative radial velocity is changed to $v_{\text{max,ua}}^{\text{PMCW}} = 46.40 \text{ m/s}$, and the achieved data rate assuming BPSK modulation in a multiuser scenario is reduced to $\mathcal{R}_{\text{BPSK}}^{\text{PMCW}} = 48.88 \text{ kb/s}$.

According to the discussion in Section 2.3, the parameters listed in Table 2.4 for the SISO case are kept if interleaved FDM is adopted in the considered OFDM-based ISAC system. However, as the number of assigned subcarriers to each of the $N_{Tx} = 4$ transmitter is equal to $N_{\text{sub}}^{\text{OFDM}}/N_{Tx} = 512$, changes are

observed w.r.t. the radar parameters listed in Table 2.5 for the SISO case. Those changes are the reduced maximum unambiguous range $R_{\max,ua}^{\text{OFDM}} = 76.80$ m and the reduced processing gain $G_p^{\text{OFDM}} = 63.24$ dB. Furthermore, in a multiuser scenario, the achieved data rate assuming QPSK modulation is also reduced to $\mathcal{R}_{\text{QPSK}}^{\text{OFDM}} = 400$ Mb/s.

Following the discussion in Section 2.4, the use of FSP as an FDM approach in the considered C-OCDM-based ISAC system yields similar changes in the signal, radar and communications performance parameters as for the interleaved FDM in the OFDM case w.r.t. to the listed parameters for the SISO case in Tables 2.4 and 2.5. Those are a reduced maximum unambiguous range $R_{\max,ua}^{\text{OCDM}} = 76.80$ m, a reduced processing gain $G_p^{\text{OCDM}} = 63.24$ dB, and a data rate $\mathcal{R}_{\text{QPSK}}^{\text{OCDM}} = 400$ Mb/s in a multiuser scenario. Additionally, as discussed in Section 2.4, an upper bound of $v_{\max,FSP}^{\text{OCDM}} = 92.71$ m/s on the maximum tolerable relative radial velocity is set to ensure that only tolerable isolation losses among the transmitters are experienced. If, instead, outer coding is adopted as a CDM approach, the required duration to transmit a block is increased to $T_{\text{block}}^{\text{OCDM}} = 10.24$ μ s and the number of transmitted blocks decreased to $M^{\text{OCDM}} = 1024$ to keep $T_{\text{frame}}^{\text{OCDM}}$ constant. This results in a reduced maximum unambiguous relative radial velocity $v_{\max,ua}^{\text{OCDM}} = 92.71$ m/s and in a reduction by a factor of $N_{\text{Tx}} = 4$ of the achieved data rate, which yields the same mentioned value in the FSP-based FDM case.

Besides the changes in the signal parameters, as well as the ones in the radar and communications performance parameters, the choice for a specific multiplexing scheme affects the robustness of the channel isolation against increasing Doppler shifts. To perform a comparative analysis among different modulation and multiplexing schemes, it is assumed that all $N_{\text{Tx}} = 4$ transmitters and a single receiver are at the same position and that there is no coupling between any of the transmitters and the aforementioned receiver. Additionally, it is assumed that a single point target with RCS $\sigma_{\text{RCS}} = 20$ dBsm is at a range of 25.5 m in front of them as previously assumed for the results in Figure 2.22. Considering that no constraints on maximum unambiguous or tolerable range are violated, Figure 2.28 presents the mean achieved dynamic range by all $N_{\text{Tx}} = 4$ transmitters assuming the same parameters from the interference analysis for radar sensing. The achieved dynamic range results are calculated as the ratio between the point target contribution to the obtained radar image and the peak of the joint transmitter-to-transmitter interference and noise contribution. The presented

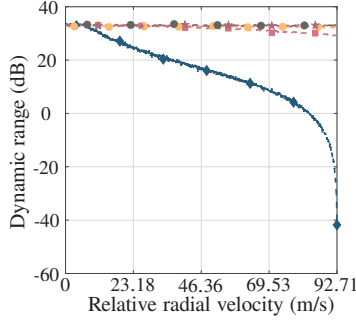


Figure 2.28. Simulated dynamic range as a function of the point target radial relative velocity for ISAC systems based on all considered modulation schemes. In the PMCW case, CDM-1 (●) represents the CDM approach based on orthogonal PRBS, while CDM-2 (◆) represents the CDM approach based on outer coding. Results are also shown for OFDM with FDM (●), C-OCDFM with CDM (■), and C-OFDFM with FDM (★). Adapted from [18], © 2022 IEEE.

results in Figure 2.28 for velocities up to 92.71 m/s, which corresponds to the maximum tolerable value in the OFDM-based ISAC system, show that most combinations of modulation and multiplexing schemes suffer nearly negligible dynamic range degradation for the whole considered velocity range. The only exception is PMCW with CDM based on outer coding, which presents severe dynamic range loss even for low relative radial velocities. The poorer performance in the aforementioned case is explained by its rather long block duration of $T_{\text{block}}^{\text{PMCW}} = 20.44 \mu\text{s}$, which makes the sum of $N_{\text{Tx}} = 4$ repetitions of the original block for decoding non-coherent due to Doppler-shift-induced phase rotations, leading to decreased power of the point target contribution to the receive signal w.r.t. to noise as discussed in [17].

2.5.2 Comparison of SM-OCDFM with OFDM and C-OCDFM

In this section, a performance analysis of discrete-Fresnel domain channel estimation for OCDFM-based radar systems is performed, including extensions of this concept for both MU or MIMO operation and ISAC operation with

the SM-OCDM scheme. Aiming at mid-range HAD applications [18], which require a maximum range of 70 m and a range resolution of 0.2 m, as well as a maximum relative radial velocity of 50 m/s and a velocity resolution of 0.2 m/s, a carrier frequency $f_c = 79$ GHz and a frequency bandwidth $B = 1$ GHz are adopted for all considered modulation schemes. Specific signal parameters are further listed for both radar and communications performance analysis, as well as the PAPR analysis later presented.

2.5.2.1 Radar performance analysis

To assess the radar performances of the investigated SISO OCDM-based radar and SISO SM-OCDM-based ISAC systems, which are both based discrete-Fresnel domain channel estimation, simulation and measurement results are presented and compared with the performance of similarly parameterized SISO C-OCDM- and OFDM-based radar systems in this subsection. In all analyses of this subsection, rectangular windowing is assumed for both range and Doppler processing, which is essentially the same as skipping windowing processing and enables a clearer comparison among the modulation schemes without influence of a particular window. Assuming that all discrete-Fresnel symbols in the frame are defined as in (2.23) for the OCDM-based radar system and that all OCDM and OFDM symbols are equal in the C-OCDM and OFDM cases, respectively, no CP is henceforth used for these systems, i.e., $N_{CP}^{OCDM} = 0$. For the proposed SISO SM-OCDM-based ISAC system, however, $N_{CP}^{OCDM} = 512$ is adopted for a meaningful analysis, as the use of this modulation scheme is only advantageous when ISAC operation is envisioned and therefore using equal OCDM symbols would be purposeless. The adopted signal parameters and the resulting radar performance parameters are listed in Table 2.6. It can be observed in this table that SM-OCDM has lower maximum range than the other schemes. The reason for that is because this scheme has a maximum tolerable range $R_{\max, CP}^{SM-OCDM} = N_{CP}^{OCDM} c_0 / (2B^{OCDM})$, while the its counterparts are only limited by their maximum unambiguous range $R_{\max, ua}^{SM-OCDM} = N_{subch}^{OCDM} c_0 / (2B^{OCDM})$ as defined in Table 2.1. Besides, processing gain and maximum unambiguous velocity, which are calculated according to the expressions in Table 2.1, are also lower in the SM-OCDM scheme due to its longer block duration T_{block}^{OCDM} caused by the use of CP.

Table 2.6: Adopted signal parameters and resulting radar performance parameters for the considered OCDM-based radar, SM-OCDM-based ISAC, C-OCDM-based radar and OFDM-based radar systems. For conciseness, the superscripts of the individual modulation schemes are omitted in the signal and radar performance parameters. © 2023 IEEE. Reprinted, with permission, from [20].

Signal and radar performance parameters				
	OCDM	SM-OCDM	C-OCDM	OFDM
Symbol length N	2048	2048	2048	2048
CP length N_{CP}	0	512	0	0
Symbols per frame M	5120	4096	5120	5120
Block duration T_{block}	2.05 μs	2.56 μs	2.05 μs	2.05 μs
Frame duration T_{frame}	10.49 μs	10.49 μs	10.49 μs	10.49 μs
Processing gain G_p	70.21 dB	69.24 dB	70.21 dB	70.21 dB
Range resolution ΔR	0.15 m	0.15 m	0.15 m	0.15 m
Max. range R_{max}	307.20 m	76.80 m	307.20 m	307.20 m
Velocity resolution Δv	0.18 m/s	0.18 m/s	0.18 m/s	0.18 m/s
Max. unamb. vel. $v_{\text{max,ua}}$	463.24 m/s	370.59 m/s	463.24 m/s	463.24 m/s

Based on the assumed parameters, simulated PPLR, PSLR, and ISLR [LMI16, 18] results are shown in Figure 2.29 to assess the distortion of the range mainlobe and sidelobes induced by Doppler shifts. Since those calculations assume a single point target, a generic subindex h is henceforth assumed for $n_{\Delta,h}$ and $k_{\Delta,h}$. In this figure, the whole unambiguous interval for range and relative radial velocity were considered, which results in $n_{\Delta,h} \in [0, 2048]$ and $k_{\Delta,h} \in [-0.5, 0.5]$. The values for $k_{\Delta,h}$ were defined based on the maximum unambiguous velocity expression from Table 2.1, the relationship $f_{\text{D,target},h} = -2v_h/\lambda_0$ between Doppler shifts and relative radial velocities described in Section 2.1, and the frequency resolution $\Delta f^{\text{OCDM}} = B^{\text{OCDM}}/N_{\text{subch}}^{\text{OCDM}}$ of the OCDM-based radar system. Consequently, $k_{\Delta,h}$ can be interpreted as a normalized Doppler shift expressed as $k_{\Delta,h} = f_{\text{D,target},h}/\Delta f^{\text{OCDM}}$. Similarly, $n_{\Delta,h}$ can be interpreted as a normalized range by the range resolution and expressed as $n_{\Delta,h} = R_h/\Delta R^{\text{OCDM}}$. Overall, degradations of PPLR, PSLR, and ISLR are mostly only observed for $|k_{\Delta,h}| > 0.1$, which is also expected, e.g., in OFDM-based radar and ISAC systems where the maximum tolerable velocity is associated with this $k_{\Delta,h}$ up-

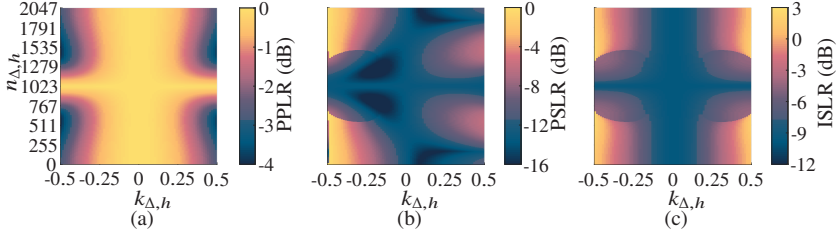


Figure 2.29: Simulated Doppler-shift tolerance: (a) PPLR, (b) PSLR, and (c) ISLR of a single target reflection with different $n_{\Delta,h} = 2R_h B^{\text{OCDM}}/c_0$ and $k_{\Delta,h} = f_{D,\text{target},h}/\Delta f^{\text{OCDM}}$ pairs obtained in an OCDM-based system with discrete-Fresnel domain radar processing. Adapted from [20], © 2023 IEEE.

per bound [NMZ18, 18]. The PPLR results in Figure 2.29(a) show a highest degradation of the processing gain G_p^{OCDM} of up to about 4 dB, e.g., for the approximate region covered by $n_{\Delta,h} \in [0, 767] \cup [1279, 2048]$ and $k_{\Delta,h} = -0.5$. Next, Figure 2.29(b) shows the PSLR results, which indicate the ratio between the highest sidelobe and the mainlobe, while the ISLR results, which indicate the ratio between the integrated sidelobe level and the integrated mainlobe level, are presented in Figure 2.29(c). Near the aforementioned region in the PPLR case, both PSLR and ISLR experience significant degradation, which can be explained by the results from (2.33) and (2.34). Besides the range-Doppler coupling in these equations, the range and Doppler dependent phase terms, which are $e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (2nk_{\Delta,h} - k_{\Delta,h}^2 + N_{\text{subch}}^{\text{OCDM}} k_{\Delta,h})}$ in (2.34), introduce changes in the shape of the estimated CIRs w.r.t. the ideally expected ones and is illustrated in more detail as follows.

Since the processing for Doppler shift estimation, and consequently relative radial velocity estimation, is the same in all considered modulation schemes, a more focused analysis of the range sidelobes is henceforth performed. Still assuming the parameters from Table 2.6, both OCDM-based radar and SM-OCDM-based ISAC systems are analyzed and compared with the considered C-OCDM- and OFDM-based radar systems. In this context, Figure 2.30 shows range cuts of range-velocity radar images obtained from measurements with a Zynq UltraScale+ RFSoc ZCU111 from Xilinx, Inc. In the presented results, $k_{\Delta,h}$ and $n_{\Delta,h}$ are defined for OFDM in a similar

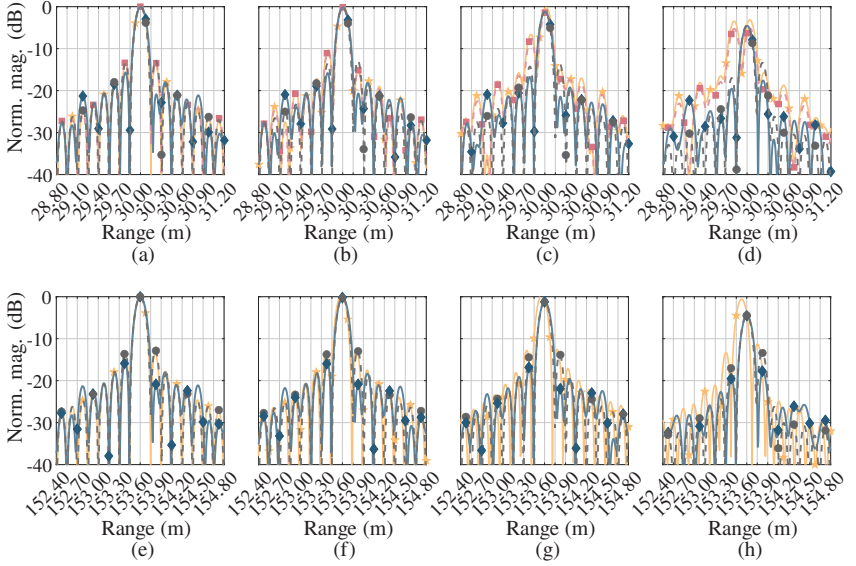


Figure 2.30: Range cuts of radar images from measurements with a simplified baseband setup for a SISO OCDM-based radar system ($\star$), an SM-OCDM-based ISAC system ($\blacksquare$), a C-OCDM-based ISAC system ($\blacklozenge$), and an OFDM-based ISAC system ($\bullet$) for a single target with relative radial velocity of (a) 0 m/s ($k_{\Delta,h} = 0$), (b) 92.71 m/s ($k_{\Delta,h} = -0.1$), (c) 231.78 m/s ($k_{\Delta,h} = -0.25$), and (d) 463.56 m/s ($k_{\Delta,h} = -0.5$). In (a), (b), (c), and (d), the actual target range is 30 m ($n_{\Delta,h} = 200$), whereas the actual target range is 153.60 m ($n_{\Delta,h} = 1024$) in (e), (f), (g) and (h). Since SM-OCDM yields processing gain around 1 dB lower than its counterparts, the magnitudes of the range profiles obtained with each modulation scheme are normalized w.r.t. their respective results at 0 m/s ($k_{\Delta,h} = 0$) for a fairer comparison. © 2023 IEEE. Reprinted, with permission, from [20].

way as for its OCDM-based counterparts, i.e., $k_{\Delta,h} = f_{D,\text{target},h}/\Delta f^{\text{OFDM}}$ and $n_{\Delta,h} = R_h/\Delta R^{\text{OFDM}}$.

The adopted SoC platform was used to emulate both the considered radar systems and transmit power $P_{\text{Tx}} = 0$ dBm, as well as the RTS described in [DKK⁺21]. Assuming a single point target at the range of 30 m ($n_{\Delta,h} = 200$), Figures 2.30(a) to 2.30(d) show range cuts of a radar image containing for target relative radial velocities of 0 m/s ($k_{\Delta,h} = 0$), 92.71 m/s ($k_{\Delta,h} = -0.1$), 231.78 m/s ($k_{\Delta,h} = -0.25$), and 463.56 m/s ($k_{\Delta,h} = -0.5$), respectively. Since

the power allocation strategy described in Section 2.4.3 was adopted, the power allocated to the active, unmodulated radar subchirp is around 3 dB lower than the total transmission power P_{Tx} . Therefore, the presented results for each modulation scheme have normalized amplitudes w.r.t. their corresponding results for zero velocity. It can be observed in these figures that the proposed OCDM-based radar system and its SM-OCDM-based ISAC yield virtually equal radar range profiles. At zero Doppler shift, i.e., $k_{\Delta,h} = 0$, they yield the same range profile result as their OFDM counterpart and a more regular sidelobe trend than the C-OCDM-based ISAC system from [18, 47], whose range sidelobes are significantly influenced by the modulated data onto its subchirps due to its correlation-based range processing. As for $k_{\Delta,h} = -0.1$, which corresponds to the maximum tolerable Doppler shift for OFDM [SW11, NMZ18], the proposed OCDM-based radar and ISAC systems present comparable performance to their OFDM counterpart, while the C-OCDM-based ISAC system from [49] still yields the worse range sidelobe pattern. Relevant differences in the range profiles obtained with the proposed OCDM-based systems w.r.t. OFDM are only observed for $k_{\Delta,h} = -0.25$ and $k_{\Delta,h} = -0.5$, namely, peak splitting and noticeable range-Doppler coupling due to the effects described in (2.33) and (2.34). However, it should be noted that these Doppler shift levels yield very high relative radial velocities and are therefore not to be expected in practical scenarios, being only shown to cover the whole maximum unambiguous velocity range of the considered OCDM-based radar system. Additionally, the combination of these Doppler shifts with the range of 30 m ($n_{\Delta,h} = 200$) leads the proposed OCDM-based radar and SM-OCDM-based ISAC systems into their most critical region in terms of PPLR, PSLR, and ISLR as shown in Figure 2.29. The results for these range and Doppler shift pairs are therefore lower bounds on the radar sensing performance of the proposed OCDM-based schemes and better results are expected for less critical combinations of range and Doppler shift values. As an example of such range and Doppler shift pairs, Figures 2.30(e) to 2.30(g) show range cuts of a radar image containing a single point target at relative radial velocities of 0 m/s ($k_{\Delta,h} = 0$), 92.71 m/s ($k_{\Delta,h} = -0.1$), 231.78 m/s ($k_{\Delta,h} = -0.25$), and 463.56 m/s ($k_{\Delta,h} = -0.5$), respectively, with a range of 153.60 m ($n_{\Delta,h} = 1024$) in all cases. It is worth mentioning that, for this specific range, the SM-OCDM-based ISAC system is not analyzed since it can only support ranges of up to 76.80 m ($n_{\Delta,h} = 512$) due to the limitation imposed by its CP length $N_{\text{CP}}^{\text{OCDM}}$ listed in Table 2.6. A similar behaviour to the OCDM-based radar system would however be expected in case higher $N_{\text{CP}}^{\text{OCDM}}$ was adopted. Since the considered range and

Doppler shift combinations for the aforementioned figures are within the central, cross-shaped region in the unambiguous normalized range-Doppler plane of best PPLR, PSLR, and ISLR performance shown in Figure 2.29, no evident degradation is observed other than the range-Doppler coupling effect for the OCDM-based radar system in Figures 2.30(g) and 2.30(h). Meanwhile, both C-OCDM- and OFDM-based radar systems show increasing irregularity in the range sidelobe pattern and decreasing normalized magnitude for the target's peak along with increasing velocity.

To support the measurement results shown in Figure 2.30, the simulated PPLR, PSLR, and ISLR performances for a single point target over the relative radial velocity range from -463.56 m/s to 463.56 m/s (i.e., $k_{\Delta,h} = 0.5$ to $k_{\Delta,h} = -0.5$) are shown in Figures 2.31(a) to 2.31(c) for 30 m ($n_{\Delta,h} = 200$) and in Figures 2.31(d) to 2.31(e) for 153.60 m ($n_{\Delta,h} = 1024$). In all these figures, it can be observed that the performances of both C-OCDM- and OFDM-based radar systems do not vary with the considered target range, since there is not range-Doppler coupling in these schemes. The OCDM-based radar and SM-OCDM-based ISAC systems, in turn, experience degradation of all three sidelobe parameters for increasing normalized Doppler shift magnitudes at 30 m ($n_{\Delta,h} = 200$) as shown in Figures 2.31(a) to 2.31(c), which mainly happens due to the peak splitting phenomenon. Even so, these two schemes outperform their C-OCDM and OFDM counterparts in terms of ISLR, which measures the overall range sidelobe level also away from the target's peak. As for the results in Figures 2.30(e) to 2.30(h), however, the SM-OCDM-based ISAC system is again not considered for 153.60 m ($n_{\Delta,h} = 1024$) due to the range limitation imposed by its CP length listed in Table 2.6. It can be seen in Figures 2.31(d) to 2.31(f) that the considered OCDM-based radar system suffers negligible degradation of PPLR, PSLR, and ISLR over the whole considered normalized Doppler shift range, always outperforming its C-OCDM and OFDM counterparts.

The results in Figures 2.30 and 2.31 show that, as in any radar modulation scheme, OCDM-based schemes with discrete-Fresnel domain channel estimation experience performance limitations. To support higher normalized Doppler shifts $k_{\Delta,h}$, and therefore higher velocities, than the aimed ones for the adopted HAD-oriented application for any normalized range, one can adjust the parameterization of the OCDM signal aiming to keep the operation points within the central, cross-shaped region of best PPLR, PSLR and ISLR shown in Fig 2.29. This can be achieved, e.g., by reducing the number of subcarriers $N_{\text{subch}}^{\text{OCDM}}$ so

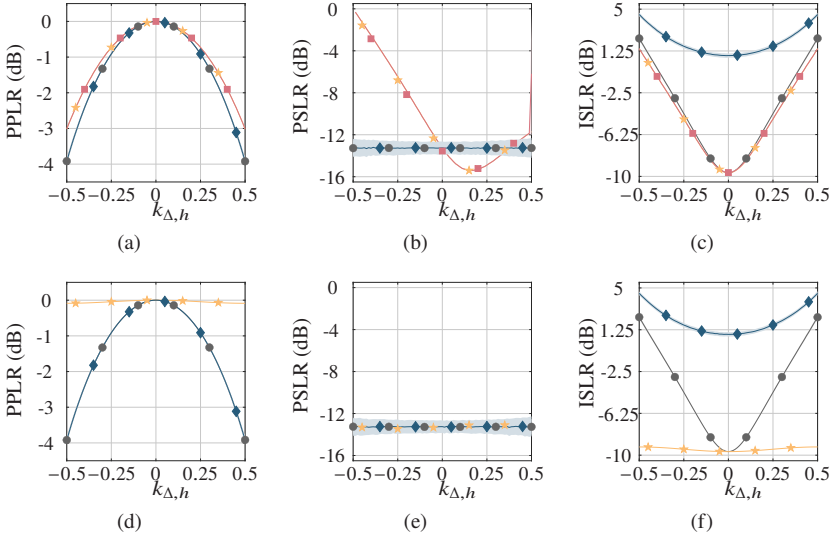


Figure 2.31: Simulated Doppler-shift tolerance comparison among SISO OCDM-based radar system (★), SM-OCDM-based ISAC system (■), C-OCDM-based ISAC system (◆), OFDM-based ISAC system (●): (a) PPLR, (b) PSRL, and (c) ISLR for target range 30 m ($n_{\Delta,h} = 200$), and (d) PPLR, (e) PSRL, and (f) ISLR for target range 153.60 m ($n_{\Delta,h} = 1024$). The continuous lines represent the mean value of the calculated parameters and the shading in the background represents the standard deviation. Adapted from [20], © 2023 IEEE.

that the normalized Doppler shifts $k_{\Delta,h}$ correspond to higher actual Doppler shifts and therefore higher velocities. Another possibility is to use Doppler shift compensation methods such as the all-cell Doppler correction (ACDC), which was originally proposed for OFDM-based radar systems [HY18] but can be applied to OCDM-based systems with little to no adjustments.

To address the FrDM multiplexing strategy for MU or MIMO operation described in Section 2.4.3, similar measurement results to the ones in Figures 2.30(a) to 2.30(d) for a target range of 30 m ($n_{\Delta,h} = 200$) are presented in Figure 2.32 for signals transmitted by $N_{Tx} = 4$ collocated transmitters and evaluated at a single receiver. For the results in this figure, the same OCDM signal parameters adopted for the SISO OCDM-based radar system as men-

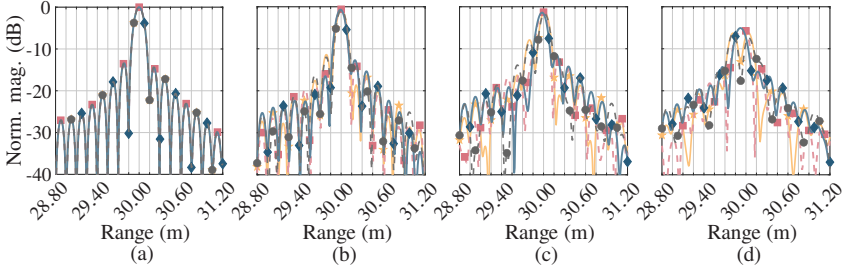


Figure 2.32: Range cuts of radar images from measurements with a simplified baseband setup for transmit channels $n_{Tx} = 0$ (★), $n_{Tx} = 1$ (■), $n_{Tx} = 2$ (◆), and $n_{Tx} = 3$ (●) in the considered MIMO OCDM-based radar system for a single target with relative radial velocity of (a) 0 m/s ($k_{\Delta,h} = 0$), (b) 92.71 m/s ($k_{\Delta,h} = -0.1$), (c) 231.78 m/s ($k_{\Delta,h} = -0.25$), and (d) 463.56 m/s ($k_{\Delta,h} = -0.5$). In all radar images the actual target range is 30 m ($n_{\Delta,h} = 200$). © 2023 IEEE. Reprinted, with permission, from [20].

tioned at the beginning of this section are adopted, leading to the same radar performance parameters listed in Table 2.6. The only exception is the maximum unambiguous range, which is reduced to $R_{\max,ua,FrDM}^{OCDM} = 76.8$ m due to the multiplexing among $N_{Tx} = 4$ transmitters. Compared to the results for the SISO OCDM-based radar system in Figures 2.30(a) to 2.30(d), which are virtually the same as for the transmitter $n_{Tx} = 0$ in the MIMO case, the results from Figure 2.32 only differ significantly for $k_{\Delta,h} = -0.25$ and $k_{\Delta,h} = -0.5$. This is due to the combined effect of the range-Doppler coupling and Doppler dependent phase terms in (2.34), which directly influence the peak splitting and range-Doppler coupling effects. As already mentioned previously, however, such high Doppler shift levels are not to be expected in practice.

Finally, to illustrate the result of the full radar signal processing chain of the considered modulation schemes, Figure 2.33 shows measured range-velocity radar images for four emulated radar point targets by the previously mentioned RTS. The considered targets have RCSs $\sigma_{RCS} = \{30, 25, 30, 27\}$ dBsm, ranges of $\{29.5, 30.3, 31, 32\}$ m, which correspond to the discrete normalized ranges $n_{\Delta,h} = \{196.67, 202, 206.67, 213.33\}$, and velocities of $\{91.15, 93.15, 92.40, 93.65\}$ m/s, which correspond to the discrete normalized Doppler shifts $k_{\Delta,h} = \{-0.0984, -0.1005, -0.0997, -0.1011\}$ and are far above the expected velocities in HAD applications [18]. In all these images, the magnitudes are normalized by the higher target peak from the OCDM

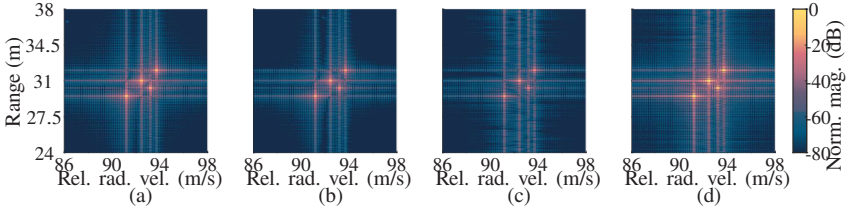


Figure 2.33: Obtained range-velocity radar images from measurements with a simplified baseband setup for (a) OCDM-based radar and (b) SM-OCDM-based ISAC systems with discrete-Fresnel domain channel estimation, as well as (c) C-OCDM- and (d) OFDM-based radar systems for four point targets with RCSs $\sigma_{\text{RCS}} = \{30, 25, 30, 27\}$ dBsm, ranges of $\{29.5, 30.3, 31, 32\}$ m ($n_{\Delta,h} = \{196.67, 202, 206.67, 213.33\}$), and velocities of $\{91.15, 93.15, 92.40, 93.65\}$ m/s ($k_{\Delta,h} = \{-0.0984, -0.1005, -0.0997, -0.1011\}$). Adapted from [20], © 2023 IEEE.

radar image in Figure 2.33(a). It is possible to see that the considered OCDM-based radar and the SM-OCDM ISAC systems present less degraded radar images with more uniform sidelobe levels than both C-OCDM and OFDM radar systems, which agrees with the previous sidelobe level analyses for this velocity range. The fact that the sidelobes for SM-OCDM ISAC systems seem even lower is actually explained by the allocation of only $(N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1)/N_{\text{subch}}^{\text{OCDM}}$ of the total power to the radar subchirp. Alone, this makes the target peaks, and therefore their associated sidelobe level, appear 3 dB lower in the radar image than in the OCDM radar case. Adding the difference in theoretical processing gain listed in Table 2.6, this difference increased to around 4 dB.

2.5.2.2 Communications performance analysis

For an analysis of the communications performance of the proposed SM-OCDM-based ISAC in Section 2.4.3, the parameters listed in Table 2.7 are adopted. For the communications analysis, the proposed SM-OCDM-based ISAC system is compared with its C-OCDM- and OFDM-based counterparts whose parameters are listed in the same table. To enable channel equalization and subsequent extraction of the receive QPSK symbols from the receive signal, the unmodulated subchirp in the SM-OCDM-based

ISAC system is used. Based on the discussion in Section 2.4.3, a total of $N_{\text{mod}}^{\text{SM-OCDM}} = N_{\text{subch}}^{\text{OCDM}} - 2N_{\text{CP}}^{\text{OCDM}} + 1$ modulation symbols is transmitted in an SM-OCDM block (i.e., OCDM symbol plus CP). As for C-OCDM and OFDM, inserted pilot subcarriers [SNZ18] and FSP-inserted pilots [OM22] is used for channel estimation, respectively. A pilot occupancy ratio of 50% for every C-OCDM and OFDM symbols disregarding CP is adopted, which is necessary to support discrete delays of up to $N_{\text{subch}}^{\text{OCDM}}/4$ and $N_{\text{subc}}^{\text{OFDM}}/4$ samples [SNZ18], respectively. The aforementioned discrete delays correspond to the respectively adopted CP lengths $N_{\text{CP}}^{\text{OCDM}}$ and $N_{\text{CP}}^{\text{OFDM}}$, and the assumed pilot occupancy ratio results in $N_{\text{mod}}^{\text{C-OCDM}} = N_{\text{subch}}^{\text{OCDM}}/2$ and $N_{\text{mod}}^{\text{OFDM}} = N_{\text{subc}}^{\text{OFDM}}/2$ for C-OCDM and OFDM, respectively. Assuming a duty cycle of 100%, the achieved communications data rates for all these three modulation schemes assuming QPSK modulation are calculated as $\mathcal{R}_{\text{QPSK}} = 2N_{\text{mod}}/T_{\text{block}}$. Note that the superscripts of the corresponding modulation schemes in $\mathcal{R}_{\text{QPSK}}$, N_{mod} and T_{block} were omitted for conciseness. Since all C-OCDM and OFDM symbols contain pilots, the absolute value of the maximum residual frequency offset that their pilots can estimate is equal to $0.5\Delta f^{\text{OCDM}} = 0.5B^{\text{OCDM}}/N_{\text{subch}}^{\text{OCDM}}$ and $0.5\Delta f^{\text{OFDM}} = 0.5B^{\text{OFDM}}/N_{\text{subc}}^{\text{OFDM}}$ [SNZ18], respectively. If lower frequency offsets are to be estimated, not all symbols must necessarily contain pilots, and the communications data rates of the C-OCDM- and OFDM-based ISAC systems can be improved. It is worth highlighting that such flexible trade-off between data rate and supported delays and frequency offsets is not possible in the SM-OCDM case, which can lead to significantly lower data rates than for C-OCDM and OFDM if other parameterizations than the assumed one are adopted.

Based on the assumed parameters, measurements over a communications channel emulated by re-purposing the previously described RTS were performed with the same SoC platform from Xilinx, Inc, ensuring time and frequency synchronization between the communications transmitter and receiver with the S&C algorithm as described in [7]. The magnitude response of the experienced communications CFR, whose frequency-selectivity is due to the sinc-shaping resulting from the use of an intermediate frequency (IF)-sampling architecture and uncalibrated effects of cables, baluns and filters, is shown in Figure 2.34(a). For all three modulation schemes, multiple channel estimates were averaged to reduce noise effect before performing equalization.

Table 2.7: Adopted signal parameters and resulting communications performance parameters for the considered SM-OCDM-, C-OCDM- and OFDM-based ISAC systems. For conciseness, the superscripts of the individual modulation schemes are omitted in the signal and radar performance parameters. © 2023 IEEE. Reprinted, with permission, from [20].

Signal and communications performance parameters			
	SM-OCDM	C-OCDM	OFDM
Symbol length N	2048	2048	2048
CP length N_{CP}	512	512	512
Block duration T_{block}	2.56 μs	2.56 μs	2.56 μs
Modul. symb. per block N_{mod}	1025	1024	1024
Modulation	QPSK	QPSK	QPSK
Data rate $\mathcal{R}_{\text{QPSK}}$	0.80 Gbit/s	0.80 Gbit/s	0.80 Gbit/s

To assess the communications performance of the considered ISAC systems, error vector magnitude (EVM) is adopted as a metric. One reason for the choice of EVM, e.g., over BER, is that S&C synchronization and channel estimation may not be possible at very low SNR regime, which would only enable a BER analysis for a reduced SNR range. A further reason is that EVM better reflects characteristics of the adopted modulation scheme in terms of amplitude and phase changes [VMR17], while BER only provides information on the ultimately experienced bit errors.

In this context, obtained normalized receive QPSK constellations for a practical SNR level are shown in Figure 2.34(b). Based on the achieved results, both SM-OCDM- and C-OCDM-based ISAC systems yield virtually the same communications performance. While the first has an estimate SNR of 29.65 dB and an EVM with mean value of -26.27 dB and standard deviation of 5.59 dB, an estimated input SNR of 29.74 dB and EVM with mean value of -26.20 dB and standard deviation of 5.53 dB were observed for the latter. As for the OFDM-based ISAC system, an estimated input SNR of 29.65 dB and EVM with mean value of -27.64 dB and standard deviation of 6.30 dB were obtained. The higher mean and value and lower standard deviation of the EVM in the OCDM-based systems can be explained by the fact that the degradation imposed by channel is spread over all subcarriers. Conversely, while a lower EVM mean value is experienced in the OFDM-based ISAC system, its higher standard deviation is due to

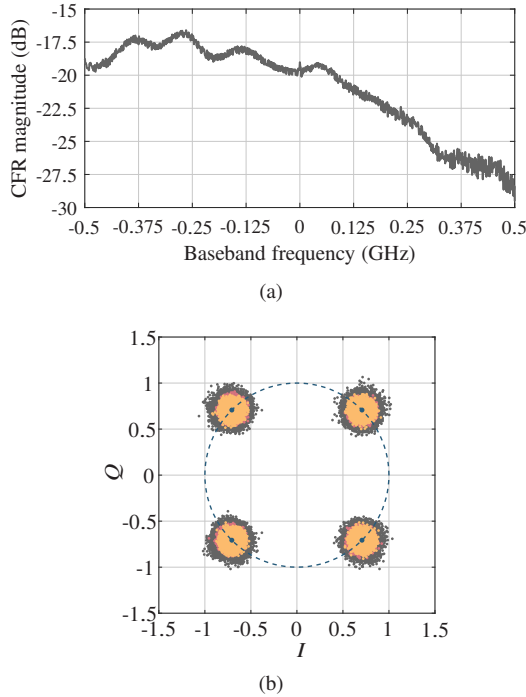


Figure 2.34: Communications measurement results with a simplified baseband setup: (a) magnitude response of the experienced communications CFR and (b) normalized receive QPSK constellations: SM-OCDM-based ISAC system (●), C-OCDM-based ISAC system (●), and OFDM-based ISAC system (●). For reference, an unit-radius circle (—) and a QPSK constellation with unit symbol energy (●) are shown. © 2023 IEEE. Reprinted, with permission, from [20].

different subcarriers experiencing different attenuation levels while propagating through the channel. More specifically, the subcarriers on the rightmost side of the spectrum experience around 10 dB more attenuation than the ones on the leftmost side (see Figure 2.34(a)), being therefore more severely degraded and having their receive QPSK symbols shifted away from the reference constellation in Figure 2.34(b). This behavior agrees with the predicted higher robustness of the OCDM-based ISAC systems in frequency-selective channels reported in the literature and leads to a better overall communications performance com-

pared to their OFDM counterpart in several aspects. Among these aspects are a constant SNR along the subcarriers in OCDM-based schemes [MCBR21], which is due to the fact that channel effects and eventual ISI is spread over all subcarriers and agrees with the more uniform constellation shape for SM-OCDM and C-OCDM in Figure 2.34(b). Meanwhile, OFDM subcarriers experience proportional SNR levels to the subchannels they are transmitted through. As a consequence, OCDM-based schemes tend to outperform their OFDM-based counterparts in terms of BER for practical SNR levels in frequency-selective channels, e.g., as supported by simulations in [OZ16a, OM21, 18] and measurements in a previous study [47]. Assuming that the communications subcarriers in both SM-OCDM and C-OCDM schemes have the same power allocation as considered in this article, they will experience equal SNR levels and deliver virtually the same BER results.

2.5.2.3 Peak-to-average power ratio analysis

Assuming the same symbol lengths N from the two previous subsections, the PAPR performance of the considered modulation schemes is analyzed in this subsection. For this analysis to be compatible to the parameterization from both radar and communications analyses, the PAPRs are calculated only within single OCDM and OFDM symbols, without considering their extended length by CPs. The simulated CCDFs of the BB PAPR for a single OCDM symbol is shown for the investigated OCDM-based radar system, its MU/MIMO variant, and the proposed SM-OCDM-based ISAC system are shown in Figure 2.35(a). For the sake of comparison, the PAPRs of considered OFDM- and C-OCDM-based ISAC systems in the previous radar and communications analysis are also shown as a baseline. As in [18], the PAPR was calculated assuming an oversampling factor of 20 to capture fast variations of time-domain signals. The particularly low BB PAPR for the considered OCDM-based radar system and its MU/MIMO variant is explained by the single active subcarrier in (2.23) and (2.36), which yields a complex exponential signal with a virtually constant envelope in the discrete-time domain after the IDFT in (2.15). As for the SM-OCDM-based ISAC system, the presence of active, modulated subcarriers in (2.39) yields nearly 6 dB higher average PAPR. Compared to the OFDM- and C-OCDM-based ISAC systems, which perform equally, the SM-OCDM-based ISAC system presents reduced PAPR by around 1.1 dB. If lower communications data rates are aimed, this difference can be increased by

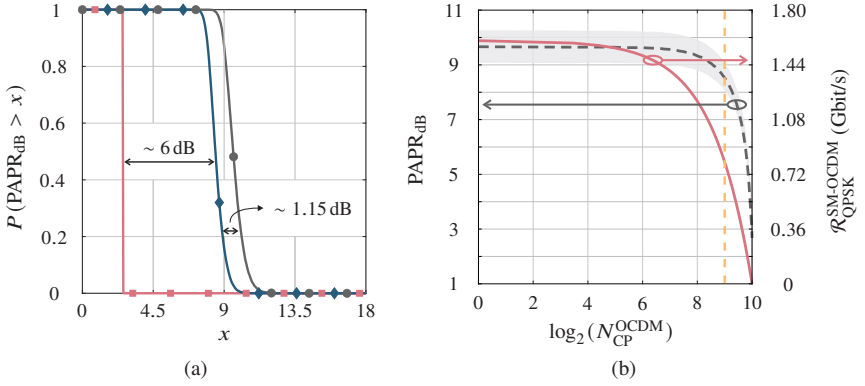


Figure 2.35: PAPR analysis: (a) simulated BB PAPR for the considered OCDM-based radar system and its MU/MIMO variant per transmit channel (■), SM-OCDM-based ISAC system (◆) and an OFDM- and C-OCDM-based ISAC systems (●), all with symbol length of $N = 2048$, and (b) SM-OCDM PAPR (—) and communications data rate (—) as functions of CP length assuming the SM-OCDM symbol structure and power allocation strategy described in Section. 2.4.3. The results for $N_{\text{CP}}^{\text{OCDM}} = 512$, which was adopted in the previous analyses, are highlighted with a yellow dashed line (—). Adapted from [20], © 2023 IEEE.

either deactivating some of the QPSK-modulated subcarriers or replacing them by specific modulation symbols, i.e., performing a similar processing to tone reservation in OFDM-based systems. While the latter approach is discussed, e.g., in [LWJJ18], the former is illustrated in Figure 2.35(b).

2.6 Chapter conclusion

This chapter presented and discussed the PMCW, OFDM, and the OCDM-based C-OCDM and SM-OCDM digital modulation schemes for ISAC. In this context, thorough descriptions of ISAC systems based on the aforementioned modulation schemes as well as of parameters for efficiently assessing both radar and communications aspects of the overall system performance were presented. Finally, a comparative performance analysis based on simulations and measurements with a RTS was presented to validate the presented discus-

sion. This analysis covered radar and communications performance parameter dependencies, a case study on a radar-centric ISAC system parameterization in a mid-range HAD scenario, interference analysis, and considerations on MIMO and multiuser operation.

The first part of the presented comparative performance analysis included only the PMCW, OFDM, and C-OCDM schemes. In this analysis, it was shown that radar-centric ISAC systems based on PMCW can achieve satisfactory radar performance and can be used for reliable, low-data-rate communications in applications such as traffic coordination and interference avoidance among ISAC nodes. Furthermore, it was shown that OFDM achieves considerably higher data rates than PMCW, while also offering simultaneous high unambiguous range and relative radial velocity values. It was also pointed out that the C-OCDM scheme achieves superior communications performance over all its counterparts, while demanding further investigation on alternative radar processing approaches to circumvent the undesired range sidelobes yielded by the correlation-based processing reported in the literature. Additionally, a comparative interference analyses focusing on radar and communications and addressing particular aspects of all considered was performed, which indicated the need for multiplexing scheme to avoid dynamic range reduction. Finally, it was shown that multiplexing schemes based on OFDM and C-OCDM tend to achieve overall better performance than their PMCW counterparts in terms of channel isolation in MIMO or multiuser operation due to multicarrier nature of the two first modulation schemes, which yields higher waveform orthogonality and enables more effective multiplexing strategies.

In the second part of the comparative performance analysis, the focus was shifted towards discrete-Fresnel domain radar channel estimation, which relies on a strategic subchirp allocation for OCDM symbols in the discrete-Fresnel domain and exploits the pulse-compression-like effect of the DF_nT. In this context, an OCDM-based radar-only system and the proposed SM-OCDM-based ISAC system based on this principle were compared with their counterparts based on OFDM and C-OCDM. The achieved results showed that the investigated discrete-Fresnel domain channel estimation for OCDM-based radar and SM-OCDM-based ISAC systems yields comparable or improved performance to C-OCDM- and OFDM-based radar/ISAC systems for the Doppler shift range associated with the radial relative velocity range of interest in all its forms, i.e., SISO and MIMO operations or sector-modulated ISAC operation. More gen-

erally, this velocity range of interest is located within the central, cross shaped region in the unambiguous normalized range-Doppler plane where the best PPLR, PSLR, and ISLR are achieved when discrete-Fresnel domain channel estimation is performed in OCDM. Significant differences, namely peak splitting and range migration, are only observed for Doppler shifts resulting from relative radial velocities that are outside the aforementioned region and therefore not to be expected in practice. Compared to the C-OCDM-based ISAC system, it was shown that all of the proposed strategies with discrete-Fresnel domain radar channel estimation yield improved radar sensing performance as their range sidelobes are not influenced by modulated data symbols onto subchirps. Additionally, the proposed strategies for SISO and MIMO OCDM-based radar sensing yield significantly lower PAPR than their OFDM and C-OCDM counterpart. As for the SM-OCDM-based ISAC system, the experienced PAPR reduction is lower, but can be improved in exchange for data rate reduction. For radar-centric systems, where very high data rates are not necessarily a requirement, the proposed SM-OCDM-based ISAC system also appears as an attractive alternative to its OFDM and C-OCDM counterparts, since it shows robustness for both radar sensing and communications.

3 Bistatic OFDM-based ISAC

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As discussed in Chapter 2, digital modulation schemes such as PMCW and OCDM can offer advantages in terms of radar sensing and communications performance, respectively, with respect to OFDM. In the context of communications-centric ISAC, however, the latter tends to be preferred. The main reason for that is its compatibility to cellular standards such as 5G NR and future 6G standards to be proposed by 3GPP. As opposed to the aforementioned alternative schemes, OFDM allows straightforward control of spectral content, therefore enabling more flexible waveform design and lowering efforts in complying to resource allocation requirements and spectral emission regulations. Out of these reasons, only OFDM will be henceforth considered and variable superscripts will be omitted for conciseness.

To exploit the inherently distributed nature of cellular networks, enabling bistatic ISAC becomes a task of high importance. In the bistatic architecture, transmitter and receiver are not co-located but rather widely separated [KPA⁺23]. In addition to enabling greater diversity in sensing measurements [TD23, BMH⁺23] and being a key building block of multistatic ISAC, this architecture avoids challenges related to full-duplex operation that are required in the monostatic case considered in Chapter 2 [BRT⁺19, BLH⁺21]. In spite of the aforementioned advantages, bistatic ISAC requires synchronization between transmitter and receiver to avoid biases to target parameter estimates and overall communications and radar sensing performance degradation.

In this context, this chapter presents a system concept and mathematical model in Section 3.1, and two signal processing strategies for bistatic OFDM-based ISAC in Section 3.2. To address the case where no hardware level synchronization between transmitter and receiver is possible, synchronization offsets

are discussed and an over-the-air (OTA) synchronization procedure is proposed in Section 3.3. Next, relevant hardware impairments are described, and countermeasures either at the system design phase or later compensated via signal processing are discussed in Section 3.4. To conclude, the chapter is summarized in Section 3.5.

3.1 System model

The considered bistatic ISAC system is constituted by two gNBs with static and known positions as depicted in Figure 3.1. Without loss of generality, scenarios with bistatic ISAC pairs constituted by two user equipments (UEs) or a UE and a gNB could also be considered. The first, gNB #1, acts as a transmitter, while the second, gNB #2, operates in the receiving mode. It is also assumed that there are P propagation paths between the gNBs labeled as $p \in \{0, 1, \dots, P-1\}$. The first, $p = 0$, is a reference path created by the gNBs via beamforming, which enables synchronization and provides a benchmark for bistatic radar sensing following the framework from [50]. This path has length R_0 that results in a propagation delay $\tau_0 = R_0/c_0$, where c_0 is the speed of light in vacuum, and Doppler shift $f_{D,0} = 0$ Hz since the gNBs are static. While this reference is assumed to be a line-of-sight (LoS) path in Figure 3.1, any dominant path, including non-line-of-sight (NLoS) ones, with constant range, Doppler shift, and angle could be used for the same purpose. If the reference path is a non-dominant NLoS one, the information that its aforementioned parameters are constant can be used to help distinguishing it from the paths associated with radar targets as discussed in [PLA⁺24], therefore enabling the compensation for any eventual synchronization and target parameter estimation biases caused by a dominant radar target. Beams are also pointed from both gNB #1 and gNB #2 towards point targets that are associated with the remaining $P-1$ propagation paths. It is assumed that the p th path has length $R_p = R_{\text{Tx-T},p} + R_{\text{T-Rx},p}$, where $R_{\text{Tx-T},p}$ and $R_{\text{T-Rx},p}$ are the ranges of the associated point target with this path w.r.t. gNB #1 and gNB #2, respectively. Due to the propagation through this path, a delay $\tau_p = R_p/c_0$ is experienced. In addition, the movement of the point target associated with the p th path results in a Doppler shift $f_{D,p} = (2v_p/\lambda_0) \cos(\gamma_p) \cos(\beta_p/2)$, where v_p is the magnitude of the three-dimensional velocity vector, γ_p is the aspect angle referenced to the bistatic bisector, and β_p is the bistatic angle, all associated with

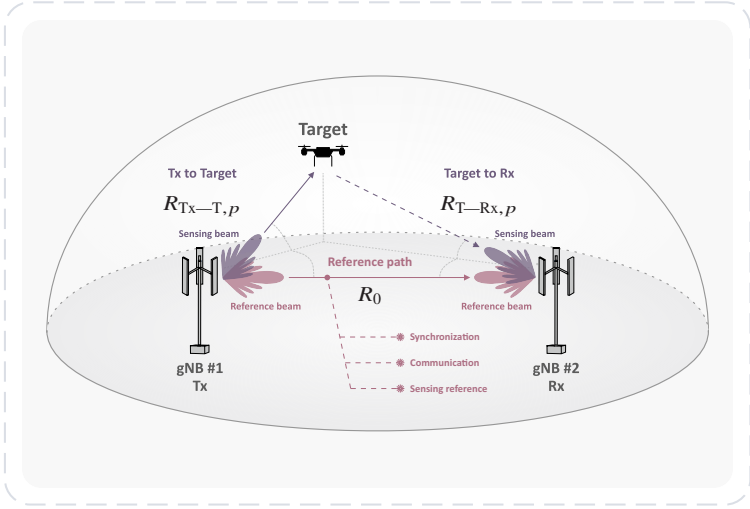


Figure 3.1: Bistatic ISAC system representation. Adapted from [24], © 2026 IEEE.

the point target [Wil08]. To ensure that targets are correctly illuminated, the beamforming directions at gNB #1 and gNB #2 can be previously estimated via hierarchical beam search [LLW⁺21] or other possibly more efficient strategies such as the extension of the sampling and reconstruction of the angular domain proposed in [MHD23] to the bistatic case [FMH⁺25, FHG⁺26]. It is worth highlighting that the assumed beam pointed in the main path direction as well as the ones pointed in the directions of the $P - 1$ point targets by gNB #1 can either be specifically generated for sensing purposes or derived from communications beams that the aforementioned gNB uses to serve UEs. Consequently, no dedicated OTA communications between gNB #1 and gNB #2 must necessarily be performed. For efficient ISAC operation, power allocation among the aforementioned beams can be performed aiming to meet communications and radar sensing performance requirements, e.g., as described in [ZHG⁺19, LRBV21].

At gNB #1, a discrete-frequency domain transmit frame $\mathbf{X} \in \mathbb{C}^{N \times M}$ with $M \in \mathbb{N}_{>0}$ OFDM symbols, each with $N \in \mathbb{N}_{>0}$ subcarriers, is generated. As depicted in Figure 3.2, it is assumed that this frame contains pilots spaced

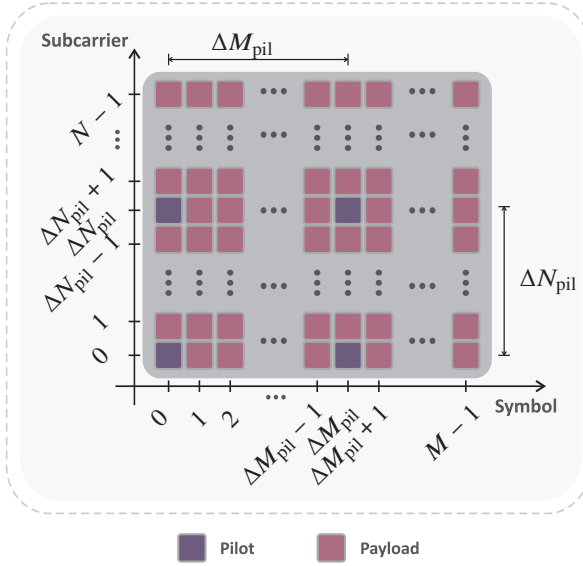


Figure 3.2: Structure of the OFDM frame $\mathbf{X}$ in the discrete-frequency domain. © 2025 IEEE. Reprinted, with permission, from [22].

by $\Delta N_{\text{pil}} \in \mathbb{N}_{>0}$ subcarriers and $\Delta M_{\text{pil}} \in \mathbb{N}_{>0}$ OFDM symbols in the OFDM frame. This leads to a total of

$$M_{\text{pil}} = \left\lfloor \frac{M}{\Delta M_{\text{pil}}} \right\rfloor + 1. \quad (3.1)$$

OFDM symbols containing

$$N_{\text{pil}} = \left\lfloor \frac{N}{\Delta N_{\text{pil}}} \right\rfloor + 1, \quad (3.2)$$

pilot subcarriers each. Each OFDM symbol of $\mathbf{X}$ undergoes an IDFT and has a CP of length $N_{\text{CP}} \in \mathbb{N}_{\geq 0}$ prepended to it. The resulting discrete-time domain sequence $x[s] \in \mathbb{C}$, $s \in \{0, 1, \dots, M(N + N_{\text{CP}}) - 1\}$, from P/S conversion on $\mathbf{X}$ then undergoes D/A conversion with sampling rate F_s , whose corresponding sampling period is $T_s = 1/F_s$. This results in a baseband transmit signal $x(t) \in \mathbb{C}$ that occupies a bandwidth $B \leq F_s$. Consequently, the subcarrier spac-

ing in the frequency domain is $\Delta f = B/N$. After conditioning by the AFE, transmission at the carrier frequency $f_c \gg B$, and reception at gNB #2, the baseband receive signal $y(t) \in \mathbb{C}$ at the receiving ISAC node in the considered system is expressed as

$$y(t) = \alpha_0 x(t - \tau_0) + \sum_{p=1}^{P-1} \alpha_p x(t - \tau_p) e^{j2\pi f_{D,p} t}. \quad (3.3)$$

In this equation, the attenuation factor α_0 , which accounts for all losses in the reference path from gNB #1 to gNB #2, is given by

$$\alpha_0 = \sqrt{\frac{G_{\text{Tx}} G_{\text{Rx}} \lambda_0^2}{(4\pi)^2 R_0^2}}, \quad (3.4)$$

where G_{Tx} and G_{Rx} are the combined single-antenna and beamforming gains at gNBs #1 and #2, respectively. In addition, $\lambda_0 = c_0/f_c$ is the wavelength associated with the carrier frequency f_c . The attenuation factor α_p for the p th propagation path, $p \in \{1, \dots, P-1\}$, associated with a point target can be expressed as

$$\alpha_p = \sqrt{\frac{G_{\text{Tx}} G_{\text{Rx}} \sigma_{\text{RCS},p} \lambda_0^2}{(4\pi)^3 R_{\text{Tx-T},p}^2 R_{\text{T-Rx},p}^2}}. \quad (3.5)$$

In this equation, $\sigma_{\text{RCS},p}$ denotes the bistatic RCS of the point target associated with the p th path, which is aspect-dependent. In other words, $\sigma_{\text{RCS},p}$ depends on the angle from which the target is illuminated by gNB #1 and the angle to which reflections are directed towards gNB #2.

Next, it is assumed that the baseband signal $y(t)$ is sampled by an ADC at gNB #2. Since the signal components traveling through the reference path tends to be much less attenuated than the ones originated by radar targets as described by their attenuation factors in (3.4) and (3.5), its expected power must be carefully considered during link budget calculations [NMZ18, BDR20], as it will likely be the one driving the ADC dynamic range and therefore the capability to detect weak target reflections [18]. However, it is important to note that, due to the significant path loss experienced, the limitation imposed by the reference path is expected to be less critical than the spillover between the transmitter and receiver in monostatic deployments [MHBW23, SHUK23], and

thus manageable. Considering that the aforementioned sampling is performed with the same sampling frequency F_s as for the DAC at the transmitter side at the time instants sT_s , the sequence $y[s] \in \mathbb{C}$ is produced. This relationship of $y[s]$ and $y(t)$ can be mathematically described as

$$y[s] = y(t) \Big|_{t=sT_s}. \quad (3.6)$$

It is further assumed that the length N_{CP} of the CP prepended to the OFDM symbols at the transmitter side is sufficient to avoid ISI due to propagation. In addition, perfect synchronization w.r.t. the reference path and absence of hardware impairments are assumed for simplicity. Consequently, a receive OFDM frame $\mathbf{Y} \in \mathbb{C}^{N \times M}$ can be formed by removing the CP and performing a DFT on each OFDM symbol extracted from the samples of $y[s]$. The element $Y_{k,m} \in \mathbb{C}$ at the k th row, $k \in \{-N/2, 1, \dots, N/2 - 1\}$, and m th column, $m \in \{0, 1, \dots, M - 1\}$, of $\mathbf{Y}$ can be expressed as [50]

$$Y_{k,m} \approx \alpha_0 X_{k,m} + \sum_{p=1}^{P-1} \alpha_p X_{k,m} e^{-j2\pi k \Delta f (\tau_p - \tau_0)} e^{j2\pi f_{D,p} [m(N+N_{\text{CP}}) + N_{\text{CP}}] / B}. \quad (3.7)$$

In this equation, $X_{k,m} \in \mathbb{C}$ is the element at the k th row and m th column of the transmit OFDM frame in the discrete-frequency domain $\mathbf{X}$.

Focusing on the radar sensing functionality of the bistatic OFDM-based ISAC system, a range-Doppler shift radar image is finally obtained after further processing on $\mathbf{Y}$. In this radar image, the reference path will appear as a dominant target with zero range and zero Doppler shift and an SNR equal to

$$\text{SNR}_{\text{image},0} = \frac{P_{\text{Tx}} \alpha_0^2 G_p}{k_B B T_{\text{therm}} \text{NF}} = \frac{P_{\text{Tx}} G_{\text{Tx}} G_{\text{Rx}} \lambda_0^2 G_p}{(4\pi)^2 R_0^2 k_B B T_{\text{therm}} \text{NF}}. \quad (3.8)$$

As for the p th path, $p \in \{1, \dots, P - 1\}$, it will appear at its relative bistatic range $R_p - R_0$ and absolute Doppler shift $f_{D,p}$, as gNBs #1 and #2 are assumed to be static. In addition, it will have an image SNR given by

$$\text{SNR}_{\text{image},p} = \frac{P_{\text{Tx}} \alpha_p^2 G_p}{k_B B T_{\text{therm}} \text{NF}} = \frac{P_{\text{Tx}} G_{\text{Tx}} G_{\text{Rx}} \sigma_{\text{RCS},p} \lambda_0^2 G_p}{(4\pi)^3 R_{\text{Tx}-\text{T},p}^2 R_{\text{T}-\text{Rx},p}^2 k_B B T_{\text{therm}} \text{NF}}. \quad (3.9)$$

As described in [JO06], surfaces of constant $\text{SNR}_{\text{image},p}$ in space are described by Cassini ovals with gNBs #1 and #2 as foci, e.g., as illustrated in Figure 3.1. On these ovals, the product $R_{\text{Tx}-\text{T},p} R_{\text{T}-\text{Rx},p}$ of the ranges between transmitter and target and receiver and target is constant.

Apart from the dynamic range limitation imposed by the reference path and further impairments, the detectability of the remaining $P - 1$ targets will depend on their image SNR described by this equation. Fixing the remaining parameters, the maximum product $(R_{\text{Tx}-\text{T},p} R_{\text{T}-\text{Rx},p})^{\max}$ that can be tolerated to achieve a minimum required image SNR, henceforth denoted as $\text{SNR}_{\text{image}}^{\min}$, for target detection can be defined. Typically, $\text{SNR}_{\text{image}}^{\min} \approx 17$ dB is assumed [MHBW23]. Based on (3.9), this relationship can be expressed as

$$(R_{\text{Tx}-\text{T},p} R_{\text{T}-\text{Rx},p})^{\max} = \sqrt{\frac{P_{\text{Tx}} G_{\text{Tx}} G_{\text{Rx}} \sigma_{\text{RCS},p} \lambda_0^2 G_{\text{p}}}{(4\pi)^3 k_{\text{B}} B T_{\text{therm}} \text{NF} \text{SNR}_{\text{image}}^{\min}}}. \quad (3.10)$$

To concretely define the maximum detectable range, the processing gain G_{p} that is defined adopted bistatic radar signal processing must be known. In addition, further impairments arising from hardware non-idealities must be characterized, as they may result in signal-to-interference-plus-noise ratio (SINR) reduction and therefore further limitation of detection capability. Those aspects are addressed in the next subsections.

3.2 Signal processing chain

In a bistatic OFDM-based ISAC system, the same transmitter processing chain as for the monostatic case in Section 2.3.2 is adopted. At the receiver side, bistatic radar signal processing is performed on the discrete-frequency domain frame $\mathbf{Y}$ obtained at gNB #2, whose element at the n th row and m th column are described by (3.7). This, however, requires either partial or full knowledge of the transmit frame $\mathbf{X}$ by gNB #1 [50]. The processing steps for the two aforementioned approaches are described in Sections 3.2.1 and 3.2.2, respectively. Finally, additional remarks to bistatic radar signal processing in general are given in Section 3.2.3.

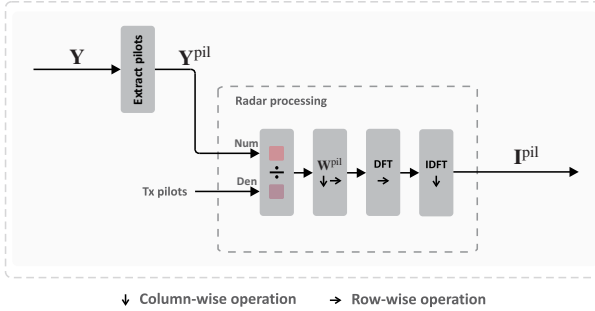


Figure 3.3: Bistatic OFDM-based ISAC receiver signal processing chain with pilot-based sensing.

3.2.1 Pilot-based bistatic radar signal processing

In the first considered bistatic radar signal processing approach, it is assumed that only known pilot subcarriers originally inserted in $\mathbf{X}$ as shown in Figure 3.2 are used for radar signal processing. In this sense, the subcarriers at the corresponding positions of the pilots are extracted from $\mathbf{Y}$, generating a frame $\mathbf{Y}^{\text{pil}} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$ that then undergoes division by the originally transmitted pilot subcarriers. Afterwards, similar processing as for the monostatic case described in Section 2.3.2 is performed. This processing starts with a two-dimensional windowing for sidelobe suppression in the radar image, which is performed via element-wise multiplication with the window matrix $\mathbf{W}^{\text{pil}} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$. Next, row-wise DFTs for Doppler shift estimation, and column-wise IDFTs to estimate time delay, which is ultimately translated into bistatic range, are respectively performed. The equivalent block diagram for the described processing gain for the pilot-based bistatic radar signal processing chain is shown in Figure 3.3. Following [SSBZ13, 50], the aforementioned processing steps in the bistatic OFDM-based ISAC system with pilot-based bistatic radar signal processing yield the range-Doppler shift radar image $\mathbf{I}^{\text{pil}} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$. This radar image is associated with a processing gain $G_{\text{p}}^{\text{pil}}$ and is associated with bistatic range resolution ΔR^{pil} , maximum unambiguous bistatic range $R_{\text{max,ua}}^{\text{pil}}$, and maximum ISI-free bistatic range $R_{\text{max,ISI}}^{\text{pil}}$, as well as Doppler shift resolution, $\Delta f_{\text{D}}^{\text{pil}}$, maximum unambiguous Doppler shift $f_{\text{D,max,ua}}^{\text{pil}}$, and maximum ICI-free Doppler shift $f_{\text{D,max,ICI}}^{\text{pil}}$ as shown in Table 3.1.

3.2.2 Full-frame bistatic radar signal processing

The second approach is based on full knowledge of the transmit frame $\mathbf{X}$ by gNB #1 for bistatic radar signal processing at gNB #2. Since no dedicated OTA communications between the aforementioned gNBs must necessarily take place via the reference path as previously discussed in Section 3.1, the estimation of $\mathbf{X}$ can happen in two ways. It is worth, however, highlighting that these possibilities will not necessarily be the recommended ones in ISAC standardization, which still has to be defined by the 3GPP. The first is through a parallel link, e.g., via the Xn interface of the next generation radio access network (NG-RAN) architecture [3GP18], which can either be used to transmit information on the transmit frame from gNB #1 to gNB #2 [SLF24] or to transmit information on both the transmit frame from gNB #1 and the receive signal at gNB #2 to a third node containing a sensing processing unit [WGM⁺23]. A second possibility, which will be henceforth assumed, is to rely on sufficient SNR at the reference path and perform communications signal processing on the received signal at gNB #2 to estimate the content of the subcarriers in the transmitted frame by gNB #1. In this approach, virtually error-free communications are desired to avoid biasing the radar sensing results, which can be accomplished via channel coding [MKJ⁺24]. The fact that there is no dedicated OTA communications link between the two gNBs prevents any processing relying on pilot subcarriers from being based on resources such as the DMRS [KGBR21]. Although it is present in sufficient density in the OFDM frame and it does not introduce communications overhead, the beamforming coefficients associated with the DMRS defined at gNB #1 may be different across the subcarriers and not known at the receiver in gNB #2 due to UE-specific precoding. An alternative solution is to reuse PRS [DSM⁺21, WWM⁺23] subcarriers, as they adopt known, constant beamforming coefficients.

Following the second approach to estimating the transmit frame $\mathbf{X}$, the processing chain depicted in Figure 3.4 is adopted. It starts by performing ZF FDE on the discrete-frequency domain receive OFDM frame $\mathbf{Y}$ via element-wise multiplication with the matrix $\mathbf{G} \in \mathbb{C}^{N \times M}$ that contains the multiplicative inverse of the estimate of the experienced communications CFR matrix. The resulting matrix then undergoes P/S conversion, and the output modulation symbol stream is then demodulated and the obtained bits decoded. This results in the receive bit stream $\mathbf{b}_{\text{Rx}} \in \{0, 1\}^{N_b^M N M \times 1}$, where $N_b^M = \log_2(\mathcal{M})$ is the number of bits per symbol of the digital modulation constellation of size $\mathcal{M} \in \mathbb{N}_{>0}$. Under

error-free communications conditions, $\mathbf{b}_{\text{Rx}}$ is a perfect estimate of the transmit bit stream $\mathbf{b}_{\text{Tx}} \in \{0, 1\}^{N_b^M N M \times 1}$. Next, an estimate of the transmit frame $\mathbf{X}$ is produced by re-encoding $\mathbf{b}_{\text{Rx}}$ and re-modulating the resulting bit stream into the receive modulation symbol stream $\mathbf{d}_{\text{Rx}} \in \mathbb{C}^{NM \times 1}$, which should ideally be a perfect estimate of the transmit modulation symbol stream $\mathbf{d}_{\text{Tx}} \in \mathbb{C}^{NM \times 1}$. $\mathbf{d}_{\text{Rx}}$ then undergoes S/P conversion and is used for the symbol-division processing to estimate the radar channel CFR as performed in both the pilot-based bistatic case in Section 3.2.1 and the monostatic case described in Section 2.3.2. The resulting matrix from the aforementioned division operation then undergoes the traditional OFDM-based radar signal processing, i.e., windowing for sidelobe suppression via element-wise multiplication with the window matrix $\mathbf{W}^{\text{full}} \in \mathbb{C}^{N \times M}$, DFT for Doppler shift estimation, and IDFT for estimation of delay and consequently bistatic range.

The described bistatic radar signal processing based on full knowledge of the transmit frame ultimately yields the range-Doppler shift radar image $\mathbf{I}^{\text{full}} \in \mathbb{C}^{N \times M}$. This radar image is associated with a processing gain G_p^{full} and is associated with bistatic range resolution ΔR^{full} , maximum unambiguous bistatic range $R_{\text{max,ua}}^{\text{full}}$, and maximum ISI-free bistatic range $R_{\text{max,ISI}}^{\text{full}}$, as well as Doppler shift resolution, Δf_D^{full} , maximum unambiguous Doppler shift $f_{D,\text{max,ua}}^{\text{full}}$, and maximum ICI-free Doppler shift $f_{D,\text{max,ICI}}^{\text{full}}$ as listed in Table 3.1.

3.2.3 Remarks on the signal processing chain

In this section, relevant aspects regarding the signal processing chain of the considered bistatic OFDM-based ISAC system are discussed. These include differences between processing approaches presented in Sections 3.2.1 and 3.2.2, cancellation of the influence of the reference path and interferers on the obtained bistatic radar images, and the influence of synchronization offsets and hardware impairments on the bistatic OFDM-based ISAC system performance.

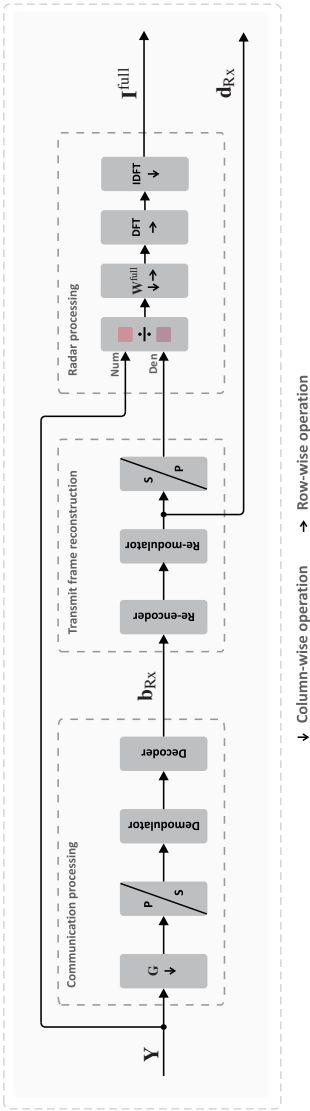


Figure 3.4: Bistatic OFDM-based ISAC receiver signal processing chain with sensing based on full transmit frame reconstruction.

Table 3.1: Performance parameters in bistatic OFDM-based ISAC systems with pilot-based and full-frame radar signal processing.

Radar performance parameters		
	Pilot-based	Full-frame
Processing gain	$G_p^{\text{pil}} = N_{\text{pil}} M_{\text{pil}}$	$G_p^{\text{full}} = N M$
Bistatic range resolution	$\Delta R^{\text{pil}} = c_0 / B$	$\Delta R^{\text{full}} = c_0 / B$
Max. ua. bistatic range	$R_{\text{max,ua}}^{\text{pil}} = N_{\text{pil}} c_0 / B$	$R_{\text{max,ua}}^{\text{full}} = N c_0 / B$
Max. tol. bistatic range	$R_{\text{max,ISI}}^{\text{pil}} = N_{\text{CP}} c_0 / B$	$R_{\text{max,ISI}}^{\text{full}} = N_{\text{CP}} c_0 / B$
Doppler shift resolution	$\Delta f_D^{\text{pil}} = B / [(N + N_{\text{CP}}) M]$	$\Delta f_D^{\text{full}} = B / [(N + N_{\text{CP}}) M]$
Max. ua. Doppler shift	$f_{D,\text{max,ua}}^{\text{pil}} = B / \lfloor 2\Delta M_{\text{pil}} (N + N_{\text{CP}}) \rfloor$	$f_{D,\text{max,ua}}^{\text{full}} = B / \lfloor 2 (N + N_{\text{CP}}) \rfloor$
Max. tol. Doppler shift	$f_{D,\text{max,ICI}}^{\text{pil}} = B / (10N)$	$f_{D,\text{max,ICI}}^{\text{full}} = B / (10N)$

3.2.3.1 Differences between bistatic radar signal processing approaches

The listed radar performance parameters in Table 3.1 show that the full-frame bistatic radar signal processing leads to higher maximum unambiguous range, maximum unambiguous Doppler shift and processing gain than in the pilot-based case as all subcarriers are used for radar signal processing. The occurrence of communications errors in spite of channel coding can, however, lead to artifacts of the obtained radar images in the full-frame approach, therefore significantly degrading the radar sensing performance. To avoid the aforementioned artifacts, bit or modulation symbol interleaving, which are commonly performed in OFDM-based communications systems, can be performed. As a consequence, the sole effect of communications errors is an increased noise plus interference floor in the radar image. It was shown in [6] that for a symbol error ratio (SER) of around 0.07 after re-encoding and re-modulation, which is associated with a considerably higher SER before demodulation, the full-frame approach results in higher image SINR than its pilot-based counterpart for the assumed set of parameters. Out of this reason, the pilot-based approach will henceforth be disregarded.

3.2.3.2 Reference path and interference cancellation

A further important point in the assumed bistatic OFDM-based ISAC system is that the reference path may mask weak target reflections in the ultimately obtained radar image in case it is a dominant path in terms of power. To solve this issue, either windowing with sufficient sidelobe suppression or target elimination techniques such as the ones described in [FKGV24] or [KHH24a] can be used. In addition, in case uplink signals from UEs in the cell of gNB #2 generate enough interference to degrade either the communications or the sensing performance of the OFDM-based bistatic ISAC system constituted by gNB #1 and gNB #2, measures such as the successive interference cancellation described in [SNZ18], on which the aforementioned approach from [KHH24a] is based, can be taken.

3.2.3.3 Synchronization offsets and hardware impairments

In the presented system model in Section 3.2, perfect synchronization between gNBs #1 and #2 and absence of hardware impairments were assumed, which resulted in the baseband receive signal expressed as in (3.3) and ultimately in the discrete-frequency domain receive OFDM frame $\mathbf{Y}$, whose elements are expressed as in (3.7). In practical deployments, however, synchronization is required to enable correctly detecting and evaluating the receive signal before processing, besides avoiding sensing bias. In studies such as [PLA⁺24] and [TMM⁺24], the need for estimating and correcting time and frequency offsets in bistatic ISAC and networked radar systems based on OFDM are discussed. The aforementioned synchronization can be performed, e.g., using the reference path and relying on fully known preamble OFDM symbols [SC97] or, as in the 5G NR case, partly known preamble OFDM symbols such as the ones in synchronization signal blocks (SSBs) [OSAA19]. The SFO, which is a further relevant synchronization offset resulting from slightly different sampling frequencies at the transmitter and receiver, has however not been considered. Further RF and BB hardware impairments known to OFDM-based communications systems are also experienced. These include antenna and array non-idealities, amplifier and mixer non-linear distortions, mixer I/Q imbalance, PN, as well as quantization, clipping, image spectra and aliasing, besides SJ in DACs and ADCs. While these impairments have been widely studied in the context of OFDM-based communications [WWD⁺16, ZZG⁺16], further analysis that also accounts for radar sensing is necessary in the ISAC context. Such analysis can help mitigate performance degradation in both communications and radar sensing functions of the considered bistatic OFDM-based ISAC system, either through compensation techniques or by minimizing the effects during system design. In this sense, Section 3.3 discusses the sources and effects of synchronization offsets, while Section 3.4 discusses RF and BB hardware impairments. In both sections, existing countermeasures are discussed or new ones are introduced.

3.3 Synchronization offsets

In order for unbiased target parameter estimation with the processing chains described in Section 3.2 to be performed, coherent processing of the receive

OFDM frame discrete-frequency domain frame $\mathbf{Y}$ obtained at gNB #2 must be performed. To ensure the required coherence, synchronization must be performed. In this context, the estimation and correction of symbol time offset (STO), carrier frequency offset (CFO), and SFO are discussed in Sections 3.3.1, 3.3.2, and 3.3.3, respectively.

3.3.1 Symbol time offset

Due to the mismatch between time references, an STO τ_Δ is experienced. Consequently, the baseband receive signal $y(t)$ at the receiving ISAC node, i.e., gNB #2, previously defined in (3.3) becomes

$$y(t) = \alpha_0 x(t - \tau_0 - \tau_\Delta) + \sum_{p=1}^{P-1} \alpha_p x(t - \tau_p - \tau_\Delta) e^{j2\pi f_{D,p}t}. \quad (3.11)$$

To avoid range bias and ISI due to delays exceeding the CP duration [HYW⁺09], the STO must be estimated and compensated. For that purpose, STO estimation techniques relying on the dominant, reference path can be used. Since, however, gNBs #1 and #2 are initially unsynchronized, techniques to estimate the STO are not be able to distinguish between the delay of the reference path τ_0 and the STO τ_Δ . Instead, their sum is estimated, leading to the STO estimate $\hat{\tau}_\Delta = \tau_0 + \tau_\Delta$. After STO compensation, this will lead to (3.11) being transformed back into (3.3) except for an additional Doppler-shift induced phase. As a result, the STO estimation will provide an estimate of the OFDM frame start sample among the samples in $y[s]$.

It is worth highlighting that while an isolated estimate of the STO is not possible, unbiased relative bistatic range estimates to the reference path can still be obtained as static ISAC nodes with known positions were assumed in Section 3.1. To obtain absolute bistatic range estimates, the known range of the reference path should be simply added to the measured relative ranges.

To obtain an estimate of the STO, the widely-known S&C algorithm is typically used in communications systems [SC97]. Prior to transmitting the OFDM frame depicted in Figure 3.2, it requires the transmission of two preamble OFDM symbols with same number of subcarriers N and CP length N_{CP} . For

STO estimation, only the first symbol is relevant. This symbol has two equal halves in the discrete time domain of length $N/2$. At the receiver side, a sliding window correlation between two consecutive windows of $N/2$ samples each is performed on $y[s]$ as

$$\gamma^{\text{STO}}[s] = \frac{\left| \sum_{\iota=0}^{N/2-1} y^*[s + \iota] y[s + \iota + N/2] \right|^2}{\sum_{\iota=0}^{N/2-1} |y[s + \iota + N/2]|^2}, \quad (3.12)$$

where a magnitude plateau of length equal to or smaller to the CP length N_{CP} is observed. This plateau provides a range of possible start samples which do not result in ISI. Not obtaining an exact start sample estimate could, however, eventually lead to range estimation biases. Another approach is to correlate the incoming sequence $y[s]$ with a reference preamble sequence. While this may provide an OFDM frame start estimate with accuracy at the integer sample level, high CFO or Doppler shifts may lead to peak power loss and sidelobe level increase in the correlation output, preventing the detection of the preamble sequence or biasing the OFDM frame start sample estimate. To avoid these issues, a three-step STO estimation procedure is proposed [50]. It relies on a frequency shift estimate provided by the S&C algorithm which is later described in Section 3.3.2 and consists of:

1. Obtaining a coarse STO estimate $\hat{\tau}_{\Delta}^{\text{coarse}}$ with the S&C algorithm, from which a coarse start point estimate is obtained.
2. Correcting the CFO and Doppler shifts with the frequency shift estimate provided by the S&C algorithm for a reduced sample range around this coarse estimate.
3. Performing a correlation of the resulting samples with one of the preamble OFDM symbols to obtain an accurate STO estimate $\hat{\tau}_{\Delta}$ and therefore a sample-accurate OFDM frame start sample estimate.

Since the proposed procedure has sample-level estimation accuracy, the pilot subcarriers shown in Fig 3.2 can be used to estimate the residual STO at the fractional sample level as described in [Sit17,SNZ18]. This is done by obtaining CIR estimates from the pilots. The delay for the dominant, reference path is then estimated, and the residual STO obtained based on the fact that the reference path should be at zero delay.

It is worth noting that, in spite of an accurate STO and consequently OFDM frame start estimation with respect to the reference path, ISI can still occur due to radar targets exceeding the maximum tolerable bistatic range defined by the CP length N_{CP} listed in Table 3.1. The first consequence is that the target power itself is reduced in the radar image. In addition, if a strong target exceeds this range, the resulting ISI may shadow weaker targets, therefore preventing their detection. In this context, it was shown in [XZZ25] that the symbol-division operation during radar processing discussed in Section 3.2 introduces randomness to the ISI, therefore making the radar processing gain effective against it. Still, ISI avoidance or compensation techniques may be needed when aiming to detect weak targets such as drones.

In this context, [WWS⁺23] proposed an ISI compensation scheme where samples after each OFDM symbol are added to their beginning to increase power aiming to increase the useful signal power and therefore SINR with respect to ISI. In addition, [OG25] presented a sliding window-based approach to extend the maximum tolerable range due to ISI by relying on pilot subcarriers. Both aforementioned techniques are, however, limited to single-target scenarios. To address this gap, alternative techniques able to cover multi-target scenarios were proposed in [GMH⁺25, XZZ25]. These compensation approaches are based on the reconstruction and cancellation of signal contributions of targets causing ISI. Furthermore, conditions on the detectability of low RCS targets were discussed based on provided closed-form expressions for SINR in both studies. While potentially effective against significant ISI from strong targets, the computational complexity of these techniques must also be considered. Alternative approaches, including the use of custom reference signals [TZWQ24] and a flexible extension of the CP length [LXM⁺24], have also been reported. However, these methods may not yet fully comply with existing standards, therefore requiring further adaptation.

3.3.2 Carrier frequency offset

The fact that gNBs #1 and #2 have distinct LOs leads to a CFO f_{Δ} and its associated phase rotation ψ_{Δ} . Disregarding STO for simplicity, the resulting

effect on the baseband receive signal $y(t)$ at the receiving ISAC node transforms (3.3) into

$$y(t) = \alpha_0 x(t - \tau_0) e^{j(2\pi f_\Delta t + \psi_\Delta)} + \sum_{p=1}^{P-1} \alpha_p x(t - \tau_p) e^{j2\pi f_{D,p} t} e^{j(2\pi f_\Delta t + \psi_\Delta)}. \quad (3.13)$$

As previously discussed in Section 3.3.1, the S&C algorithm is used for CFO estimation. Since synchronization in the considered bistatic OFDM-based ISAC system relies on the dominant, reference path and gNBs #1 and #2 are static, the true CFO can be estimated. This is done via a two-step procedure relying on the same preamble symbol used for STO estimation and an additional preamble OFDM symbol that is transmitted right after [SC97]. The first step consists of obtaining a fractional CFO estimate $\hat{f}_\Delta^{\text{frac}} \in [-0.5\Delta f, 0.5\Delta f]$. This is done by evaluating the phase φ^{CFO} that is given by

$$\varphi^{\text{CFO}} = \arg \left(\sum_{\iota=0}^{N/2-1} y^*[s + \iota] y[s + \iota + N/2] \right), \quad (3.14)$$

where $\arg(\cdot)$ denotes the argument or phase of a complex number. Note that the input of $\arg(\cdot)$ in (3.14) is also used to calculate the numerator of (3.12) and is therefore calculated based on the first preamble OFDM symbol. To obtain a fractional CFO estimate, φ^{CFO} is evaluated at the point that is first taken as a coarse start point estimate for the OFDM frame. Once φ^{CFO} is obtained, the fractional CFO estimate $\hat{f}_\Delta^{\text{frac}}$ is calculated as

$$\hat{f}_\Delta^{\text{frac}} = \varphi^{\text{CFO}} / (\pi N T_s). \quad (3.15)$$

Afterwards, $\hat{f}_\Delta^{\text{frac}}$ is corrected for the second preamble OFDM symbol, which is then further processed to obtain an integer CFO estimate $\hat{f}_\Delta^{\text{int}}$. For that purpose, the element at the k th subcarrier of first and second preamble OFDM symbols in the discrete-frequency domain at the receiver side are denoted as $Y_k^{\text{pb},1} \in \mathbb{C}$ and $Y_k^{\text{pb},2} \in \mathbb{C}$. Knowing that

$$D_k^{\text{pb}} = X_k^{\text{pb},2} / X_k^{\text{pb},1}, \quad (3.16)$$

where $X_k^{\text{pb},1}$ and $X_k^{\text{pb},2}$ denote the originally transmit content of the k th sub-carrier of the first and second preamble OFDM symbols, respectively, the parameter $\gamma^{\text{CFO}}[\kappa]$ is calculated via correlation-based processing as

$$\gamma^{\text{CFO}}[\kappa] = \frac{\sum_{k=-N}^N Y_{k+2\kappa}^{\text{pb},1*} D_k^{\text{pb}*} Y_{k+2\kappa}^{\text{pb},2}}{2 \sum_{k=-N}^N |Y_k^{\text{pb},2}|^2}. \quad (3.17)$$

Based on κ that maximizes $|\gamma^{\text{CFO}}[\kappa]|$, an integer CFO estimate $\hat{f}_{\Delta}^{\text{int}}$ such that $(\hat{f}_{\Delta}^{\text{int}}/\Delta f) \in \mathbb{Z}$ can be calculated as

$$\hat{f}_{\Delta}^{\text{int}} = \frac{2}{NT_s} \arg \max_{\kappa} |\gamma^{\text{CFO}}[\kappa]|, \quad (3.18)$$

where $\arg \max_{\kappa}(\cdot)$ returns the value of κ at which the input expression attains its maximum value.

After combining $\hat{f}_{\Delta}^{\text{frac}}$ and $\hat{f}_{\Delta}^{\text{int}}$ into a full estimate as

$$\hat{f}_{\Delta} = \hat{f}_{\Delta}^{\text{frac}} + \hat{f}_{\Delta}^{\text{int}}, \quad (3.19)$$

the CFO can be corrected for the whole receive sequence $y[s]$, and therefore the entire OFDM frame, by multiplying it with a complex exponential associated with the frequency $-\hat{f}_{\Delta}$ in the discrete-time domain, generating a sequence $y^{\text{corr}}[s] \in \mathbb{C}$. Mathematically, this is expressed as

$$y^{\text{corr}}[s] = y[s] e^{-j2\pi \hat{f}_{\Delta} s T_s}. \quad (3.20)$$

It is worth highlighting that this may not compensate for the phase ψ_{Δ} , which, however, results in no impairments for either communications or radar signal processing. As mentioned in Section 3.3.1, the CFO correction is only performed for the samples around the first preamble OFDM symbol for reduced computational complexity. Only after a more accurate estimate of the OFDM frame start sample is obtained, the CFO is corrected for all samples belonging to the frame. As in the STO case from Section 3.3.1, a residual CFO estimation can be performed using the pilot subcarriers shown in Fig 3.2. As described in [Sit17, SNZ18], this is done via the same DFT-based processing used to

estimate Doppler shifts in ISAC systems based on the waveforms described in Chapter 2.

Similarly to the occurrence of ISI despite accurate STO estimation and compensation as described in Section 3.3.1, ICI can also happen even under accurate estimation and compensation of the CFO with respect to the reference path. This occurs due to targets exhibiting high Doppler shifts, leading not only to reduced target power in the radar image but also to an elevated interference floor due to ICI. If the reflections from such targets are sufficiently strong, they may mask weaker ones. In such cases, countermeasures may be required.

In this context, an ICI compensation technique, named ACDC, was proposed in [HY18]. In this technique, linearly-dependent OFDM symbols are transmitted across the frame. This allows compensating for the relationship among OFDM symbols, performing Doppler shift estimation based on DFTs, and then compensating for the frequency shifts introduced by Doppler shifts. Next, each OFDM symbol is transformed into the discrete-frequency domain via DFT, and the resulting symbol undergoes symbol division and IDFT to perform range estimation and finally produce a radar image. While ACDC is capable of compensating for the ICI associated with multiple radar targets, the required linear dependence of OFDM symbols makes it unsuitable for ISAC systems where full communications capability must be preserved.

A further approach to handle ICI was proposed in [KWK21]. In this study, an algorithm capable of detecting targets and estimating their range, Doppler and angle parameters under the effect of ICI was proposed. This algorithm allows using frequency shift estimates to resolve Doppler shift ambiguities, besides allowing to exploit ICI to enable resolving even targets at the same range, Doppler shift, and angle in the radar image. Alternatively, the super-resolution radar image generation algorithm proposed in [SZFL25] can be used. In this approach, the contributions of initially detected targets are reconstructed via two-dimensional (2D) surface fitting and then subtracted from the originally received signal. By doing so, ICI from strong, moving targets is compensated and the effective sensing dynamic range is increased, therefore potentially allowing to detect weaker targets.

While proven effective via simulations, the suitability of the discussed approaches for practical deployments, particularly under high target density and additional impairments, must be carefully assessed. Before resorting to ICI

compensation techniques, it must also be considered that the randomness of transmitted modulation symbols in the OFDM frame ultimately results in random ICI [Sma12]. Consequently, it can be suppressed by the radar processing gain, as is the case with ISI [XZZ25].

3.3.3 Sampling frequency offset

If the adopted sampling rate F_s^{Rx} to perform analog-to-digital conversion on $y(t)$ to produce $y[s]$ at the receiver side is different than F_s^{Tx} , then SFO occurs. The experienced SFO is given by

$$\text{SFO} = F_s^{\text{Rx}} - F_s^{\text{Tx}}, \quad (3.21)$$

and the normalized SFO by the transmit sampling rate F_s^{Tx} is defined as $\delta \in \mathbb{R}$ and expressed as

$$\delta = \frac{\text{SFO}}{F_s^{\text{Tx}}} = \frac{F_s^{\text{Rx}} - F_s^{\text{Tx}}}{F_s^{\text{Tx}}}. \quad (3.22)$$

For a non-zero normalized SFO δ and disregarding STO and CFO for simplicity, $y[s]$ becomes

$$y[s] = y(t) \Big|_{t=sT_s^{\text{Tx}}(1-\delta)}. \quad (3.23)$$

Consequently, the receive OFDM frame $\tilde{\mathbf{Y}}$ becomes $\mathbf{Y}$. For small SFOs that do not lead to ISI, the element $Y_{n,m} \in \mathbb{C}$ at the k th row and m th column of $\mathbf{Y}$ can be expressed as [Sma12, CTL12]

$$Y_{k,m} = \alpha_k^{\text{SFO}} (\tilde{Y}_{k,m}) e^{j\psi_{k,m}^{\text{SFO}}} + \xi_{k,m}^{\text{SFO,ICI}}, \quad (3.24)$$

where $\alpha_k^{\text{SFO}} \in \mathbb{R}_{\geq 0}$ is an amplitude modulation expressed as

$$\alpha_k^{\text{SFO}} = \frac{\sin(\pi\delta k)}{N \sin\left(\frac{\pi\delta k}{N}\right)}, \quad (3.25)$$

$\psi_{k,m}^{\text{SFO}} \in \mathbb{R}$ is a phase rotation expressed as

$$\psi_{k,m}^{\text{SFO}} = -\frac{2\pi\delta k}{N} [m(N + N_{\text{CP}}) + N_{\text{CP}}] - \frac{\pi\delta k}{N} (N - 1), \quad (3.26)$$

and $\xi_{k,m}^{\text{SFO,ICI}} \in \mathbb{C}$ is the ICI term expressed as

$$\xi_{k,m}^{\text{SFO,ICI}} = \sum_{l=-N/2, l \neq k}^{N/2-1} \tilde{Y}_{l,m} \frac{\sin(\pi[(1+\delta)l-k])}{N \sin\left(\frac{\pi[(1+\delta)l-k]}{N}\right)} e^{j2\pi \frac{m(N+N_{\text{CP}})+N_{\text{CP}}}{N} \delta l} \times e^{j\pi \frac{N-1}{N} [(1+\delta)l-k]}. \quad (3.27)$$

Rearranging the first term of (3.26) and resorting to the DFT shift theorem, it is concluded that the SFO causes a delay that is equal to

$$\tau_m^{\text{SFO}} = \delta [m(N + N_{\text{CP}}) + N_{\text{CP}}] T_s \quad (3.28)$$

at the m th OFDM symbol. Since τ_m^{SFO} changes with m , a delay migration defined as

$$\Delta\tau_m^{\text{SFO}} = \tau_m^{\text{SFO}} - \tau_0^{\text{SFO}} = \delta [m(N + N_{\text{CP}})] T_s \quad (3.29)$$

is experienced along the OFDM symbols in $\mathbf{Y}$. Comparing (3.24) with the effect of a CFO on the receive OFDM symbol as described in [Sma12], it is also concluded that the normalized SFO δ causes a frequency shift equal to

$$f_k^{\text{SFO}} = (\delta k) \Delta f \quad (3.30)$$

at the k th subcarrier of every OFDM symbol, which corresponds to the k th row of $\mathbf{Y}$. As f_k^{SFO} is a function of k , a frequency shift migration

$$\Delta f^{\text{SFO}} = f_{N/2-1}^{\text{SFO}} - f_{-N/2}^{\text{SFO}} \quad (3.31)$$

is experienced from the first to the last OFDM subcarrier in each OFDM symbol contained in $\mathbf{Y}$.

By estimating the previously described SFO-induced impairments to the reference path, the SFO itself can be estimated. For that purpose, either preamble OFDM symbols as considered for the STO and CFO cases in Sections 3.3.1 and 3.3.2 as well as in [6, 50], or pilot subcarriers can be used. In both cases, the condition that they are known at the receiver side is assumed. The first approach, however, leads to undesirable overhead, reducing the communications data rate. By using only pilot subcarriers, a possible approach to estimate the SFO is by observing the frequency shift migration along consecutive at pilot

subcarriers along the OFDM symbols. To improve the effective SNR for SFO estimation, the delay migration can be observed instead. This approach consists of obtained CFR estimates from the pilot subcarriers in $\mathbf{Y}$ [HW98, CEPB02], and transforming them into CIR estimates via IDFTs. From these CIR estimates, the delay migration, and consequently, the SFO can be estimated. This approach benefits from a processing gain that is similar to the one experienced during range processing in OFDM-based radar and ISAC systems as discussed in Section 2.3.2 and [18]. Consequently, higher effective SNR and better accuracy for SFO estimation is experienced than in the approach based on frequency shift migration that operates on the subcarrier level.

An SFO estimation approach based on delay migration has been proposed by Wu et al. [WZL12], which however has not provided accuracy bounds to their estimator. In this sense, Section 3.3.3.1 describes an SFO estimation approach based on [WZL12], highlighting its limitations and providing accuracy bounds. To ensure higher robustness at SFO-induced ISI and reduce computational complexity at higher SFOs, Section 3.3.3.2 presents the tilt inference of time offset (TITO) method. Finally, SFO correction strategies to be performed after estimation are discussed in Section 3.3.3.3.

3.3.3.1 Delay migration-based SFO estimation

Assuming sufficient SNR regime, negligible SFO-induced ICI and no SFO-induced ISI at every m th OFDM symbol, one can use the pilot subcarriers in $\mathbf{Y}$ to obtain CIR estimates. For simplicity, the aforementioned pilot subcarriers are henceforth assumed to be uniformly distributed as depicted in Figure 3.2. In practical implementations such as 5G NR, the fact that there is no dedicated OTA communications link between the two gNBs prevents any processing relying on pilot subcarriers from being based on resources such as DMRS [KGBR21]. Although it is present in sufficient density in the OFDM frame and does not introduce communications overhead, the beamforming coefficients associated with DMRS defined at gNB #1 may be different across the subcarriers and not known at the receiver in gNB #2 due to UE-specific precoding. An alternative solution is to repurpose PRS [DSM⁺21, WWM⁺23] subcarriers for SFO estimation, as it adopts known, constant beamforming coefficients. Although SSBs could appear as an alternative, their low density in the OFDM

frame would not allow accurate SFO estimation [PTG⁺24], especially under high SFO regime.

The obtained CIR estimates from the pilot subcarriers in $\mathbf{Y}$ form a delay-slow time profile represented by the matrix $\mathbf{D} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$. To ensure fine delay granularity in the contained CIR estimates in $\mathbf{D}$ and therefore improve the accuracy of estimated delays, the required IDFT processing can be zero padded. The m_{pil} th column, $m_{\text{pil}} \in \{0, 1, \dots, M_{\text{pil}} - 1\}$, of the $\mathbf{D}$ contains one CIR estimate obtained from the N_{pil} pilot subcarriers at the corresponding pilot OFDM symbol. Based on equations (3.1) and (3.2) as in Section 3.1, spacing ΔN_{pil} and ΔM_{pil} between pilot subcarriers in the column and row directions in the OFDM frame is assumed. Since the assumed uniform pilot subcarrier spacing may not be observed in practical deployments, it is also worth highlighting that it is not required for delay migration and consequent SFO estimation.

To estimate the SFO, the proposed method, TITO, is applied to the reference path, which is taken as a benchmark and has its delay progression along the successive CIR estimates contained in the columns of $\mathbf{D}$ analyzed. For that purpose, the delay migration of the aforementioned path is estimated at every m_{pil} th column of $\mathbf{D}$, yielding $\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}}$. Knowing that the correspondence between column indexes from $\mathbf{D}$ and $\mathbf{Y}$ is given by

$$m = m_{\text{pil}} \Delta M_{\text{pil}}, \quad (3.32)$$

it holds that

$$\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}} = \widehat{\Delta\tau}_{m_{\text{pil}} \Delta M_{\text{pil}}}^{\text{SFO}}. \quad (3.33)$$

In this equation, $\widehat{\Delta\tau}_{m_{\text{pil}} \Delta M_{\text{pil}}}^{\text{SFO}}$ denotes an estimate of $\Delta\tau_{m_{\text{pil}} \Delta M_{\text{pil}}}^{\text{SFO}}$. By feeding all M_{pil} delay migration estimates $\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}}$ to a linear regression model based on (3.29), a least squares (LS) estimate $\widehat{\delta}$ of the normalized SFO is finally obtained. This is achieved by taking the partial derivative of the sum of the squared differences to be minimized in

$$\widehat{\delta} = \arg \min_{\delta} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \left(\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}} - \left\{ \delta \left[(m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) \right] T_s \right\} \right)^2 \quad (3.34)$$

with respect to δ , setting it to zero and solving to yield $\widehat{\delta}$, which is expressed as in (3.35).

Due to the linear regression nature of the SFO estimation, the estimate $\widehat{\delta}$ has a standard deviation $\sigma_{\widehat{\delta}}$ given by (3.36). The expression in (3.36) shows that $\sigma_{\widehat{\delta}}$ depends on the standard deviation $\sigma_{\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}}$ of the delay migration estimates at each m_{pil} th column of $\mathbf{D}$. Based on the Cramér–Rao lower bound (CRLB) of delay migration estimation in OFDM-based systems at high SNR [BSJ11], the lower bound for $\sigma_{\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}}$ assuming negligible SFO-induced ICI and ISI is

$$\sigma_{\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}} \geq \sqrt{\frac{6}{\text{SNR}_0(N_{\text{pil}}^2 - 1)N_{\text{pil}}} \left(\frac{1}{2\pi\Delta f\Delta N_{\text{pil}}} \right)^2}. \quad (3.37)$$

The SNR of the reference path in this equation is calculated assuming that the receive signal in the considered bistatic OFDM-based ISAC system is impaired by AWGN. It is equivalent to (3.8) without the radar processing gain, being therefore given by

$$\text{SNR}_0 = \frac{P_{\text{Tx}}\alpha_0^2}{k_{\text{B}} B T_{\text{therm}} \text{NF}}, \quad (3.38)$$

where k_{B} is the Boltzmann constant, T_{therm} is the standard room temperature in Kelvin, and NF is the overall receiver noise figure. Based on (3.36) and (3.37), a lower bound for $\sigma_{\widehat{\delta}}$ at high SNR and still assuming negligible SFO-induced ICI and ISI can be defined as in (3.39). While this lower bound can only be reached by unbiased estimators such as multiple signal classification (MUSIC). In practical deployments, however, if the processing based on zero-padded IDFTs is used as a maximum likelihood estimator (MLE), then bias occurs due to the limited granularity of the axis of possible delays that can be estimated: This axis is solely composed by integer multiples of $1/(\eta B)$, where $\eta \in \mathbb{N}_{>0}$ is the FDZP factor. Consequently, a more accurate lower bound for $\sigma_{\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}}$ at high SNR is given by [BSJ11]

$$\sigma_{\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}} \geq \frac{\sqrt{3}}{6\eta N_{\text{pil}}\Delta f\Delta N_{\text{pil}}}. \quad (3.40)$$

$$\hat{\delta} = \frac{M_{\text{pil}} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \left[(m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right] \widetilde{\Delta \tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}} - \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \left[(m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right] \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \widetilde{\Delta \tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}}}{M_{\text{pil}} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \left[(m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right]^2 - \left[\sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} (m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right]^2} \quad (3.35)$$

$$\sigma_{\hat{\delta}} \approx \frac{\sigma_{\widetilde{\Delta \tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}}}}{\Delta \tau_{\text{pil}, m_{\text{pil}}}} \sqrt{M_{\text{pil}} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} \left[(m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right]^2 - \left[\sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} (m_{\text{pil}} \Delta M_{\text{pil}}) (N + N_{\text{CP}}) T_s \right]^2} \quad (3.36)$$

$$\sigma_{\hat{\delta}} \gtrapprox \sqrt{\frac{\{6/[\text{SNR}(N^2 - 1)N]\}[1/(2\pi\Delta f\Delta N_{\text{pil}})]^2}{M_{\text{pil}} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} [(m_{\text{pil}}\Delta M_{\text{pil}})(N + N_{\text{CP}})T_s]^2 - \left[\sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} (m_{\text{pil}}\Delta M_{\text{pil}})(N + N_{\text{CP}})T_s\right]^2}}. \quad (3.39)$$

$$\sigma_{\hat{\delta}} \gtrapprox \sqrt{\frac{\sqrt{3}/(6\eta N_{\text{pil}}\Delta f\Delta N_{\text{pil}})}{M_{\text{pil}} \sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} [(m_{\text{pil}}\Delta M_{\text{pil}})(N + N_{\text{CP}})T_s]^2 - \left[\sum_{m_{\text{pil}}=0}^{M_{\text{pil}}-1} (m_{\text{pil}}\Delta M_{\text{pil}})(N + N_{\text{CP}})T_s\right]^2}}. \quad (3.41)$$

Consequently, a lower bound for $\sigma_{\hat{\delta}}$ for the case in which $\hat{\delta}$ is obtained based on MLE delay migration estimates $\widehat{\Delta\tau}_{\text{pil},m_{\text{pil}}}^{\text{SFO}}$ can be defined as in (3.41). Since the MLE error is actually uniformly distributed due to the quantized zero-padded IDFT bins, the aforementioned lower bound is only a simplified approximation.

In order for the SFO estimation to work as intended, i.e., without relevant performance degradation or bias, the experienced normalized SFO δ should not induce ICI and ISI that result in relevant degradation of the reference path in the CIRs estimates in the columns of $\mathbf{D}$. Some remarks on these issues are presented as follows.

Remark 1. It is ideally expected that the normalized SFO δ only results in frequency shifts f_k^{SFO} that cause negligible ICI.

Elaboration: To avoid non-negligible ICI that will eventually lead to degradation of the delay and consequently SFO estimation and its subsequent correction, the SFO-induced frequency shift f_k^{SFO} at every k th OFDM subcarriers must be significantly lower than the subcarrier spacing Δf . Assuming the typical frequency shift upper bound for OFDM systems of one tenth of Δf [HYW⁺09, 18],

$$f_n^{\text{SFO}} < \Delta f / 10 \quad (3.42)$$

should hold to result in negligible ICI. Combining (3.42) with (3.30) results in $(\delta k)\Delta f < \Delta f / 10$. Applying (3.42) to the highest absolute SFO-induced ICI, which is experienced at $k = \pm N/2$, results in

$$|\delta| < \delta_{\text{max}}^{\text{ICI}}, \quad (3.43)$$

where

$$\delta_{\text{max}}^{\text{ICI}} = 1/(5N). \quad (3.44)$$

Note that, while subcarrier $k = -N/2$ exists, $k = N/2$ is only approached when performing zero-padding.

Remark 2. It is ideally expected that the normalized SFO δ only results in delays τ_m^{SFO} that do not cause ISI.

Elaboration: To avoid ISI-induced performance degradation of the delay and SFO estimation and therefore of the reflective intelligent surface (RIS)-assisted bistatic ISAC system as a whole, ISI should not happen at any of the OFDM symbols in the frame. If the SFO-induced delay τ_m^{SFO} is negative, ISI will necessarily occur. For positive delays, however, τ_m^{SFO} at all OFDM symbols, regardless of whether they contain pilots or not, should be lower than the CP-free delay $\tau_{\max}^{\text{CP}} = N_{\text{CP}}/B$. Therefore, it should hold that

$$\tau_{\max}^{\text{CP}} > \tau_{m_{\text{pil}}\Delta M_{\text{pil}}}^{\text{SFO}} \geq 0. \quad (3.45)$$

Consequently, the delay τ_{M-1}^{SFO} at the $(M-1)$ th column of $\mathbf{Y}$ should be such that

$$N_{\text{CP}}/B > \tau_{M-1}^{\text{SFO}} \geq 0. \quad (3.46)$$

Based on (3.28), this results in

$$\delta_{\max}^{\text{ISI}} > \delta \geq 0, \quad (3.47)$$

where

$$\delta_{\max}^{\text{ISI}} = \frac{N_{\text{CP}}}{B [(M-1)(N+N_{\text{CP}}) + N_{\text{CP}}] T_s}. \quad (3.48)$$

3.3.3.2 TITO method for ISI-aware SFO estimation

The approximate lower bounds for the SFO estimates in (3.39) and (3.41) so far are only accurate under the assumption of negligible SFO-induced ICI and ISI. The aforementioned ICI is expected to be handled by the previously mentioned processing gain of $10 \log_{10}(N_{\text{pil}})$ dB at the delay migration estimation. The SFO-induced ISI, however, can severely degrade the delay migration estimation [WWS⁺23] and, consequently, the SFO estimation. In order to avoid this, the TITO method for SFO estimation is proposed. This method consists of performing the described SFO estimation based on the slope (or tilt) estimation (or inference) of the delay (or time offset) migration of the reference path along CIR estimates contained in the columns of the CIR estimate matrix $\mathbf{D}$, which ultimately allows obtaining an estimate $\hat{\delta}$ of the normalized SFO based on the knowledge of its relationship with the delay migration described by (3.29). The novelty of TITO compared to approaches such as the one based on the method by Wu et al. [WZL12] lies in the adaptive choice of the number

$M_{\text{pil}}^{\text{TITO}} \leq M_{\text{pil}}$ of columns of $\mathbf{D}$ to be used for delay migration and subsequent SFO estimation with the aim of avoiding using CIR estimates that originate from ISI-contaminated pilot OFDM symbols, which is explained as follows. First, it is important to highlight that ISI only happens at the pilot OFDM symbols of index m_{pil} where the SFO-induced delay $\tau_{m_{\text{pil}}}^{\text{SFO}}$ is either negative or positive and exceeds the CP duration $N_{\text{CP}}T_s^{\text{Tx}}$. In both cases, the ISI-induced SINR degradation of the reference path in the aforementioned CIR estimates depends on the SFO-induced excess delay causing the ISI. Considering these issues, only the $M_{\text{pil}}^{\text{TITO}}$ first columns of $\mathbf{D}$ where tolerable SINR degradation is experienced are used to obtain the required delay migration estimates for the SFO estimation with the TITO method. This also results in lower computational complexity compared with the method by Wu et al. [WZL12] in cases with high SFO, since only a limited number of columns of $\mathbf{D}$ are used for CIR and subsequent SFO estimation. Based on (3.29), $M_{\text{pil}}^{\text{TITO}}$ is chosen such that

$$\begin{aligned}
 M_{\text{pil}}^{\text{TITO}} = \max \quad & m_{\text{pil}} + 1 \\
 \text{s.t.} \quad & \left| \frac{\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}} - \widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}-1}^{\text{SFO}}}{\Delta M_{\text{pil}}(N + N_{\text{CP}})T_s^{\text{Tx}}} \right| \leq (1 + \varepsilon)|\delta_{\text{max}}|.
 \end{aligned} \tag{3.49}$$

In (3.49), $|\delta_{\text{max}}| \in \mathbb{R}_{\geq 0}$ is the maximum absolute normalized SFO (positive or negative) that is expected in the bistatic OFDM-based ISAC system. Furthermore, ε denotes an empirically defined margin that accounts for factors that influence the delay migration estimation inaccuracies, such as the chosen FDZP factor η in the MLE estimation and the SNR. If the constraint in (3.49) is violated, it is considered that the SFO-induced ISI causes significant SINR degradation that in turn leads to non-negligible bias in the delay migration estimates at pilot OFDM symbols of index $m_{\text{pil}} = M_{\text{pil}}^{\text{TITO}}$ or higher. The CIRs contained in $\mathbf{D}$ which correspond to these pilot OFDM symbols are therefore discarded and only the $M_{\text{pil}}^{\text{TITO}}$ first columns of $\mathbf{D}$ are considered for the delay migration estimation and subsequent SFO estimation. While this degrades the accuracy of the SFO estimation according to (3.39) and (3.41), it avoids biasing the SFO estimate $\hat{\delta}$ due to ISI as later shown in the results in Section 4.1.3. Finally, it is worth highlighting that, forcing $M_{\text{pil}}^{\text{TITO}} = M_{\text{pil}}$, the TITO SFO estimation method becomes equivalent to the one by Wu et al. [WZL12].

3.3.3.3 SFO correction

Once the estimate $\hat{\delta}$ has been obtained with (3.35) using a limited number of pilot OFDM symbols defined according to (3.49), resampling is performed on the discrete-time domain samples associated with the receive frame including CP. Typically, a Farrow filter [Far88] with previous cubic interpolation [EGH93] is employed. After resampling, the CP is removed and the OFDM symbols are once again transformed into the discrete-frequency domain for further communications processing to estimate the transmit frame $\mathbf{X}$ and subsequent bistatic radar sensing processing as described in Section 3.2 and [50].

While the resampling approach is able to correct both low and high SFOs, a simplified correction can be performed for the case where the SFO is so low that negligible ICI and ISI occur. In this case, a ZF equalization of $Y_{k,m}$ is performed to yield an estimate of $\tilde{Y}_{k,m}$. This estimate is given by $\hat{\tilde{Y}}_{k,m} = Y_{k,m} / (\hat{\alpha}_k^{\text{SFO}} e^{j\hat{\psi}_{k,m}^{\text{SFO}}})$, where $\hat{\alpha}_k^{\text{SFO}}$ and $\hat{\psi}_{k,m}^{\text{SFO}}$ are estimates of α_k^{SFO} and $\psi_{k,m}^{\text{SFO}}$ calculated by using $\hat{\delta}$ instead of its true value δ in (3.25) and (3.26), respectively. Since this method is assumed to be used for very low SFO only, $\alpha_k^{\text{SFO}} \approx 1$ is expected for all k , and the described equalization can be simplified to $\hat{\tilde{Y}}_{k,m} = Y_{k,m} / e^{j\hat{\psi}_{k,m}^{\text{SFO}}}$ or, alternatively,

$$\hat{\tilde{Y}}_{k,m} \approx \tilde{Y}_{k,m} + W_{k,m} / e^{j\hat{\psi}_{k,m}^{\text{SFO}}}. \quad (3.50)$$

The N rows and M columns filled with elements $\hat{\tilde{Y}}_{k,m}$ form a receive frame $\hat{\tilde{\mathbf{Y}}} \in \mathbb{C}^{N \times M}$, which can finally be used for the previously mentioned further processing. Since this approach only requires dividing each subcarrier in the discrete-frequency domain by a phase term, it results in lower computational complexity than imposed by resampling approach. This happens since the latter requires resampling the whole sequence $y[s]$, including samples belonging to the CPs of the OFDM symbols. In addition, a DFT must be performed again for each OFDM symbol after S/P conversion and CP removal to reconstruct the OFDM frame with corrected SFO.

A further alternative to correct the SFO is using the Radon-Fourier transform (RFT)-based approach introduced in [APW⁺25], which works by correcting the SFO-induced range and Doppler shift migration of targets in a radar image. While it can significantly reduce computational complexity compared

to resampling, the RFT-based SFO correction cannot properly compensate the SFO-induced ISI and subcarrier-dependent ICI, which may lead to severe impairments in medium and high SFO regimes.

3.3.4 Synchronization procedure overview

Based on the STO, CFO, and SFO estimation and correction algorithms described in Sections 3.3.1, 3.3.2, and 3.3.3, respectively, a full synchronization procedure is defined as depicted in Figure 3.5.

The proposed synchronization procedure starts by obtaining a coarse STO estimate $\hat{\tau}_{\Delta}^{\text{coarse}}$ and a fractional CFO estimate $\hat{f}_{\Delta}^{\text{frac}}$ from the receive sequence $y[s]$ using information extracted from the first preamble OFDM symbol by the S&C algorithm. These are used to identify the samples belonging to the second preamble OFDM symbol. Next, the discrete-frequency domain OFDM symbols $\mathbf{Y}^{\text{pb},1} \in \mathbb{C}^{N \times 1}$ and $\mathbf{Y}^{\text{pb},2} \in \mathbb{C}^{N \times 1}$ that correspond to the first and second preamble OFDM symbols are used together with $\hat{f}_{\Delta}^{\text{frac}}$ to estimate the integer CFO $\hat{f}_{\Delta}^{\text{int}}$, ultimately providing a full CFO estimate $\hat{f}_{\Delta}$. Next, CFO correction based on $\hat{f}_{\Delta}$ is performed for the samples of $y[s]$ around the region of the first preamble OFDM symbol, producing the sequence $y^{\text{corr}}[s] \in \mathbb{C}$. This sequence is then correlated with the originally transmit symbol in the discrete-time domain, providing a sample-accurate STO estimate $\hat{\tau}_{\Delta}$. Based on this, the samples of $y[s]$ corresponding to the full OFDM frame are selected and CFO correction is performed on them, resulting in the sequence $y^{\text{corr}'}[s] \in \mathbb{C}$.

The next step of the synchronization procedure is to estimate the SFO. For that, $y^{\text{corr}'}[s]$ undergoes S/P conversion, generating the discrete-time domain OFDM frame $\mathbf{y}_{\text{CP}}^{\text{corr}'} \in \mathbb{C}^{(N+N_{\text{CP}}) \times M}$. After CP removal on $\mathbf{y}_{\text{CP}}^{\text{corr}'}$, which produces, $\mathbf{y}^{\text{corr}'} \in \mathbb{C}^{N \times M}$, DFTs are performed on all OFDM symbols to produce the discrete-frequency domain OFDM frame $\mathbf{Y}^{\text{corr}'} \in \mathbb{C}^{N \times M}$. Pilot subcarriers are then extracted from this frame, resulting in the OFDM frame $\mathbf{Y}^{\text{corr}',\text{pil}} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$ that is used for SFO estimation with the TITO algorithm as discussed in Section 3.3.3. Once the SFO estimate $\hat{\delta}$ has been obtained, SFO correction is performed by resampling $y^{\text{corr}'}[s]$ to generate the sequence $y^{\text{corr}''}[s] \in \mathbb{C}$.

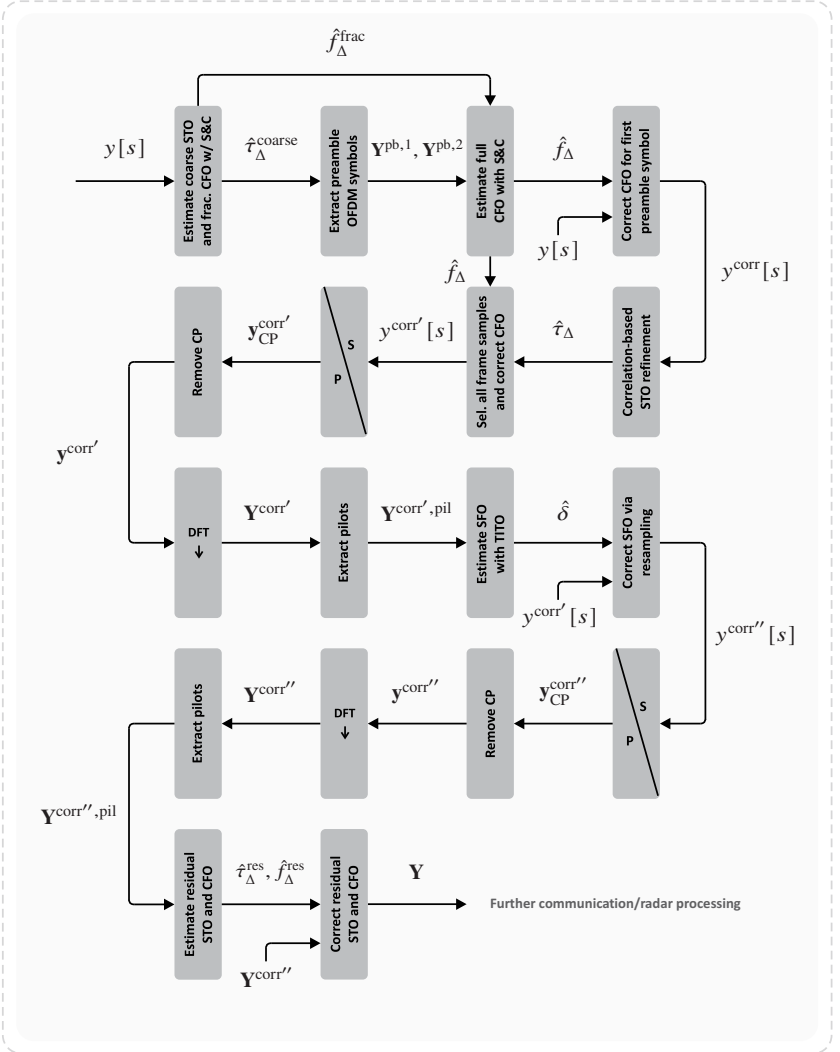


Figure 3.5: Full synchronization procedure for bistatic OFDM-based ISAC.

After S/P conversion on $y^{\text{corr}''}[s]$, which results in the discrete-time domain OFDM frame $\mathbf{y}_{\text{CP}}^{\text{corr}''} \in \mathbb{C}^{(N+N_{\text{CP}}) \times M}$, CP removal and DFTs along the OFDM symbols are performed. This respectively generates the discrete-time and discrete-frequency domain OFDM frames $\mathbf{y}^{\text{corr}''} \in \mathbb{C}^{N \times M}$ and $\mathbf{Y}^{\text{corr}''} \in \mathbb{C}^{N \times M}$, respectively. From the latter frame, pilot subcarriers are extracted to produce the frame $\mathbf{Y}^{\text{corr}'', \text{pil}} \in \mathbb{C}^{N_{\text{pil}} \times M_{\text{pil}}}$, which is used to estimate residual STO and CFO following [Sit17, SNZ18] as discussed in Sections 3.3.1 and 3.3.2. After correcting the aforementioned residual offsets, the discrete-frequency domain OFDM frame $\mathbf{Y}$ is obtained and can be used for further communications and bistatic radar signal processing as previously discussed in Section 3.2.

In the specific case of residual CFO estimation and correction, it is worth highlighting that it becomes relevant when high SFO is experienced and a heterodyne architecture with a digital IF is adopted. The reason for this is the SFO-induced frequency shift described by (3.30), which will result in a bias to the CFO estimate as it happens before SFO correction. Since this additional SFO-induced CFO will be estimated by the S&C algorithm and corrected, a frequency shift with same magnitude but opposite sign will be experienced after SFO correction. Consequently, this frequency shift must be estimated along with the actual residual CFO via pilots.

3.4 Hardware impairments

In this section, relevant hardware impairments in the considered bistatic OFDM-based ISAC system previously mentioned in Section 3.2.3.3 are discussed. In this sense, Figure 3.6 shows simplified heterodyne architectures for transmitter and receiver units of the considered OFDM-based ISAC system. In this figure, the hardware components where the most relevant impairments are experienced are highlighted. These are namely antennas, PA and LNA, mixers, oscillators, and DACs and ADCs, and the experienced impairments or synchronization offsets raised by each of those components as well as countermeasures are discussed in Sections 3.4.2 to 3.4.5, respectively. For simplicity, all the aforementioned impairments are individually addressed. In the cases where the discussed hardware impairment is mathematically described, a simplified formulation disregarding synchronization offsets and the remaining impairments is presented for conciseness.

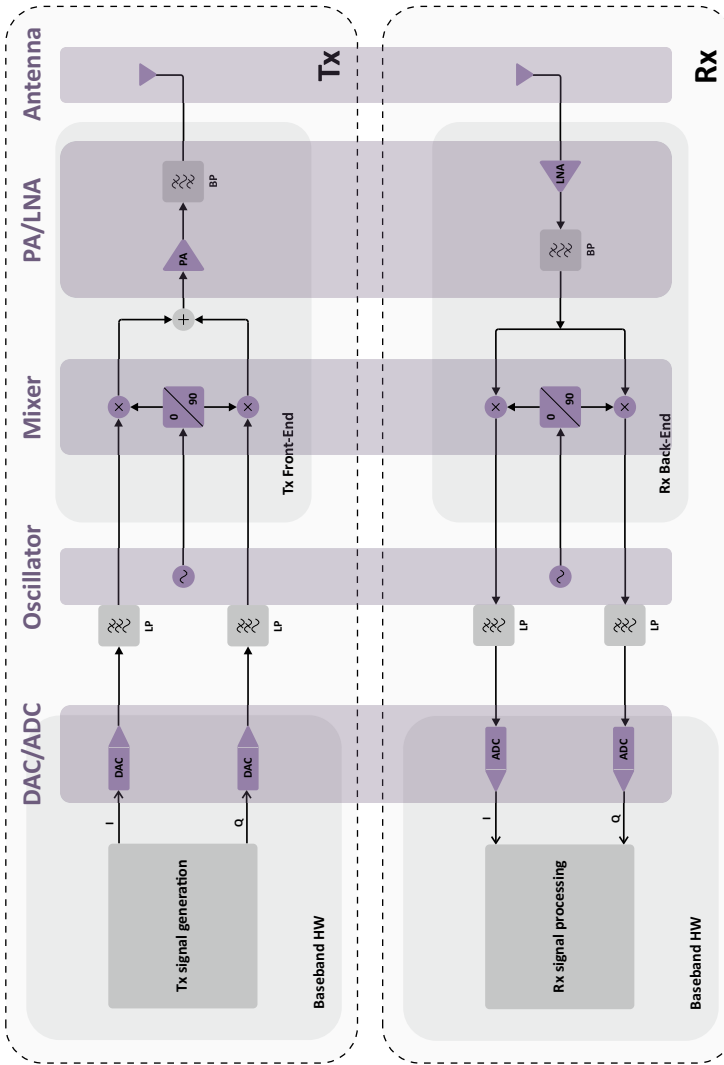


Figure 3.6: Simplified transmitter and receiver architectures with highlighted hardware components where impairments take place.

3.4.1 Antennas and antenna arrays

Antenna arrays and the antenna elements that constitute them are subject to a series of impairments. Particularly in the context of cellular-based, millimeter wave (mmWave) ISAC, where directional transmission with multiple transmit beams is expected to be performed for power efficiency and constraints related to physical layer security (PLS) and interference [BLR⁺20, MOS25], beam squint and element coupling are among the main effects that can impair both communications and radar sensing performances if not avoided or compensated. These impairments and their implications in the considered bistatic OFDM-based ISAC system are further discussed as follows.

3.4.1.1 Beam squint

Beam squint is a common hardware impairment in mmWave MIMO systems due to the use of wideband signals that becomes specially critical for increased antenna array sizes. This happens since slightly different delays are experienced at the different BB and RF chains leading to the individual antennas. Although small, these delays may lead to a significant variation of the resulting beam direction along the signal bandwidth B if it is relatively wide compared to the carrier frequency f_c [AA22]. Consequences of beam squint in ISAC include angular estimation bias and also inter-beam interference in case multiple beams are used.

Possible solutions to beam squint include the use of frequency-selective phase shifter networks, true-time delay (TTD) networks, or RF lens-based arrays [EMCE24, PKK⁺23], all of which require changes at the hardware level. Approaches relying on signal processing include beam broadening and phase correction techniques [EMCE24]. Although both these approaches solely rely on signal processing, the first results in angular resolution and array gain loss, while the latter approach performs better in scenarios with digital beamforming or hybrid beamforming with high RF chain count. In the bistatic OFDM-based ISAC verification measurements presented in Section 4.3, beamforming is performed based on RISs [LBE⁺21], which prevents applying any signal-processing-based beam squint compensation techniques. Since the bistatic MIMO-OFDM-based ISAC measurements discussed in Section 5.4 assume digital beamforming at both transmitter and receiver, the phase correction ap-

proach is henceforth assumed to be applied. It is simultaneously implemented with a frequency-selective amplitude and phase correction of the transmit and receive channels based on calibration measurements, as later described in Sections 5.2.1 and 5.2.2.3.

3.4.1.2 Element coupling

Although no full-duplex operation is required in bistatic ISAC as opposed to monostatic ISAC [BLH⁺21], which eliminates the coupling between transmit and receive antenna arrays, the coupling among the antenna elements within an array also affects the ISAC system design and performance. The consequences of mutual coupling among the antenna elements in an array include distortion of the ideal array patterns in both transmitting and receiving arrays, besides more severe spectral regrowth due to the PA nonlinearities discussed in Section 3.4.2 in the case of transmitting arrays [CZL18]. As countermeasures can only be applied during the array design phase, mutual coupling between antenna elements is not further investigated in this dissertation.

3.4.2 Power amplifier and low-noise amplifier

Amplifiers are transistor-based circuits designed to increase signal power. This is crucial to enable both communications and radar sensing, especially at mmWave frequencies where high path losses are experienced and targets with RCS are to be detected, respectively [SBH⁺23]. In modern radar and communications systems, as well as in ISAC [AMCA24], the two types of amplifiers that play the most relevant roles in driving system performance are PAs and LNAs. The first are employed to enhance signal power before transmission, while the latter are used at the input of the receiver aiming to amplify weak incoming signals while avoiding noise enhancement. In both PAs and LNAs, non-linear distortions of the input signal may happen, leading to not only lower amplification than the design target, but also interference. These issues are described as follows.

3.4.2.1 Non-linear distortion

Non-linear distortion at amplifiers is particularly an issue when the input signal to the amplifier covers a wide bandwidth and has high PAPR, which e.g., is the case of OFDM in 5G NR [CTZ⁺20].

To predict the behavior of PAs and LNAs and therefore allow either avoiding non-linear distortion at the design phase or developing further countermeasures to it, a wide range of models has been proposed in the literature. These models aim to obtain the amplitude-to-amplitude modulation (AM-AM) and amplitude-to-phase modulation (AM-PM) characteristics of the amplifier, both as functions of the input signal magnitude. Denoting the complex baseband equivalent of the amplifier output signal as $x_{\text{out}}^{\text{amp}}(t) \in \mathbb{C}$, it can be mathematically expressed as a function of the input signal as

$$x_{\text{out}}^{\text{amp}}(t) = \text{AM}(|x(t)|) + e^{j[\arg(x(t)) + \text{PM}(|x(t)|)]}. \quad (3.51)$$

In this equation, $\text{AM}(\cdot)$ and $\text{PM}(\cdot)$ return the amplifier output signal as functions of the input based on its AM-AM and AM-PM characteristics, respectively. Early models used to ultimately obtain the AM-AM and AM-PM profiles of amplifiers include the one proposed by Saleh for traveling-wave tube amplifiers [Sal81] and the one proposed by Rapp for solid-state PAs [Rap91]. For amplifiers designed for applications with wider bandwidths, more accurate models using Volterra series or polynomial models, either with or without memory [Sma12, VLT⁺19, SKB⁺24], are usually adopted.

It is henceforth assumed that second-order intermodulation products of the OFDM subcarriers, which are associated with the second-order terms in the aforementioned polynomial models, can be filtered out at both transmitter and receiver. Consequently, the main distortion sources will come from how close the amplifier is being operated to the saturation or compression as well as the third-order harmonics. The linearity parameters used to assess these two points are the input-referred 1-dB compression point (IP1dB) and input-referred third-order intercept point (IIP3), or their output-referred counterparts associated output-referred 1 dB compression point (OP1dB) and output-referred third-order intercept point (OIP3) [BSBF21]. Focusing on the input-referred parameters, the IP1dB indicates the input-referred instantaneous power operating point at which the amplifier gain is 1 dB below than its nominal value. To

avoid distortion of the output signal by the amplifier due to compression, an input back-off (IBO) from the IP1dB is usually required. The IBO measured in decibels is mathematically defined as [Sma12]

$$\text{IBO} = \frac{\text{IP1dB}}{\langle |x(t)|^2 \rangle}, \quad (3.52)$$

where the IP1dB is defined in Watts and

$$\langle |x(t)|^2 \rangle = \lim_{\tau \rightarrow \infty} \frac{1}{\tau} \int_{-\tau/2}^{\tau/2} |x(t)|^2 dt \quad (3.53)$$

is the average power of $x(t)$. As for the IIP3, which are usually associated with higher instantaneous input levels than the IP1dB, indicates the input-referred point where the third-order harmonics reach the power of the fundamental signal, which is the desired output of the amplifier. In order to avoid distortion due to third-order harmonics, which lead to spectral regrowth [Sma12], it is also important to ensure a back-off from the IIP3.

Henceforth, a conservative approach of ensuring sufficient IBO relative to the IP1dB during the design phase is assumed as the adopted countermeasure against amplifier distortion. The influence of amplifier non-linear distortion, more specifically compression and spectral regrowth, on both communications and sensing performances of the considered bistatic OFDM-based ISAC system as a function of the IBO will be later analyzed in Section 4.2.1. Although not further investigated, there are other countermeasures against amplifier distortion in OFDM-based ISAC and radar systems. These include, e.g., digital pre-distortion (DPD) techniques at the transmitter side [MSHW23, FBB24], equalization at the receiver side [KHH24b], and appropriate design of OFDM symbols aiming to reduce their PAPR [RM13, BFB24]. The latter can, however, limit the possibility of using these symbols to perform communications and therefore prevent ISAC operation.

3.4.3 Mixer

Similar to amplifiers, mixers are also transistor-based RF circuit components. Their primary function is, however, to shift the center frequency of a signal by multiplying it with an LO signal. At the transmitter side of the OFDM-based

ISAC system, mixers are used to up-convert OFDM signals from the BB to the RF carrier. In practical deployments, OFDM signals are typically up-converted from an IF rather than directly from BB to avoid issues such as severe attenuation around 0 Hz, i.e., direct current (DC), due to the use of baluns for differential to single-ended conversion or LO leakage. Conversely, mixers at the receiver side are used to do the inverse, i.e., down-convert OFDM signals from the RF carrier into the BB or an IF. Due to their significant importance, non-linear effects of mixers must be understood and either avoided or compensated. The most relevant for the considered bistatic OFDM-based ISAC system among the aforementioned non-linear effects are non-linear distortion due to compression and I/Q imbalance, which are discussed in what follows.

3.4.3.1 Non-linear distortion

As previously mentioned, mixers, like amplifiers, are transistor-based circuits. Out of this reason, non-linear distortion due to compression and spectral regrowth may also be experienced depending on the operation point. For conciseness, further explanation analysis is spared since these effects are already described in Section 3.4.2.

It is, however, worth mentioning that, in the case of down-conversion mixers at the receiver side, second-order harmonic distortion cannot be filtered out as assumed for amplifiers in Section 3.4.2. This is particularly relevant in both homodyne receiver architectures or heterodyne receiver architectures with relatively low IF compared to the signal bandwidth [Sma12], as may be the case of OFDM-based ISAC systems. For conciseness, this impairment will not be further investigated.

3.4.3.2 Frequency-selective I/Q imbalance

Also known as quadrature mixers, I/Q mixers, are often employed due to their capability of suppressing unwanted sidebands while preserving phase and amplitude information [Sma12]. This is possible since an I/Q mixer basically consists of two mixers with phase-shifted LO signals by 90° , which results in an inherent image frequency rejection capability without requiring additional filtering. Ideally, signals traveling through the in-phase (I) and quadrature (Q)

signal branches that terminate at the inputs or outputs of an I/Q mixer should be phase-shifted by $\pi/2$ rad to ensure full suppression of the unwanted sideband. In practical deployments, however, differences between the resulting CIRs of I and Q branches are experienced due to non-ideal transfer functions of analog LP filters as well as DACs at the transmitter side or ADCs at the receiver side. The combined effect of these non-idealities is called I/Q imbalance, and it becomes frequency-selective in systems with considerably wide bandwidth [VRK01b]. Denoting the amplitude and phase imbalances as $\alpha^{I/Q\text{-imbal}} \in [0, 1]$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} \in [-\pi, \pi)$, respectively, the complex baseband representation $x_{\text{out}}^{I/Q\text{-imbal}}(t) \in \mathbb{C}$ of the transmit RF signal output by the mixer under I/Q imbalance is expressed as [SKSW20]

$$x_{\text{out}}^{I/Q\text{-imbal}}(t) = b_1^{I/Q\text{-imbal}}(t) * x(t) + b_2^{I/Q\text{-imbal}}(t) * x^*(t), \quad (3.54)$$

where $*$ is the linear convolution operator,

$$b_1^{I/Q\text{-imbal}}(t) = \left[\delta(t) + \left(1 - \alpha_{\Delta}^{I/Q\text{-imbal}}\right) e^{-j\varphi_{\Delta}^{I/Q\text{-imbal}}} h_{\Delta}^{I/Q\text{-imbal}}(t) \right] / 2, \quad (3.55)$$

and

$$b_2^{I/Q\text{-imbal}}(t) = \left[\delta(t) - \left(1 - \alpha_{\Delta}^{I/Q\text{-imbal}}\right) e^{j\varphi_{\Delta}^{I/Q\text{-imbal}}} h_{\Delta}^{I/Q\text{-imbal}}(t) \right] / 2. \quad (3.56)$$

In these equations, $h_{\Delta}^{I/Q\text{-imbal}}(t)$ is the relative CIR between the I and Q branches of the mixer.

Under I/Q imbalance occurrence, limited sideband suppression is experienced [Nus21]. Ultimately, this will lead to degradation of communications and radar sensing performance in the considered bistatic OFDM-based ISAC system. Regarding OFDM-based communications, I/Q imbalance is known to cause ICI due to incomplete sideband suppression, besides spreading and rotating constellation symbols. As for radar sensing, it may result in dynamic range reduction due to the aforementioned ICI and even occurrence of ghost targets due to aliased sidebands that were not fully suppressed.

Although I/Q imbalance suppression measures can be taken at the mixer design phase [KTU24], further measures are required for further suppression due to the high sensitivity of radar sensing in OFDM-based ISAC. In the context of OFDM-based communications, several studies have proposed signal pro-

cessing techniques to compensate I/Q imbalance [VRK01a, AVR07, AVR08b], often also in combination with further impairments happening at the mixer such as frequency offsets and PN [XSL05, TCV⁺05]. A negative aspect of some of these techniques is that they require at least partial knowledge of the transmit data to obtain accurate channel estimates, often with known preambles or favorable pilot subcarrier placements that may not necessarily occur in practical systems [KAZV12]. This is particularly an issue in ISAC as the contribution of weak radar targets may not be noticeable in typical channel estimates used in OFDM-based communications [SKSW20]. To circumvent this issue, [SKSW20] proposed a blind I/Q imbalance estimation and correction approach for OFDM-based radar systems, which can be directly applied to the considered bistatic OFDM-based ISAC system. This approach is based on the blind adaptive filtering technique proposed in [AVR08a], but has a time-varying filter update step that allows the filter coefficients to converge to a state that can accurately compensate frequency-selective I/Q imbalance without oscillating due to the presence of multiple targets with different Doppler shifts. It is worth highlighting that this technique only compensates for I/Q imbalance at the receiver side. In fact, the algorithm is specifically designed to avoid compensating the combined effect of transmit and receive I/Q imbalances, as it could lead to still allow for the occurrence of ghost targets. To compensate for transmit I/Q imbalance, DPD techniques such as the one in [KZHI17] can be used. Therefore, it is henceforth assumed that transmit I/Q imbalance is compensated alongside with other hardware non-idealities via DPD, e.g., as later described in Section 5.2.1.

3.4.4 Oscillator

Besides the non-linear distortion and frequency-selective I/Q imbalance at the mixing stage, a further impairment is also experienced due to non-idealities in the LO signal that is input to the mixers at both transmit and receiver sides. This impairment is the PN, and it consists of phase variations in the LO signal.

PN becomes particularly relevant at the mmWave frequency range and above, and it has been widely investigated for communications systems [Sma12, KKP⁺14]. In addition, it is also known to degrade sensing performance. For example, PN imposes challenges to important ISAC tasks such as clutter removal for effective sensing, e.g., in indoor environments [HMG⁺23]. In FMCW radar

systems, PN introduces random frequency deviations to the chirp signals, which ultimately lead to SNR loss and distort the IF signal spectrum [ASB⁺16]. In the context of digital waveforms for radar sensing, the influence of PN on PMCW was also analyzed for radar networks in [KBBP24] and [WAA⁺24]. The latter study demonstrated that PN leads to an SNR reduction due to degradation of the correlation properties of the adopted PRBS and a stripe along the Doppler shift direction for each target in the radar image. In [SSK⁺18b, AWJ⁺24], these effects were also briefly discussed for OFDM radar and their causes attributed to a PN-induced common phase error (CPE) and ICI. Further studies such as [KWK23, KME⁺23, KKW⁺24] derive CRLBs on the achievable accuracy of range and velocity, which is related to the Doppler shift, of a single radar target under PN influence. In [KWK23], a strategy to estimate the PN parameters and use them to improve the accuracy of range and velocity estimates is also proposed. A detailed analysis of the PN influence on the SNR loss and distortion of radar images in both range and Doppler shift directions for different choices of OFDM waveform parameters and scenarios has, however, not been reported in the literature so far. To fill this gap, a mathematical formulation of PN and its effects in mmWave OFDM-based ISAC is described in this section, and a numerical analysis is later presented in Section 4.2.3. Given their widespread adoption in state-of-the-art systems, phase-locked loop (PLL)-based LOs are assumed. To better highlight the differences of the PN effects in the considered bistatic OFDM-based ISAC system to its monostatic counterpart, a general mathematical formulation is presented to allow analyzing both architectures.

3.4.4.1 Oscillator phase noise

In an OFDM-based ISAC system, up-converting $x(t)$ to the carrier frequency $f_c \gg B$ with an I/Q mixer results in the RF analog signal $x_{\text{RF}}(t) \in \mathbb{C}$, which is expressed as

$$x_{\text{RF}}(t) = x(t) e^{j[2\pi f_c t + \psi_{\text{Tx}} + \theta_{\text{Tx}}^{\text{PN}}(t)]}. \quad (3.57)$$

It is worth highlighting that $x_{\text{RF}}(t)$ is only modeled as a complex-valued signal for simplicity. In practice, it should be real-valued and both the frequency band of interest centered at f_c and its mirror image centered at $-f_c$ should be considered when describing $x_{\text{RF}}(t)$ and the RF signals derived from it. Furthermore, the term $\psi_{\text{Tx}} \in \mathbb{R}$ in (3.57) denotes the transmitter carrier phase offset, and $\theta_{\text{Tx}}^{\text{PN}}(t) \in \mathbb{R}$ represents the transmitter PN that has a double-sided power spec-

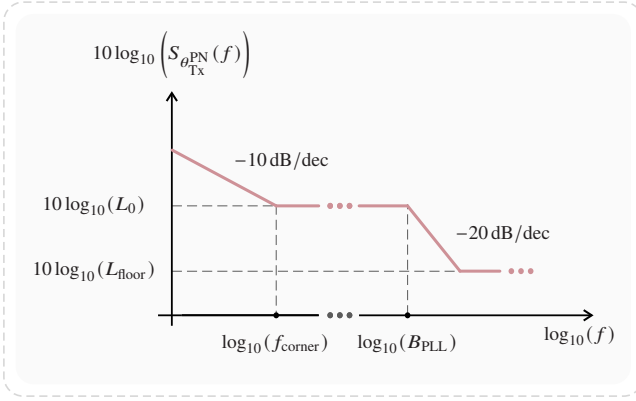


Figure 3.7: Double-sided PSD profile of the assumed PN model for an PLL-based LO [Sma12, KKP⁺14, SSK⁺18b]. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

tral density (PSD) $S_{\theta_{\text{Tx}}^{\text{PN}}}(f) \in \mathbb{R}_{\geq 0}$ [vS23]. The profile of the aforementioned PSD is shown in Figure 3.7 assuming that the LO is implemented by using a PLL [Sma12, KKP⁺14, SSK⁺18b]. In this figure, f_{corner} and B_{PLL} denote the flicker corner frequency and the PLL loop bandwidth, respectively. It is assumed that the PN PSD has a slope of -10 dB/dec until f_{corner} . Between f_{corner} and B_{PLL} , the PSD assumes the value L_0 , which is defined as the in-band PN level. After that interval, the PSD decays at a rate of -20 dB/dec until it reaches the PN floor L_{floor} . Both L_0 and L_{floor} are measured in rad^2/Hz . When the base-10 logarithm of the PSD of the PN is taken, the resulting unit becomes dBc/Hz. Based on these parameters, the transmit PN double-sided PSD can be mathematically described by [Sma12, SSK⁺18b]

$$S_{\theta_{\text{Tx}}^{\text{PN}}}(f) = \frac{B_{\text{PLL}}^2 L_0}{B_{\text{PLL}}^2 + f^2} \left(1 + \frac{f_{\text{corner}}}{f} \right) + L_{\text{floor}}. \quad (3.58)$$

After being amplified by a PA and radiated by the transmit antenna of the considered ISAC system, $x_{\text{RF}}(t)$ propagates through P paths labeled as $p \in \{0, 1, \dots, P-1\}$. As in Section 3.1, it is assumed that the p th path is associated with attenuation α_p , delay τ_p , and Doppler shift $f_{\text{D},p}$. Consequently, the

receive captured signal $y_{\text{RF}}(t) \in \mathbb{C}$ by the receive antenna of the considered ISAC system can be expressed as

$$y_{\text{RF}}(t) = \sum_{p=0}^{P-1} \alpha_p x_{\text{RF}}(t - \tau_p) e^{j2\pi f_D p t}. \quad (3.59)$$

By substituting (3.57) into (3.59),

$$y_{\text{RF}}(t) = \sum_{p=0}^{P-1} \alpha_p \left[x(t - \tau_p) e^{j[2\pi f_c(t - \tau_p) + \psi_{\text{Tx}} + \theta_{\text{Tx}}^{\text{PN}}(t - \tau_p)]} \right] e^{j2\pi f_D p t} \quad (3.60)$$

is obtained. After $y_{\text{RF}}(t)$ is amplified by a LNA, it undergoes bandpass (BP) filtering and is down-converted with an I/Q mixer. The resulting baseband signal $y(t)$ from the aforementioned down-conversion can be expressed as

$$y(t) = y_{\text{RF}}(t) e^{-j[2\pi f_c t + \psi_{\text{Rx}} + \theta_{\text{Rx}}^{\text{PN}}(t)]}, \quad (3.61)$$

where $\psi_{\text{Rx}} \in \mathbb{R}$ and $\theta_{\text{Rx}}^{\text{PN}}(t) \in \mathbb{R}$ denote the receiver carrier phase offset and the receiver PN time series in radians, respectively. It is henceforth assumed that the receiver LO is also derived from a PLL and that its PN has a double-sided PSD $S_{\theta_{\text{Rx}}^{\text{PN}}}(f) \in \mathbb{R}_{\geq 0}$ with the same profile as its counterpart at the transmitter side given by (3.58) and shown in Figure 3.7. By substituting (3.60) into (3.61), the latter equation can be expanded as

$$\begin{aligned} y(t) &= \left\{ \sum_{p=0}^{P-1} \alpha_p \left[x(t - \tau_p) e^{j[2\pi f_c(t - \tau_p) + \psi_{\text{Tx}} + \theta_{\text{Tx}}^{\text{PN}}(t - \tau_p)]} \right] e^{j2\pi f_D p t} \right\} \\ &\quad \times e^{-j[2\pi f_c t + \psi_{\text{Rx}} + \theta_{\text{Rx}}^{\text{PN}}(t)]} \\ &= \sum_{p=0}^{P-1} \alpha_p x(t - \tau_p) e^{j[\psi_{\text{Tx}} - \psi_{\text{Rx}} - 2\pi f_c \tau_p]} e^{j2\pi f_D p t} e^{j[\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)]}. \end{aligned} \quad (3.62)$$

It is next assumed that $\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t) \ll 1$, which is reasonable for typical PN levels in practice. Therefore, it holds that $\cos[\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)] \approx 1$

and $\sin [\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)] \approx \theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)$, which allows further simplifying

$$e^{j[\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)]} = \cos [\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)] + j \sin [\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)] \quad (3.63)$$

as

$$e^{j[\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)]} \approx 1 + j [\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)]. \quad (3.64)$$

If, in addition, the phase terms associated with transmitter, receiver and the delay of the p th path are jointly represented by ψ_p such that

$$\psi_p = \psi_{\text{Tx}} - \psi_{\text{Rx}} - 2\pi f_c \tau_p, \quad (3.65)$$

then (3.62) can be rewritten as

$$y(t) = \left[\sum_{p=0}^{P-1} \alpha_p x(t - \tau_p) e^{j\psi_p} e^{j2\pi f_{\text{D},p} t} \right] \{1 + j[\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)]\}. \quad (3.66)$$

After LP filtering, the receive BB signal $y(t)$ undergoes A/D conversion with sampling rate F_s , which is associated with the sampling period $T_s = 1/F_s$. If $F_s > B$, it is assumed that digital LP filtering and downsampling to B is performed. Disregarding AWGN, this yields the discrete-time domain sequence $y[s]$ as discussed in Section 3.1. By performing S/P conversion on $y[s]$ and discarding the CP at each OFDM symbol, the discrete-time domain frame $\mathbf{y} \in \mathbb{C}^{N \times M}$ is obtained. Based on (3.66) and (3.6), the sample at the n th row, $n \in \{0, 1, \dots, N-1\}$, and m th column of $\mathbf{y}$ can be expressed as

$$y_{n,m} = \left(\sum_{p=0}^{P-1} \alpha_p x_{\langle n - n_{\Delta,p} \rangle_{N,m}} e^{j\psi_p} e^{j2\pi f_{\text{D},p} [m(N + N_{\text{CP}}) + N_{\text{CP}} + n]/B} \right) \times \left[1 + j \left(\theta_{\text{Tx}}^{\text{PN}}([m(N + N_{\text{CP}}) + N_{\text{CP}} + n - n_{\Delta,p}]/B) - \theta_{\text{Rx}}^{\text{PN}}([m(N + N_{\text{CP}}) + N_{\text{CP}} + n]/B) \right) \right], \quad (3.67)$$

assuming that the Doppler-shift induced ICI is negligible for all P targets for simplicity. In this equation, $x_{n,m}$ is the element at the n th row and m th column

of the discrete-time domain transmit OFDM frame $\mathbf{x} \in \mathbb{C}^{N \times M}$. Without loss of generality, $n_{\Delta,p} = \tau_p B$ in this equation is assumed to be an integer to simplify the notation. In addition, $\langle \cdot \rangle_N$ is the modulo N operator. Next, the OFDM symbols contained in $\mathbf{y}$ undergo DFTs. This results in the discrete-frequency domain OFDM frame $\mathbf{Y}$, whose sample at k th subcarrier, $k \in \{-N/2, 1, \dots, N/2 - 1\}$, at the m th OFDM symbol is expressed

$$\begin{aligned}
 Y_{k,m} = & \left[\sum_{p=0}^{P-1} \alpha_p X_{k,m} e^{j\psi_p} e^{-j2\pi k \tau_p B/N} e^{j2\pi f_{D,p} [m(N+N_{CP})+N_{CP}]/B} \right] \\
 & + \frac{j}{N} \sum_{\kappa=-N/2}^{N/2} \left[\sum_{p=0}^{P-1} \alpha_p X_{\kappa,m} e^{j\psi_p} e^{-j2\pi \kappa \tau_p B/N} e^{j2\pi f_{D,p} [m(N+N_{CP})+N_{CP}]/B} \right] \\
 & \times \sum_{n=0}^{N-1} \left(\theta_{\text{Tx}}^{\text{PN}} ([m(N+N_{CP})+N_{CP}+n-n_{\Delta,p}]/B) \right. \\
 & \left. - \theta_{\text{Rx}}^{\text{PN}} ([m(N+N_{CP})+N_{CP}+n]/B) \right) e^{j2\pi(\kappa-k)n/B}. \quad (3.68)
 \end{aligned}$$

Considering a single path, i.e., $P = 1$, and omitting the subscript p for the sake of simplicity, the OFDM frame $\mathbf{Y}$ is split into the PN-free OFDM frame $\tilde{\mathbf{Y}} \in \mathbb{C}^{N \times M}$ and the PN-induced distortion to the OFDM frame $\boldsymbol{\xi}^{\text{PN}} \in \mathbb{C}^{N \times M}$. Consequently, the sample at the k th row and m th column of the PN-free OFDM frame $\tilde{\mathbf{Y}}$ is expressed as

$$\tilde{Y}_{k,m} = \alpha X_{k,m} e^{j\psi} e^{-j2\pi k \tau B/N} e^{j2\pi f_D [m(N+N_{CP})+N_{CP}]/B}, \quad (3.69)$$

while the distortion at the same position of ξ^{PN} is expressed as

$$\begin{aligned} \xi_{k,m}^{\text{PN}} = & \frac{j}{N} \sum_{\kappa=-N/2}^{N/2} \alpha X_{\kappa,m} e^{j\psi} e^{-j2\pi\kappa\tau B/N} e^{j2\pi f_D [m(N+N_{\text{CP}})+N_{\text{CP}}]/B} \\ & \times \sum_{n=0}^{N-1} \left(\theta_{\text{Tx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n-n_{\Delta}]/B) \right. \\ & \left. - \theta_{\text{Rx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n]/B) \right) e^{j2\pi(\kappa-k)n/N}. \quad (3.70) \end{aligned}$$

Splitting (3.70) into the cases where the sum index κ is equal to or different than the analyzed subcarrier index k , i.e., $\kappa = k$ or $\kappa \neq k$, $\xi_{k,m}^{\text{PN,CPE}} \in \mathbb{C}$ and $\xi_{k,m}^{\text{PN,ICI}}$ are respectively obtained. These variables represent the PN-induced CPE and the PN-induced ICI, which are expressed as

$$\begin{aligned} \xi_{k,m}^{\text{PN,CPE}} = & \frac{j}{N} \tilde{Y}_{k,m} \sum_{n=0}^{N-1} \left(\theta_{\text{Tx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n-n_{\Delta}]/B) \right. \\ & \left. - \theta_{\text{Rx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n]/B) \right) \quad (3.71) \end{aligned}$$

and

$$\begin{aligned} \xi_{k,m}^{\text{PN,ICI}} = & \frac{j}{N} \sum_{\kappa=-N/2, \kappa \neq k}^{N/2} \tilde{Y}_{\kappa,m} \\ & \times \sum_{n=0}^{N-1} \left(\theta_{\text{Tx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n-n_{\Delta}]/B) \right. \\ & \left. - \theta_{\text{Rx}}^{\text{PN}} ([m(N+N_{\text{CP}})+N_{\text{CP}}+n]/B) \right) e^{j2\pi(\kappa-k)n/N}, \quad (3.72) \end{aligned}$$

respectively [Gar01].

According to (3.71), the CPE causes a phase rotation due to the multiplication of $\tilde{Y}_{k,m}$ by $j\Phi_m^{\text{PN,CPE}}$, where $\Phi_m^{\text{PN,CPE}} \in \mathbb{R}$ is expressed as

$$\Phi_m^{\text{PN,CPE}} = \frac{1}{N} \sum_{n=0}^{N-1} \left(\theta_{\text{Tx}}^{\text{PN}} ([m(N + N_{\text{CP}}) + N_{\text{CP}} + n - n_{\Delta}]/B) - \theta_{\text{Rx}}^{\text{PN}} ([m(N + N_{\text{CP}}) + N_{\text{CP}} + n]/B) \right) \quad (3.73)$$

and does not depend on the subcarrier index k . Consequently, all subcarriers within the same OFDM symbol suffer the same phase rotation. While this can be corrected to a certain extent for communications by performing channel estimation and equalization with pilot subcarriers strategically placed in the OFDM frame, it may become an issue for radar sensing. This is due to the fact that the goal of radar sensing is to perform highly accurate channel estimation over consecutive OFDM symbols to allow estimating range and Doppler shift of targets [18]. While a phase rotation does not degrade the range profiles obtained at each OFDM symbol, it significantly impairs the Doppler shift estimation. This happens since the aforementioned process consists of converting the Doppler-shift induced phases into a Doppler shift estimate by performing one DFT over multiple OFDM symbols for. Since CPE alters the phase of each OFDM symbol, the obtained Doppler shift profiles will be distorted, presenting possible loss of the target peak power and increase of the sidelobe levels along the Doppler shift direction at the target range bin of the ultimately obtained range-Doppler shift radar image. The PN-induced ICI, in turn, already degrades each OFDM symbol individually and not only when analyzing the whole frame as in the CPE case. The ICI causes loss of orthogonality between the subcarriers, reducing their effective SINR and degrading the performance of both OFDM-based communications and also sensing, resulting in peak power loss and sidelobe level increase in the obtained range profiles at each OFDM symbol and ultimately degrading the obtained range-Doppler shift radar image. According to [Gar01, Sma12, SSK⁺18b], it is assumed that the CPE is the dominant PN-induced effect if $\Delta f/2 > B_{\text{PLL}}$, while ICI becomes the most relevant effect for $\Delta f/2 < B_{\text{PLL}}$. Since B_{PLL} values of 50 kHz [DvZM⁺17], through around 150 kHz [Key25], and even up to 200 kHz [SSK⁺18b] are reported for mmWave, both aforementioned cases can happen in practice, e.g., when subcarrier spacings ranging from 15 kHz to 120 kHz are used according to 3GPP specifications [3GPP20].

The influence of PN on the OFDM-based ISAC system performance will depend on whether $\theta_{\text{Tx}}^{\text{PN}}(t)$ and $\theta_{\text{Rx}}^{\text{PN}}(t)$ are correlated. While correlated PN is observed in monostatic radar sensing due to the use of the same oscillator by both transmitter and receiver, the use of distinct oscillators results in non-correlated PN for bistatic radar sensing and communications. The aforementioned scenarios are depicted in Figure 3.8, where a gNB with full-duplex capability [BRT⁺19, BLH⁺21] performs monostatic sensing in Figure 3.8(a) and two gNBs perform simultaneous communications and bistatic sensing following the system concept from [6, 50] in Figure 3.8(b).

Besides quantifying the PN-induced impairments, there is also a growing interest in the literature regarding PN estimation and compensation. In OFDM-based optical communications, for instance, a dedicated CW tone can be simultaneously transmitted with the OFDM signal to fully estimate the PN, i.e., both CPE and ICI [HPN12]. A CW-based approach has also been proposed in [WAA⁺24], but it is only applicable to PMCW-based radar sensing or ISAC. Further PN compensation strategies were proposed in [CBY02, ZTS07, SVTR09, SSFI09, SV10, WMTH16, LY17]. Although these methods allow fully estimating the PN, they either require knowledge of the channel or perform joint PN and channel estimation. In ISAC applications, this would lead to rather high computational complexity as a dedicated estimate for each OFDM symbol would be needed, especially in scenarios with multiple targets with different Doppler shifts. In addition, the PN estimation would be biased towards estimating the PN of a dominant, reference path. This is, e.g., the case in bistatic sensing or communications with a LoS component between transmitter and receiver, as depicted in Figure 3.8(b). In monostatic sensing, a reference path is usually present due to coupling between transmit and receive antennas. Nevertheless, since this path has virtually zero range, the correlation between $\theta_{\text{Tx}}^{\text{PN}}(t)$ and $\theta_{\text{Rx}}^{\text{PN}}(t)$ will be maximum for it, and therefore the PN influence on this path will be significantly suppressed. Consequently, the transmit-receive coupled path cannot be used as a reference for CPE estimation. Using other targets at higher ranges also not a reliable strategy, as it cannot be ensured that they are static and their Doppler-shift-induced phases may also potentially bias the PN estimate.

While the PN-induced ICI cannot be efficiently combated, CPE can be estimated and corrected to a certain extent [OK20]. This partial PN compensation method is particularly efficient when the PN varies slowly within an OFDM symbol duration, which leads to dominant CPE over PN-induced ICI [AC98]. In

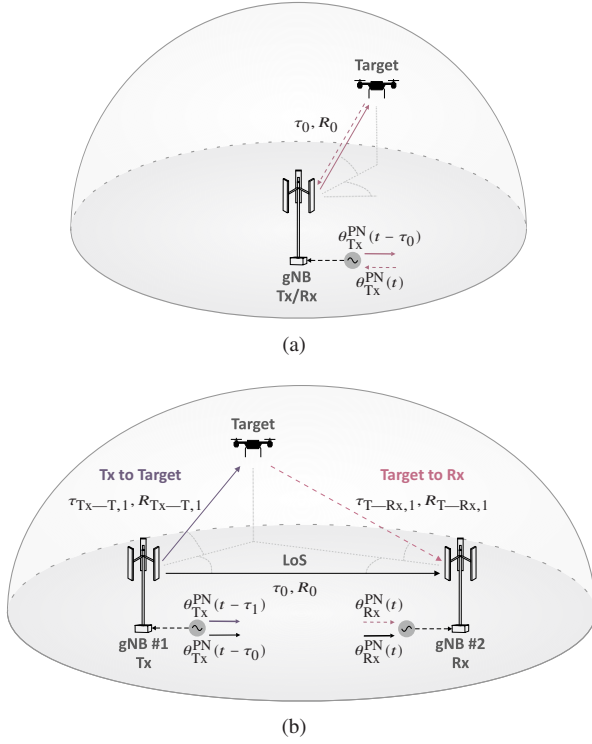


Figure 3.8: Monostatic (a) and bistatic (b) ISAC scenarios. In the monostatic scenario in (a), a gNB performs monostatic sensing to detect a single target of index $p = 0$ at a range R_0 , which results in a single propagation path ($P = 1$) with propagation delay $\tau_0 = R_0/c_0$. In this scenario, the effective transmit PN contribution is $\theta_{\text{Tx}}^{\text{PN}}(t - \tau_0)$, while the receive PN is $\theta_{\text{Rx}}^{\text{PN}}(t) = \theta_{\text{Tx}}^{\text{PN}}(t)$. In the bistatic ISAC scenario in (b), two gNBs communicate with one another through an LoS path of index $p = 0$, physical length R_0 , and propagation delay $\tau_0 = R_0/c_0$. In addition, the aforementioned gNBs also perform bistatic sensing of a single target of index $p = 1$ at a range R_1 , and with corresponding propagation delay $\tau_1 = R_1/c_0$. It holds that $\tau_1 = \tau_{\text{Tx}-\text{T},1} + \tau_{\text{T}-\text{Rx},1}$, where $\tau_{\text{Tx}-\text{T},1} = R_{\text{Tx}-\text{T},1}/c_0$, $\tau_{\text{T}-\text{Rx},1} = R_{\text{T}-\text{Rx},1}/c_0$, and $R_{\text{Tx}-\text{T},1}$ and $R_{\text{T}-\text{Rx},1}$ are respectively the ranges between transmitting gNB and radar target, and radar target and receiving gNB. In the bistatic case, the transmit PN contributions for the LoS path ($p = 0$) and the path associated with the radar target are $\theta_{\text{Tx}}^{\text{PN}}(t - \tau_0)$ and $\theta_{\text{Tx}}^{\text{PN}}(t - \tau_1)$, respectively. Unlike in the monostatic case, the receive PN contribution for both paths, $\theta_{\text{Rx}}^{\text{PN}}(t)$, has no correlation with $\theta_{\text{Tx}}^{\text{PN}}(t)$ in bistatic ISAC as different oscillators are used by the transmitting and receiving gNBs. Adapted from [21], licensed under CC BY-NC-ND 4.0.

OFDM-based communications, CPE estimation can be performed by using pilot subcarriers [RK95], at which the observed phase of the modulation symbol at the receiver side is compared with the one of the originally transmitted symbol. The aforementioned procedure can be mathematically described as

$$\widehat{\Theta}_m^{\text{PN,CPE}} = \arg \left\{ \sum_{k=-N/2}^{N/2-1} Y_{k,m} X_{k,m}^* \right\}, \quad (3.74)$$

where $\widehat{\Theta}_m^{\text{PN,CPE}} \in \mathbb{R}$ is the CPE estimate for the m th OFDM symbol, $(\cdot)^*$, and $\arg\{\cdot\}$ returns the argument of a complex number. This estimate is then used for CPE correction on the receive frame $\mathbf{Y}$, producing the frame $\mathbf{Y}^{\text{corr}} \in \mathbb{C}^{N \times M}$ whose element at the k th and m th column is expressed as

$$Y_{k,m}^{\text{corr}} = Y_{k,m} e^{-j\widehat{\Theta}_m^{\text{PN,CPE}}}. \quad (3.75)$$

In 5G NR, for instance, the phase tracking reference signal (PT-RS) is used to estimate, track, and allow correcting PN-induced CPE [QHN⁺18]. The discussed CPE estimation and correction can improve the constellation shape and therefore communications and sensing performances. It is, however, only suitable when there is a dominant reference path, and therefore only suitable for bistatic ISAC architectures as previously discussed. As for the discussed full PN estimation strategies, the effectiveness of the CPE correction for sensing may also be limited if the PN changes rapidly, as multiple targets with rather low RCS and at different ranges other than the one of the reference path are expected [ZHA⁺23, ZWZ⁺24]. These targets will experience significantly different PN realizations and consequently CPEs. In such cases, CPE estimation and correction is still desirable as it will improve the communications performance of the bistatic ISAC system. To decouple the CPE of the reference path and other targets, the CPE at a given OFDM symbol can be estimated based on the phase of the peak associated with the reference path after obtaining a CIR estimate from the pilot subcarriers.

3.4.5 Digital-to-analog and analog-to-digital converters

When producing the baseband transmit signal $x(t)$ from the samples in $x[s]$ at the transmitter side via D/A conversion, and when performing A/D conversion on the baseband receive signal $y(t)$ to produce the discrete-time domain sequence $y[s]$, a series of non-idealities arise. These include quantization, clipping, the occurrence of image spectra and aliasing, as well as SJ. Each of these impairments is described in further detail in the following.

3.4.5.1 Quantization

Due to finite resolution of the DAC, the samples of $x[s]$ are represented by a limited number $N_{\text{bit}}^{\text{DAC}}$ of DAC bits. Consequently, only a discrete set of voltage levels can be generated when producing the analog signal $x(t)$ during D/A conversion. Similarly, the accuracy with which the sequence $y[s]$ represents the analog signal $y(t)$ after A/D conversion is defined by the number of ADC bits $N_{\text{bit}}^{\text{ADC}}$, as it determines the set of discrete values the samples of $y[s]$ can assume. The limited resolution in both D/A and A/D conversions introduces deviations from the ideally expected outputs, which are referred to as quantization noise.

Based on the discussion in Section 2.1 and [VGG⁺14], it holds for the experienced full-scale signal-to-quantization-noise ratio (SQNR) in OFDM-based systems is a function of the number of ADC bits $N_{\text{bit}}^{\text{ADC}}$ that [Nus21]

$$10 \log_{10}(\text{SQNR}) = 6.02 N_{\text{bit}}^{\text{ADC}} + 4.77 - 10 \log_{10}(\text{PAPR}). \quad (3.76)$$

If oversampling is performed during A/D conversion, quantization noise within the signal band of interest is reduced. According to [Kes08], (3.76) becomes

$$10 \log_{10}(\text{SQNR}) = 6.02 N_{\text{bit}}^{\text{ADC}} + 4.77 - 10 \log_{10}(\text{PAPR}) + 10 \log_{10} \left(\frac{F_s}{B} \right). \quad (3.77)$$

Since, due to the central limit theorem, the quantization noise experienced in the sampled discrete-time domain signal eventually becomes Gaussian noise after DFT and subsequent processing steps used for communications and radar sensing in the considered bistatic OFDM-based ISAC system, the assumption

that the radar processing gain is effective against quantization noise made in (2.7) holds true. Following the discussion from Section 2.1 for the monostatic case, the required number of ADC bits $N_{\text{bit}}^{\text{ADC}}$ for the bistatic OFDM-based ISAC system can be calculated based on the effective ADC dynamic range after signal processing $\mathcal{D}_{\text{image}}$ such that

$$N_{\text{bit}}^{\text{ADC}} = (10 \log_{10}(\mathcal{D}_{\text{image}} \cdot \text{PAPR} \cdot B / (G_p \cdot F_s)) - 4.77) / 6.02. \quad (3.78)$$

Contrary to the monostatic case, where $\mathcal{D}_{\text{image}}$ is defined based on the spillover between transmitter and receiver, $\mathcal{D}_{\text{image}}$ is defined based on the power of the reference path in the considered bistatic OFDM-based ISAC system as it has dominant power. Consequently, $\mathcal{D}_{\text{image}}$ will assume the same value as the image SNR for the reference path in (3.8).

3.4.5.2 Clipping

If a voltage output beyond the capabilities of a DAC is demanded or when the voltage of the analog signal to be sampled by an ADC exceeds its input range, clipping occurs. The experienced clipping in DACs and ADCs happens as soon as the input or output ranges of the converter are exceeded, respectively, with no gradual compression. This effect is known as hard clipping.

In amplifiers, soft clipping happens in the transition between its linear and saturation regions. Only when in full saturation, amplifiers cause hard clipping of signals. Due to the similarity of the hard clipping in DACs and ADCs and the effects caused by amplifiers in full saturation, similar effects are expected to happen during A/D and D/A conversion. As discussed in Section 3.4.2, these effects include in-band interference for both DAC and ADC clipping, as well as spectral regrowth and resulting OOB emissions specifically in the case of DAC clipping. These effects will be further analyzed in Section 4.2.1 for amplifiers, and further investigations into DAC and ADC clipping are left open for future study.

3.4.5.3 Image spectra and aliasing

In OFDM-based systems in general, oversampling is usually desired at both DACs and ADCs. In the case of the former, a higher sampling rate than the OFDM signal bandwidth, i.e., $F_s \gg B$, is used to ensure that the series of image spectra resulting from the D/A conversion are pushed away from the signal of interest. Due to the combined effect of the sinc-like attenuation during D/A conversion and the use of LP filters, this ensures sufficient suppression of the image spectra, e.g., as shown in Figure 2.9. By using $B = 0.85F_s$, which corresponds to an oversampling ratio of $F_s/B = 1.18$, a ratio of -120 dB between the power of the image spectra and the one of the signal of interest is already observed [Nus21]. With state-of-the-art high-speed DACs, however, considerably higher D/A oversampling ratios can be achieved.

At the receive side, in addition to an increased SQNR as previously discussed, oversampling during A/D conversion allows preventing aliasing from both the transmit image spectra discussed earlier and foreign signals occupying frequencies adjacent to the OFDM signal band. Furthermore, it relaxes the requirements on the analog anti-aliasing filters, since efficient digital filtering can then be employed to suppress these unwanted components.

3.4.5.4 Sampling jitter

In practice, the sampling clocks that determine the sampling instants of DACs and ADCs are non-ideal, primarily due to PN introduced by imperfections in the oscillators from which they are derived. This leads to SJ, which refers to deviations from the ideal sampling instants. According to [LF07], SJ is composed of both aperture jitter, which represents the sampling time variations caused by noise in the sample-and-hold circuit of an ADC, and clock jitter, which is caused by the clock generator used to feed the ADC with a clock signal. The analysis in the aforementioned study shows that, for short signal block durations, as is typically the case in mmWave OFDM-based systems, clock jitter is the dominant effect. In addition, it is shown that increasing oversampling is ineffective against clock jitter-induced interference. Unlike aperture jitter that is spread over the whole digitization band, this happens since clock jitter interference is concentrated within the frequency band occupied by

the input signal to an ADC. For this reason, numerous studies consider only the effect of clock jitter.

When experienced in OFDM-based systems with BB sampling, SJ leads to ICI [Sma12]. For the case where BP sampling is adopted, [PF09] and [SV10] presented an analysis of ADC SJ on an OFDM-based communications system. These studies showed that, besides ICI, SJ results in an oscillator PN-like effect due to the use of a digital IF, leading to CPE. For the BB sampling case, [MT03, YA10] investigated the effect of oversampling on the ICI induced by white ADC SJ, where the SJ at different sampling time instants is uncorrelated as opposed to colored SJ which has non-flat PSD, without considering DAC SJ. Furthermore, [OLS06] derived closed-form expressions and upper bounds for the ICI power due to white and colored ADC SJ only, also considering the effect of oversampling. It was shown that, for colored SJ, oversampling is not as effective as for the white SJ case, and that a residual ICI is always present.

Shifting the focus from the receiver to the transmitter side, [HMAM15] presented an analysis of white DAC SJ, showing that it leads to OOB radiation and investigating spectral sidelobe suppression techniques to compensate for it. Finally, [GGT21], analyzed the effects of white and colored SJ at both ADC and DAC for hybrid OFDM-digital filter multiple access passive optical networks. Afterwards, the authors proposed a SJ simulation model consisting of oversampling, a fractional delay filter at the increased sampling rate, and downsampling. Based on the obtained results, it was concluded that the effects of SJ are virtually the same regardless of whether they take place at a DAC or an ADC. In addition, it was observed that oversampling at the ADC is ineffective against DAC jitter.

In future 6G ISAC systems, which are expected to operate at increasingly higher frequencies and require tighter synchronization margins, understanding the effects of SJ and its implications on not only communications, but also on the radar sensing performance becomes crucial. In addition, to better understand how sampling jitter can affect the performance of OFDM-based ISAC, the effect of OFDM signal parameters, i.e., number of subcarriers and modulation alphabet, than in the aforementioned studies needs to be considered. Furthermore, a SJ model that better reflects the behavior of state-of-the-art hardware must be adopted. Such model should, e.g., consider the use of PLL-based oscillators to derive DAC and ADC sampling clocks, which leads to colored SJ. In this sense, the impact of colored SJ raised by the use of a PLL-

based oscillator to generate the sampling clock of OFDM-based ISAC systems is investigated in the following, and later numerically analyzed in Section 4.2.4.

It is henceforth assumed that FDZP is performed on the discrete-frequency domain OFDM symbols contained in $\mathbf{X}$ to perform oversampling with factor $\eta \in \mathbb{N}_{>0}$. After further processing as described in Section 3.1, the discrete-time domain sequence $x[s]$ is ultimately generated. It then undergoes D/A conversion to produce in $x(t)$. If an ideal, bandlimited DAC to $f \in [-\eta F_s/2, \eta F_s/2]$ is used, and only the samples of $x[s]$ associated to a single OFDM symbol are considered for simplicity, this process can be mathematically described as

$$x(t) = \sum_{s=-\eta N_{\text{CP}}}^{\eta N-1} x[s] \frac{\sin(\pi(t - sT_s/\eta)/(T_s/\eta))}{\pi(t - sT_s/\eta)/(T_s/\eta)} \quad (3.79)$$

for $-N_{\text{CP}}T_s \leq t < NT_s$. In (3.79), $s \in \{-\eta N_{\text{CP}}, -\eta N_{\text{CP}} + 1, \dots, \eta N - 1\}$ is considered. More specifically, $s \in \{-\eta N_{\text{CP}}, \dots, -1\}$ are the discrete-time domain CP samples of the considered oversampled OFDM symbol, while $s \in \{0, \dots, \eta N - 1\}$ correspond to the remaining samples.

The expression in (3.79) is only valid for ideal sampling instants. It is henceforth assumed that the DAC sampling clock is derived from a PLL-based oscillator, which has a PN $\theta_{\text{DAC}}^{\text{PN}}(t) \in \mathbb{R}$ with double-sided PSD $S_{\theta_{\text{DAC}}^{\text{PN}}}(f) \in \mathbb{R}_{\geq 0}$ [Sma12, KKP⁺14, SSK⁺18b, vS23]. The presence of the PN $\theta_{\text{DAC}}^{\text{PN}}(t)$ in the oscillator used to derive the sampling clock ultimately leads to a SJ $\delta_{\text{DAC}}^{\text{SJ}}(t) \in \mathbb{R}$, whose relationship to the aforementioned PN can be expressed as [Sma12]

$$\delta_{\text{DAC}}^{\text{SJ}}(t) = \theta_{\text{DAC}}^{\text{PN}}(t)/(2\pi\eta F_s). \quad (3.80)$$

Considering the effect of SJ, $x(t)$ becomes $x^{\text{SJ}}(t) \in \mathbb{C}$, and (3.79) can be rewritten as

$$x^{\text{SJ}}(t) = \sum_{s=-\eta N_{\text{CP}}}^{\eta N-1} x[s] \frac{\sin\left(\pi\left[t - \left(sT_s/\eta + \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)\right)}{\pi\left[t - \left(sT_s/\eta + \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)}. \quad (3.81)$$

Assuming an ideal single-path channel, the receive BB signal $y(t)$ can simply be expressed as

$$y(t) = x^{\text{SJ}}(t). \quad (3.82)$$

After sampling at an ADC with SJ $\delta_{\text{ADC}}^{\text{SJ}}(t) \in \mathbb{R}$ also resulting from the use of a PLL-based oscillator to derive the sampling clock, the s' th sample $y[s']$, $s' \in \{-\eta N_{\text{CP}}, -N_{\text{CP}} + 1, \dots, \eta N - 1\}$, of the discrete-time domain receive sequence is obtained as

$$\begin{aligned} y[s'] &= y\left(s'T_s/\eta + \delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta)\right) \\ &= x^{\text{SJ}}\left(s'T_s/\eta + \delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta)\right). \end{aligned} \quad (3.83)$$

To expand (3.83), (3.82) and (3.81) are used, which results in

$$\begin{aligned} y[s'] &= \sum_{s=-N_{\text{CP}}}^{\eta N-1} x[s] \\ &\times \frac{\sin\left(\pi\left[\left(s'T_s/\eta + \delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta)\right) - \left(sT_s/\eta + \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)\right)}{\pi\left[\left(s'T_s/\eta + \delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta)\right) - \left(sT_s/\eta + \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)}. \end{aligned} \quad (3.84)$$

Next, the CP is removed from $y[s']$, and a DFT is performed to produce the discrete-frequency domain receive OFDM symbol $\mathbf{R} \in \mathbb{C}^{N \times 1}$, whose l th element, $l \in \{-\eta N/2, -\eta N/2 + 1, \dots, \eta N/2 - 1\}$, is expressed as

$$R_l = \sum_{s'=0}^{\eta N-1} y[s'] e^{j2\pi \frac{ls'}{\eta N}}. \quad (3.85)$$

The expression in (3.84) is then substituted into (3.85) to yield

$$\begin{aligned} Y_l &= \sum_{s'=0}^{\eta N-1} \left(\sum_{s=-\eta N_{\text{CP}}}^{\eta N-1} x[s] \right. \\ &\times \frac{\sin\left(\pi\left[\left(s' - s\right)T_s/\eta + \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)\right)}{\pi\left[\left(s' - s\right)T_s/\eta + \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]/(T_s/\eta)} \Bigg) \\ &\times e^{j2\pi ls'/(\eta N)}. \end{aligned} \quad (3.86)$$

Assuming $\omega = \pi/(T_s/\eta)$, the fraction in (3.86) can be rewritten as the following function $\varphi(s, s')$ of s and s'

$$\varphi(s, s') = \frac{\sin\left(\omega\left[(s' - s)T_s/\eta + \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]\right)}{\omega\left[(s' - s)T_s/\eta + \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta)\right)\right]}. \quad (3.87)$$

Expanding (3.87) via Taylor series and keeping only the first two terms yields

$$\begin{aligned} \varphi(s, s') &= \frac{\sin(\pi(s' - s))}{\pi(s' - s)} \\ &\quad - \left(\frac{\pi(s' - s) \cos(\pi(s' - s)) - \sin(\pi(s' - s))}{\pi^2(s' - s)^2} \right) \\ &\quad \times \omega \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta) \right). \end{aligned} \quad (3.88)$$

The latter equation can be further rewritten as

$$\varphi(s, s') = \begin{cases} 1, & s' = s \\ \frac{\cos(\pi(s' - s))}{\pi(s' - s)} \omega \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta) \right), & s' \neq s \end{cases}, \quad (3.89)$$

which allows rewriting (3.86) as

$$\begin{aligned} R_l &= X_l + \sum_{s'=0}^{\eta N-1} \left(\sum_{s=-N_{\text{CP}}, s \neq s'}^{\eta N-1} x[s] \right. \\ &\quad \times \left. \frac{\cos(\pi(s' - s))}{\pi(s' - s)} \omega \left(\delta_{\text{ADC}}^{\text{SJ}}(s'T_s/\eta) - \delta_{\text{DAC}}^{\text{SJ}}(sT_s/\eta) \right) \right) e^{j2\pi l s' / (\eta N)}. \end{aligned} \quad (3.90)$$

After discarding the subcarriers added at the transmitter side for FDZP, the discrete-frequency domain receive OFDM symbol $\mathbf{Y} \in \mathbb{C}^{N \times 1}$ is obtained. Note that although $\mathbf{Y}$ was used to denote the discrete-frequency domain OFDM frame containing multiple symbols in the previous subsections, this variable is repurposed here for simplicity. Since $(\eta - 1)N$ inactive subcarriers are added at the transmitter side for FDZP, $(\eta - 1)N/2$ to the left and $(\eta - 1)N/2$ to

the right of the discrete-frequency domain spectrum, the k th element $Y_k \in \mathbb{C}$, $k \in \{-N/2, -N/2 + 1, \dots, N/2 - 1\}$, of $\mathbf{Y}$ is expressed as

$$R_k = Y_k. \quad (3.91)$$

An analysis of the obtained expressions in (3.90) and (3.91) reveals that the non-ideal sampling results in an ICI term that depends on all discrete-time domain samples, including the ones in the CP, as well as their mutual leakage due to SJ at both DAC and ADC. If only SJ at the DAC is considered, it is known that unwanted OOB radiation will occur [HMAM15]. Combined with ADC SJ, this will not only result in ICI among the subcarriers of interest, but in more severe ICI due to the leakage into the subcarriers originally used for FDZP at the transmitter side that were supposed to be inactive. The aforementioned issues are the only impairments induced by SJ when BB sampling is performed. When BP sampling is adopted instead, the transmit samples undergo digital up-conversion (DUC) to an IF before D/A conversion at the transmitter side, and the receive samples undergo digital down-conversion (DDC) from the IF into the baseband after A/D conversion at the receiver side. Consequently, not only OOB radiation due to DAC SJ and ICI due to both DAC and ADC SJ occur, but also a PN-like effect is induced to the IF as discussed in [PF09], which in turn results in more ICI and CPE as discussed in Section 3.4.4 and [21].

3.5 Chapter conclusion

This chapter presented a system concept for bistatic OFDM-based ISAC based on a reference path between transmitter and receiver. This path is used to enable OTA synchronization, as well as to be used as a sensing reference with respect to which radar target parameters are estimated. After a comprehensive mathematical formulation of the considered system, two radar signal processing strategies were proposed. These are, namely, pilot-based processing, which relies solely on pilot subcarriers to enable sensing, and full-frame processing, which delivers superior radar sensing performance compared to the former approach, at the cost of requiring estimation of the full transmit frame through communications processing.

To ensure unbiased sensing and avoid overall performance degradation of the ISAC system, an OTA synchronization procedure capable of handling STO, CFO, and SFO was proposed. While part of this procedure is based on the use of conventional synchronization techniques such as S&C algorithm and correlation with known preamble OFDM symbols, a strategic combination of these algorithms is proposed for STO and CFO estimation and correction. In addition, and a novel algorithm for SFO estimation, namely TITO, was introduced. As a result, accurate synchronization can be achieved to not only meet typical communications system requirements but also the stricter demands of unbiased sensing.

Finally, an overview of hardware impairments caused by non-idealities in antennas and antenna arrays, amplifiers, mixers, oscillators, and DACs and ADCs is presented, and corresponding countermeasures are described. Some of these countermeasures are applied during the system and transmit signal design phase of the bistatic OFDM-based ISAC system, while others can be addressed through signal processing at the receiver side. Among the former are antenna element coupling in arrays, non-linear distortion in PAs, LNAs, and mixers, as well as quantization noise, clipping, the occurrence of image spectra and aliasing, and SJ during D/A and A/D conversion. As for the impairments that are at least partially compensable through signal processing techniques, they include beam squint, which must also be addressed at the transmitter side, frequency-selective I/Q imbalance, and PN. To complete the discussion presented in this chapter, a numerical analysis and a measurement-based verification are provided in Chapter 4.

4 Performance evaluation of bistatic OFDM-based ISAC

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In this chapter, the performance of the bistatic OFDM-based ISAC system presented in Chapter 3 under the effects of synchronization offsets and hardware impairments is analyzed. For conciseness, not all aspects discussed in the previous chapter will be analyzed as they have already been addressed in the literature. These are namely:

- **A performance analysis of the pilot-based and full-frame bistatic radar signal processing approaches.** Besides the differences in the achievable radar performance parameters listed in Table 3.1, a detailed investigation was reported in [6], and relevant findings therein were already discussed in Section 3.2.3.
- **A performance analysis of STO and CFO estimation in the proposed synchronization procedure.** The S&C algorithm [SC97] for coarse STO and CFO estimation has been already widely studied in the literature and even considered for 5G NR systems based on the primary synchronization signal (PSS) contained in the SSB [OSAA19]. STO fine tuning via correlation, as well as further STO and CFO refinement based on pilot subcarriers, has also been subject of extensive investigation [Sit17, SNZ18]. Since the synchronization procedure proposed in Section 3.3 simply combines these techniques, no further analyses are performed. Associated impairments with time and frequency offsets, namely ISI and ICI, are, however, analyzed as they may also happen not only due to synchronization offsets, but also due to delay and Doppler shift of targets.

- **A performance analysis of the bistatic OFDM-based system under the influence of several hardware impairments discussed in Section 3.4.**

These are beam squint, antenna element coupling within an array, as well as quantization noise, clipping, and issues arising from image spectra and aliasing in D/A and A/D conversion. Although relevant, the understanding of their effects on radar sensing performance is either straightforward and already explained in the aforementioned section or requires more detailed investigation that is out of the scope of this dissertation.

The performed investigations in this chapter assume the full-frame bistatic radar signal processing approach described in Section 3.2.2 and include an analysis of synchronization offsets and resulting impairments in Section 4.1 as well as an analysis of hardware impairments in Section 4.2. More specifically, brief but important remarks on ISI and ICI, besides an analysis of SFO, amplifier non-linear distortion, frequency-selective I/Q imbalance, PN, and SJ are presented as either no specific analysis was presented in the context of ISAC systems in the literature or not all relevant aspects have been addressed. Afterwards, a measurement setup and results are presented and discussed in Section 4.3 to verify the performance of the bistatic OFDM-based ISAC system with the proposed signal processing and synchronization concepts under occurrence of the previously discussed synchronization offsets and hardware impairments in Section 4.3. Finally, concluding remarks are provided in Section 4.4.

4.1 Numerical analysis of synchronization offsets and impairments

In this section, an analysis of synchronization offsets and related impairments is performed based on simulation results. In the context of time and frequency offsets, the influence of their associated impairments, namely ISI and ICI, on radar sensing performance are analyzed. Since these effects have already been investigated in the literature to a certain extent, only brief analyses of ISI and ICI are presented in Sections 4.1.1 and 4.1.2, respectively. Regarding SFO, a thorough analysis of its effects and resulting radar sensing performance degradation is presented in Section 4.1.3.

Table 4.1: Adopted OFDM signal parameters for the synchronization offset and impairment analyses. © 2025 IEEE. Reprinted, with permission, from [22].

Carrier frequency (f_c)	26.2 GHz
Frequency bandwidth (B)	500 MHz
No. of subcarriers (N)	2048
Subcarrier spacing (Δf)	244.14 kHz
CP length (N_{CP})	512
No. of OFDM symbols (M)	4096
Pilot spacing ($\Delta N_{pil}, \Delta M_{pil}$)	2, 4
No. of pilot subc. per pilot OFDM symbol (N_{pil})	1024
No. of OFDM symbols w/ pilots (M_{pil})	1024
Modulation alphabet	QPSK
Channel coding and code rate	LDPC, 2/3

For the conducted analyses, the OFDM signal parameters listed in Table 4.1 are adopted. According to Table 3.1 in Section 3.2, these parameters result in the communications and radar performance parameters listed in Table 4.2 assuming full-frame bistatic radar signal processing.

4.1.1 Intersymbol interference analysis

As discussed in Section 3.3.1, STO may lead to ISI if the excess delay goes beyond the CP duration and no countermeasures are implemented. In this case, two impairments are experienced: (i) the ISI itself due to samples from the previous OFDM symbol occupying the start of sample interval assigned to the current one, and (ii) the ISI-induced ICI due to the missing samples of the evaluated OFDM symbol at the start of its sample interval.

Based on the expressions from [WWS⁺23,GMH⁺25,XZZ25] and assuming that the full radar processing gain $G_p^{\text{full}} = 69.24$ dB is experienced against ISI and ISI-induced ICI due to their random nature, Figure 4.1 shows the normalized magnitudes for the desired signal, ISI, and ISI-induced ICI as a function of the

Table 4.2: Resulting communications and radar performance parameters with full-frame bistatic radar signal processing for the synchronization offset and impairment analyses. © 2025 IEEE. Reprinted, with permission, from [22].

Comm. data rate (100% duty cycle, $\mathcal{R}_{\text{comm}}$)	0.40 Gbit/s
Processing gain (G_p^{full})	69.24 dB
Range resolution (ΔR^{full})	0.60 m
Max. rel. unamb. range ($R_{\text{max,ua}}^{\text{full}}$)	1227.95 m
Max. rel. ISI-free range ($R_{\text{max,ISI}}^{\text{full}}$)	306.99 m
Doppler shift resolution (Δf_D^{full})	47.68 Hz
Max. unamb. Doppler shift ($f_{D,\text{max,ua}}^{\text{full}}$)	97.66 kHz
Max. ICI-free Doppler shift ($f_{D,\text{max,ICI}}^{\text{full}}$)	24.41 kHz

relative bistatic range. The obtained results assume a nearly ideal propagation through a single path, being delay associated to range the only experienced channel effect. As expected, the results show a decreasing desired signal power and increasing interference power after the maximum relative ISI-free range listed in Table 4.2, i.e., $R_{\text{max,ISI}}^{\text{full}} = 306.99$ m. At the maximum relative unambiguous range $R_{\text{max,ua}}^{\text{full}} = 1227.95$ m, these effects are the most severe, with a desired signal power loss of 12.04 dB and ISI and ISI-induced ICI levels after radar signal processing of -70.49 dB and -76.51 dB, respectively, with respect to the desired signal power level without ISI. Combining all effects under the simplified assumption that ISI and ISI-induced ICI are uncorrelated, an SIR of 57.48 dB is experienced for the considered path.

If ideal synchronization is assumed for the reference path, ISI will happen due to targets with relative bistatic range higher than $R_{\text{max,ISI}}^{\text{full}}$. For all targets at such ranges, desired target reflection power reduction, ISI, and ISI-induced ICI will be experienced. Consequently, the target reflection power in the ultimately obtained radar image will be reduced, and the overall interference level will increase based on the combination of ISI and ISI-induced ICI from all aforementioned targets. Due to the high processing gain experienced against these interference contributions, however, the most relevant ISI-induced effect is the desired target reflection power reduction. The increased interference level in

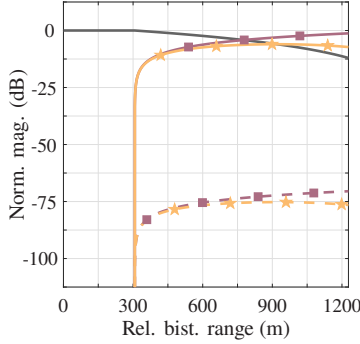


Figure 4.1. Calculated desired signal (—), ISI (■), and ISI-induced ICI (★) levels as functions of the relative bistatic range. For both ISI and ISI-induced ICI, continuous (—) and dashed (---) lines denote the normalized magnitude before and after experiencing the radar processing gain G_p^{full} , respectively.

the radar image only becomes an issue in scenarios where targets with highly different ranges and/or RCSs are to be detected [GMH⁺25].

It is worth highlighting that, compared to the monostatic case, reflections off targets whose range result in relevant ISI are less likely to arrive at the receiver of the considered bistatic OFDM-based ISAC system with relevant power. This is the case since the attenuation experienced by the aforementioned reflections is proportional to the actual bistatic range $R_p = R_{\text{Tx-T},p} + R_{\text{T-Rx},p}$, and not the relative bistatic range $R_p - R_0 = R_{\text{Tx-T},p} + R_{\text{T-Rx},p} - R_0$ as discussed in Section 3.1. Considering (3.9) and (3.10) and depending on the range R_0 of the reference path, it can be concluded that some targets reflections may experience severe attenuation if their relative bistatic ranges are higher than $R_{\text{max,ISI}}^{\text{full}}$. Consequently, even more negligible interference level increase due to ISI and ISI-induced ICI associated with such targets is experienced in the radar image, being the only relevant impairment a more severe peak power loss of these target reflections in the radar image as previously discussed.

4.1.2 Inter-carrier interference analysis

In Section 3.3.2, it was discussed that a CFO leads to ICI, which ultimately results in Doppler shift estimation bias and distortion of the obtained range-

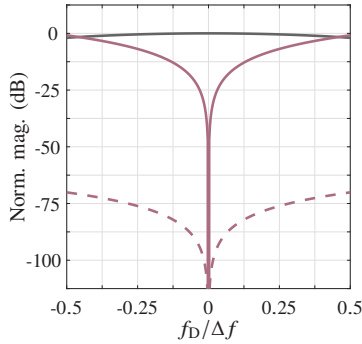


Figure 4.2. Calculated normalized magnitude of useful signal (—) and ICI before (— and ■) and after (— and ■) experiencing the radar processing gain G_p^{full} , respectively.

Doppler shift radar image after bistatic radar signal processing. In addition, ICI is also experienced due to the Doppler shift of targets even under perfect synchronization, which eliminates the CFO with respect to the reference path. Similarly to the ISI case from Section 4.1.1, this leads to peak power loss of the target reflection and interference level increase in the obtained range-Doppler shift radar image, which can also impact the ability of the bistatic OFDM-based ISAC system to detect other targets.

The severity of the aforementioned ICI-induced effects is analyzed in Figure 4.2 as functions of the normalized Doppler shift by the subcarrier spacing, i.e., $f_D / \Delta f$. The obtained results, i.e., the normalized magnitude of the useful signal and the ICI level before and after experiencing the full radar processing gain $G_p^{\text{full}} = 69.24$ dB, were obtained based on the closed-form expression for the ICI from [Sma12]. In OFDM-based communications systems, $|f_D / \Delta f| = 0.1$ is usually assumed as the maximum tolerable ICI, and this value is often assumed for OFDM-based radar sensing [NMZ18], resulting in the maximum tolerable relative radial velocity for the monostatic case listed in Table 2.2 in Chapter 3 and in the maximum tolerable Doppler shift for the bistatic case listed in Table 3.1 in Chapter 4. The results in Figure 4.2 show that this Doppler shift results in a negligible magnitude loss of only 0.07 dB for the useful signal component, but an ICI level only 14.83 dB below the ideal useful signal level, which can indeed result in significant communications performance degradation. In the radar sensing case, however, processing gain is experienced against ICI due to its

random nature, which decreased the ICI level to -84.07 dB with respect to the ideal useful signal level for the assumed parameters in Tables 4.1 and 4.2. For $|f_D/\Delta f| = 0.5$, which is the highest maximum unambiguous Doppler shift that can be experienced and according to Table 3.1 is only achieved for $\Delta M_{\text{pil}} = 1$ and $N_{\text{CP}} = 0$, a useful signal attenuation of only 1.96 dB is experienced. The ICI level, however, increased to -70.09 dB with respect to the ideal useful signal level. Consequently, it can be concluded that ICI only becomes a major factor for scenarios with target reflections that are already weak or when significant difference between receive power of target reflections is experienced and the strong targets are associated with Doppler shifts. In the latter case, the ICI caused by the Doppler shift of the strong targets may shadow the reflections of weaker targets in the radar image, therefore preventing their detection.

4.1.3 Sampling frequency offset analysis

In this section, the performance of the bistatic OFDM-based ISAC system with SFO correction based on an estimate obtained with the introduced TITO method is analyzed.

4.1.3.1 SFO-induced impairments

Based on the adopted parameters, the SFO converted from Hertz to δ , which represents the normalized SFO to the Nyquist sampling rate $F_s = B$, is shown in Figure 4.3. In this figure, the maximum normalized SFO of ± 123 ppm expected with the adopted measurement setup later described in Section 4.3 based on two distinct Zynq UltraScale+ RFSoc ZCU111 SoC platforms is also shown. This number derives from the stability of ± 61.5 ppm of the programmable clocks at each board, from which the sampling clocks are derived, and it represents a common normalized SFO expected in practice.

Next, the SFO-induced impairments to the discrete-frequency domain OFDM frame $\mathbf{Y}$ apart from ISI discussed in Section 3.3.3, namely the amplitude modulation α_k^{SFO} , the delay migration $\Delta \tau_m^{\text{SFO}}$, and the subcarrier frequency shift f_k^{SFO} are shown in Fig 4.4 for $|\delta| \in \{0, 0.1, 1, 10, 100, 1000\}$ ppm. It can be seen in Figure 4.4(a) that all considered SFO values except for $|\delta| = 1000$ ppm yield negligible amplitude modulation along the subcarrier index k . In Fig-

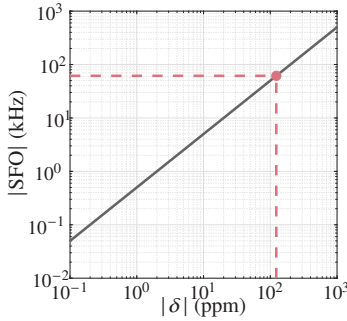


Figure 4.3: Equivalent SFO in kHz to the normalized SFO to the Nyquist sampling rate $F_s = B$ (—). For reference, the maximum expected absolute value of the SFO in the bistatic setup from Section 4.3 based on two Zynq UltraScale+ RFSoc ZCU111 SoC platforms is also shown (●). © 2025 IEEE. Reprinted, with permission, from [22].

ures 4.4(b) and (c), however, it is seen that $|\delta| \in \{100, 1000\}$ ppm yield delays and frequency shifts that exceed both the ICI- and ISI-free SFO limits defined according to (3.43) and (3.47), respectively. Based on the OFDM signal parameters listed in Table 4.1, the aforementioned SFO ranges for approximately ICI- and ISI-free operation are $-97.66 \text{ ppm} < \delta < 97.66 \text{ ppm}$ and $0 \text{ ppm} < \delta < 48.83 \text{ ppm}$, respectively. The achieved results show that only $|\delta| \in \{100, 1000\}$ ppm among the considered normalized SFO values exceed both ICI- and ISI-free SFO ranges due to their resulting delays and frequency shifts, respectively, as previously discussed and shown in Figures 4.4(b) and (c). Since the achieved ICI- and ISI-free SFO ranges do not fully cover the expected SFOs of up to $\pm 123 \text{ ppm}$ that may occur in the aforementioned measurement setup that will later be considered in Section 4.3, the SFO correction via the frequency-domain equalization from (3.50) is henceforth not considered and the resampling approach explained in Section 3.3.3.3 is adopted instead.

To clearly illustrate the effects of SFO on the receive QPSK constellation, simulation results for an AWGN channel with $\text{SNR} = 20 \text{ dB}$ and an SFO of $\delta = 1 \text{ ppm}$ are shown in Figure 4.5. The assumed channel model agrees with the assumption in Section 3.1 that the reference path between transmitter and receiver is much stronger than the propagation paths associated with radar targets and that perfect time and frequency synchronization with respect to this main path is performed. In the aforementioned figure, the ICI effect described in (3.27) and shown in Figure 4.4(c) can be observed for the OFDM symbol of

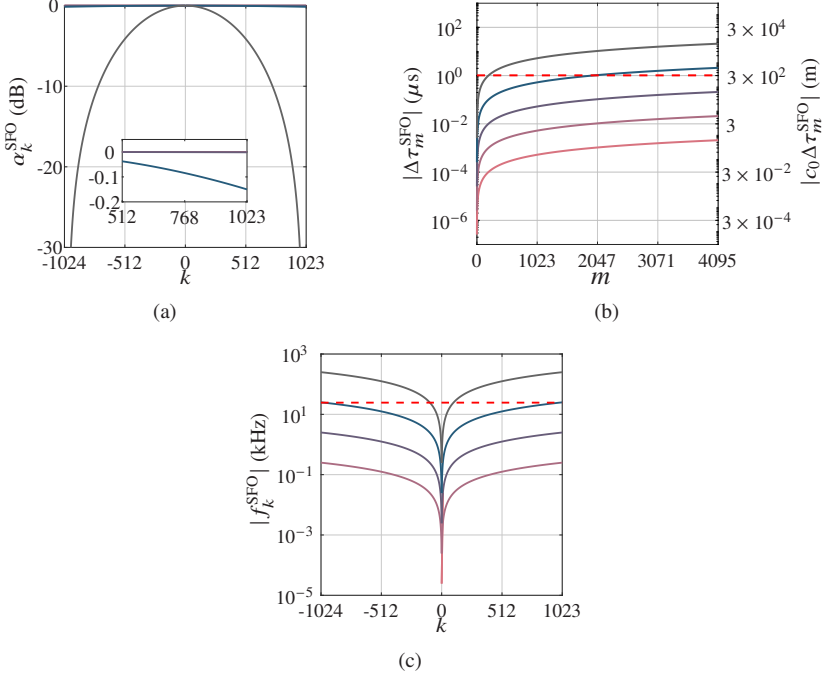


Figure 4.4: SFO effects: calculated (a) amplitude modulation α_k^{SFO} along the subcarriers, (b) absolute value of the delay migration $\Delta \tau_m^{\text{SFO}}$ along the OFDM symbols, and (c) absolute value of the subcarrier frequency shift f_k^{SFO} along the subcarriers, all evaluated for $|\delta| = 0.1$ ppm (—), $|\delta| = 1$ ppm (—), $|\delta| = 10$ ppm (—), $|\delta| = 100$ ppm (—), and $|\delta| = 1000$ ppm (—). For comparison with the achieved values, the maximum ISI-free delay of N_{CP}/B and range of $c_0 N_{\text{CP}}/B$ for the positive δ case, as well as the maximum tolerable frequency shift of $\Delta f/10$ defined in (3.42) which causes negligible ICI are marked (—) in (b) and (c), respectively. Adapted from [22], © 2025 IEEE.

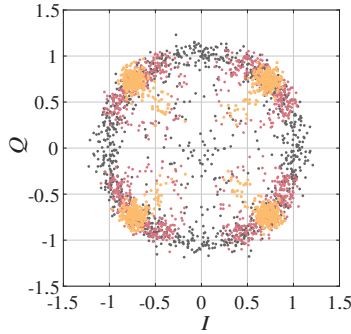


Figure 4.5: Simulated normalized receive QPSK constellations obtained from simulations for an AWGN channel with $\text{SNR} = 20$ dB and $\delta = 1$ ppm at OFDM symbols $m = 0$ (●, $\text{EVM} = -21.39$ dB), $m = 64$ (●, $\text{EVM} = -13.05$ dB), and $m = 4094$ (●, $\text{EVM} = 0.18$ dB). © 2025 IEEE. Reprinted, with permission, from [22].

index $m = 0$ in the form of spreading of some constellation points towards the origin, while the constellation for $m = 64$ is still clearly affected by ICI but also presents more relevant phase rotation due to the dependence of $\psi_{k,m}^{\text{SFO}}$ in (3.26) on the OFDM symbol index m . As for OFDM symbol $m = 4094$, both effects are present, but the higher SFO-induced delay results in more accentuated rotation of the QPSK constellation.

Regarding sensing performance, bistatic radar images assuming $\text{SNR} = 20$ dB before radar processing and perfect time and frequency synchronization with respect to the reference path are shown in Figure 4.6 for $\delta = 0$ ppm and $\delta = 1$ ppm. For these radar images, a radar target with $\text{SNR} = -10$ dB before radar processing, a relative range of 5 m, and a relative Doppler shift of 5 kHz with respect to the main path were assumed. In both range and Doppler shift directions, rectangular windowing was used to avoid that the SFO effects are masked by the use of a window to reduce sidelobe level. For simplicity, the presented radar images are obtained by processing all subcarriers. In practice, this would require pilot-based channel estimation and equalization to correctly estimate the QPSK data in payload subcarriers and only then allow full-frame bistatic radar signal processing as described in Section 3.2.2 and [6, 50]. For the presented analysis, however, *genie-aided* decoding is assumed for simplicity. The results in these figures show that range and Doppler shift migration occur for $\delta = 1$ ppm, spreading both the reference path and the radar

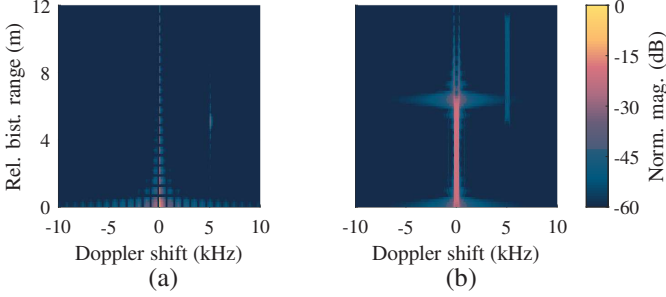


Figure 4.6: Simulated range-Doppler bistatic radar images with one reference path and a target at a relative bistatic range of 5 m and Doppler shift of 5 kHz for (a) $\delta = 0$ ppm and (b) $\delta = 1$ ppm. © 2025 IEEE. Reprinted, with permission, from [22].

target in the relative bistatic range and Doppler shift directions and reducing their magnitude and consequently SINR compared with the case with $\delta = 0$ ppm. These effects are respectively caused by delay and frequency shift migrations described by (3.29) and (3.31). A zoom in on the reference path for both cases is shown in Figure 4.7. In this figure, a range migration from 0 m to $c_0\{\delta[M(N + N_{CP})]T_s\} = 6.29$ m is observed, which agrees with the expected delay migration that can be calculated based on (3.29). Similarly, a Doppler shift migration from $\delta(-N/2)\Delta f = -0.25$ kHz to $\delta(N/2)\Delta f = 0.25$ kHz is experienced as predicted in (3.30) and (3.31). Besides range and Doppler shift migration, both the reference path and the radar target experience SINR loss due to the SFO-induced ICI. ISI, however, is not experienced since $\delta = 1$ ppm lies in the ISI-free SFO range $0 \text{ ppm} < \delta < 48.83 \text{ ppm}$ defined by (3.47) for the OFDM parameterization in the considered RIS-assisted bistatic ISAC system.

The presented QPSK constellations and radar images for $\delta = 1$ ppm show that even small SFOs can greatly impair both communications and radar sensing performances of bistatic ISAC systems. For the expected SFO of up to $\delta = \pm 123$ ppm in the measurement setup later described in Section 4.3, even higher degradation is expected since ISI can occur even for positive SFOs as indicated by Figure 4.4(b). To prevent this, the experienced SFO must be accurately estimated and corrected, which is addressed in Section 4.1.3.2.

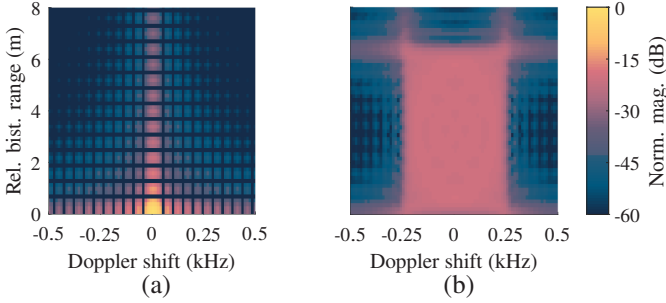


Figure 4.7: Zoom-in on reference path of the simulated radar images from Figure 4.6 for (a) $\delta = 0$ ppm and (b) $\delta = 1$ ppm. © 2025 IEEE. Reprinted, with permission, from [22].

4.1.3.2 SFO estimation

To perform SFO estimation, the approach based on the method by Wu et al. [WZL12], which is equivalent to the TITO method when forcing $M_{\text{pil}}^{\text{TITO}} = M_{\text{pil}}$, first estimates the delay migration, and then its slope along the pilot OFDM symbols. Only then the SFO is estimated via (3.35). Therefore, as discussed in Section 3.3.3, the delay migration estimates must be accurate in order for an accurate SFO estimate to be obtained. In this context, the root mean squared error (RMSE) of the SFO estimates obtained with the approach based on the method by Wu et al. [WZL12] are shown in Figure 4.8 for $\delta \in \{0.1, 1, 10, 100, 1000\}$ ppm. These results were obtained from the SFO estimates based on delay migration estimation with FDZP factor $\eta = 20$ for SNR = -30 dB through 30 dB. To reduce the computational complexity associated with the high FDZP factor η , the chirp Z-transform (CZT) can be used instead of the IDFT only around the expected reference path tap position in the columns of $\mathbf{D}$ [BMG⁺19].

The obtained RMSE results show a decreasing SFO estimation accuracy with increasing $|\delta|$. The only exception is $\delta = 0.1$ ppm, which yields worse RMSE than, e.g., $\delta = 1$ ppm and $\delta = 10$ ppm for SNR values above around -12 dB. This happens due to the limited delay estimation accuracy, which suffers from quantization depending on the FDZP factor η and becomes more critical due to the small delay migration induced by such a low SFO. As the SFO is based on delay migration, the accuracy of the SFO estimates is therefore also reduced.

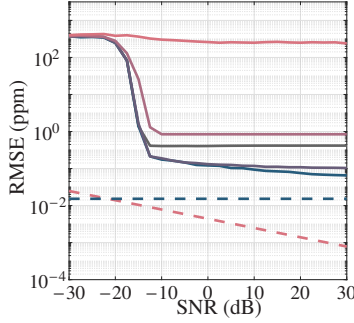


Figure 4.8: Simulated RMSE of delay-based SFO estimation based on [WZL12] for SNR = -30 dB through 30 dB and $\delta = 0.1$ ppm (—), $\delta = 1$ ppm (—), $\delta = 10$ ppm (—), $\delta = 100$ ppm (—), and $\delta = 1000$ ppm (—). For comparison, the approximate CRLB from (3.39) (—) and the lower bound for the MLE estimator with ZP from (3.41) (—) are also shown. Adapted from [22], © 2025 IEEE.

For all other SFO values, the RMSE degradation is primarily caused by the SINR reduction of the peaks corresponding to the reference path in the delay-slow time profile matrix $\mathbf{D}$ that are used to obtain the required delay migration estimates for SFO estimation. This reduction happens due to SFO-induced ICI and ISI as discussed in Section 4.1.3.1 and illustrated in Figure 4.9 assuming the same SNR conditions for the reference path and the radar target as for the results in Figure 4.6.

To analyze the SINR degradation more closely, the simulated SINR of the reference path estimated with each m_{pil} th pilot OFDM symbol, i.e., along the CIR estimates in the columns of the delay-slow time profile matrix $\mathbf{D}$, is shown in Figure 4.10 for SNR = 20 dB and $\delta \in \{\pm 0.1, \pm 1, \pm 10, \pm 100, \pm 1000\}$ ppm. To avoid biasing the SINR results, a Chebyshev window with 100 dB sidelobe suppression was used before forming the delay profiles at each column of $\mathbf{D}$, and the SINR was calculated as the ratio between the peak corresponding to the reference path and the region outside its mainlobe. The obtained results show that the SINR at $m_{\text{pil}} = 0$ decreases with increasing $|\delta|$. This is solely due to the SFO-induced ICI for positive δ , since the CP prevents ISI due to delay migration at this point, and due to both SFO-induced ICI and ISI for negative δ . A relevant SINR decrease, however, is only observed for $|\delta| = 1000$ ppm, while all the other SFOs result in negligible interference and therefore close SINR value to the expected SNR of 48.70 dB. This value is equal to the input SNR of 20 dB,

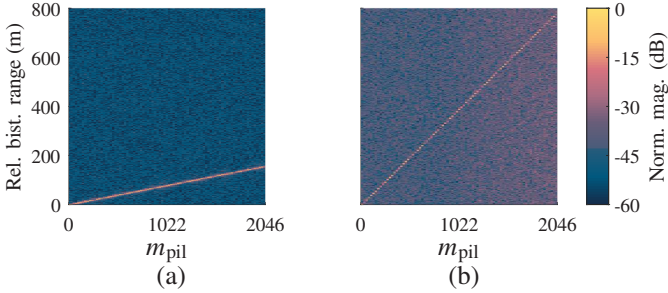


Figure 4.9: Simulated normalized delay-slow time profile assuming a reference path with $\text{SNR} = 20$ dB and a radar target with $\text{SNR} = -10$ dB at a relative bistatic range of 5 m and Doppler shift of 5 kHz for (a) $\delta = 100$ ppm and (b) $\delta = 500$ ppm. While a constant normalized magnitude ratio between the reference path and the noise floor is observed over all m_{pil} for $\delta = 100$ ppm, a smooth degradation is observed for $\delta = 500$ ppm starting at $m_{\text{pil}} \approx 1022$. © 2025 IEEE. Reprinted, with permission, from [22].

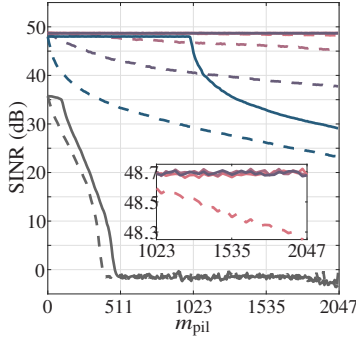


Figure 4.10: Simulated SINR of the corresponding peak to the estimated reference path in the CIR estimate at the m_{pil} th column of $\mathbf{D}$ used for obtaining the delay migration estimate $\widehat{\Delta\tau}_{\text{pil}, m_{\text{pil}}}^{\text{SFO}}$. Results are shown for $|\delta| = 0.1$ ppm (—), $|\delta| = 1$ ppm (—), $|\delta| = 10$ ppm (—), $|\delta| = 100$ ppm (—), and $|\delta| = 1000$ ppm (—), with continuous lines (—) for positive δ and dashed lines for negative δ (---). © 2025 IEEE. Reprinted, with permission, from [22].

plus the processing gain of $10 \log_{10}(N_{\text{pil}}) = 30.10$ dB, minus the windowing SNR loss of around 1.40 dB. To provide a more thorough analysis of the effect of the SFO-induced ICI, Figure 4.11 shows the SINR at $m_{\text{pil}} = 0$ as a function of δ . Since, as mentioned before, ISI already happens at $m_{\text{pil}} = 0$ for negative

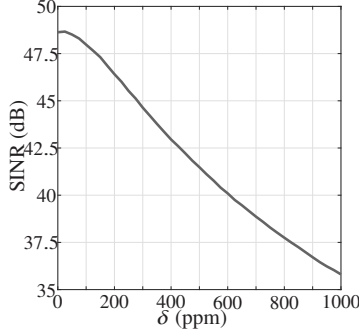


Figure 4.11: Simulated SINR as a function of the SFO δ for the corresponding peak to the estimated reference path in the CIR estimate at the first column of $\mathbf{D}$ ($m_{\text{pil}} = 0$), based on which $\widehat{\Delta\tau}_{\text{pil},0}^{\text{SFO}}$ is estimated. © 2025 IEEE. Reprinted, with permission, from [22].

δ among the considered values, only positive δ is analyzed for the presented results in this figure. An SINR degradation due to SFO-induced ICI of 1 dB is only experienced at 125 ppm, which is already slightly above the maximum of 123 ppm expected in the setup later considered in Section 4.3. Moreover, a SINR degradation of 3 dB or higher is only experienced for $\delta \geq 250$ ppm. This allows concluding that only rather high SFOs lead to significant SINR loss due to the induced ICI. Back to Figure 4.10, it is observed that the SINR continuously drop for negative δ along with m_{pil} , since all OFDM symbols in the frame are affected by the aforementioned SFO-induced ISI, which becomes more severe with increasing m_{pil} as the absolute value of the delay caused by SFO from (3.28) increases. Under influence of positive δ , however, the SINR only starts to drop after the SFO-induced delay exceeds the CP duration, only then causing ISI.

The discussed results so far have indicated that the SFO estimates with the approach based on the method by Wu et al. [WZL12] may be significantly biased even without relevant influence of SFO-induced ICI when ISI eventually happens, as shown in Figure 4.10. This is caused by the use of the CIR estimates in the columns of the delay-slow time profile matrix $\mathbf{D}$ that correspond to pilot OFDM symbols of index m_{pil} that are affected by ISI. As discussed in Section 3.3.3.2, this is circumvented with the TITO method by setting an upper bound to the step of the reference path delay migration between two consecutive estimates. Once this threshold is exceeded, the corresponding pilot

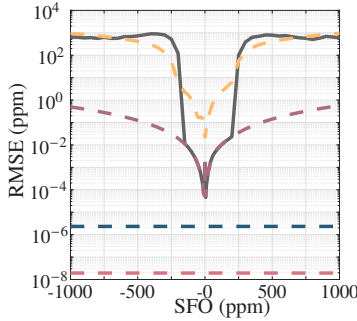


Figure 4.12: Simulated RMSE of SFO estimates at SNR = 20 dB for $\delta = -1000$ ppm through 1000 ppm with the proposed TITO method (—), the approach based on the method by Wu et al. [WZL12] (---), and the method by Tsai et al. [TKC05] (---). For further comparison, the CRLB from (3.39) (---) and the lower bound for the MLE estimator with ZP from (3.41) (---) are also shown. © 2025 IEEE. Reprinted, with permission, from [22].

OFDM symbol m_{pil} and all next ones are assumed to be under strong SFO-induced ISI and are therefore disregarded, leading to a total number of used pilot OFDM symbols $M_{\text{pil}}^{\text{TITO}} \leq M_{\text{pil}}$ for SFO estimation with the TITO method.

Assuming SNR = 20 dB as in the previous analysis in this section, Figure 4.12 shows the achieved RMSE of the SFO estimates with both the approach based on the method by Wu et al. [WZL12] and the TITO method assuming $\delta_{\text{max}} = 1000$ ppm and $\varepsilon = 0.1$ for $\delta = -1000$ ppm through $\delta = 1000$ ppm. For comparison, the performance of the method by Tsai et al. [TKC05] is also shown. The achieved results show that, between around $\delta = -150$ ppm and $\delta = 200$ ppm, TITO yields equivalent performance to the delay-based method based on Wu et al. [WZL12], while the method by Tsai et al. [TKC05] performs the worst. When analyzing the reference path in the obtained CIR estimates stored in the columns of $\mathbf{D}$ in the former two methods, it can be seen that the experienced processing gain of $10 \log_{10}(N_{\text{pil}}) = 30.10$ dB is sufficient to avoid severe degradation of the reference path SINR and consequently of the SFO estimates due to the SFO-induced ISI in this SFO region. As discussed in Section 4.1.1, this is the case as ISI does not immediately introduces severe degradation, but rather becomes more relevant with increasing excess delay or range. The method by Tsai et al. [TKC05], however, does not profit from this processing gain as it operates on the subcarrier level. Outside this SFO

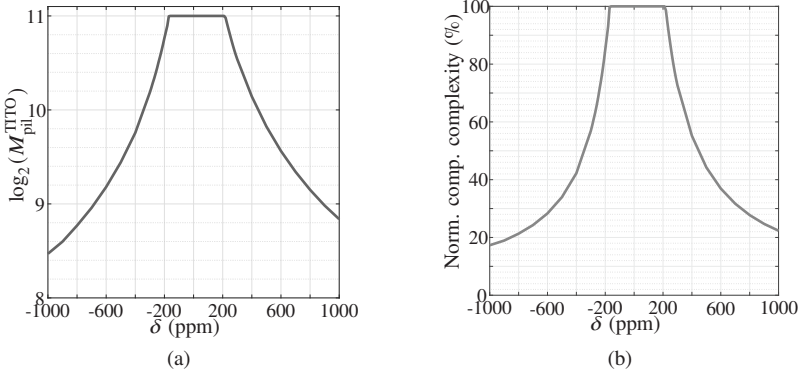


Figure 4.13: Simulated reduction in the number of used pilot OFDM symbols for SFO estimation with the TITO method at SNR = 20 dB: (a) total number $M_{\text{pil}}^{\text{TITO}}$ of used pilot OFDM symbols as a function of the SFO δ and (b) normalized computational complexity compared with the approach based on the method by Wu et al. [WZL12]. Between around $\delta = -150$ ppm and $\delta = 200$ ppm, there is a plateau with $M_{\text{pil}}^{\text{TITO}} = M_{\text{pil}}$. For these SFO values, all pilot OFDM symbols are used for the SFO estimation with the TITO method and, consequently, no computational complexity reduction is experienced. © 2025 IEEE. Reprinted, with permission, from [22].

region, the RMSE of the method based on Wu et al. [WZL12] is significantly worsened by the ISI-contaminated pilot OFDM symbols used to estimate the aforementioned CIRs. While the Tsai et al. [TKC05] method achieves equal or better performance, it is also affected by SFO-induced ISI and its RMSE is still rather high, i.e., 5.6 ppm or more. The TITO method, in turn, circumvents the ISI issue with its choice of $M_{\text{pil}}^{\text{TITO}} \leq M_{\text{pil}}$ pilot OFDM symbols used for the delay migration estimation required to estimate the SFO as described by (3.49) and illustrated in Figure 4.13(a). In Figure 4.13(b), the resulting computational complexity experienced as less pilot OFDM symbols are processed at high SFO values is shown. The obtained results allow concluding that the performance of TITO is mainly affected by the SFO-induced ICI and the reduced number of used pilot OFDM symbols, which affects the SFO estimation accuracy as illustrated in Figure 4.14. Still, TITO outperforms both its counterparts for the adopted OFDM signal parameters, presenting RMSEs as low as 0.01 ppm for $|\delta| = 150$ ppm and around 0.5 ppm for $|\delta| = 1000$ ppm.

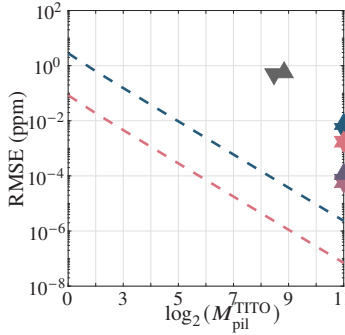


Figure 4.14: Simulated RMSE of the SFO estimates at SNR = 20 dB and the total number $M_{\text{pil}}^{\text{TITO}}$ of used pilot OFDM symbols for SFO estimation with the TITO method for $\delta = -1000$ ppm ($\blacktriangledown$), $\delta = -100$ ppm ($\blacktriangledown$), $\delta = -10$ ppm ($\blacktriangledown$), $\delta = -1$ ppm ($\blacktriangledown$), $\delta = -0.1$ ppm ($\blacktriangledown$), $\delta = 0.1$ ppm ($\blacktriangle$), $\delta = 1$ ppm ($\blacktriangle$), $\delta = 10$ ppm ($\blacktriangle$), $\delta = 100$ ppm ($\blacktriangle$), and $\delta = 1000$ ppm ($\blacktriangle$). For comparison, the CRLB from (3.39) (—) and the lower bound for the MLE estimator with ZP from (3.41) (—) as functions of $M_{\text{pil}}^{\text{TITO}}$ are also shown. © 2025 IEEE. Reprinted, with permission, from [22].

4.2 Numerical analysis of hardware impairments

In this section, a numerical analysis of the hardware impairments discussed in Section 3.4 is presented. More specifically, both the receive constellation distortion and the radar sensing performance degradation as consequences of amplifier non-linear distortion, mixer I/Q imbalance, oscillator PN, as well as DAC and ADC SJ are analyzed via simulations in Sections 4.2.1 to 4.2.4, respectively. Finally, remarks on the obtained results are presented in Section 4.2.5.

4.2.1 Amplifier non-linear distortion analysis

This section presents a brief analysis of the effects of amplifier non-linear distortion on the communications and sensing performances of the considered bistatic OFDM-based ISAC system assuming ideal synchronization and disregarding all other hardware impairments and AWGN. For that purpose, the

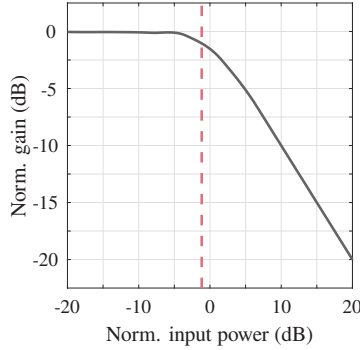


Figure 4.15. Simulated amplifier normalized gain as a function of the normalized input power derived from its AM-AM characteristic. For convenience, the IP1dB is highlighted (—).

memoryless amplifier model based on a lookup table (LUT) from [Mat12] was extrapolated to generate AM-AM and AM-PM characteristics resulting in the curves in Figures 4.15 and 4.16 is adopted. Instead of absolute input powers and gains, a generic model with normalized values was adopted. In addition, the same OFDM signal parameters as for the SFO analysis in Section 4.1.3, which are listed in Table 4.1, are adopted. The only differences are a reduced number of $M = 128$ OFDM symbols and that *genie-aided* decoding, i.e., full knowledge of the transmit data, is assumed. The first point is to keep the same number of OFDM symbols used for further hardware impairment analyses in the next sections. The lower number of OFDM symbols is assumed to reduce the rather high computational power associated with the SJ simulations presented later in Section 4.2.4. As for the second point, it results in ideal reconstruction of the transmit frame for the full-frame bistatic radar signal processing described in Section 3.2.2 and [6, 50] is assumed.

4.2.1.1 Out-of-band emission

The following analysis is performed for different IBO values rather than input powers, which are calculated based on (3.52). For the IBO values of -10 dB, 0 dB, and 10 dB, which correspond to operation points deep in saturation, at exactly the IP1dB, and in the linear region, the transmit signal spectra shown

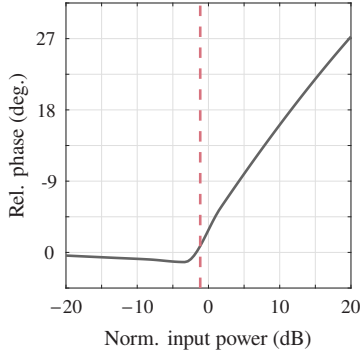


Figure 4.16. Simulated amplifier AM-PM characteristic showing the relationship between relative phase and normalized input power. For convenience, the IP1dB is highlighted (—).

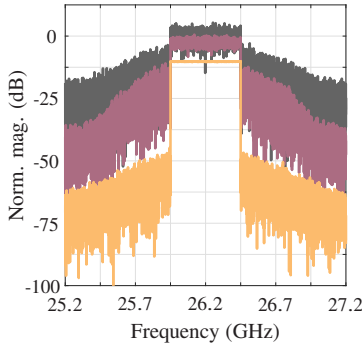


Figure 4.17. Simulated transmit signal spectrum under amplifier non-linearity for IBO = -10 dB (—), IBO = 0 dB (—), and IBO = 10 dB (—).

in Figure 4.17 are obtained. This figure shows that IBO = 10 dB results in a clean in-band spectral content OFDM signal, besides the lowest OOB level. At IBO = 0 dB, an amplification gain of 10 dB with respect to the previous case can be observed. In-band interference and increased OOB emission are, however, observed. As for IBO = -10 dB, no further amplification is experienced, being more pronounced in-band and OOB interference the only observed effects. While considering OOB emission is important for assessing amplifier efficiency and interference management, further analysis in this section will focus on in-band interference caused by amplifier non-linear distortion.

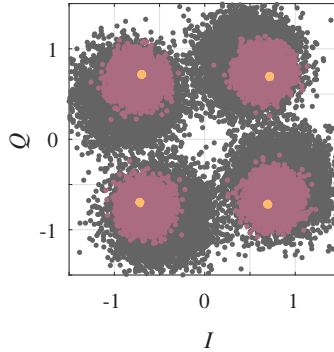


Figure 4.18. Simulated receive QPSK constellations for IBO = -10 dB ($\bullet$), IBO = 0 dB ($\bullet$), and IBO = 10 dB ($\bullet$).

4.2.1.2 Constellation distortion

Next, Figure 4.18 shows receive QPSK constellations for the same previously considered IBO values assuming an ideal channel for the reference path only. The results show negligible constellation distortion for IBO = 10 dB, and more spread but still mostly separated clusters for IBO = 0 dB. As for operation in the saturation region with IBO = -10 dB, more severe spreading of the constellation clusters is observed, besides rotation due to the AM-PM characteristic of the amplifier previously shown in Figure 4.16. When considering further hardware impairments and attenuation due to propagation, not only this operation point, but also the latter with IBO = 0 dB result in non tolerable constellation distortion. This distortion is quantitatively assessed using the mean EVM as a metric, as shown in Figure 4.19. The obtained results show that an IBO between around 10 dB and 15 dB, which is on the same order of magnitude as the OFDM signal PAPR for the adopted parameters as previously analyzed in Figure 2.21 in Section 2.5.1.1, is sufficient to ensure tolerable constellation distortion. This is supported by the obtained mean EVM values between -34.65 dB and -37.98 dB for this IBO range.

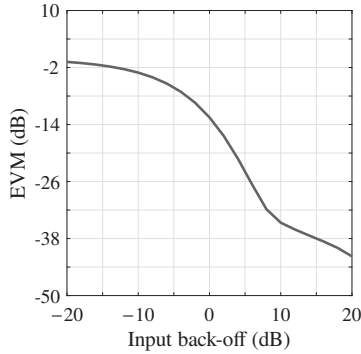


Figure 4.19. Simulated mean EVM as a function of the IBO.

4.2.1.3 Effects on radar sensing

While the observed communications performance degradation aligns with expectations based on existing studies on OFDM-based systems, there is limited insight into sensing performance degradation in the literature. In this sense, Figure 4.20 shows obtained bistatic range-Doppler shift radar images considering only the reference path under the influence of amplifier non-linear distortion for the IBO values of -10 dB, 0 dB, and 10 dB. To focus only on the interference introduced by amplifier non-linear distortion, both range and Doppler shift sidelobes were suppressed by Chebyshev windowing with 100 dB sidelobe suppression. It can be seen that, although lower IBO leads to higher interference floor in the radar image, a dynamic range of around 60 dB is still obtained in the worst case, i.e., $\text{IBO} = -10$ dB. This is explained by the fact that amplifier non-linear distortion tends to be random due to the random nature of modulation symbols carried by the OFDM subcarriers. Consequently, the radar processing gain is effective against amplifier non-linear distortion. This is confirmed by the mean image SIR results shown in Figure 4.21, where $N = 256$ and $N = 16384$ are also considered besides the already adopted number of subcarriers $N = 2048$. The image SIR provides a measure between a target mainlobe peak and the interference induced by the amplifier non-linear distortion in the radar image. The results show that the image SIR increases by approximately 9 dB when N is increased from $N = 2048$ to $N = 16384$, and similarly from $N = 256$ to $N = 2048$. This aligns with the expected radar processing gain in-

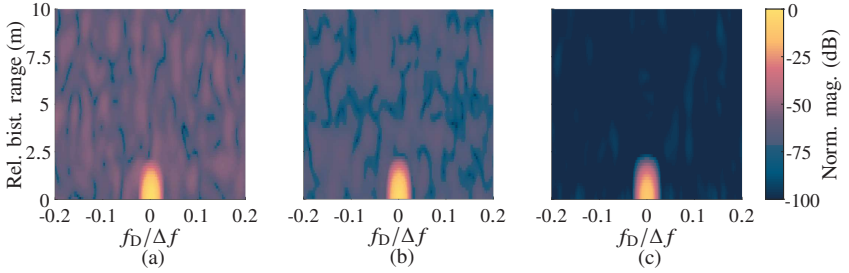


Figure 4.20: Simulated bistatic range-Doppler radar images under the influence of amplifier non-linear distortion obtained with $N = 2048$, $N_{CP} = 512$, $M = 128$, QPSK modulation, and Chebyshev windowing with 100 dB sidelobe suppression in both range and Doppler shift directions. Only the reference path at a relative bistatic range of 0 m and with a Doppler shift of 0 kHz was considered. The considered IBOs in the simulations were (a) IBO = -10 dB, (b) IBO = 0 dB, and (c) IBO = 10 dB.

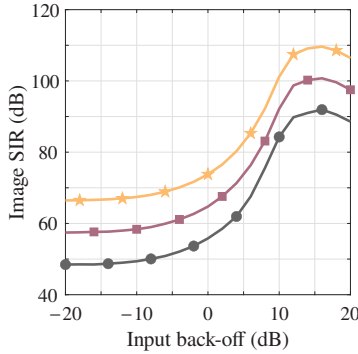


Figure 4.21: Simulated mean image SIR as a function of the IBO for $N = 256$ (●), $N = 2048$ (■), and $N = 16384$ (★).

crease of $10 \log_{10}(16384/2048) = 10 \log_{10}(2048/256) = 9.03$ dB. As for the communications case, the obtained results show that IBOs between around 10 dB and 15 dB prevent relevant sensing performance degradation due to amplifier non-linear distortion. At this region, the maximum values are obtained for the mean image SIR, with around 100 dB for $N = 2048$. The subsequent slight decrease in the mean image SIR is due to phase variations seen in the linear region of the amplifier AM-PM characteristic in Figure 4.16, which is due to small inaccuracies of the adopted model.

4.2.2 Mixer I/Q imbalance analysis

Although the effects of frequency-selective I/Q balance are well study for OFDM-based communications system in the literature and some analyses have been presented for OFDM-based radar [SSK⁺18b, SKSW20], this section presents a concise analysis covering both communications and sensing aspects of the considered bistatic OFDM-based ISAC system. As in Section 4.2.1, the OFDM signal parameters Table 4.1 with the exception of $M = 128$ OFDM symbols are adopted, and full-frame bistatic radar signal processing with *genie-aided* decoding is assumed. Furthermore, only the reference path is henceforth considered for both communications and radar analyses as it is expected to be the strongest among all paths in the considered bistatic OFDM-based ISAC system. This results in the fact that the relative CIR between I and Q branches of the mixer is equal to $h_{\Delta}^{I/Q\text{-imbal}}(t) = \delta(t)$. In addition, ideal synchronization is assumed and no further hardware impairments or AWGN are considered. The presented analysis in the following show the effects of frequency-selective I/Q imbalance both before and after applying the blind estimation and compensation approach from [SKSW20].

4.2.2.1 Constellation distortion

Figure 4.22(a) shows simulated receive QPSK constellations obtained without I/Q imbalance compensation for three cases. These are namely high amplitude imbalance and no phase imbalance ($\alpha_{\Delta}^{I/Q\text{-imbal}} = 3 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^{\circ}$), no amplitude imbalance and high phase imbalance ($\alpha_{\Delta}^{I/Q\text{-imbal}} = 0 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^{\circ}$), as well as low amplitude and phase imbalances ($\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 1^{\circ}$). In the latter case, the shift of the constellation points from their expected positions is negligible, making it a suitable baseline for comparison with the other two cases. As for the first case, i.e., $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^{\circ}$, the constellation points remain in the correct decision region as no phase imbalance is experienced. They are, however, shifted away from their expected positions. The second combination of imbalances, i.e., $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^{\circ}$, leads to the worst distortion of the receive QPSK constellation. While the receive constellation for this case seems less distorted than in the former case at first

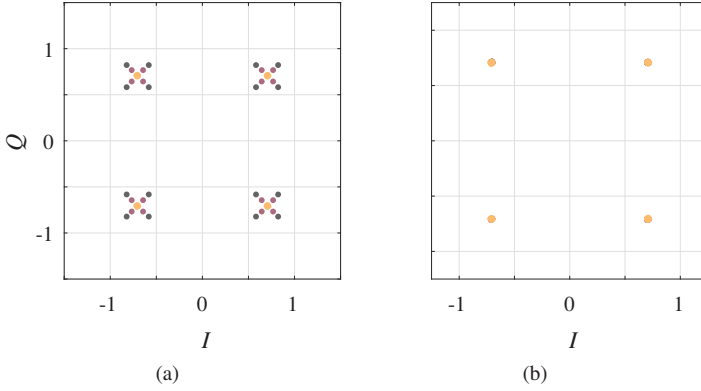


Figure 4.22. Simulated receive QPSK constellations (a) without and (b) with compensation I/Q imbalance for $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^\circ$ (●), $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^\circ$ (●), as well as $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1 \text{ dB}$ and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 1^\circ$ (●).

sight, the observed points actually originate from clusters in neighboring decision regions. This happens since the phase imbalance $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^\circ$ is higher than the angular distance of 90° between neighbor clusters. For all considered cases, virtually ideal constellations are obtained after I/Q imbalance compensation as seen in Figure 4.22(b).

To provide a quantitative analysis of the constellation distortion induced by I/Q imbalance, Figure 4.23 shows the EVM as a function of the phase imbalance $\varphi_{\Delta}^{I/Q\text{-imbal}}$ for several amplitude imbalances $\alpha_{\Delta}^{I/Q\text{-imbal}}$. The obtained results show that smaller EVMs than 30 dB are only obtained for $\alpha_{\Delta}^{I/Q\text{-imbal}}, \varphi_{\Delta}^{I/Q\text{-imbal}} \leq 1 \text{ dB}$ and phase imbalances of units of degrees. For higher amplitude imbalances or $\varphi_{\Delta}^{I/Q\text{-imbal}} \geq 10^\circ$, non-negligible constellation distortion is indicated by obtained EVMs of -20 dB or higher. For all considered amplitude imbalances, i.e., $\alpha_{\Delta}^{I/Q\text{-imbal}} \in \{0 \text{ dB}, 0.1 \text{ dB}, 1 \text{ dB}, 3 \text{ dB}\}$, and $\varphi_{\Delta}^{I/Q\text{-imbal}} \leq 10^\circ$, the blind I/Q estimation and compensation approach from [SKSW20] is able to ensure EVMs equal to or smaller than -30 dB and therefore negligible constellation distortion. While this does not apply for $\varphi_{\Delta}^{I/Q\text{-imbal}} \geq 10^\circ$, it is worth highlighting that such high phase imbalances are not expected in practice.

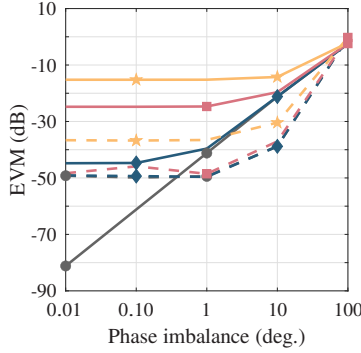


Figure 4.23. Simulated EVM as a function of the phase imbalance $\varphi_{\Delta}^{I/Q\text{-imbal}}$ for $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0$ dB (●), $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1$ dB (◆), $\alpha_{\Delta}^{I/Q\text{-imbal}} = 1$ dB (■), and $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3$ dB (★). Continuous and dashed lines show the results without (—) and with (---) I/Q imbalance compensation, respectively.

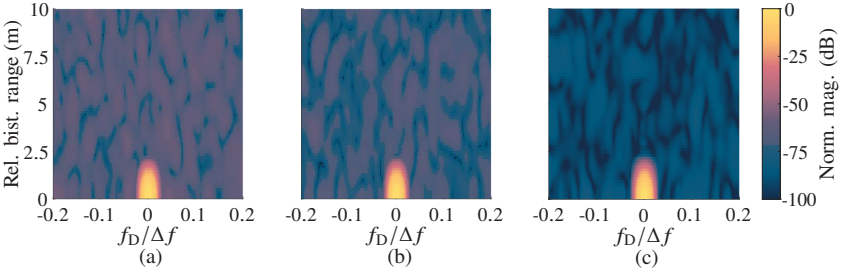


Figure 4.24: Simulated bistatic range-Doppler radar images under the influence of I/Q imbalance obtained with $N = 2048$, $N_{CP} = 512$, $M = 128$, QPSK modulation, and Chebyshev windowing with 100 dB sidelobe suppression in both range and Doppler shift directions. Only the reference path at a relative bistatic range of 0 m and with a Doppler shift of 0 kHz was considered. The considered amplitude and phase imbalances in the simulations were (a) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^{\circ}$, (b) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^{\circ}$, and (c) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 1^{\circ}$.

4.2.2.2 Effects on radar sensing

To analyze the effects of I/Q imbalance on the radar sensing performance of the bistatic OFDM-based ISAC system, Figure 4.24 shows range-Doppler shift obtained for the same amplitude and phase imbalance pairs considered

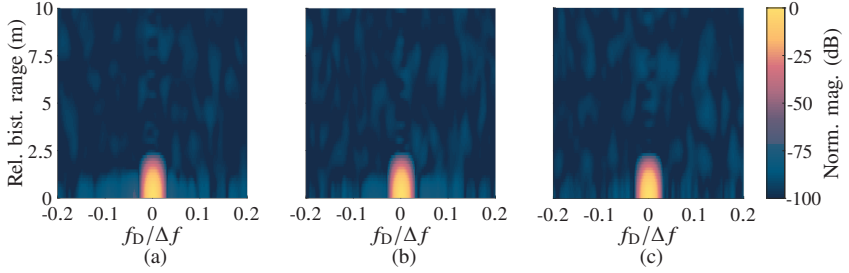


Figure 4.25: Simulated bistatic range-Doppler radar images obtained after I/Q imbalance compensation with $N = 2048$, $N_{CP} = 512$, $M = 128$, QPSK modulation, and Chebyshev windowing with 100 dB sidelobe suppression in both range and Doppler shift directions. Only the reference path at a relative bistatic range of 0 m and with a Doppler shift of 0 kHz was considered. The considered amplitude and phase imbalances in the simulations were (a) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^\circ$, (b) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^\circ$, and (c) $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 1^\circ$.

in the communications analysis. As in Section 4.2.1.3, Chebyshev windowing with 100 dB sidelobe suppression was used during both range and Doppler shift processing to focus on the interference introduced by I/Q imbalance. The highest interference floor among the obtained images is observed for $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 0^\circ$, followed by the case with $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 100^\circ$. For low amplitude and phase imbalances, which is the case for $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} = 1^\circ$, the lowest interference floor is observed. After I/Q imbalance compensation, the region outside the reference path mainlobe in the radar images is primarily dominated by the suppressed sidelobes resulting from Chebyshev windowing as shown in Figure 4.25. Only minor remnants of I/Q imbalance-induced interference are still visible.

The analysis of radar sensing performance degradation due to I/Q imbalance is complemented by the dependency of image SIR on amplitude and phase imbalances shown in Figure 4.26. The obtained results show that an image SIR of 90 dB or higher is obtained for $\alpha_{\Delta}^{I/Q\text{-imbal}} \leq 0.1$ dB and $\varphi_{\Delta}^{I/Q\text{-imbal}} \leq 1^\circ$. If either amplitude or phase imbalances are increased, significant image SIR degradation is observed. By applying the blind I/Q estimation and compensation approach from [SKSW20], however, image SIRs equal to or higher than 90 dB can be ensured for all considered amplitude imbalances, namely

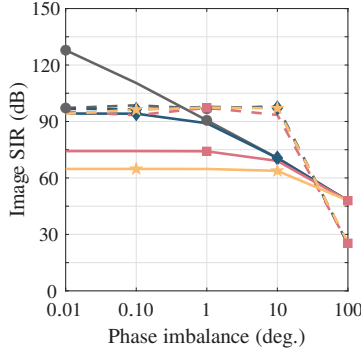


Figure 4.26. Simulated image SIR as a function of the phase imbalance $\varphi_{\Delta}^{I/Q\text{-imbal}}$ for $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0$ dB (●), $\alpha_{\Delta}^{I/Q\text{-imbal}} = 0.1$ dB (◆), $\alpha_{\Delta}^{I/Q\text{-imbal}} = 1$ dB (■), and $\alpha_{\Delta}^{I/Q\text{-imbal}} = 3$ dB (★). Continuous and dashed lines show the results without (—) and with (---) I/Q imbalance compensation, respectively.

$\alpha_{\Delta}^{I/Q\text{-imbal}} \in \{0 \text{ dB}, 0.1 \text{ dB}, 1 \text{ dB}, 3 \text{ dB}\}$, and $\varphi_{\Delta}^{I/Q\text{-imbal}} \leq 10^{\circ}$. As discussed in Section 4.2.2.1, this covers commonly expected impairments in practice.

4.2.3 Oscillator phase noise analysis

In this section, the performance of mmWave OFDM-based ISAC under PN is analyzed. Both monostatic and bistatic architectures are considered, and the PN-induced degradation of both communications and radar sensing performances is investigated. For that purpose, a corner frequency of $f_{\text{corner}} = 10$ kHz and a PLL loop bandwidth of $B_{\text{PLL}} = 150$ kHz were assumed, besides the in-band PN level of $L_0 = -105$ dBc/Hz and the PN floor $L_{\text{floor}} = -155$ dBc/Hz. The aforementioned parameters were chosen in order to simulate a PN PSD that is in between the specified ones of the E8257D PSG Microwave Analog Signal Generator from Keysight Technologies [Key25] for 20 GHz and 40 GHz. Those frequencies cover the initial part of the 5G NR FR2, which was used in the setups for ISAC measurements in previous works [13, 22]. In addition, the simulations in this sections consider the listed OFDM signal parameters in Table 4.3, with the specific values used mentioned in each individual analysis.

Table 4.3: Adopted OFDM signal parameters for PN analysis. Adapted from [21], licensed under CC BY-NC-ND 4.0.

Carrier frequency (f_c)	26.2 GHz
Bandwidth (B)	500 MHz
No. of subcarriers (N)	{256, 512, 1024, 2048, 4096, 8192, 16384}
CP length (N_{CP})	{0, $N/4$, N }
No. of OFDM symbols (M)	{32, 128, 512, 2048}
Modulation alphabet	{QPSK, 16-QAM, 64-QAM, 256-QAM}

Based on the assumed parameters, the resulting double-sided PN PSD, which is assumed to be the same for both $S_{\theta_{Tx}^{PN}}(f)$ and $S_{\theta_{Rx}^{PN}}(f)$, and the probability density function (PDF) of the PN time series are shown in Figures 4.27(a) and 4.27(b), respectively. It is worth highlighting that this model is assumed for both transmitter and receiver. However, the interaction of both aforementioned PN terms will depend on whether a monostatic or bistatic ISAC architecture is assumed. In addition, the results in Figure 4.28 reveal the relevance of B_{PLL} and its interplay with the subcarrier spacing Δf . Figure 4.28(a) shows the integrated PN level as a function of the frequency offset, revealing that a PN level of -52.78 dBc, i.e., around 4.88 dB below the total PN level of -47.90 dBc, is accumulated by a frequency offset equal to B_{PLL} as expected from the discussion in Section 3.4.4. This level is calculated by integrating the PN PSD occupied bandwidth assumed alongside the parameters listed in Table 4.3, i.e., $B = 500$ MHz. In Figure 4.28(b), the subcarrier spacing Δf is shown as a function of the assumed numbers of subcarriers N in Table 4.3. For better visualization, the N axis is shown in $\log_2$ scale, which results in $\log_2(\{256, 512, 1024, 2048, 4096, 8192, 16384\}) = \{8, 9, 10, 11, 12, 13, 14\}$. It is seen that only $N \leq 2048$ yield higher Δf than B_{PLL} . A more relevant ratio, which is the one between half of the subcarrier spacing, i.e., $\Delta f/2$, and B_{PLL} as discussed in Section 3.4.4, is shown in Figure 4.28(c). It can be observed that $N \in \{256, 512, 1024\}$ yield $(\Delta f/2)/B_{PLL} > 1$, which implies that CPE is the dominant PN-induced effect for these numbers of subcarriers. However, for $N \in \{2048, 4096, 8192, 16384\}$, it holds that $(\Delta f/2)/B_{PLL} < 1$. Consequently, the PN-induced ICI becomes the dominant effect. This happens since the subcarrier spacing Δf decreases with increasing N , leading to more

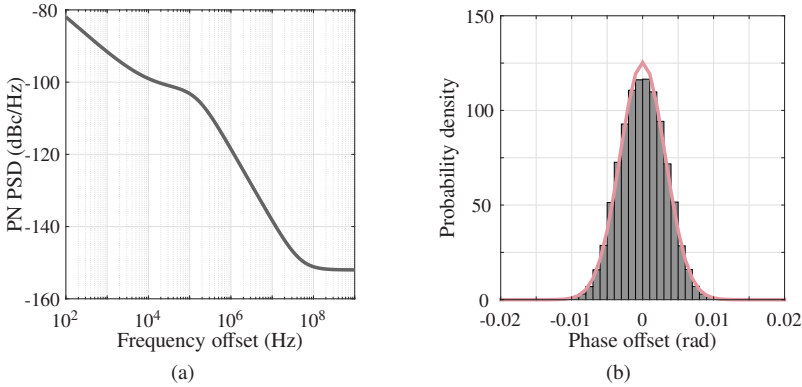


Figure 4.27: Adopted PN model: (a) double-sided PN PSD and (b) PDF of the PN in the time domain. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

significant leakage of each subcarrier, which follows the shape of the PSD from Figure 4.27(a), on the frequency bins of its neighbors.

To evaluate the actual effect of PN on the performance of OFDM-based ISAC systems, simulation results focusing on the distortion of the constellation and radar sensing performance are presented in Sections 4.2.3.1 and 4.2.3.2, respectively. Finally, remarks on the obtained simulation results in the two aforementioned sections are given in Section 4.2.3.3.

4.2.3.1 Constellation distortion

For the analysis on the constellation distortion in an OFDM-based ISAC system under PN influence, a communications scenario where transmit and receive PN are uncorrelated is assumed. Assuming that the aforementioned PN contributions have the same level, their incoherent combination results in a 3 dB higher PN level as they are assumed to be uncorrelated Gaussian processes. In addition, an ideal channel model with a single path ($P = 1$) and null delay and Doppler shift (i.e., $\tau_0 = 0$ s and $f_{D,0} = 0$ Hz) is adopted. Since no ISI happens in this scenario, $N_{CP} = 0$ is assumed.

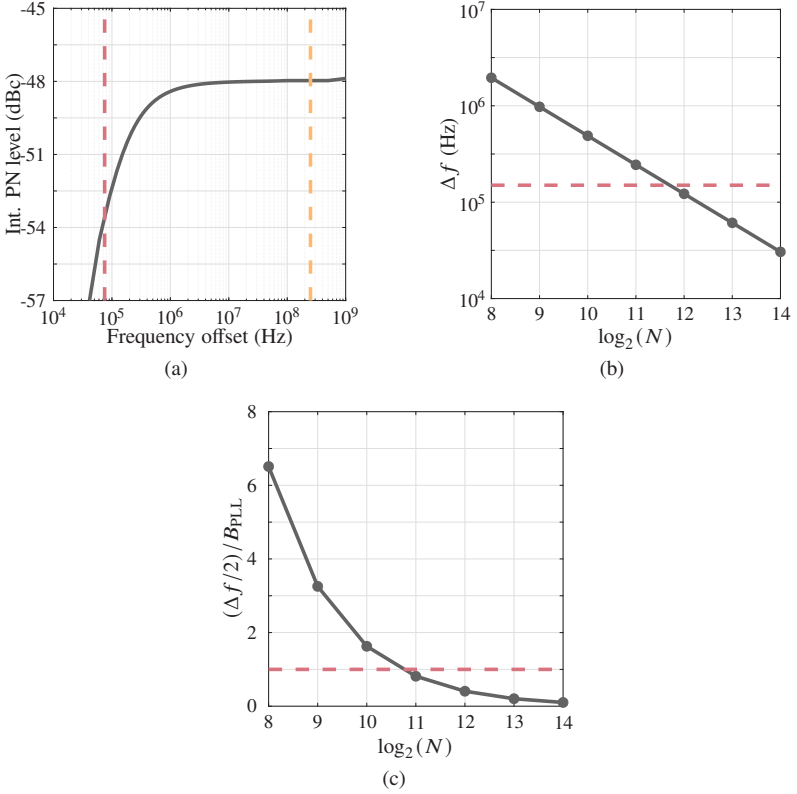


Figure 4.28: Transition from CPE to ICI as dominant PN-induced effect: (a) simulated integrated PN level as a function of frequency offset (double-sided) for the frequency range of interest, (b) subcarrier spacing Δf as a function of the number of subcarriers N , and (c) ratio between half of Δf and the PLL loop bandwidth B_{PLL} as a function of N (●). In (a) and (b), B_{PLL} is highlighted (---), whereas in (b), the ratio $(\Delta f/2)/B_{\text{PLL}} = 1$ below which ICI becomes more relevant than CPE is highlighted with the same line type. In addition, the offset of $B/2 = 250$ MHz is also highlighted in (a) (---), as this figure shows integrated double-sided PSD and therefore this offset corresponds to the bandwidth $B = 500$ GHz listed in Table 4.3. Adapted from [21], licensed under CC BY-NC-ND 4.0.

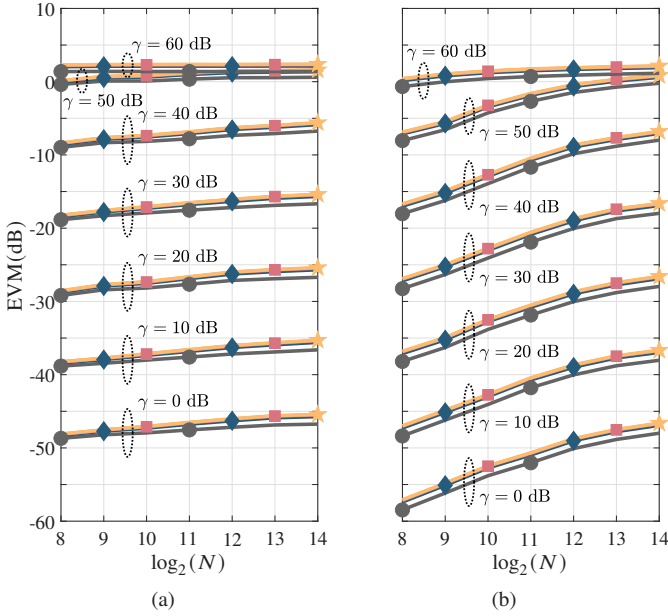


Figure 4.29: Simulated EVM for $B = 500$ MHz as in Table 4.3, $M = 128$ OFDM symbols, different numbers of subcarriers N and combined transmit and receive PN levels of -44.90 dBc, -34.90 dBc, -24.90 dBc, -14.90 dBc, -4.90 dBc, 4.90 dBc, and 14.90 dBc ($\gamma = 0$ dB, 10 dB, 20 dB, 30 dB, 40 dB, 50 dB, and 60 dB, respectively). In all cases, and QPSK ($\bullet$), 16-quadrature amplitude modulation (QAM) ($\blacklozenge$), 64-QAM ($\blacksquare$), and 256-QAM ($\star$) modulations were considered. In (a), both ICI and CPE are entirely present, while in (b) CPE was estimated and corrected. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

Disregarding AWGN, Figure 4.29(a) shows the achieved mean EVM over multiple PN realizations as a function of the number of subcarriers N for all modulation alphabets listed in Table 4.3. For the achieved results, the PN model from Figure 4.27 was considered, and its original level of -47.90 dBc was increased by a factor $\gamma \in \mathbb{R}_{\geq 0}$ to showcase the induced degradation by PN at higher levels. Focusing on the $\gamma = 0$ dB case, i.e. the original PN model from Figure 4.27 with a combined transmit and receive PN level of -44.90 dBc, it can be seen that the EVM degrades, i.e., increases, linearly along with increasing N , which is due to the decreasing subcarrier spacing Δf and resulting increased ICI effect.

This becomes more pronounced for higher-order modulations, i.e., 16-QAM, 64-QAM, and 256-QAM, when compared to QPSK. This is due to the fact that the constellation points are relatively closer to one another in these modulations, and also because they do not have a constant envelope. Consequently, neighbor subcarriers may not have the same energy, leading to a more severe ICI effect. For PN increase factors of $\gamma = \{0, 10, 20, 30, 40, 50\}$ dB, which result in combined transmit and receive PN levels of -44.90 dBc, -34.90 dBc, -24.90 dBc, -14.90 dBc, -4.90 dBc, 4.90 dBc, respectively, the EVM is linearly degraded, i.e., increased, by the same factor γ . The EVM degradation only stops following the previously described pattern at a combined transmit and receive PN level of 14.90 dBc, which corresponds to $\gamma = 60$ dB. At this PN level, the EVM is constant regardless of the adopted number of subcarriers. In addition, the EVM increase trend observed for lower γ is not present for $\gamma = 60$ dB, as the EVM difference w.r.t. $\gamma = 50$ dB is smaller than 10 dB for all considered N values. This happens since even the highest Δf cannot prevent severe ICI at such PN level. Still considering an ideal channel model as well as the ideal case where all subcarriers are known at the receiver side, the EVM after CPE estimation and correction based on (3.74) and (3.75), respectively, is shown as a function of the number of subcarriers N is shown in Figure 4.29(b). It is observed that a significant EVM improvement of up to around 9.78 dB is observed for decreasing N in comparison with the results without CPE correction in Figure 4.29(a). This happens since CPE is more relevant for lower N , while higher N , and therefore lower subcarrier spacing Δf , leads to more pronounced PN-induced ICI that cannot be compensated with the adopted method.

Next, Figure 4.30 shows the experienced SIR as a function of the combined transmit and receive PN level. For the performed simulations, $N \in \{256, 2048, 16384\}$, as well as QPSK and 256-QAM modulations were adopted. If the CPE is not compensated, all considered N values and modulations yield similar performance, with SIR decreasing linearly from 45.72 dB on average at the PN level of -45 dBc until an average of 1.5 dB at the PN level of 0 dBc. In this zone, $N = 256$ performs around 1.5 dB better than $N = 2048$ and 3 dB better than $N = 16384$. At PN levels slightly higher than the latter, the linear behavior of the SIR is no longer observed, and a constant level is reached. In the case where the CPE is estimated and corrected, a higher SIR is observed for lower N in the linear SIR zone, which was already expected based on the EVM results from Figure 4.29(b). A significant SIR improvement, however, is only observed for $N = 256$ among

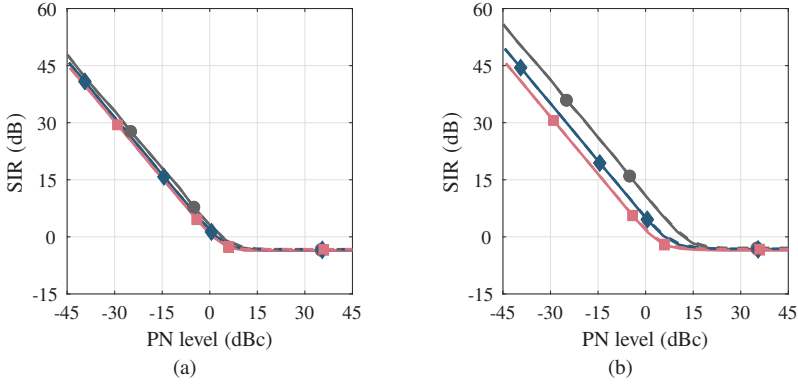


Figure 4.30: Simulated SIR as a function of the combined transmit and receive PN level (a) without and (b) with CPE estimation and correction. Results are shown assuming $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■). Results are also shown for 256-QAM and the aforementioned number of subcarriers, i.e., $N = 256$ (--- and ●), $N = 2048$ (--- and ◆), and $N = 16384$ (--- and ■). In both cases, $B = 500$ MHz as in Table 4.3 and $M = 128$ OFDM symbols were considered. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

the considered values. The maximum SIR improvement in the linear zone, which is of around 7.90 dB, is observed for $N = 256$. As for the other N values, a lower SIR improvement is observed after CPE estimation and correction, with a negligible difference observed for $N = 16384$. As in the case of Figure 4.29(b), the better performance improvement for lower N is due to the fact that the subcarrier spacing is higher and the ICI less pronounced in these cases. For all considered settings, an SIR of at least 15 dB is observed for combined transmit and receive PN levels equal to or lower than -15 dBc. At this PN level, reasonable communications performance in OFDM-based systems can still be achieved, e.g., as shown in [IGBS23]. For higher PN levels, critically low SIR may be experienced, leading to significant communications performance degradation. This is especially detrimental for bistatic ISAC, where reliable communications are required to reconstruct the transmit data and enable high-performance sensing [6, 50].

Examples on how the EVM and SIR degradation is related to changes in the constellation shape are shown for QPSK modulation considering -14.90 dBc ($\gamma = 30$ dB) in Figures 4.31 and 4.32, which respectively address the cases

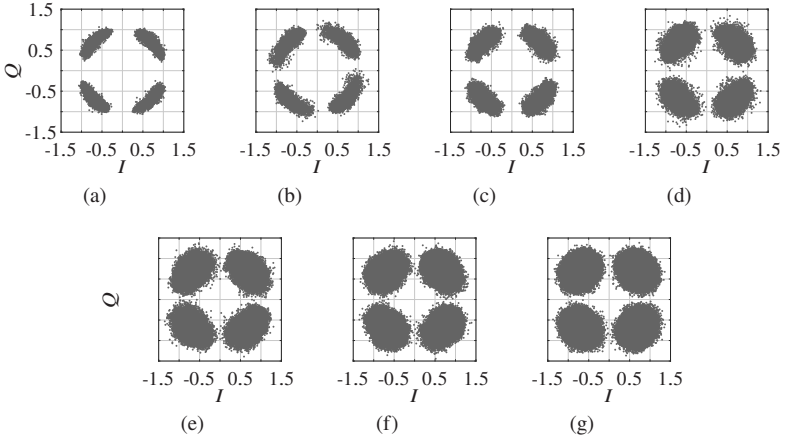


Figure 4.31: Simulated normalized receive QPSK constellations obtained for a combined transmit and receive PN level of -14.90 dBc ($\gamma = 30$ dB) and different numbers of subcarriers: (a) $N = 256$, (b) $N = 512$, (c) $N = 1024$, (d) $N = 2048$, (e) $N = 4096$, (f) $N = 8192$, and (g) $N = 16384$. In all cases, $B = 500$ MHz as in Table 4.3 and $M = 128$ OFDM symbols were considered. Adapted from [21], licensed under CC BY-NC-ND 4.0.

without and with CPE estimation and correction. As expected from the results in Figure 4.28, the obtained constellations show that the CPE is the dominating PN-induced degradation for $N \in \{256, 512, 1024, 2048\}$, whereas $N \in \{4096, 8192, 16384\}$ results in more significant ICI, which is evidenced by a more spurious, noise-like spreading of the constellation clusters.

4.2.3.2 Effects on radar sensing

Considering the interaction of the PN at both transmitter and receiver as discussed in Section 4.2.3, the total level of combined transmit and receive PN is shown in Fig 4.33 for both monostatic and bistatic ISAC architectures previously depicted in Figure 3.8. Figure 4.33(a) shows the combined PN level for monostatic radar sensing as a function of the monostatic range. For monostatic radar sensing, the transmit PN $\theta_{\text{Tx}}^{\text{PN}}(t)$ and the receive PN $\theta_{\text{Rx}}^{\text{PN}}(t)$ are correlated. As discussed in Section. 4.2.3, the transmit PN contribution to the overall PN is $\theta_{\text{Tx}}^{\text{PN}}(t - \tau)$, where $\tau = 2R/c_0$ is the propagation delay associated with the

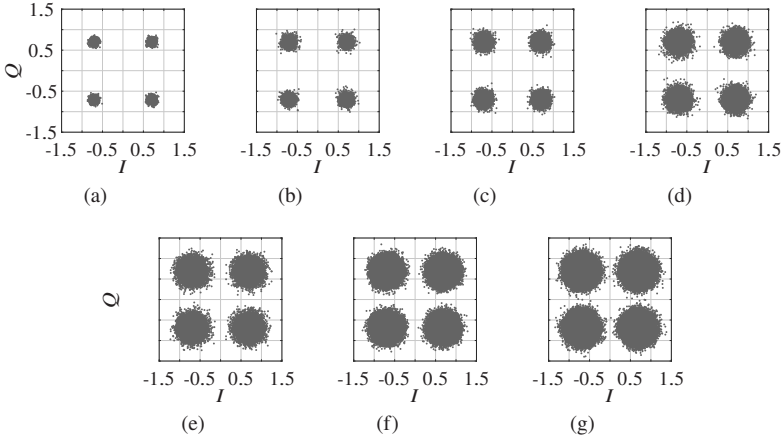


Figure 4.32: Simulated normalized receive QPSK constellations obtained after CPE estimation and correction for a combined transmit and receive PN level of -14.90 dBc ($\gamma = 30$ dB), $B = 500$ MHz as in Table 4.3, $M = 128$ OFDM symbols, and number of subcarriers equal to: (a) $N = 256$, (b) $N = 512$, (c) $N = 1024$, (d) $N = 2048$, (e) $N = 4096$, (f) $N = 8192$, and (g) $N = 16384$. Adapted from [21], licensed under CC BY-NC-ND 4.0.

monostatic range R . The receive PN contribution is in turn $\theta_{\text{Rx}}^{\text{PN}}(t) = \theta_{\text{Tx}}^{\text{PN}}(t)$. This results in a combined transmit and receive PN level that tends to 0 rad^2 or $-\infty$ dBc for decreasing ranges or propagation delays, as

$$\lim_{\tau \rightarrow 0} [\theta_{\text{Tx}}^{\text{PN}}(t - \tau) - \theta_{\text{Tx}}^{\text{PN}}(t)] = 0. \quad (4.1)$$

For increasing ranges or delays, however, the $\theta_{\text{Tx}}^{\text{PN}}(t)$ and $\theta_{\text{Rx}}^{\text{PN}}(t)$ tend to become uncorrelated and their PN levels, which are both equal to -47.90 dBc, are simply added, resulting in the same combined transmit and receive PN level of -44.90 dBc observed in Section 4.2.3.1. The bistatic sensing case in Figure 4.33(b) shows a different behavior. Since $\theta_{\text{Tx}}^{\text{PN}}(t)$ and $\theta_{\text{Rx}}^{\text{PN}}(t)$ are uncorrelated as in the communications case discussed in Section 4.2.3.1, since distinct oscillators are used by transmitter and receiver, the transmit and receive PN level add to -44.90 dBc regardless of the experienced range or propagation delay. For comparison, the achievable maximum unambiguous range for both monostatic and bistatic radar sensing, as a function of the number of subcarriers

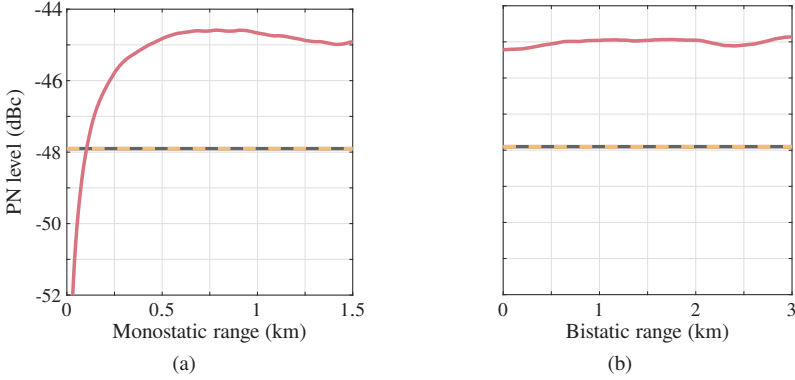


Figure 4.33: Simulated combined (—) transmit (—) and receive (—) PN level as a function of range: (a) monostatic sensing with correlated PN at transmitter and receiver and (b) bistatic sensing with uncorrelated PN. For communications, the same PN level is obtained as for bistatic sensing. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

N and assuming $B = 500$ MHz as in Table 4.3 is shown in Figure 4.34. The maximum unambiguous range is defined as $R_{\max, \text{ua}} = N\Delta R$, where ΔR is the range resolution. For monostatic radar sensing, the maximum unambiguous range is defined as $R_{\max, \text{ua}}^{\text{monost}} = N\Delta R^{\text{monost}} = N[c_0/(2B)]$ [18], while for bistatic radar sensing it is $R_{\max, \text{ua}}^{\text{bist}} = N\Delta R^{\text{bist}} = N(c_0/B)$ [50]. The achieved results in Figure 4.34 show that the maximum unambiguous range spans from 0.08 km to 4.91 km for the monostatic case, and from 0.15 km to 9.82 km for the bistatic case. This reveals that, for $N \leq 4096$, the whole achieved monostatic maximum unambiguous ranges profit from the transmit and receive PN correlation to a certain extent. Consequently, lower combined transmit and receive PN levels are experienced than in both the monostatic case for $N > 4096$ and the bistatic case regardless of N .

Next, to quantify the PN-induced degradation on monostatic and bistatic sensing in OFDM-based ISAC systems, the PPLR, PSLR, and ISLR parameters are used [LMI16, LLL⁺24, 18]. The first of the aforementioned parameters expresses the ratio between the mainlobe power under PN influence and the ideally expected mainlobe power without PN [OJUY20] after experiencing the radar processing gain of ideally NM under absence of PN [18, 50]. The latter parameters,

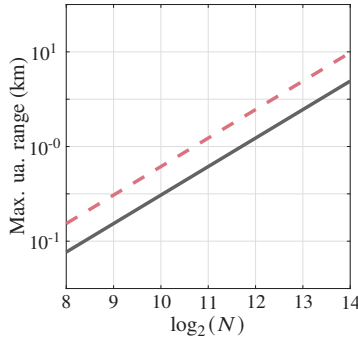


Figure 4.34: Calculated maximum unambiguous range as a function of the number of subcarriers N : monostatic sensing (—) and bistatic sensing (---). Adapted from [21], licensed under CC BY-NC-ND 4.0.

i.e., PSLR and ISLR, express the ratio between the peak power of the highest sidelobe and the mainlobe and the ratio between the integrated sidelobe and mainlobe powers [LMI16], respectively. All three parameters are used to evaluate the PN effect on target reflections in a radar image under different settings. In addition, an analysis using image SIR, which provides a measure between a target peak and the PN-induced interference in the radar image, as a performance parameter are also presented. The obtained results are discussed in Sections 4.2.3.2.1 through 4.2.3.2.6, for which parameter sets Table 4.3 are considered. The specifically adopted parameter values are explicitly mentioned in each of the aforementioned sections. Since the constellation distortion and SIR were analyzed in detail in Section 4.2.3.1 for all N and modulation alphabets from Table 4.3, only $N \in \{256, 2048, 16384\}$ as well as QPSK and 256-QAM are henceforth considered for conciseness. This covers the cases $\Delta f/2 \gg B_{\text{PLL}}$, $\Delta f/2 \approx B_{\text{PLL}}$, and $\Delta f/2 \ll B_{\text{PLL}}$, respectively, as well as the two extremes of the considered modulation alphabets.

To focus only on the PN-induced impairments, self-interference is disregarded in the monostatic sensing case, and full knowledge of the transmit OFDM frame as well as perfect time, frequency and sampling frequency synchronization is assumed for bistatic sensing. For a more insightful analysis, PPLR and ISLR are calculated in both range and Doppler shift directions and, as in Section 4.2.3.1, AWGN is not considered to highlight the PN effects. As in Section 4.2.3.1, multiple PN level increase factors γ were considered in the simulations.

4.2.3.2.1 Main- and sidelobe distortion for different numbers of subcarriers and modulation alphabets

To enable analyzing the radar sensing performance as a function of the number of subcarriers N and the modulation alphabet for a wide range of combined transmit and receive PN level, Figure 4.35 shows the PPLR, range PS LR and ISLR, as well as Doppler shift PS LR and ISLR obtained for a static target. Since the monostatic and bistatic cases only differ in the transmit and receive PN correlation, which ultimately yields different combined PN levels as shown in Figure 4.33, a generic case was simulated and the PN level was varied. This is later also done in Sections 4.2.3.2.2 and 4.2.3.2.3. In addition, $N \in \{256, 2048, 16384\}$, $N_{CP} = N$ to ensure that the maximum unambiguous range is measurable, as well as QPSK and 256-QAM were considered, besides $M = 128$ as in Section 4.2.3.1. The obtained results in Figure 4.35(a) reveal that the combined PN level result in negligible PPLR until around -9.25 dBc. Afterwards, the PPLR degrades linearly until around 20.26 dBc, reaching a constant level that depends on the adopted N and modulation. These are -33.67 dB, -42.13 dB, and -50.66 dB for $N = 256$, $N = 2048$, and $N = 16384$, respectively, and QPSK modulation. For 256-QAM, the obtained PPLR values for high combined PN level were 5.28 dB higher, i.e., -28.47 dB, -36.85 dB, and -45.38 dB for $N = 256$, $N = 2048$, and $N = 16384$, respectively. From $N = 256$ to $N = 2048$ as well as from $N = 2048$ to $N = 16384$, the PPLR degrades 8.47 dB on average, i.e., it is worsened by a factor of approximately 7. This is close to the ratio 8 between $N = 16384$ and $N = 2048$, and also $N = 2048$ and $N = 256$, which also indicates the factor by which the subcarrier spacing Δf is decreased when increasing N , leading to more significant PN-induced ICI effect. An exact PPLR degradation by a factor of 8 is only not observed due to the form in which the PN level is spread over frequency as shown in Figure 4.27(a).

Regarding the range PS LR, Figure 4.35(b) shows that apart from a transition zone of a combined PN level between approximately 10.21 dBc and 25.73 dBc, where 256-QAM starts degrading at PN levels around 3 dB lower than QPSK, the choice of the modulation order does not influence the achieved PS LR. For combined transmit and receive PN levels from 25.73 dBc, $N = 256$, $N = 2048$, and $N = 16384$ result in constant range PS LR values of -3.42 dB, -2.97 dB, and -2.28 dB, respectively. In terms of range ISLR, a similar transition zone to the aforementioned one is observed in Figure 4.35(c). Afterwards, the

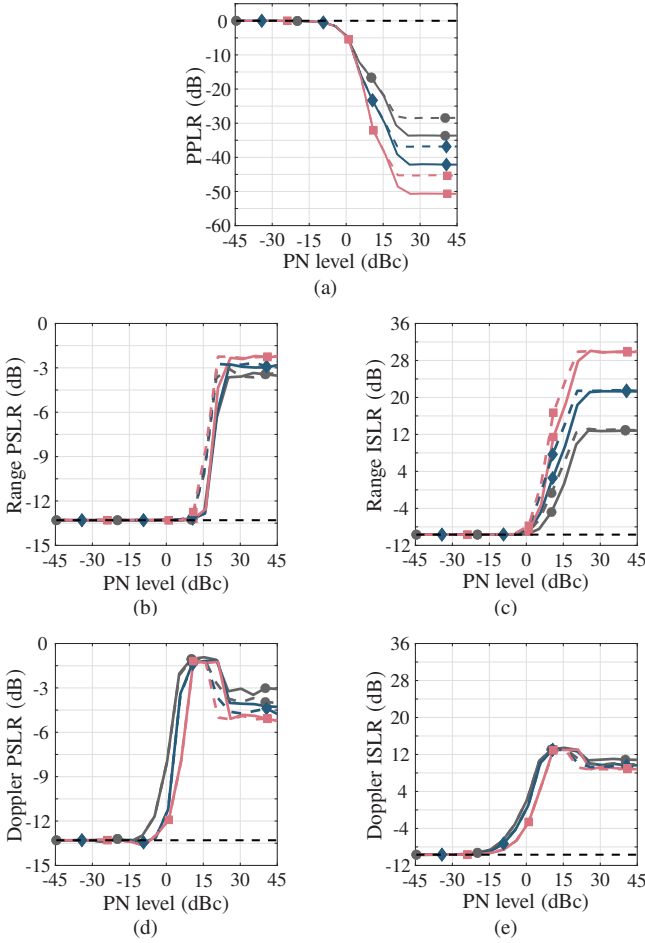


Figure 4.35: Simulated PPLR (a), range PSLR (b) and ISLR (c), and Doppler shift PSLR (e) and ISLR (d) as functions of the combined transmit and receive PN level. Simulation results are shown for a static target assuming $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■). In addition, results are also shown for 256-QAM with $N = 256$ (--- and ●), $N = 2048$ (--- and ◆), and $N = 16384$ (--- and ■). In all subfigures, $B = 500$ MHz as in Table 4.3, a CP length of $N_{CP} = N$ and $M = 128$ OFDM symbols were considered. In addition, the PPLR, PSLR and ISLR values for simulations without PN are also shown for comparison (---) and are the same regardless of the adopted modulation. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

same result is achieved regardless of the adopted modulation, and $N = 256$, $N = 2048$, and $N = 16384$ yield ISLR values of around 12.99 dB, 21.53 dB, and 30.02 dB, respectively. The difference of about 8.5 dB, which corresponds to worse ISLR by a factor of 7 between $N = 16384$ and $N = 2048$, as well as between $N = 2048$ and $N = 256$, is due to the increase in the number of samples used to represent the range axis of the radar image. Since the mainlobe width is the same for all three aforementioned cases but the amount of samples containing sidelobes increases, the ISLR naturally degrades. This is, however, only observed since the high PN level results in significant ICI that severely increases the sidelobe level. Otherwise approximately the same ISLR is observed for all considered N values as is the case for combined PN levels equal to or smaller than about 9.65 dBc in Figure 4.35(c). The achieved values show that the range PSLR and ISLR are only affected by rather high PN levels. This indicates that the PN-induced ICI can be suppressed by the radar processing gain as it is a Gaussian noise-like effect, and that CPE is the dominating source of PN-induced degradation of the sensing performance in OFDM-based ISAC for reasonable PN levels.

Finally, in terms of Doppler shift PSLR and ISLR, better performance is achieved by QPSK and smaller N in the region between -14.55 dBc and 11 dBc, where both PSLR and ISLR linearly degrade. The influence of the adopted modulation order on this tendency is small and becomes negligible with increasing N . Afterwards, both parameters improve with increasing combined transmit and receive PN level of up to about 26.06 dBc, where a constant level is achieved. This is illustrated in Figure 4.36, where cuts along the Doppler shift direction of the radar images used to generate the results from Figure 4.35 are shown. It can be observed that a combined PN level of 10 dBc ($\gamma = 54.90$ dB) results in higher sidelobes and a slightly narrower mainlobe due to interaction with the sidelobes than in the case for 35 dBc ($\gamma = 79.90$ dB). As a result, this ultimately leads to worse PSLR and ISLR performances. The unexpected improved radar sensing performance with increasing PN level is due to the fact that the resulting PN PDF presents phase values outside the $(-\pi, \pi)$ range as shown in Figure 4.37. Consequently, the PDF folds around and becomes a wrapped Gaussian distribution. This results in a changing PSLR and ISLR performance until the folding of the PDF becomes strong enough to give the PN PDF an uniform shape, which is when the PSLR and ISLR parameters present a constant level. It is, however, worth mentioning that such high combined transmit and receive PN levels should not be expected in practice, especially

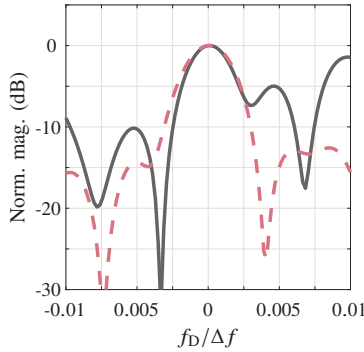


Figure 4.36: Cut along the Doppler shift direction of bistatic range-Doppler radar images simulated for a static target assuming $B = 500$ MHz as in Table 4.3, $N = 2048$, $N_{\text{CP}} = 512$, $M = 128$, QPSK modulation, and rectangular windowing in both range and Doppler shift directions. Results are shown for 10 dBc ($\gamma = 54.90$ dB) (—) and 35 dBc ($\gamma = 79.90$ dB) (---). Since the considered PN levels lead to different PPLR and therefore peak heights in the radar image, the magnitudes are normalized to the respective static target peaks located at $f_D / \Delta f = 0$. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

for mmWave systems. The aforementioned achieved PSLR levels are of approximately -3.02 dB, -4.07 dB, and -5.07 dB for $N = 256$, $N = 2048$, and $N = 16384$, respectively. As for ISLR, the aforementioned N values yield 10.88 dB, 9.57 dB, and 8.97 dB, respectively. The slightly better PSLR and ISLR along with N is due to the increasing ICI relevance in this case, which tends to make the PN-induced CPE less severe, although still significant, therefore resulting in less degradation of the DFT-based Doppler shift estimation that relies on the phases of consecutive OFDM symbols resulting from the target's Doppler shift.

4.2.3.2.2 Main- and sidelobe distortion for different numbers of OFDM symbols

Setting $N = N_{\text{CP}} = 2048$ and focusing only on QPSK modulation, Figure 4.38 shows results on the radar sensing performance degradation due to PN for $M \in \{32, 128, 512, 2048\}$ OFDM symbols. The obtained results show that increasing M results in a worse overall performance. A slight improvement in range PSLR and ISLR can indeed be seen in Figures 4.38(b) and 4.38(c),

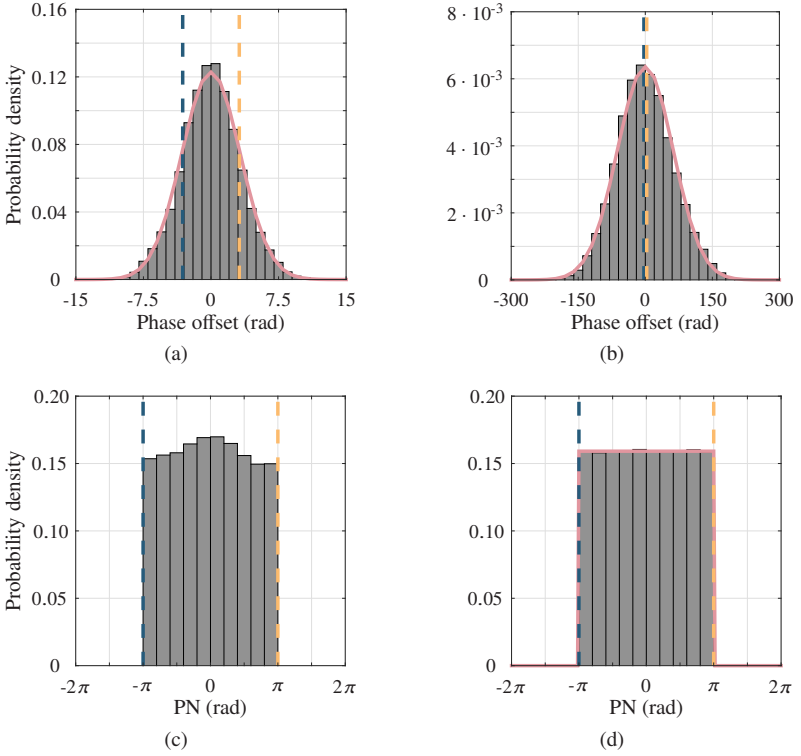


Figure 4.37: Simulated PDF of the PN in the time domain: (a) 9 dBc ($\gamma = 53.90$ dB) and (b) 34 dBc ($\gamma = 78.90$ dB). In (c), and (d), the same cases as in (a) and (b) are respectively shown, after wrapping the phase between $-\pi$ ($\leftarrow$) and π ($\rightarrow$). While the PDFs in both (a) and (b) can be properly fitted by Gaussian distributions, and (d) by an uniform distribution, the PDF in (c) rather presents the form of a wrapped Gaussian distribution. For convenience, all obtained distribution fits are marked ($\rightarrow$). Reprinted from [21], licensed under CC BY-NC-ND 4.0.

which is due to a higher radar processing gain against PN-induced ICI, which is the same for every OFDM symbol. However, with increasing M a higher number of OFDM symbols with PN-induced CPE is processed, which leads to a degradation of the PPLR proportional to M as observed in Figures 4.38(a). After a linear degradation zone from around -19.87 dBc, the PPLR reaches a constant level at 26.14 dBc. This level is equal to -36.58 dB, -42.08 dB,

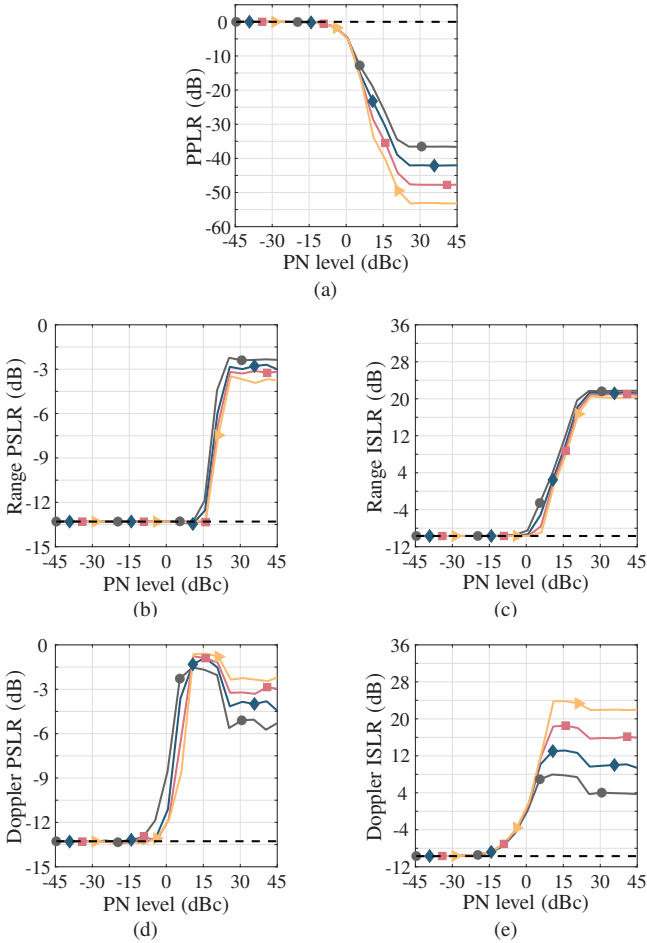


Figure 4.38: Simulated PPLR (a), range PSLR (b) and ISLR (c), and Doppler shift PSLR (e) and ISLR (d) as functions of the combined transmit and receive PN level. Simulation results are shown for a static target assuming $N = 2048$, QPSK modulation, and $M = 32$ (●), $M = 128$ (◆), $M = 512$ (■), and $M = 2048$ (▲). In all subfigures, $B = 500$ MHz as in Table 4.3 and a CP length of $N_{CP} = N$ were considered, and the PPLR, PSLR and ISLR values for simulations without PN are also shown (---). Reprinted from [21], licensed under CC BY-NC-ND 4.0.

−47.76 dB, and −53.21 dB for $M = 32$, $M = 128$, $M = 512$, and $M = 2048$,

respectively. This corresponds to an average PPLR degradation of 5.58 dB (factor of 3.61) from an M value to the next higher one, and it is close to the ratio of 4 between them. In addition, the Doppler shift PSLR and ISLR also degrade with increasing M due to increased CPE effect as seen in Figures 4.38(d) and 4.38(e). It can be seen that both curves present a similar behavior as their counterparts in Figure 4.35. In terms of PSLR, constant levels of -5.10 dB, -3.85 dB, -3.3 dB, and -2.24 dB is observed for the considered M values in crescent order. The PSLR difference between consecutive values is, in all cases, smaller than 2 dB, and the PSLR values do not increase linearly with the ratio in which M is increased. Regarding Doppler shift ISLR, however, an average increasing step of 6.24 dB (factor 4.21) is observed for increasing M , which is close to the ratio of 4 between the considered consecutive M values. The observed ISLR values at the high PN level regime for $M = 32$, $M = 128$, $M = 512$, and $M = 2048$ are 3.85 dB, 10.01 dB, 15.80 dB, and 21.89 dB, respectively.

4.2.3.2.3 Main- and sidelobe distortion for moving targets

In Figure 4.39, radar sensing performance results are shown for $N = N_{CP} = 2048$, QPSK modulation, and Doppler shifts of $f_D \in \{0 \text{ Hz}, \pm 0.1\Delta f, \pm 0.5\Delta f\}$ to also cover scenarios with moving targets. Besides the static target case for $f_D = 0 \text{ Hz}$, which was already investigated in Figure 4.35 and is only shown here as a reference, Doppler shifts of $f_D = \pm 0.1\Delta f$ are considered as they are assumed to result in tolerable Doppler-shift-induced ICI [18]. The Doppler shifts of $f_D = \pm 0.5\Delta f$, in turn, correspond to the maximum achievable unambiguous Doppler shift interval, which is obtained in the case where $N_{CP} = 0$ [18, 50]. Since the difference w.r.t. the range and Doppler shift PSLR and ISLR results from Figure 4.35 was negligible, only PPLR is shown in Figure 4.39. The achieved results show that the PN-induced impairments combined have significantly more influence than the Doppler-shift-induced ICI on the radar sensing performance of the OFDM-based ISAC system. Only a slightly better PPLR is shown for $f_D = \pm 0.5\Delta f$, which is due to the fact that the Doppler-shift induced ICI may not always add constructively to its PN-induced counterpart. This difference, however, only becomes visible for relatively high combined PN levels, i.e., above 5.70 dBc, and it is of around 4 dB when a constant PPLR level is achieved for a PN level of 25.74 dBc or higher.

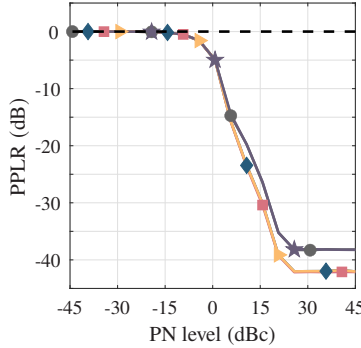


Figure 4.39: Simulated PPLR as a function of the combined transmit and receive PN level. Simulation results are shown for a moving target assuming $N = 2048$, QPSK modulation, and $f_D = -0.5\Delta f$ (●), $f_D = -0.1\Delta f$ (◆), $f_D = 0$ (■), $f_D = 0.1\Delta f$ (▶), and $f_D = 0.5\Delta f$ (★). A bandwidth $B = 500$ MHz as in Table 4.3, a CP length of $N_{CP} = N$, and $M = 128$ OFDM symbols were considered. For comparison, the PPLR values for simulations without PN or Doppler shift are also shown (—). Reprinted from [21], licensed under CC BY-NC-ND 4.0.

4.2.3.2.4 Main- and sidelobe distortion for multiple moving targets

The previous subsections show how PN degrades the PPLR, as well as range and Doppler shift PSLR and ISLR, of a single target under different settings. These parameters allow quantifying both the power loss at the mainlobe and the sidelobe level increase in range and Doppler shift directions. To illustrate how this impacts the radar sensing performance of an OFDM-based ISAC system in multi-target scenarios, Figure 4.40 shows range-Doppler shift radar images for one static and one moving target. In addition $N = 2048$, $N_{CP} = 512$, $M = 128$, and QPSK modulation were assumed. No AWGN was considered, and rectangular windowing was used in both range and Doppler shift processing. It can be seen that the sinc-shaped sidelobe pattern expected even without PN influence is clearly observable for a combined transmit and receive PN level of -44.90 dBc ($\gamma = 0$ dB). For the considered higher combined PN levels of -14.90 dBc ($\gamma = 30$ dB) and 4.90 dBc ($\gamma = 50$ dB), the mainlobe of both targets suffer from power loss as expected from the PPLR, and the Doppler shift sidelobes become much more relevant than their counterparts in the range direction, as expected from the analyses in the previous subsections.

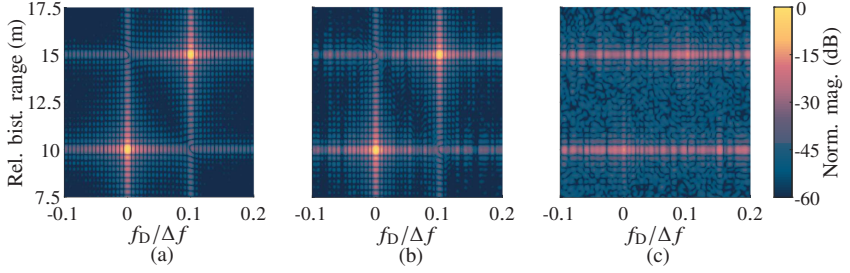


Figure 4.40: Simulated bistatic range-Doppler radar images with $B = 500$ MHz as in Table 4.3, $N = 2048$, $N_{CP} = 512$, $M = 128$, QPSK modulation, and rectangular windowing in both range and Doppler shift directions. Two targets were considered, the first at 10 m and with a Doppler shift of 0 kHz and the second at 15 m and with a Doppler shift of $0.1\Delta f$ (which results in negligible ICI [18]). The considered combined transmit and receive PN levels in the simulations were (a) -44.90 dBc ($\gamma = 0$ dB), (b) -14.90 dBc ($\gamma = 30$ dB), and (c) 4.90 dBc ($\gamma = 50$ dB). Adapted from [21], licensed under CC BY-NC-ND 4.0.

Next, Figure 4.41 shows range-Doppler shift radar images obtained with the same parameters adopted for Figure 4.40, but applying Chebyshev windows with 100 dB sidelobe suppression in both range and Doppler shift directions. It can be seen in Figure 4.41 that the windowing is only effective in range direction, as the PN-induced effect that plays a major role in range processing is ICI. As discussed in the previous subsections, the ICI produces a Gaussian noise-like effect that can be suppressed by the radar processing gain to some extent. Since this effect is additive, it does not reduce the effectiveness of windowing during range processing. As for windowing in the Doppler shift direction, it becomes little effective with increasing combined transmit and receive PN level as shown for -14.90 dBc ($\gamma = 30$ dB) and 4.90 dBc ($\gamma = 50$ dB). The reason for this is the severe CPE, which prevents the DFT-based estimation to generate a sinc-shaped Doppler shift profile even with rectangular windowing, e.g., as previously shown in Figure 4.36. Consequently, the distorted, non-decreasing Doppler shift sidelobe pattern cannot be effectively suppressed by Chebyshev windowing.

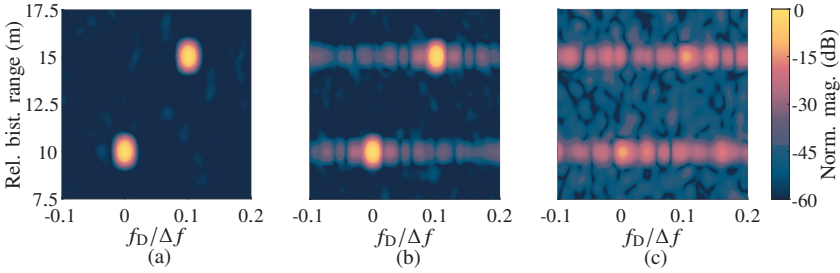


Figure 4.41: Simulated bistatic range-Doppler radar images obtained with $B = 500$ MHz as in Table 4.3, $N = 2048$, $N_{CP} = 512$, $M = 128$, QPSK modulation, and Chebyshev windowing with 100 dB sidelobe suppression in both range and Doppler shift directions. Two targets were considered, the first at 10 m and with a Doppler shift of 0 kHz and the second at 15 m and with a Doppler shift of $0.1\Delta f$ (which results in negligible ICI [18]). The considered combined transmit and receive PN levels in the simulations were (a) -44.90 dBc ($\gamma = 0$ dB), (b) -14.90 dBc ($\gamma = 30$ dB), and (c) 4.90 dBc ($\gamma = 50$ dB). Adapted from [21], licensed under CC BY-NC-ND 4.0.

4.2.3.2.5 Image signal-to-interference ratio

To quantify the PN-induced radar image distortion as observed in Section 4.2.3.2.4, the image SIR as a function of the combined transmit and receive PN level is analyzed in Figure 4.42. For simplicity, a single-target scenario is considered, and the same parameters from Section 4.2.3.2.1 are assumed. For the obtained results, a Chebyshev window with 100 dB sidelobe suppression was used and the SIR was calculated as the magnitude ratio between the target reflection peak and the region outside the mainlobe. In addition, both mean and minimum SIR values were presented. These correspond to the average and minimum SIR calculated between the peak and each point outside the mainlobe, respectively.

The obtained mean SIR is shown in Figure 4.42(a). In this figure, a similar trend to what is experienced when performing communications over an ideal channel as shown in Figure 4.30 is observed. However, the relationship between the SIR at the subcarrier level and the image SIR is not fully linear, as the PN-induced CPE and ICI are differently suppressed by the radar processing gain. Back to Figure 4.42(a), an approximately linear degradation is seen until a combined transmit and receive PN level of around 0 dBc is observed in Figure 4.42(a). Afterwards, the SIR decreases more steeply until a 0 dB

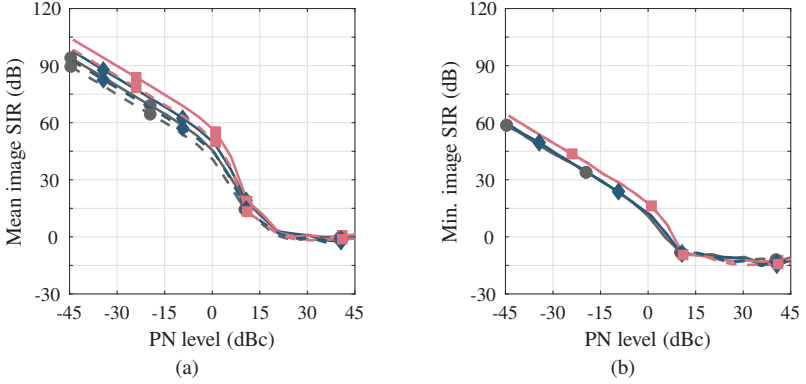


Figure 4.42: Simulated image SIR: (a) mean and (b) minimum values. In the performed simulations, a single static target was considered and $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■) were assumed. Results were also obtained for 256-QAM with $N = 256$ (--- and ●), $N = 2048$ (--- and ◆), and $N = 16384$ (--- and ■). In all subfigures, $B = 500$ MHz as in Table 4.3, a CP length of $N_{\text{CP}} = N$ and $M = 128$ OFDM symbols were considered. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

floor is reached at around 20 dBc PN level. As expected, the minimum SIR shown in Figure 4.42(b) tends to be significantly lower than the mean SIR in Figure 4.42(a), including the minimum SIR floor of around -15 dB at high PN levels. This happens since the minimum SIR is dominated by the Doppler shift sidelobes that are more severely degraded by CPE than the range sidelobes are by ICI, e.g., as discussed in Section 4.2.3.2.4. Compared to the results on mainlobe and sidelobe distortion in the previous subsections, it is seen that the image SIR is reduced even for rather low PN levels, while PPLR and range and Doppler PSLR and ISLR do not degrade before -15 dBc. At this PN level, however, mean and minimum image SIR values of approximately 60 dB and 30 dB or higher are respectively observed, which already represents non-negligible sensing performance degradation.

4.2.3.2.6 Sensing performance under 3GPP specifications

In the 3GPP technical report TR 38.803 [3GP22], PN models for gNBs and UEs at 30 GHz were proposed. Based on these models, the double-sided PN

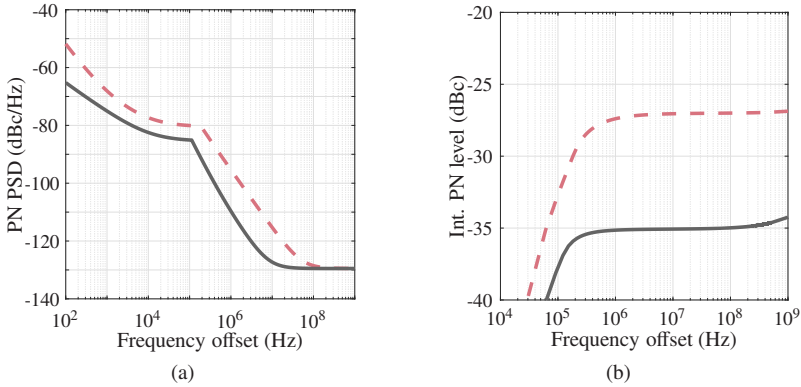


Figure 4.43: Simulated PN characteristics based on 3GPP TR 38.803 model for a gNB (—) and a UE (---) at 30 GHz: (a) double-sided PN PSD and (b) integrated PN level as a function of frequency offset (also double-sided) for the frequency range of interest. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

PSDs shown in Figure 4.43(a) and the integrated PN level were obtained. To analyze the sensing performance degradation imposed by these PN PSD profiles, four settings are considered. These are namely monostatic sensing at a single gNB, bistatic sensing between two gNBs, bistatic sensing with a gNB acting as a transmitter and a UE as a receiver, and bistatic sensing between two UEs, which are henceforth named Settings #1 to #4, respectively. For all settings, a single static target was considered, and the adopted OFDM signal parameters are listed in Table 4.4. The parameters for Setting #1 were suggested in [MHBW23], while those for Settings #2 to #4 were used for measurement-based ISAC demonstrations in [HMG⁺23, HMGt24, 53]. Moreover, all of the adopted parameter sets comply with 3GPP specifications. Since there is only a slight difference between the adopted carrier frequencies and 30 GHz, the PN models from Figure 4.43(a) were still used for Settings #1 to #4.

The resulting transmit, receive, and combined transmit and receive PN levels for each setting, as well as the sensing performance degradation measured by PPLR, range PSLR and ISLR, Doppler shift PSLR and ISLR, as well as mean an minimum image SIR obtained for all four settings are listed in Table 4.5. The achieved results show that the lowest combined transmit and receive PN level was -308.60 dBc in Setting #1, which is due to the fact that monostatic

Table 4.4: Adopted 5G-compliant OFDM signal parameters for the settings of monostatic sensing at a single gNB (#1), bistatic sensing between two gNBs (#2), bistatic sensing with a gNB as a transmitter and a UE as a receiver (#3), and bistatic sensing between two UEs (#4). Adapted from [21], licensed under CC BY-NC-ND 4.0.

	Setting #1	Setting #2	Setting #3	Setting #4
Carrier frequency (f_c)	28 GHz	27.4 GHz	27.4 GHz	27.4 GHz
Frequency bandwidth (B)	1.6 GHz	190 MHz	190 MHz	190 MHz
No. of subcarriers (N)	12672	1584	1584	1584
CP length (N_{CP})	1600	112	112	112
No. of OFDM symbols (M)	384	1120	1120	1120
Digital modulation	QPSK	QPSK	QPSK	QPSK

sensing is performed and the transmit and receive PN correlation reduces the effectively experienced PN level. The achieved PN level was -23.98 dBc in Setting #4, where bistatic sensing between two UEs is performed.

Overall, all considered settings yielded combined transmit and receive PN levels are all below -15 dBc, below which only negligible main- and sidelobe level degradation is expected according to the presented results in Sections 4.2.3.2.1 to 4.2.3.2.4. This is confirmed by the remaining results in Table 4.5, which show that negligible PPLR, range PSLR, and ISLR degradation in Settings #2 to #4 w.r.t. to the virtually PN-free case in Setting #1 is observed. Regarding Doppler shift sidelobes, only a small variation in terms of PSLR is observed, besides a slight increase in ISLR from Settings #1 to #4 due to the increasing PN level. In terms of mean image SIR, high values are observed in all settings, which means that only negligible noise floor increase in radar images is expected. Regarding minimum SIR, only Setting #1 presents no relevant SIR degradation due to non-suppressed Doppler shift sidelobes, which is simply explained by the rather low PN level in this case. As for Settings #2 to #4, the obtained minimum SIR values of 56.92 dB, 53.72 dB, and 51.51 dB show that non-negligible degradation in the Doppler shift direction is expected, although not severe.

Since only tolerable sensing performance degradation is observed in all four considered settings, it can be concluded that OFDM-based ISAC systems in the lower FR2 bands with 3GPP-compliant OFDM signal parameterization and

Table 4.5: Resulting simulated PN-induced sensing performance degradation for Settings #1, #2, #3, and #4. Adapted from [21], licensed under CC BY-NC-ND 4.0.

	Setting #1	Setting #2	Setting #3	Setting #4
Tx PN level	−34.45 dBc	−35.07 dBc	−35.07 dBc	−27.01 dBc
Rx PN level	−34.45 dBc	−35.07 dBc	−27.01 dBc	−27.01 dBc
Comb. Tx and Rx PN level	−308.60 dBc	−32.09 dBc	−26.43 dBc	−23.98 dBc
PPLR	0 dB	0 dB	−0.01 dB	−0.02 dB
Range PSLR	−13.30 dB	−13.30 dB	−13.30 dB	−13.30 dB
Range ISLR	−9.68 dB	−9.68 dB	−9.68 dB	−9.68 dB
Doppler PSLR	−13.30 dB	−13.27 dB	−13.27 dB	−13.26 dB
Doppler ISLR	−9.68 dB	−9.67 dB	−9.65 dB	−9.62 dB
Mean image SIR	148.28 dB	92.23 dB	86.15 dB	83.53 dB
Min. image SIR	100 dB	56.92 dB	53.72 dB	51.51 dB

PN levels equal to or lower than the models from TR 38.803 [3GP22] present PN-robust sensing performance in both monostatic and bistatic architectures.

4.2.3.2.7 Sensing performance with CPE compensation

To analyze the effectiveness of the CPE estimation and correction according to (3.74) and (3.75) sensing performance, a bistatic scenario with a dominant, reference path only is assumed due to the reasons discussed in Section 3.4.4. In addition, the same parameters adopted for the results in Figure 4.35 discussed in Section 4.2.3.2.1 are considered. The obtained results in Figure 4.44. Comparing with the results without CPE compensation from Figure 4.35, it is seen that the PPLR becomes negligible for combined transmit and receive PN levels below 0 dBc, whereas a PPLR of −5 dB is achieved without CPE compensation. For higher PN levels, a PPLR floor is still reached, but less severe peak power losses are observed than in Figure 4.35. On the one hand, it can also be seen that the range PSLR is significantly improved, with better nearly ideal results for QPSK and smaller degradation than without CPE compensation for 256-QAM. On the other hand, the range ISLR is not significantly improved as it is mainly caused by PN-induced ICI, which is not compensated. Finally, both Doppler shift PSLR and ISLR experience significant improvement, also with

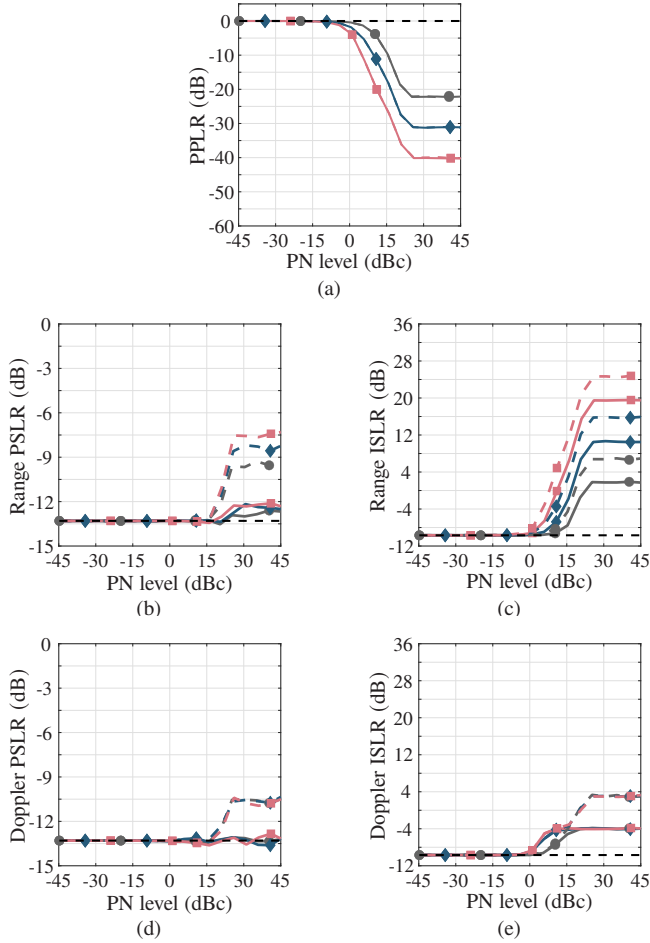


Figure 4.44: Simulated PPLR (a), range PSLR (b) and ISLR (c), and Doppler shift PSLR (e) and ISLR (d) as functions of the combined transmit and receive PN level considering CPE estimation and correction from (3.74) and (3.75), respectively. The same scenarios as in Figure 4.35 are considered, i.e., a single static target, $B = 500$ MHz as in Table 4.3, CP length of $N_{CP} = N$, and $M = 128$ OFDM symbols are assumed, as well as $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■). Results are also shown for 256-QAM with $N = 256$ (--- and ●), $N = 2048$ (--- and ◆), and $N = 16384$ (--- and ■). As in Figure 4.35, the ideal PPLR, PSLR and ISLR values without PN are also shown (---). Reprinted from [21], licensed under CC BY-NC-ND 4.0.

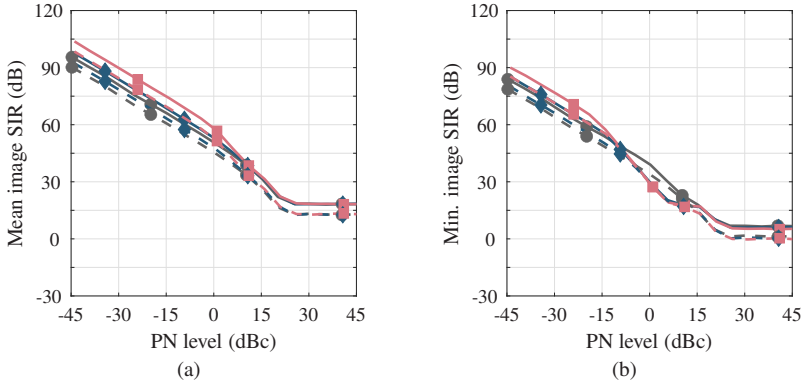


Figure 4.45: Simulated image SIR: (a) mean and (b) minimum values after CPE estimation and correction CPE estimation and correction from (3.74) and (3.75), respectively. As for Figure 4.45, a single static target was considered and $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■) were assumed. Results were also obtained for 256-QAM with $N = 256$ (--- and ●), $N = 2048$ (--- and ◆), and $N = 16384$ (--- and ■). In addition, $B = 500$ MHz as in Table 4.3, a CP length of $N_{CP} = N$, and $M = 128$ OFDM symbols were considered. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

better performance for QPSK compared to 256-QAM. As in the range case, some degradation is still experienced since the ICI also plays a role, although not dominant, in the Doppler shift estimation.

As for the SIR performance with the adopted CPE estimation and correction approaches in the same scenario, the mean and minimum SIR results from Figures 4.45(a) and 4.45(b) are respectively obtained. The results in Figure 4.45(a) show that a slower SIR degradation is observed after 0 dBc than without CPE compensation previously shown in Figure 4.42(a). In addition, a mean image SIR floor of around 15 dB instead of 0 dB in the case without CPE compensation is reached at around 20 dBc. In Figure 4.45(b), a significant SIR improvement is also observed, with at least 15 dB higher values than in the case without CPE correction in Figure 4.42(b).

When performing CPE estimation and correction in multi-target scenarios, the approaches described by (3.74) and (3.75) cannot be indiscriminately applied. Instead, CIR estimates can be obtained, e.g., based on pilot subcarriers, and

the CPE for each target can be estimated as discussed in Section 3.4.4. In practical applications, however, the radar targets tend to be rather weak and not detectable without the additional processing gain of M provided by Doppler shift estimation. In addition, if the aforementioned targets are moving, their CPE will not be distinguishable from their Doppler-shift-induced phase. A solution is then to use the dominant, reference path to estimate the CPE and reuse this estimation for the whole OFDM symbol, which encompasses all possible targets. While this can improve the communications performance of the OFDM-based ISAC system, the CPE compensation effectiveness will be reduced for targets at higher ranges than the reference path, which will appear at 0 m in the bistatic radar image [6, 22, 50]. The reason for this is that such targets will experience a different realization of the combined transmit and receive PN $\theta_{\text{Tx}}^{\text{PN}}(t - \tau_p) - \theta_{\text{Rx}}^{\text{PN}}(t)$, which will lead to different CPEs. This is illustrated in Figure 4.46, where the root mean squared difference between the CPE estimates for the reference path and a second target is analyzed and compared to the standard deviations of the individual CPE estimates. To obtain these results, Chebyshev windowing with 100 dB sidelobe suppression was applied to separate the reference path and the radar target in the obtained CIR estimates. In addition, only bistatic ranges between 5% and 95% of the maximum unambiguous range $R_{\text{max,ua}}^{\text{bist}} = N\Delta R^{\text{bist}} = N(c_0/B)$ were considered for the radar target to ensure that it is distinguishable from the reference path. Finally, the same SNR was assumed for both the radar target at all ranges and the reference path in the obtained CIRs. Although not realistic, this ensures the same CPE estimation accuracy for both of them to simplify the present analysis. The presented results in Figure 4.46 show that, for short OFDM symbols, i.e., $N = 256$, the PN variation is fast enough to yield non-negligibly biased CPE estimates for the radar target even for short ranges in case the estimate for the reference path is reused. For higher N values, less CPE performance degradation is observed since the induced CPE, which is related with the term in (3.73) that tends to zero with increasing N as the transmit and receive PN terms also do, is reduced.

4.2.3.3 Remarks on simulation results

It is worth highlighting that the assumed PN model shown in Figure 4.27 presents a comparatively low PN level compared to typical hardware. If communications or sensing is performed by a UE or between a gNB and a UE instead of the gNB-only scenarios depicted in Figure 3.8, considerably higher PN levels

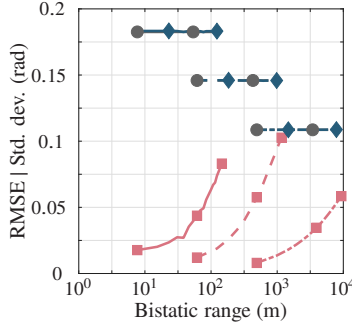


Figure 4.46: Simulated root mean squared difference between CPE estimates of reference path and target reflection as a function of the range for $N = 256$ (— and ■), $N = 2048$ (--- and ■), and $N = 16384$ (- · - and ■). For comparison, the CPE estimate standard deviation is shown for the reference path and $N = 256$ (— and ●), $N = 2048$ (--- and ●), and $N = 16384$ (- · - and ●). In addition, the CPE estimate standard deviation of the target reflection is also shown for $N = 256$ (— and ◆), $N = 2048$ (--- and ◆), and $N = 16384$ (- · - and ◆). As for the results shown in Figures 4.44 and 4.45, $B = 500$ MHz as in Table 4.3, a CP length of $N_{CP} = N$, and $M = 128$ OFDM symbols were considered. Reprinted from [21], licensed under CC BY-NC-ND 4.0.

can be observed as UE hardware is usually less advanced. This can be observed for lower FR2 bands below 30 GHz with 3GPP-compliant parameterization as discussed Section 4.2.3.2.6, where a combined transmit and receive PN level of -23.98 dBc ($\gamma = 20.92$ dB above the PN level from Figure 4.27) was observed between two UEs in a bistatic ISAC architecture. One example where even higher PN level is observed is the 77 GHz setup in [WAA⁺24], whose PN PSD is somewhat similar to what is obtained increasing the level of the original PN model from Figure 4.27 by $\gamma = 30$ dB. In this case, non-negligible EVM values are observed, which is graphically represented by the constellations in Figure 4.31.

Based on the results from Sections 4.2.3.2.1 to 4.2.3.2.6, it can be concluded that non-negligible PN-induced sensing performance degradation in terms of mainlobe and sidelobe distortion is only observed for combined transmit and receive PN levels above -15 dBc. At this point, PPLR, Doppler shift PSLR, and Doppler shift ISLR are the first sensing performance parameters that start to degrade, e.g., as seen in Figure 4.35(a), (d), and (e). Although slightly more robust, the range ISLR can also start degrading at this PN level depending on

the OFDM signal parameterization, e.g., as shown in Figure 4.38(c). As for the range PSLR, a significant higher robustness against PN compared to the other sensing performance parameters is observed, which non-negligible degradation being only observed for PN levels above around 11 dBc. At this point, the range processing gain N [18, 50] is not anymore sufficient to sufficiently suppress the PN-induced ICI, resulting in the fact that the highest sidelobe may either be significantly distorted or no longer be the nearest one to the mainlobe, which ideally results in a PSLR of -13.30 dB.

Still regarding PN-induced mainlobe and sidelobe level degradation, Section 4.2.3.2 showed that, while the ICI may introduce a certain degree of degradation, the dominating impairment for sensing tends to be the CPE, which is evidenced by the fact that Doppler shift PSLR and ISLR start degrading at lower combined transmit and receive PN levels than range PSLR and ISLR in all investigated settings. As a consequence, target sidelobes in the Doppler shift direction tend to present irregular shape and notably higher level than in the range direction of obtained radar images. As previously mentioned, however, this effect is only observed for combined transmit and receive PN levels from around -15 dBc ($\gamma = 29.90$ dB) onwards. In addition, it becomes more severe with increasing number M of OFDM symbols used for radar processing. This happens since PN is not an additive noise, being therefore not linearly suppressed by a radar processing gain as it is the case for AWGN [18, 50].

If SIR is analyzed instead of mainlobe and sidelobe degradation, it can be seen that mean and minimum image SIR are limited to around 60 dB and 30 dB at a PN level of -15 dBc. Although no degradation of PPLR, range PSLR and ISLR, and Doppler shift PSLR and ISLR at this PN level is observed as previously discussed, non-negligible sensing performance degradation is experienced especially in the Doppler shift direction, where unwanted sidelobes that cannot be suppressed by windowing drive the minimum SIR value up. This is, e.g., illustrated by Figure 4.41(b), where windowing is no longer effective against Doppler shift sidelobes at a similar PN level, i.e. -14.90 dBc. Considering CPE estimation and correction approaches from (3.74) (3.75), respectively, significant sensing performance improvement was observed. In terms of mainlobe and sidelobe levels, only negligible degradation w.r.t. the PN-free case was observed for combined transmit and receive PN levels of 0 dBc or lower, which covers all possible practical scenarios. In terms of SIR, the most relevant difference was observed for the aforementioned PN level range,

where at least 15 dB minimum SIR improvement was observed w.r.t. the case without CPE compensation, ensuring an SIR of at least 30 dB for all assumed sets of OFDM signal parameters.

It is worth mentioning that, if different PN PSD shapes are considered, which is the case when the parameters of either the transmit, receive or both PN contributions are different then the assumed ones, the most relevant parameters to evaluate the performance of OFDM-based ISAC systems will still be the combined transmit and receive PN level as well as the PLL loop bandwidth B_{PLL} , which defines the transition from CPE to ICI as the dominant PN-induced effect. Related analysis were in [KKP⁺14] to assess the performance of communications systems and [SSK⁺18b], where a brief discussion on the effects of PN on stepped-carrier OFDM radar was presented.

4.2.4 Sampling jitter analysis

This section presents a performance analysis of OFDM-based ISAC system under the influence of colored SJ at both ADC and DAC. In this context, both BB and BP sampling strategies are considered, and the OFDM signal parameters in Table 4.6 are adopted, and the specifically considered parameter values are mentioned in the following subsections. The digital IFs $f_{\text{IF}} = 0$ Hz and $f_{\text{IF}} = 1$ GHz are for BB and BP sampling, respectively.

Besides the aforementioned OFDM signal parameters, the PLL-based oscillator PN model discussed in [Sma12] and in Section 3.4.4 is adopted. Its parameters were chosen to match the PN PSD reported for the PLL contained in the Texas Instruments LMX2594 RF synthesizer [Tex19] for the closest frequency to the 122.88 MHz reference oscillator. Both the aforementioned PLL and sampling clock frequency are used for the reference oscillator from which the sampling clocks of DAC and ADC tiles are generated in the Zynq UltraScale+ RFSoc ZCU111 SoC platform [AMD21]. In reality, however, the oscillator signal passes through a Texas Instruments LMK04208 low-noise clock jitter cleaner with dual loop PLLs [Tex16], which shapes the PN PSD and eventually reduces the resulting SJ. For simplicity, the jitter cleaner LMK04208 is disregarded from the following analyses, and the root mean square (RMS) SJ associated with the LMX2594 is varied instead to allow analyzing the effect of a wider

Table 4.6: Adopted OFDM signal parameters for SJ analysis. Reprinted from [23], licensed under CC BY 4.0.

Carrier frequency (f_c)	26.2 GHz
Bandwidth (B)	500 MHz
Critical sampling period (T_s)	2 ns
Digital IF (f_{IF})	{0 Hz, 1 GHz}
Oversampling factor (η)	{1, 2, 4, 8}
No. of subcarriers (N)	{256, 512, 1024, 2048, 4096, 8192, 16384}
CP length (N_{CP})	{0, $N/4$, N }
No. of OFDM symbols (M)	128
Modulation alphabet	{QPSK, 16-QAM, 64-QAM, 256-QAM}

range of SJ RMS values on the communications and radar sensing performance in the considered OFDM-based ISAC system.

The adopted PN model parameters, which were based on data from the LMX2594 data sheet [Tex19], result in the double-sided PN PSD shown in Figure 4.47(a) and the integrated PN level shown in Figure 4.47(b). A PN time series was then generated based on the aforementioned PSD, and the resulting SJ from the PLL-based oscillator PN calculated according to (3.80). The resulting SJ PDF is shown in Figure 4.47(d). Due to the fact that the considered PN PSD from Figure 4.47(a) is not flat, the SJ is colored. Moreover, since individual RF synthesizers and therefore PLLs are used for the DAC and ADC tiles in the Zynq UltraScale+ RFSoc ZCU111, the DAC and ADC SJs $\delta_{DAC}^{SJ}(t)$ and $\delta_{ADC}^{SJ}(t)$ are henceforth assumed to be uncorrelated regardless of whether a monostatic or bistatic ISAC architecture is assumed and communications or sensing is performed.

4.2.4.1 Communications performance analysis

The SJ-induced communications performance degradation in an OFDM-based ISAC system over an ideal, noiseless channel as assumed in Section 3.4.4 is analyzed. For that, EVM and SIR are adopted as performance parameters.

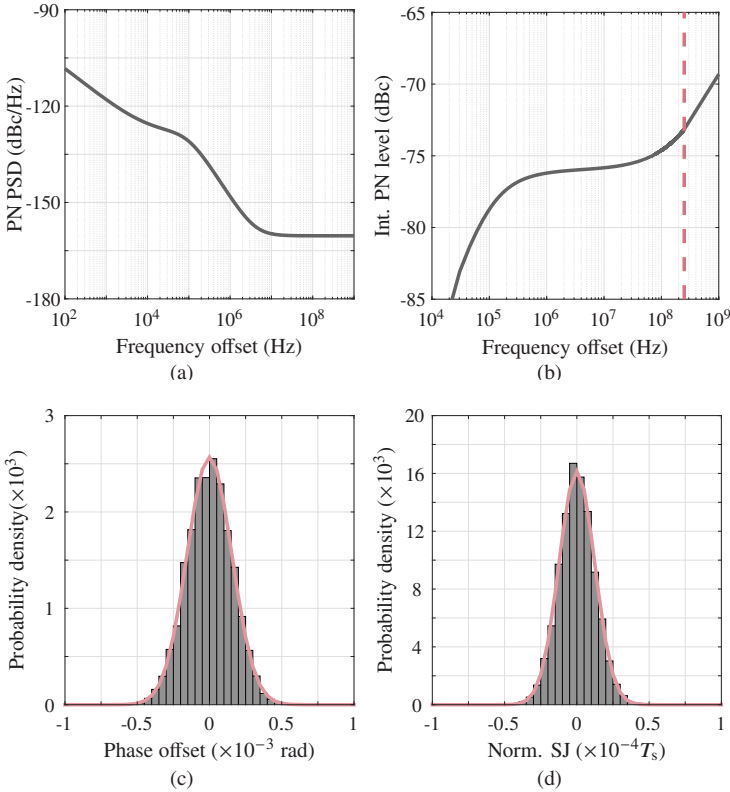


Figure 4.47. Simulated PN characteristics with a PN model based on the Texas Instruments LMX2594 RF synthesizer [Tex19]. The simulated double-sided PN PSD is shown in (a), while the integrated PN level as a function of frequency offset (double-sided) for the frequency range of interest assuming BB sampling and $\eta = 1$ is presented in (b). In (b), $B/2 = 250$ MHz is highlighted ($-$), as this offset corresponds to the bandwidth $B = 500$ MHz listed in Table 4.6. Finally, the PDF of the PN time series is shown in (c) and the corresponding SJ PDF in (d). Since a critical sampling period of $T_s = 2$ ns was considered, the PDF in (d) is associated with an RMS SJ of 49.44 fs. Reprinted from [23], licensed under CC BY 4.0.

Results and discussions are presented for both BB and BP sampling strategies in Sections 4.2.4.1.1 and 4.2.4.1.2, respectively, and remarks on the presented simulation results are given in Section 4.2.4.1.3.

4.2.4.1.1 Baseband sampling

In Figure 4.48, the achieved mean EVM is shown as a function of the number of subcarriers N for all modulation alphabets from Table 4.6, namely QPSK, 16-QAM, 64-QAM and 256-QAM. In addition, the case where no FDZP was performed, i.e., $\eta = 1$, as well as the cases with FDZP oversampling factors of $\eta \in \{2, 4, 8\}$ were analyzed. The achieved results assume the use of the SJ model with PDF depicted in Figure 4.47(d), which was derived from the PN PSD in Figure 4.47(a), but with varying RMS values to better analyze the effects of different SJ levels. The adopted SJ RMS values, which were assumed to be the same for both DAC and ADC, were 10^{-14} s, 10^{-13} s, 10^{-12} s, 10^{-11} s, and 10^{-10} s. Since $T_s = 1/B = 2$ ns, these numbers correspond to $0.5 \times 10^{-5}T_s$, $0.5 \times 10^{-4}T_s$, $0.5 \times 10^{-3}T_s$, $0.5 \times 10^{-2}T_s$, and $0.5 \times 10^{-1}T_s$, respectively. It can be seen in Figure 4.48(a) that all considered parameterizations result in the same EVM of -24.13 dB. When $\eta = 2$ is adopted in Figure 4.48(b), a significantly improved EVM is obtained in comparison to the case for $\eta = 1$ in Figure 4.48, especially for RMS SJ values equal to or smaller than 10^{-11} s (i.e., $0.5 \times 10^{-2}T_s$). When the oversampling factor is further increased to $\eta = 4$, an improvement of around 2 dB or less for $N \leq 1024$ is observed for an RMS SJ of 10^{-11} s. For RMS SJ values equal to or lower than 10^{-12} s, an improvement of up to 3 dB is observed for the whole considered N range. As for the oversampling factor $\eta = 8$, similar performance to the case with $\eta = 4$ is experienced, with exception for RMS SJ values equal to or lower than 10^{-12} s (i.e., $0.5 \times 10^{-3}T_s$). At the aforementioned SJ levels, up to around 1.4 dB higher EVM is experienced with $\eta = 8$. This is due to the fact that the experienced deviations from the correct sampling points are more relevant in this case than for $\eta = 4$, where the two times higher oversampled sampling period is experienced, i.e., $T_s^\eta = T_s/4$ instead of $T_s^\eta = T_s/8$. This result aligns with the fact that colored SJ makes oversampling less effective against the SJ than it would be in a case with white DAC and ADC SJ.

For all oversampling factors other than $\eta = 1$, i.e., $\eta \in \{2, 4, 8\}$, increasing the number of subcarriers N tends to result in lower EVM for RMS SJ equal to or lower than 10^{-11} s (i.e., $0.5 \times 10^{-2}T_s$). This is due to the smaller subcarrier spacing $\Delta f = B/N$, which leads to a more pronounced ICI effect and less relevant CPE. Since, however, the ICI is rather low for these RMS SJ values, the greater robustness against CPE with increasing N stands out. In addition, an EVM degradation along with N of between approximately 13 dB and 15 dB

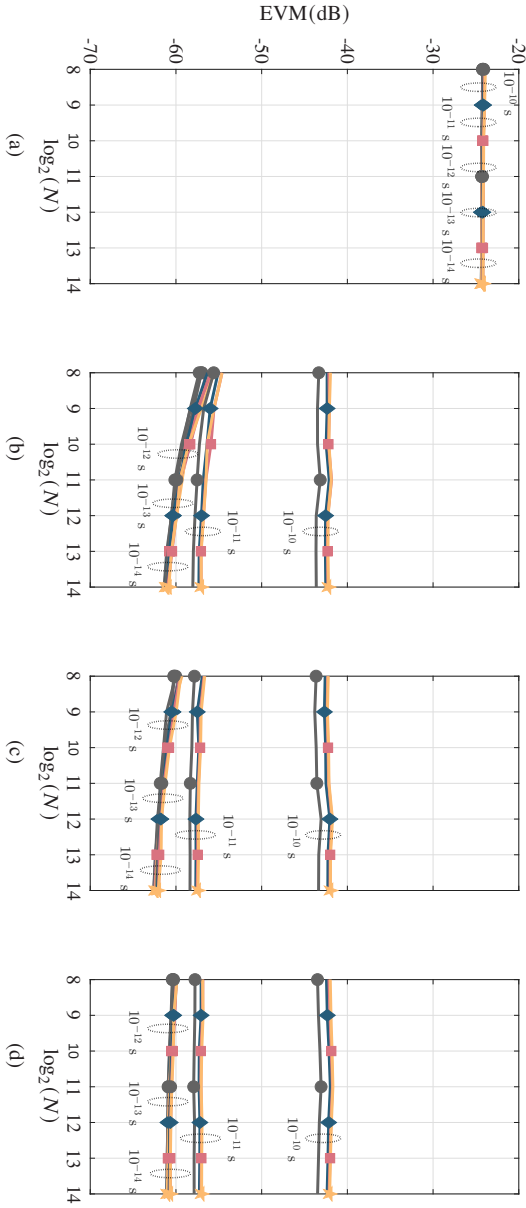


Figure 4.48. Simulated mean EVM obtained with BB sampling as a function of the number of subcarriers N . The performed simulations considered $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, and SJ RMS values of 10^{-14} s, 10^{-13} s, 10^{-12} s, 10^{-11} s, and 10^{-10} s at both transmitter and receiver. These values correspond to RMS SJs $0.5 \times 10^{-5} T_s$, $0.5 \times 10^{-4} T_s$, $0.5 \times 10^{-3} T_s$, $0.5 \times 10^{-2} T_s$, and $0.5 \times 10^{-1} T_s$, respectively. In addition, QPSK (●), 16-QAM (◆), 64-QAM (■), and 256-QAM (★) modulations were considered. For the results in (a), (b), (c), and (d), $\eta = 1$, $\eta = 2$, $\eta = 4$, and $\eta = 8$ were used, respectively. Reprinted from [23], licensed under CC BY 4.0.

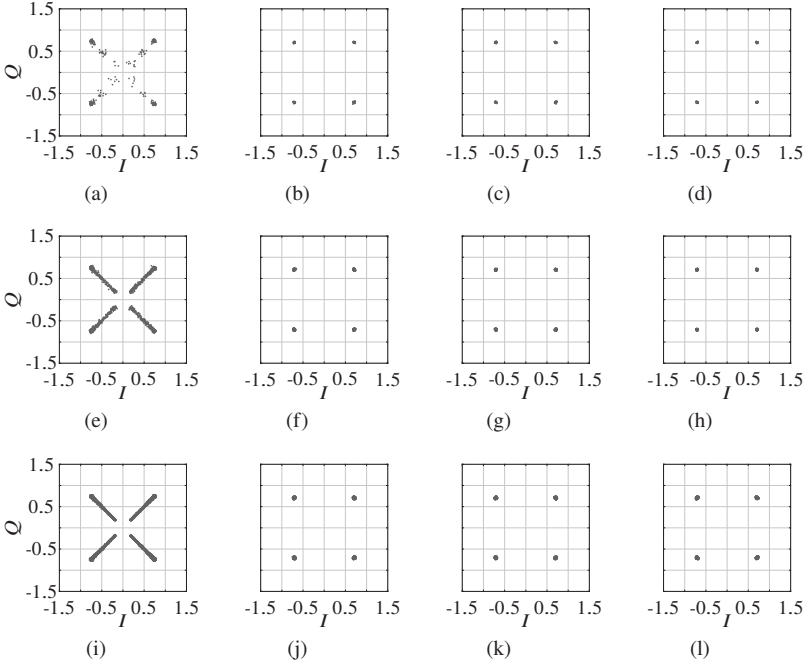


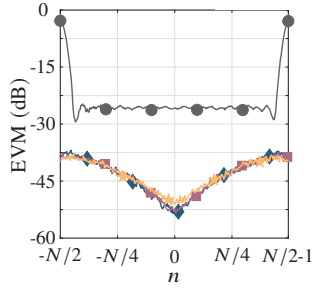
Figure 4.49. Simulated normalized receive QPSK constellations obtained with BB sampling, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, and an RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$). Results are shown for the same bandwidth of $B = 500$ MHz from Table 4.6 but different numbers of subcarriers and ZP factors, namely $N = 256$ and (a) $\eta = 1$, (b) $\eta = 2$, (c) $\eta = 4$, (d) $\eta = 8$, $N = 2048$ and (e) $\eta = 1$, (f) $\eta = 2$, (g) $\eta = 4$, (h) $\eta = 8$, and $N = 16384$ and (i) $\eta = 1$, (j) $\eta = 2$, (k) $\eta = 4$, (l) $\eta = 8$. Reprinted from [23], licensed under CC BY 4.0.

is experienced between the RMS SJ values of 10^{-11} s and 10^{-10} s (i.e., $0.5 \times 10^{-2} T_s$ and $0.5 \times 10^{-1} T_s$), showing that the RMS SJ point where performance degradation starts becoming relevant lies between these two values. At this point, one relevant factor for performance degradation is the fact that increasing DAC SJ leads to increased OOB radiation [HMAM15], which will ultimately result in additional ICI after sampling at an ADC also impaired by SJ.

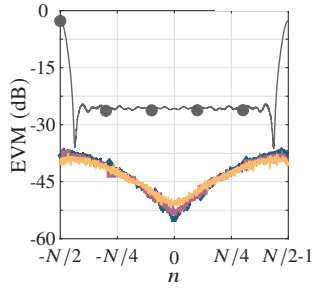
To illustrate the influence of DAC and ADC SJ on the receive constellations, Figure 4.49 shows normalized QPSK constellations obtained with the trans-

mission of OFDM symbols with $N \in \{256, 2048, 16384\}$, $N_{\text{CP}} = 0$ since an ideal channel is assumed, and $\eta \in \{1, 2, 4, 8\}$. For all cases, $M = 128$ OFDM symbols and an RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$) were considered. The obtained constellations confirm that the highest EVM values are observed for $\eta = 1$, besides an increasing EVM trend from $\eta = 2$ to $\eta = 8$, while N plays no significant role. While this would be different for lower RMS SJ values as previously discussed, rather low EVM values would be experienced and their effect would not be clearly visible in the QPSK constellations. To enable a better understanding of the mean EVM behavior and the constellation shapes, Figure 4.50 shows the EVM per subcarrier for the same parameters assumed for the results presented in Figure 4.49. The obtained results reveal that $\eta = 1$ results in pronounced distortion at the edge subcarriers due to the use of the critical sampling frequency. It can also be seen that the EVM pattern is clearly improved with oversampling factors $\eta \in \{2, 4, 8\}$. The obtained values are still frequency-selective, which, however, is due to the considered colored SJ. As expected from the results in Figure 4.48 for the RMS SJ of 10^{-10} s, no significant difference is observed for different N values for the considered RMS SJ value.

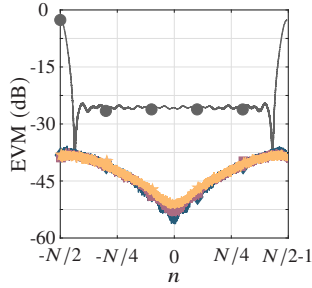
To further quantify the SJ-induced communications performance degradation in the considered OFDM-based ISAC system, Figure 4.51 shows the obtained SIR as a function of the RMS SJ, which is assumed to be equal at both DAC and ADC. For the performed simulations, the same numbers of subcarriers N , CP length, number of OFDM symbols, digital modulation schemes, and oversampling factors were adopted as for the results in Figures 4.49 and 4.50, i.e., $N \in \{256, 2048, 16384\}$, $N_{\text{CP}} = 0$, $M = 128$, QPSK and $\eta \in \{1, 2, 4, 8\}$. The achieved results show a constant SIR of 15.92 dB for $\eta = 1$ regardless of the number of subcarriers N , which is mainly due to the severe distortion of the edge subcarriers. For $\eta = 2$, in turn, a maximum SIR of around 56.44 dB is observed, whereas $\eta = 4$ and $\eta = 8$ yield maximum SIR values of approximately 57.27 dB and 57.33 dB, respectively. The obtained results agree with the fact that oversampling has limited effectiveness against ADC SJ and the OOB radiation due to DAC SJ, as previously discussed in Section 3.4.5.4. With increasing RMS SJ after 10^{-11} s (i.e., $0.5 \times 10^{-2} T_s$), however, the performance of the cases with higher η among $\eta \in \{2, 4, 8\}$ tend to start degrading. In addition, slightly higher robustness against increasing RMS SJ is observed for a higher number of subcarriers N for these oversampling factors, as was the case in the previous EVM analyses. Next, Figure 4.52 shows the obtained SIR considering different



(a)



(b)



(c)

Figure 4.50. Simulated EVM at the n th subcarrier obtained with BB sampling and an RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$), $B = 500$ MHz as in Table 4.6, QPSK modulation, and: (a) $N = 256$, (b) $N = 2048$, and (c) $N = 16384$. In all cases, results are shown for $\eta = 1$ (—), $\eta = 2$ (—), $\eta = 4$ (—), and $\eta = 8$ (—). Adapted from [23], licensed under CC BY 4.0.

RMS SJ values for the DAC and the ADC, $N = 2048$, and QPSK modulation.

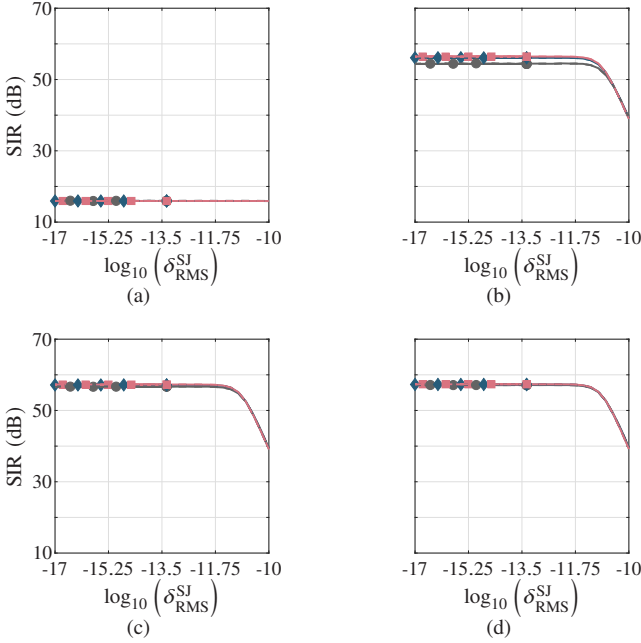


Figure 4.51. Simulated SIR as a function of the RMS SJ obtained with BB sampling, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, QPSK modulation, and (a) $\eta = 1$, (b) $\eta = 2$, (c) $\eta = 4$, and (d) $\eta = 8$. In all cases, results are shown for $N = 256$ (●), $N = 2048$ (◆), and $N = 16384$ (■). Adapted from [23], licensed under CC BY 4.0.

It can be seen that, apart from $\eta = 1$, which achieves the worst SIR performance at low RMS SJs, a slight SIR improvement is observed when increasing η from 2 to 4, but no further noticeable enhancement is observable for $\eta = 8$, which supports the previous results. In addition, the obtained results reveal that the adopted colored SJ model leads to similar effects when applied at either DAC or ADC, which agrees with the result from (3.90).

4.2.4.1.2 Bandpass sampling

In BP sampling, the transmit signal must first undergo oversampling before DUC to the digital IF. The inverse is done at the receiver side, i.e., DDC followed by downsampling. In this subsection, a digital IF of $f_{\text{IF}} = 1$ GHz and

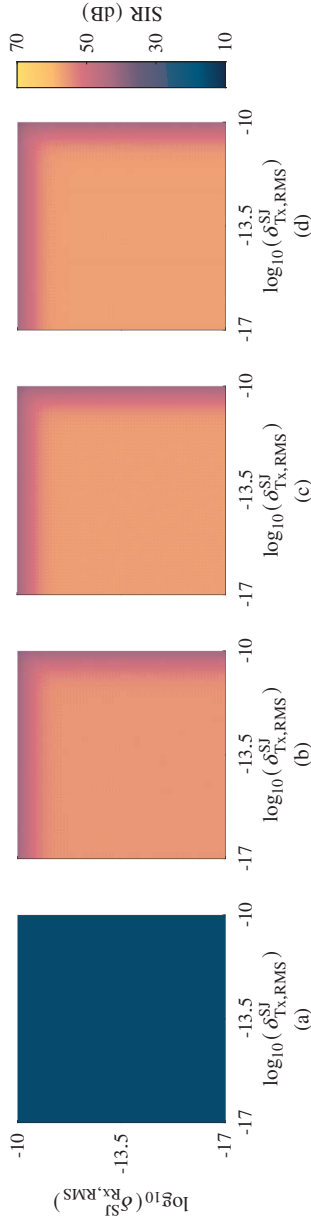


Figure 4.52. Simulated SIR obtained with BB sampling, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, QPSK modulation, and different RMS SJ values at the DAC and ADC. Results are shown for $N = 128$ and (a) $\eta = 1$, (b) $\eta = 2$, (c) $\eta = 4$, and (d) $\eta = 8$. Adapted from [23], licensed under CC BY 4.0.

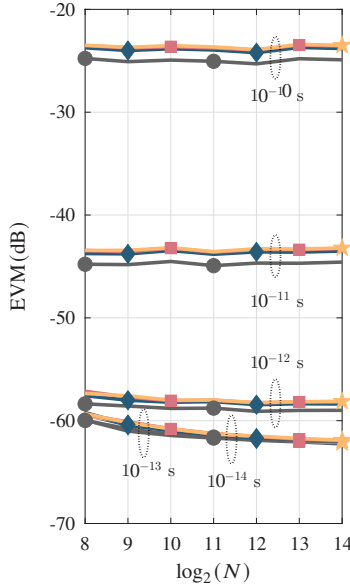


Figure 4.53. Simulated mean EVM obtained with BP sampling at a digital IF $f_{IF} = 1$ GHz as a function of the number of subcarriers N . For the obtained results, $\eta = 8$, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, and SJ RMS values of 10^{-14} s, 10^{-13} s, 10^{-12} s, 10^{-11} s, and 10^{-10} s (i.e., $0.5 \times 10^{-5}T_s$, $0.5 \times 10^{-4}T_s$, $0.5 \times 10^{-3}T_s$, $0.5 \times 10^{-2}T_s$, and $0.5 \times 10^{-1}T_s$, respectively) at both the DAC and the ADC were considered. As in Figure 4.48, QPSK (●), 16-QAM (◆), 64-QAM (■), and 256-QAM (★) modulations were considered. Reprinted from [23], licensed under CC BY 4.0.

an oversampling factor $\eta = 8$ used during FDZP are adopted to analyze the effect of SJ on the communications performance of an OFDM-based ISAC system with BP sampling. Given these parameters, Figure 4.53 shows the achieved mean EVM as a function of the number of subcarriers for the same digital modulations and RMS SJ values considered for the results for BB sampling in Figure 4.48. Compared to the results for $\eta = 8$ in the BB sampling case, considerably higher EVM values are achieved with BP sampling.

The aforementioned results are illustrated by the normalized receive QPSK constellations in Figure 4.54 for the RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1}T_s$). The observed results show no difference in the constellations densities for

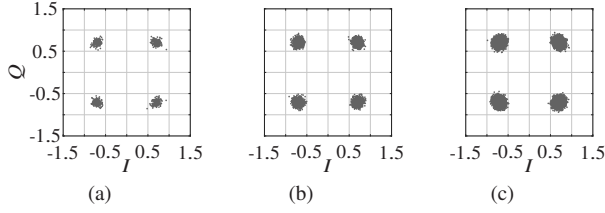


Figure 4.54. Simulated normalized receive QPSK constellations obtained with BP sampling at a digital IF $f_{IF} = 1$ GHz, $\eta = 8$, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, and an RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$) at both the DAC and the ADC. Results are shown for (a) $N = 256$, (b) $N = 2048$, and (c) $N = 16384$ total number of subcarriers. Reprinted from [23], licensed under CC BY 4.0.

different N as expected from Figure 4.53. The only observable difference is the presence of fewer constellation points for $N = 16384$ as the same number of OFDM symbols, i.e., $M = 128$, was assumed in all cases. Compared to the constellations in the BB case with $\eta = 8$ in Figure 4.49, a clear degradation of the constellation is observed both in the form of spreading due to ICI and slight rotation of some points due to CPE, which is mainly due to two factors. The first is the more severe ICI effect. This happens since the OFDM signal is transmitted at higher frequencies, where the PN PSD from which the SJ is derived accumulates to a higher level as seen in Figure 4.47(b). The second factor, which is the source of the constellation rotation, is the CPE that is experienced due to the PN-like effect induced by the SJ on the digital IF used to transmit the OFDM signal as discussed in [PF09]. These results are linked to the ones in Figure 4.55, where the obtained EVM per subcarrier for the same signal parameters are shown. Unlike the results for BB sampling in Figure 4.50, a smaller EVM variation with an overall higher level is observed in all cases with BP sampling, which is due to the aforementioned accumulated PN level at that region and SJ-induced CPE to the digital IF.

Next, Figure 4.56 shows the experienced SIR as a function of the RMS SJ for QPSK modulation and the same remaining signal parameters assumed for the results in Figures 4.54 and 4.55. The obtained results show that a maximum SIR of around 55.82 dB is achieved for all considered N values, which is around 1.51 dB lower than in the BB sampling case with $\eta = 8$ in Figure 4.51(d). For $N = 16384$, an even lower SIR of around 45.01 dB is achieved at the same aforementioned RMS SJ. If CPE estimation and correction is performed, e.g.,

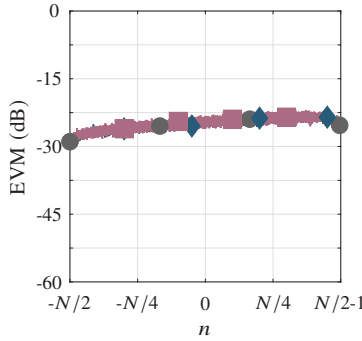


Figure 4.55. Simulated EVM at the n th subcarrier obtained with BP sampling at a digital IF $f_{\text{IF}} = 1$ GHz, $\eta = 8$, $B = 500$ MHz as in Table 4.6, $M = 128$ OFDM symbols, QPSK modulation and an RMS SJ of 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$) at both the DAC and the ADC. Results are shown for $N = 256$ (●), $N = 2048$ (◆), and $N = 16384$ (■) total number of subcarriers. Reprinted from [23], licensed under CC BY 4.0.

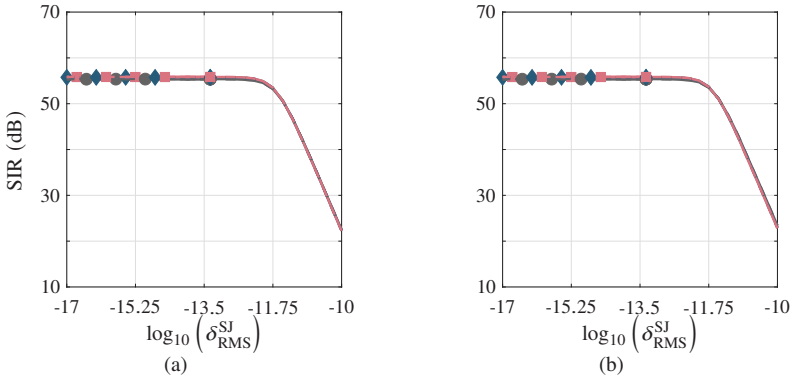


Figure 4.56. Simulated SIR as a function of the RMS SJ (same at both the DAC and the ADC) obtained with BP sampling at a digital IF $f_{\text{IF}} = 1$ GHz, $\eta = 8$, $B = 500$ MHz as in Table 4.6, and $M = 128$ OFDM symbols (a) without and (b) with CPE correction. In all cases, results are shown for $N = 256$ (●), $N = 2048$ (◆), and $N = 16384$ (■), as well as QPSK (—) and 256-QAM (---) modulations. In the obtained results, an SIR improvement of only around 0.57 dB is seen when CPE compensation is performed. In addition, no significant difference is observed between the experienced SIR with QPSK and 256-QAM. Adapted from [23], licensed under CC BY 4.0.

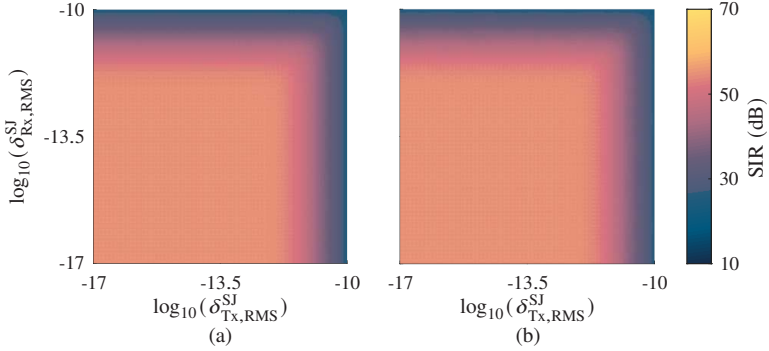


Figure 4.57. Simulated SIR as a function of the RMS SJ (same at both the DAC and the ADC) obtained with BP sampling at a digital IF of $f_{\text{IF}} = 1$ GHz, $\eta = 8$, $B = 500$ MHz as in Table 4.6, $N = 2048$ subcarriers, and $M = 128$ OFDM symbols, and QPSK modulation (a) without and (b) with CPE correction. As in Figure 4.56, only around 0.57 dB SIR improvement due to CPE compensation was observed. Adapted from [23], licensed under CC BY 4.0.

as described in Section 3.4.4, the SIR is improved by only around 0.57 dB for RMS SJs above 10^{-11} s (i.e., $0.5 \times 10^{-2} T_s$), which is where the SIR starts degrading. The reason for this is that, when CPE becomes relevant, the ICI induced by SJ is already severe enough to impair the CPE estimation and make its correction less effective. Results assuming different RMS SJ values at the DAC and ADC are finally shown for a reduced RMS SJ range in Figure 4.57 for the cases with and without CPE compensation. As in the BB sampling case for $\eta = 8$ in Figure 4.52(d), similar effects are observed for both DAC and ADC SJs. With BP sampling, however, a significant SIR degradation is experienced as previously discussed, and the CPE correction presents the same slight SIR improvement of only around 0.57 dB for RMS SJs above 10^{-11} s. This allows concluding that, for the adopted colored SJ model, ICI is the dominant impairment induced by SJ.

4.2.4.1.3 Remarks on simulation results

As evidenced by the obtained expressions (3.90) and (3.91), besides related studies in the literature, it was shown that SJ exclusively causes ICI if BB sampling is adopted. If BP sampling is adopted instead, SJ also leads to CPE and

additional ICI due to a PN-like effect on the digital IF. Consequently, BP tends to result in more severely degraded communications performance, yielding higher EVM and lower SIR than in the BB case as the receive constellations are not only spread due to ICI, but also rotated due to CPE. It was also shown that the absence of oversampling in the BB sampling case, i.e., $\eta = 1$, leads to the distortion of the edge subcarriers and therefore the highest EVM and lowest SIR. While performing oversampling prevents that issue, the results have shown that adopting increasing oversampling factors among the considered values $\eta \in \{2, 4, 8\}$ only leads to improvement at rather low RMS SJ at both DAC and ADC. For higher RMS SJ, however, more significant OOB radiation due to DAC SJ happens, which combined with the ICI induced by both DAC and ADC SJ, leads to more degraded communications performance, both in terms of EVM and SIR, than with a lower oversampling factor among the considered ones. Finally, it can be concluded that RMS SJ values of up to around 10^{-10} s (i.e., $0.5 \times 10^{-1} T_s$) at both DAC and ADC result in tolerable communications performance degradation for BB sampling with all considered oversampling factors, ensuring EVM below -40 dB and maximum SIR of around 56 dB to 57 dB. The reason why better values are not obtained with even lower RMS SJ values is due to limited accuracy in the SJ simulation, which is done via resampling with a Farrow filter [Far88] preceded by cubic interpolation [EGH93]. For the BP sampling scenario at an IF of $f_{IF} = 1$ GHz and with oversampling $\eta = 8$, the same RMS SJ upper bound applies.

4.2.4.2 Radar sensing performance analysis

In this section, the effects of SJ on the sensing performance of OFDM-based ISAC systems are analyzed. For that purpose, the same SJ model from Section 4.2.4.1 is adopted. To analyze how the EVM and SIR degradation ultimately translate into degradation of target reflections in a radar image, several performance parameters are used. These include PPLR, PSLR, and ISLR [LMI16, LLL⁺24, 18], as well as SIR [21]. To better understand the effects of SJ, the PSLR and ISLR parameters are evaluated for both range and Doppler shift directions of the simulated radar images. Finally, SIR is used as a measure of the ratio between a target peak and the interference floor raised by SJ in the radar image.

The obtained results assuming BB and BP sampling are discussed in Sections 4.2.4.2.1 and 4.2.4.2.2, respectively, with remarks on the presented simulation results given in Section 4.2.4.2.3. For the performed simulations, the parameters from Table 4.6 are considered. For conciseness, however, only $N \in \{256, 2048, 16384\}$, besides QPSK and 256-QAM are henceforth considered. In addition, a scenario with a single static target was considered and all other typical impairments in OFDM-based ISAC systems such as AWGN, external interference and self-interference were neglected to focus on the SJ effects. Since the assumed SJ model is uncorrelated in both monostatic and bistatic ISAC architectures, the same degradation due to SJ is experienced in both cases. Therefore, the following radar sensing performance analyses assume a bistatic OFDM-based ISAC with perfect synchronization and full knowledge of the transmit OFDM frame for radar signal processing. Finally, the same SJ is assumed for both DAC and ADC. This is done since, as also demonstrated in [GGT21] and by the results in Section 4.2.4.1, both DAC and ADC SJ yield the same effects. The expression in (3.90) also confirms this, showing that the combined DAC and ADC SJ is the defining factor for the OFDM-based ISAC system performance.

4.2.4.2.1 Baseband sampling

Assuming BB sampling for the considered OFDM-based ISAC system, PPLR, range and Doppler shift PSLR, as well as range and Doppler shift ISLR were calculated as functions of the normalized RMS SJ values between 10^{-17} s and 10^{-10} s (i.e., $0.5 \times 10^{-8}T_s$ and $0.5 \times 10^{-1}T_s$, respectively) for oversampling factors $\eta \in \{1, 2, 4, 8\}$. In the performed simulations, the considered RMS SJ values were assumed to be equal at both DAC and ADC, and only rectangular windowing was performed during both range and Doppler processing. Since negligible difference between QPSK and 256-QAM, among different N values, and for different RMS SJ values within the considered range was observed, the obtained results are shown in Table 4.7 as a function of the oversampling factor η only. The obtained results show negligible PPLR for all cases, and similar overall performance. The most pronounced differences, which are still rather small, are observed for range PSLR and ISLR with $\eta = 1$. With this oversampling factor, the worst values are achieved due to the distortion of the edge subcarriers even with low RMS SJ. It is worth highlighting that, since CPE due to SJ is not experienced in the BB sampling case and the fixed number

Table 4.7: Simulated PPLR, range and Doppler shift PSLR, and range and Doppler shift ISLR for BB sampling and RMS SJ between 10^{-17} s and 10^{-10} s ($0.5 \times 10^{-8} T_s$ and $0.5 \times 10^{-1} T_s$). Reprinted from [23], licensed under CC BY 4.0.

η	PPLR	Range PSLR (dB)	Doppler PSLR (dB)	Range ISLR (dB)	Doppler ISLR (dB)
1	0 dB	-13.44 dB	-13.30 dB	-10.84 dB	-9.68 dB
2	0 dB	-13.30 dB	-13.30 dB	-9.68 dB	-9.68 dB
4	0 dB	-13.30 dB	-13.30 dB	-9.69 dB	-9.68 dB
8	0 dB	-13.27 dB	-13.30 dB	-9.67 dB	-9.68 dB

$M = 128$ of OFDM symbols is assumed, the same Doppler shift PSLR and ISLR is observed in all cases.

While PPLR, PSLR, and ISLR do not clearly showcase the influence of SJ on the radar sensing performance, the image SIR distinctly captures this effect as illustrated in Figures 4.58 and 4.59. These figures show the achieved mean and minimum image SIR, respectively, as functions of the RMS SJ. These parameters are defined as the average and minimum power ratio, respectively, that is calculated between the target peak and each point outside the mainlobe. The performed simulations assumed equal RMS SJ at both DAC and ADC, as well as the remaining parameters used for the previously discussed PPLR, PSLR, and ISLR results. Overall, a better performance is obtained with QPSK compared to 256-QAM, with a more pronounced difference in the mean image SIR case ranging from 3 dB to 6 dB. In addition, increasing numbers of subcarriers N and oversampling factors η also result in higher SIR, although through different mechanisms. While a higher N increases the processing gain achieved against the SJ-induced ICI during range processing, a higher η results in more inactive FDZP subcarriers, which absorb part of the SJ-induced leakage discussed in Section 3.4.5.4 and are discarded at the receiver instead of corrupting the active subcarriers. It can also be seen that the mean image SIR values are from around 12 dB to 37 dB higher than the minimum image SIR values, with exact values depending on η , N , and the modulation alphabet. This is due to the SJ-induced ICI, which may lead to artifacts that cannot always be suppressed by windowing as later shown in the results in Figures 4.60 and 4.61. For all considered η values, the SIR values only start degrading at RMS SJ equal to or

higher than 10^{-11} s (i.e., $0.5 \times 10^{-2} T_s$), which was also the case in the communications analysis in Section 4.2.4.1. Unlike in the communications case, a more pronounced image SIR improvement is observed when increasing the oversampling factor from $\eta = 2$ to $\eta = 4$, and from $\eta = 4$ to $\eta = 8$. This is particularly visible for mean SIR with $N \in \{256, 2048\}$ (comparing Figures 4.58(b) and 4.58(c)) in the former case, and with $N \in \{256\}$ (comparing Figures 4.58(c) and 4.58(d)) in the latter. In addition, the fact that maximum observed mean SIR is limited at around 136 dB, as seen for $N = 16384$ and QPSK modulation in Figure 4.58(b), (c), and (d) is due to the previously discussed limited accuracy in the SJ simulation.

To illustrate the influence of the degradation of the previously analyzed parameter on obtained radar images, Figures 4.60 and 4.61 show obtained bistatic range-Doppler shift radar images with $\eta = 1$ and $\eta = 8$, respectively. Further considered parameters are $N = 2048$, $N_{CP} = 2048$, $M = 128$, and QPSK. In addition, two targets were simulated. The first was at 10 m with a Doppler shift of 0 kHz, and the second at 15 m with a Doppler shift of $0.1 \Delta f$. In both figures, radar images are shown for the cases where rectangular windowing or Chebyshev windowing with 100 dB sidelobe suppression are adopted for both range and Doppler shift processing, with RMS SJ values at both DAC and ADC of 10^{-16} s, 10^{-13} s, and 10^{-10} s (i.e., $0.5 \times 10^{-7} T_s$, $0.5 \times 10^{-4} T_s$, and $0.5 \times 10^{-1} T_s$, respectively) for both cases. While rectangular windowing results in high range and Doppler shift sidelobes, which partly mask the effect of SJ as illustrated by the previously presented discussion on PPLR, PSLR and ISLR results, Chebyshev windowing allows clearly observing the expected SIR degradation along with increasing RMS SJ as previously seen in Figures 4.58 and 4.59.

4.2.4.2.2 Bandpass sampling

For the case where BP sampling is performed at a digital IF of $f_{IF} = 1$ GHz with oversampling factor $\eta = 8$, Table 4.8 shows the simulated PPLR, range and Doppler shift PSLR, as well as range and Doppler shift ISLR. Simulations were performed for both QPSK and 256-QAM, $N \in \{256, 2048, 16384\}$ and $M = 128$, within the RMS SJ range from 10^{-17} s to 10^{-10} s (i.e., $0.5 \times 10^{-8} T_s$ to $0.5 \times 10^{-1} T_s$, respectively). As in the BB case in Section 4.2.4.2.1, negligible differences between all these settings were observed, which is why a single result is shown for each performance parameter. Compared to the BB results

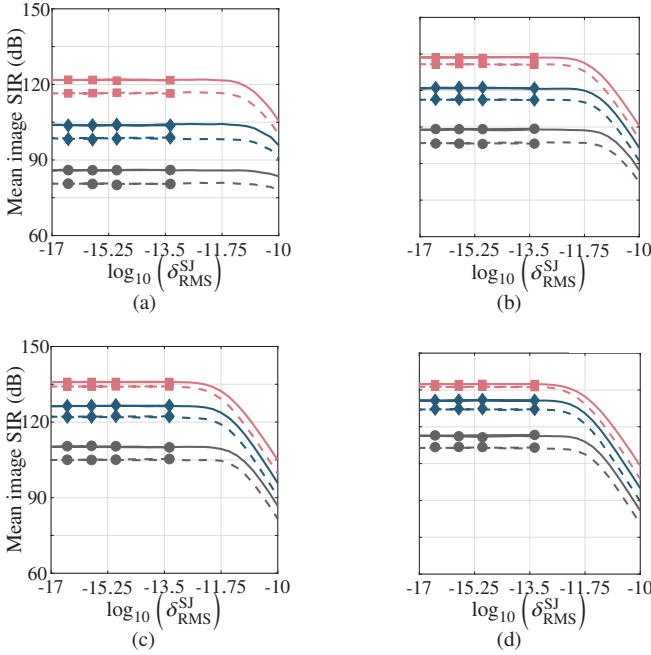


Figure 4.58. Simulated mean image SIR as a function of the normalized RMS SJ by the critical sampling period T_s (same at both the DAC and the ADC) jitter obtained with BB sampling, $B = 500$ MHz as in Table 4.6, and $M = 128$ OFDM symbols. The simulated mean image SIR is shown for (a) $\eta = 1$, (b) $\eta = 2$, (c) $\eta = 4$, and (d) $\eta = 8$. In all cases, results are shown for $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (--- and ◆), and $N = 16384$ and QPSK (— and ■), as well as $N = 256$ and 256-QAM (— and ●), $N = 2048$ and 256-QAM (--- and ◆), and $N = 16384$ and 256-QAM (— and ■). Adapted from [23], licensed under CC BY 4.0.

for $\eta = 8$ in Table 4.7, only negligible differences of 0.06 dB and 0.03 dB were observed in range PSLR and ISLR, respectively.

Next, the mean and minimum image SIR obtained as a function of the RMS SJ are shown in Figures 4.62(a) and 4.62(b), respectively. Since negligible improvement is experienced with CPE correction as previously demonstrated by the results in Figure 4.56, only results without CPE correction are shown. As in the BB case in Figures 4.58 and 4.59, equal RMS SJ was assumed for both DAC and ADC, and the same set of parameters except for a fixed oversampling factor $\eta = 8$ is adopted. Overall, a similar behavior as in the BB case with $\eta = 8$

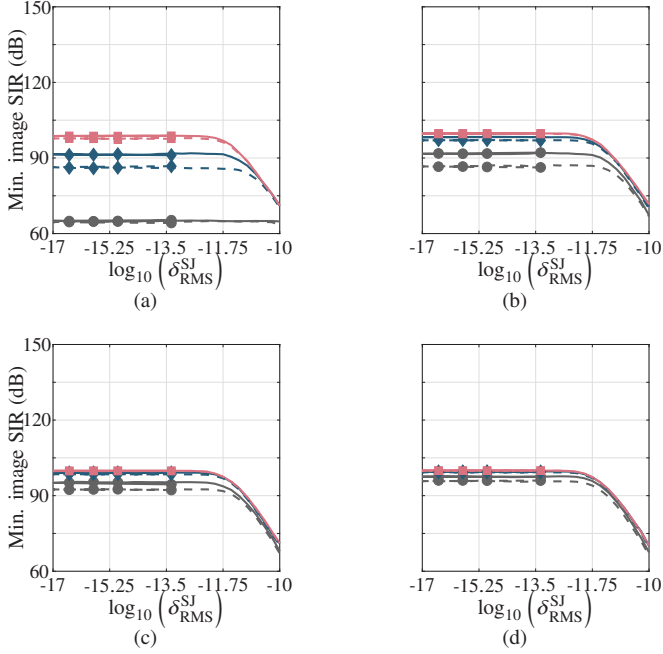


Figure 4.59. Simulated minimum image SIR as a function of the normalized RMS SJ by the critical sampling period T_s (same at both the DAC and the ADC) jitter obtained with BB sampling, $B = 500$ MHz as in Table 4.6, and $M = 128$ OFDM symbols. The simulated minimum image SIR is shown for (a) $\eta = 1$, (b) $\eta = 2$, (c) $\eta = 4$, and (d) $\eta = 8$. In all cases, results are shown for $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ♦), and $N = 16384$ and QPSK (— and ■), as well as $N = 256$ and 256-QAM (— and ●), $N = 2048$ and 256-QAM (— and ♦), and $N = 16384$ and 256-QAM (— and ■). Adapted from [23], licensed under CC BY 4.0.

Table 4.8: Simulated PPLR, Range and Doppler Shift PSLR, and Range and Doppler Shift ISLR for BP Sampling and RMS SJ between 10^{-17} s and 10^{-10} s ($0.5 \times 10^{-8} T_s$ and $0.5 \times 10^{-1} T_s$). Reprinted from [23], licensed under CC BY 4.0.

η	PPLR	Range PSLR (dB)	Doppler PSLR (dB)	Range ISLR (dB)	Doppler ISLR (dB)
8	0 dB	−13.33 dB	−13.30 dB	−9.70 dB	−9.68 dB

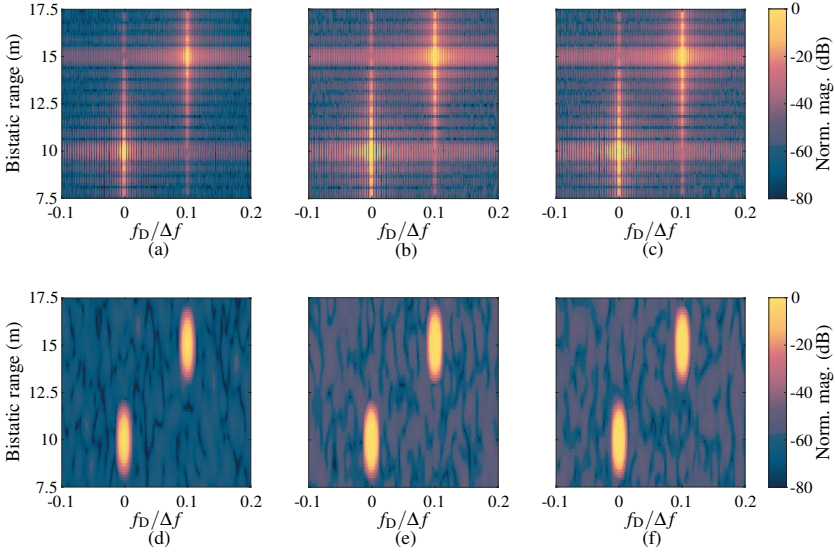


Figure 4.60. Simulated bistatic range-Doppler shift radar images obtained with BB sampling, $B = 500$ MHz as in Table 4.6, $N = 2048$, $N_{CP} = 2048$, $M = 128$, QPSK modulation, and $\eta = 1$. A target at 10 m with a Doppler shift of 0 kHz and another at 15 m with a Doppler shift of $0.1\Delta f$ were considered. Results are shown for rectangular windowing for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (a) 10^{-16} s, (b) 10^{-13} s, and (c) 10^{-10} s (i.e., $0.5 \times 10^{-7}T_s$, $0.5 \times 10^{-4}T_s$, and $0.5 \times 10^{-1}T_s$, respectively), as well as for and Chebyshev windowing with 100 dB sidelobe suppression for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (d) 10^{-16} s, (e) 10^{-13} s, and (f) 10^{-10} s. Reprinted from [23], licensed under CC BY 4.0.

is observed, i.e., higher N yields higher image SIR and QPSK outperforms 256-QAM. Lower image SIR values are obtained in the BP sampling case due to the fact that CPE is experienced besides ICI. Comparing Figure 4.62(a) with the BB sampling results from Figure 4.58(d), lower mean image SIR values by 7.11 dB, 4.93 dB, and 1.79 dB are experienced considering QPSK modulation and $N = 256$, $N = 2048$, and $N = 16384$, respectively. This aligns with the fact that lower N results in more pronounced CPE. For 256-QAM, these values are increased to 7.24 dB, 5.90 dB, and 2.36 dB as higher modulation order results in more severe effect for the same CPE. In terms of minimum image SIR, however, the results for BP sampling in Figure 4.62(b) are worse than their BB

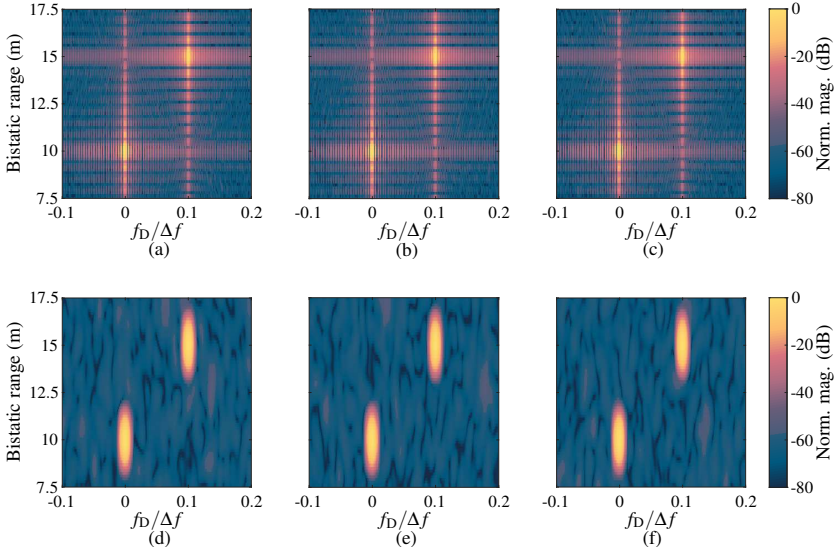


Figure 4.61. Simulated bistatic range-Doppler shift radar images obtained with BB sampling, $B = 500$ MHz as in Table 4.6, $N = 2048$, $N_{CP} = 2048$, $M = 128$, QPSK modulation, and $\eta = 8$. The same targets as for the radar images shown in Figure 4.60 were considered. Results are shown for rectangular windowing for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (a) 10^{-16} s, (b) 10^{-13} s, and (c) 10^{-10} s (i.e., $0.5 \times 10^{-7}T_s$, $0.5 \times 10^{-4}T_s$, and $0.5 \times 10^{-1}T_s$, respectively), as well as for Chebyshev windowing with 100 dB sidelobe suppression for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (d) 10^{-16} s, (e) 10^{-13} s, and (f) 10^{-10} s. Reprinted from [23], licensed under CC BY 4.0.

counterparts in Figure 4.59(d) only for $N = 256$. More specifically, 2.32 dB and 3.59 dB lower values are obtained for QPSK and 256-QAM, respectively.

To illustrate the influence of the previously analyzed mainlobe and sidelobe degradation, measured by PPLR, PSLR, and ISLR, as well as of the dynamic range reduction measured by the SIR, on the obtained radar images, Figure 4.63 shows simulated bistatic range-Doppler shift radar images with BP sampling at a digital IF of $f_{IF} = 1$ GHz and with $\eta = 8$. In addition, $N = 2048$, $N_{CP} = 2048$, $M = 128$, and QPSK modulation were considered. As for windowing, results are shown for both rectangular window and Chebyshev window with 100 dB sidelobe suppression applied during both range and Doppler shift processing. The same two targets as for the BB case in Figures 4.60 and 4.61 are considered,

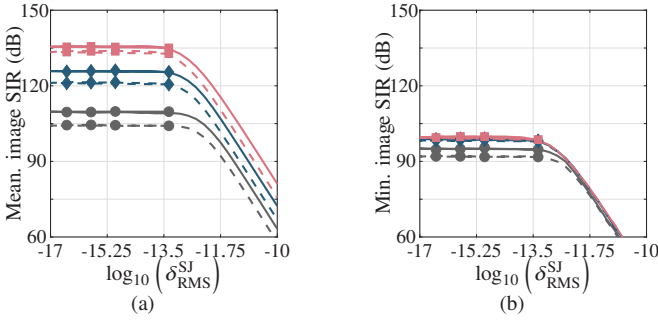


Figure 4.62. Simulated mean (a) and minimum (b) image SIR as a function of the normalized RMS SJ by the critical sampling period T_s (same at both the DAC and the ADC) jitter obtained with BP sampling at a digital IF $f_{IF} = 1$ GHz, $\eta = 8$, $B = 500$ MHz as in Table 4.6, and $M = 128$ OFDM symbols. In all cases, results are shown for $N = 256$ and QPSK (— and ●), $N = 2048$ and QPSK (— and ◆), and $N = 16384$ and QPSK (— and ■), as well as $N = 256$ and 256-QAM (--- and ●), $N = 2048$ and 256-QAM (--- and ◆), and $N = 16384$ and 256-QAM (— · — and ■). Adapted from [23], licensed under CC BY 4.0.

namely one at 10 m with a Doppler shift of 0 kHz, and the other at 15 m with a Doppler shift of $0.1\Delta f$. Furthermore, RMS SJ values of 10^{-16} s, 10^{-13} s, and 10^{-10} s (i.e., $0.5 \times 10^{-7}T_s$, $0.5 \times 10^{-4}T_s$, and $0.5 \times 10^{-1}T_s$, respectively) were considered, with the same value being adopted for both DAC and ADC at each simulation. The resulting range-Doppler shift radar images from simulations with the aforementioned parameters confirm the findings from the previous results on radar sensing with BP sampling in the considered OFDM-based ISAC system. More specifically, they show that both ICI and CPE strongly influence the quality of the obtained radar images, which become severely degraded at significantly lower RMS SJ than in the BB case. While the ICI can be suppressed to a certain extent due to the radar processing gain, the CPE effect is not effectively attenuated, leading to stripes in the Doppler shift direction at the range of the true targets. This is especially visible in Figure 4.63(f) where 10^{-10} s RMS SJ and Chebyshev windowing were considered. Since, in contrast to the range sidelobes, these artifacts cannot be effectively suppressed by windowing, they can eventually lead to the detection of ghost targets.

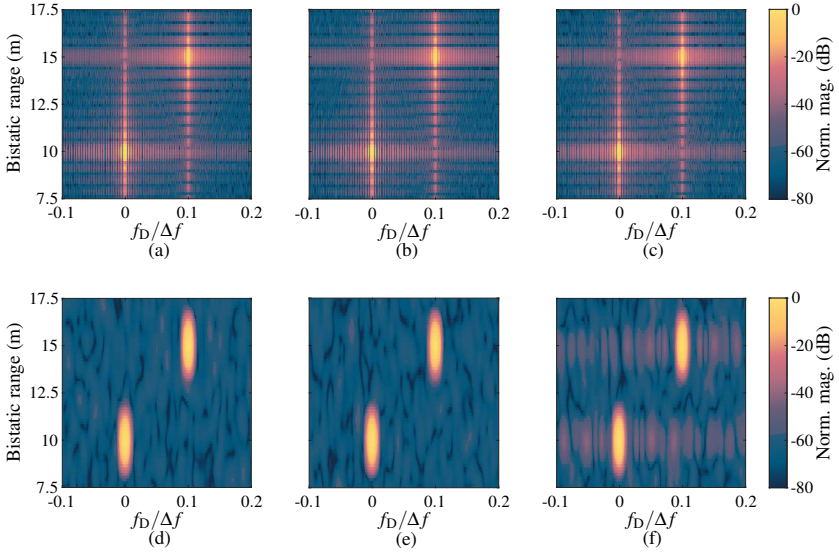


Figure 4.63. Simulated bistatic range-Doppler shift radar images obtained with BP sampling at a digital IF $f_{IF} = 1$ GHz, $N = 2048$, $N_{CP} = 2048$, $B = 500$ MHz as in Table 4.6, $M = 128$, QPSK modulation, $\eta = 8$ and Chebyshev windowing with 100 dB sidelobe suppression for both range and Doppler shift processing. The same targets as for the radar images shown in Figures 4.60 and 4.61 were considered. Results are shown for rectangular windowing for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (a) 10^{-16} s, (b) 10^{-13} s, and (c) 10^{-10} s (i.e., $0.5 \times 10^{-7}T_s$, $0.5 \times 10^{-4}T_s$, and $0.5 \times 10^{-1}T_s$, respectively), as well as for and Chebyshev windowing with 100 dB sidelobe suppression for both range and Doppler shift processing and RMS SJ values at both DAC and ADC of (d) 10^{-16} s, (e) 10^{-13} s, and (f) 10^{-10} s. Reprinted from [23], licensed under CC BY 4.0.

4.2.4.2.3 Remarks on simulation results

The obtained results have shown that neither the mainlobe, measured by PPLR, nor the range and Doppler shift sidelobes, measured by PSLR and ISLR, of a target in a radar image experience significant degradation for the considered RMS SJ range between 10^{-17} s and 10^{-10} s (i.e., $0.5 \times 10^{-8}T_s$ and $0.5 \times 10^{-1}T_s$, respectively) at both DAC and ADC for any of the considered numbers of subcarriers, i.e., $N \in \{256, 2048, 16384\}$, and oversampling factors, i.e., $\eta \in \{1, 2, 4, 8\}$, both for BB and BP sampling.

Although no relevant main or sidelobe degradation is measured by the PPLR, PSLR and ISLR parameters, it was shown that the described ICI and CPE effects lead to non-negligible image SIR degradation for RMS SJ above 10^{-11} s (i.e., $0.5 \times 10^{-2} T_s$). Notably, it was verified that the CPE in the BP sampling case may lead to stripes in the Doppler shift direction at the range of the true targets in the radar image. This can lead to ghost targets and, in scenarios with multiple targets, mask weaker target reflections.

4.2.5 Remarks on analyses of hardware impairments

To analyze the communications performance degradation introduced by the different hardware impairments, the EVM was used in the last subsections. It is calculated as [VMR17]

$$\text{EVM} = \sqrt{\frac{\frac{1}{NM} \sum_{l=0}^{NM-1} |d_{\text{Rx},l} - d_{\text{Tx},l}|^2}{\overline{\mathbf{d}_{\text{Tx}}}}}, \quad (4.2)$$

where $\overline{\mathbf{d}_{\text{Tx}}} \in \mathbb{R}_{\geq 0}$ is the normalized energy of the transmit modulation alphabet. For a normalized constellation, it holds that $\overline{\mathbf{d}_{\text{Tx}}} = 1$. Disregarding actual AWGN and assuming that the considered hardware impairments cause AWGN-like interference and that the interference components associated with different impairments are independent from one another, the effective EVM is given by [LLCW06]

$$\text{EVM}_{\text{eff}}^{\text{HW}} = \sqrt{\sum_{\iota \in \mathcal{J}_{\text{HW}}} (\text{EVM}_{\iota}^{\text{HW}})^2}, \quad (4.3)$$

where $\mathcal{J}_{\text{HW}}$ is the set of considered hardware impairments. To obtain the resulting EVM in an AWGN channel, the EVM contribution of noise can be combined with the contributions $\text{EVM}_{\iota}^{\text{HW}}$ of the different hardware impairments in the same way as in (4.3).

As the difference between the receive and transmit modulation symbols in the numerator of (4.2) happens due to AWGN-like interference and considering high NM , the relationship between the EVM and the experienced SIR can be described based on [SRI06, MA09]. Considering this and the fact that the reference path is associated with most of the receive power as discussed in Section 3.1, the SIR of the aforementioned path against the interference intro-

duced by hardware impairments, which was also adopted as a communications performance parameter in the previous subsections, is given by

$$\text{EVM}_{\text{eff}}^{\text{HW}} \approx \sqrt{\frac{1}{\text{SIR}_0}}. \quad (4.4)$$

As the interference components are assumed to behave as AWGN, the full radar processing gain G_p is experienced again them. Consequently, the reference path will have image SIR equal to

$$\text{SIR}_{\text{image},0} = \text{SIR}_0 G_p. \quad (4.5)$$

4.3 Measurement-based verification

This section presents a measurement setup and results to demonstrate the feasibility of bistatic OFDM-based ISAC system concept, signal processing chain, and synchronization procedure under the influence of the synchronization offsets and hardware impairments. In this context, a RIS-assisted bistatic ISAC system concept and a measurement setup are presented in Section 4.3.1, and measurement results are presented and discussed in Section 4.3.2 to demonstrate the feasibility of the proposed synchronization and signal processing chains also under NLoS conditions.

4.3.1 RIS-assisted bistatic ISAC system concept and setup

To not only demonstrate the functionality of bistatic OFDM-based ISAC in practice but also show that it is possible even in NLoS conditions, the RIS-assisted bistatic ISAC system concept depicted in Figure 4.64 is proposed. In this concept, it is assumed that the LoS path between the gNBs is obstructed, and that a relay RIS is used to steer the incoming signals from gNB #1 in the directions of gNB #2 [LBE⁺21].

The corresponding measurement setup to the proposed RIS-assisted bistatic ISAC system concept shown in Figure 4.65 was assembled in the anechoic cham-

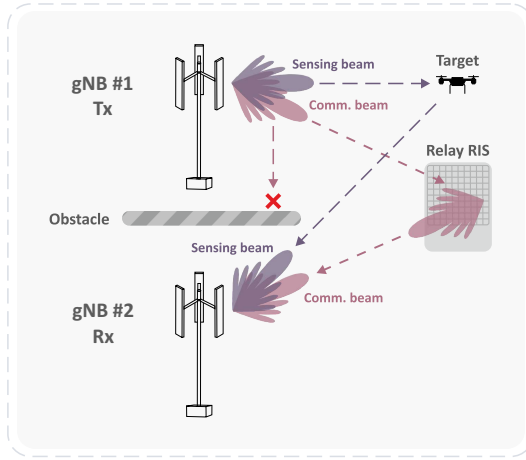
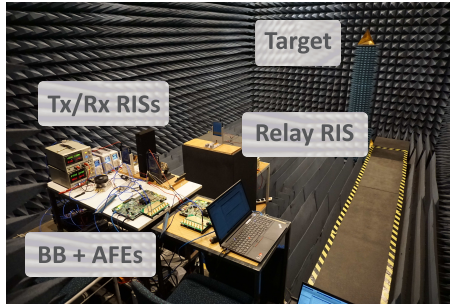


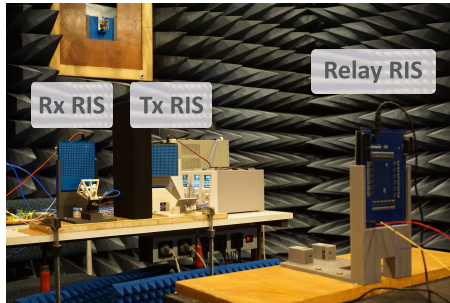
Figure 4.64. RIS-assisted bistatic ISAC system concept. © 2025 IEEE. Reprinted, with permission, from [22].

ber of the Institute of Radio Frequency Engineering and Electronics (IHE) at the Karlsruhe Institute of Technology (KIT). Its BB part consisted of two Zynq UltraScale+ RFSoc ZCU111 SoC platforms from Xilinx, Inc, one acting as a transmitter and another one as a receiver. For the DACs and ADCs in the boards, a sampling rate of $F_s = 4$ GHz was adopted. In addition, $N_{\text{bit}}^{\text{DAC}} = N_{\text{bit}}^{\text{ADC}} = 12$ bits were used, leading to an SQNR of 74 dB that poses no significant impairment for the considered scenario. Although exact numbers were not provided, the data converter specifications of the ZCU111 [AMD21] mentions the use of jitter cleaning circuits alongside the PLLs mentioned in Section 4.2.4, which allows assuming that the experienced RMS SJ is in the femtosecond level and therefore does not result in significant performance degradation for communications or radar sensing.

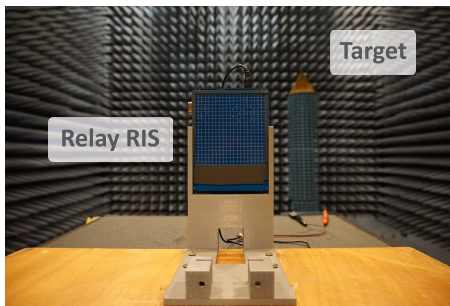
The transmitter and receiver boards were connected to the same RF AFE and ABE, respectively, used in [LBE⁺21, 42] via coaxial cables, which then transmit/receive OFDM signals with the parameters listed in Table 4.1 to/from the transmit and receive RISs that were used as the arrays of gNB #1 and #2, respectively. Due to the use of LP filters and the oversampling factor of 8, which is observed comparing the OFDM signal bandwidth B with the adopted



(a)



(b)



(c)

Figure 4.65. Assembled measurement setup for the RIS-based bistatic ISAC system in the IHE anechoic chamber at the KIT. © 2025 IEEE. Reprinted, with permission, from [22].

sampling frequency $F_s = 4$ GHz, no relevant aliasing was experienced. Before transmission, the digital signals at the transmitter side were normalized to avoid clipping at the DACs. This was done by setting the power of the strongest OFDM transmit signal sample to 1 dBm, which is the maximum DAC output power [Die23]. This resulted in a reduction of the average transmit power of the OFDM signal by 13.10 dB, which corresponds to the PAPR of the adopted OFDM signal. Consequently, an IBO of only 5.10 dB with respect to the AFE IP1dB of -7 dBm [Li24] would be ensured. To avoid non-linear distortion effects at amplifiers and mixers as discussed in Sections 3.4.2 and 3.4.3, and illustrated in Section 4.2.1, a 10 dB attenuator was used. The effective IBO for the AFE was therefore equal to 15.10 dB, which is higher than the 13.10 dB OFDM signal PAPR.

To generate the LO for the RF AFEs ensuring a stable reference with low PN, the E8257D PSG Microwave Analog Signal Generator from Keysight Technologies [Key25] was adopted. According to the analyses in Section 4.2.3, a combined transmit and receive PN level of -45 dBc was therefore experienced, leading to negligible communications and radar sensing performance degradation without the need for CPE estimation and correction.

Since the LoS link between transmit and receive RISs was obstructed by an RF absorber, a third RIS with known position was used as a relay. Beamforming was then performed for transmit and receive RISs so that they both had one beam pointing at the relay RISs and another beam pointing at a corner reflector placed at the other side of the anechoic chamber to serve as a radar target. While it is reasonable to assume that the beamforming directions towards the relay RIS are known, the exact beamforming directions from transmit and receive RISs towards the radar target were defined via exhaustive search. In practical deployments, more efficient approaches such as the ones from [MHD23, FMH⁺25, 13] can be adapted to the bistatic sensing case to both perform initial target detection as mentioned in Section 3.1. Due to the use of RISs, beam squint compensation could not be performed as previously discussed in Section 3.4.1.

A coordinate system was adopted for the measurement setup shown in Figure 4.65 assuming that the origin $[0, 0, 0]$ m was between the phase center of the transmitter and receiver RISs, which were located at $[-0.15, 0, 0]$ m and $[0.15, 0, 0]$ m, respectively. Furthermore, the phase center of the relay RIS was located at $[0, 1, 0]$ m and the center of the corner reflector at $[0.75, 4.50, 0.9]$ m. Considering that the reference antennas of both transmitter and receiver were

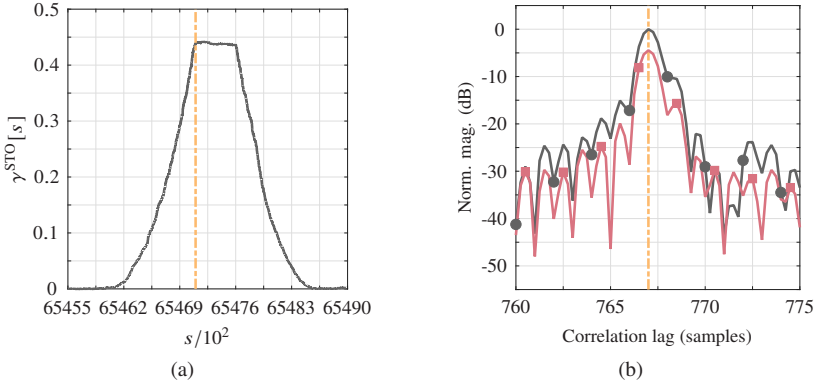


Figure 4.66. Measured synchronization results: (a) S&C sliding window correlation result $\gamma^{STO} [s]$ as a function of the sample index s of the receive sample stream (—) and (b) result of correlation of receive sample stream with preamble OFDM symbol before (— and ■) and after frequency shift correction (— and ●) with estimate from the S&C algorithm. In both cases, the ultimately obtained estimate of the start sample of the OFDM frame, which corresponds to the position of the correlation peak after frequency shift correction in (b), is highlighted (—).

located at a distance of 0.15 m of their respective RISs, the bistatic range ground truth values for the reference path enabled by the relay RIS and the radar target were 2.32 m and 9.63 m, respectively. Assuming perfect synchronization w.r.t. the reference path, this should result in a relative bistatic range of 7.31 m for the radar target.

4.3.2 Measurement results and discussion

Using the described measurement setup under the aforementioned conditions, 2 preamble OFDM symbols were transmitted before the OFDM frame to enable coarse time and frequency synchronization with the S&C algorithm [SC97]. Afterwards, frequency shift correction based on the CFO estimate $\hat{f}_\Delta = 138.46$ kHz, i.e., $0.57\Delta f$, obtained with the S&C algorithm was performed, and a fine time synchronization was performed via cross-correlation with the first preamble OFDM symbol. This procedure is illustrated in Figure 4.66. The visible sidelobes appear due to the fact that the initial time and

frequency synchronization based on the S&C algorithm and correlation are performed on the receive sample stream with sampling frequency $F_s = 2$ GHz, which corresponds to an oversampling factor of 4 when compared with the bandwidth $B = 500$ MHz. Note that the sampling frequency $F_s = 2$ GHz is half of the sampling frequency of the ADCs $F_s = 4$ GHz, and it is achieved via decimation taking place on the SoC platform before samples are provided for post-processing.

After this point, downsampling was performed, followed by SFO estimation. To demonstrate the superiority of the TITO method over alternative SFO estimation approaches from the literature considered in Section 4.1.3, results obtained with the latter are also presented in the following analysis. Using the TITO method with $\delta_{\max} = 123$ ppm, which is equal to the absolute value of maximum expected SFO in the adopted setup, the same SFO estimate of -104.29 ppm was obtained as with the algorithm based on Wu et al. [WZL12]. This was the case since only tolerable SINR degradation due to SFO-induced ICI and ISI was experienced, and therefore (3.49) resulted in $M_{\text{pil}}^{\text{TITO}} = M_{\text{pil}}$ pilot OFDM symbols used by the TITO method to estimate the SFO with (3.35), meaning that both methods are equivalent for the adopted set of parameters and hardware. Since, however, the TITO method achieves equal or better performance compared to the method based on Wu et al. [WZL12], the latter will be excluded from the subsequent analyses. The method by Tsai et al. [TKC05], however, yielded a biased SFO estimate of 11.60 ppm, which ultimately resulted in a BER of 0.41 even with low-density parity-check (LDPC) decoding and produced a bistatic radar image where no targets could be detected. Out of this reason, the method by Tsai was alternatively implemented with 20 dedicated preamble OFDM symbols for SFO estimation transmitted right before the OFDM frame. Although this incurs in intolerable overhead in communications operations, only so an SFO estimate of -103.73 ppm could be obtained. Since the SFO estimation with the Tsai algorithm based on preamble symbols is not accurate enough for bistatic sensing, the latter method was also combined with a residual SFO estimation and subsequent correction via discrete-time domain shifting of the OFDM symbols as described in [BJT⁺21, 50]. While the aforementioned residual SFO correction compensates the delay migration, it does not alleviate the effect of the SFO-induced ICI.

After SFO estimation and correction via resampling, besides blind I/Q imbalance estimation and correction as discussed in Section 3.4.3 and analyzed in

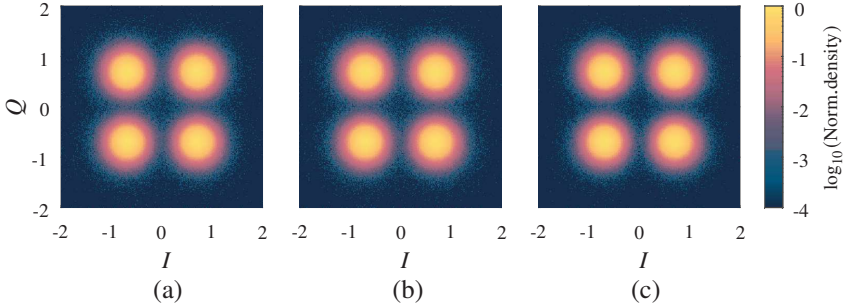


Figure 4.67: Measured normalized QPSK constellations assuming SFO estimation and correction based on (a) the method by Tsai et al. without residual SFO correction, (b) the method by Tsai et al. with residual SFO correction, and (c) the TITO method. © 2025 IEEE. Reprinted, with permission, from [22].

Section 4.2.2, further communications processing as described in Section 3.2.2 and [6, 50] was performed. As a result, the QPSK constellations shown in Figure 4.67 were obtained. The fact that the constellations are not particularly compact happens because the used RF AFEs were originally designed for narrow-band data communications over bandwidths of only a few MHz [LBE⁺21, 42], which resulted in much more pronounced frequency selectivity and attenuation for the adopted 500 MHz as observed in the experienced CFR shown in Figure 4.68. The obtained results show virtually the same performance regardless of the SFO estimation method. This is evidenced by the EVM with mean value of -14.84 dB and standard deviation of 5.76 dB obtained in the case where SFO estimation was performed with the TITO method. Simultaneously, a mean EVM of -14.69 dB and a standard deviation of 5.79 dB were obtained for the case where SFO estimation was done with the method by Tsai et al. [TKC05] without residual SFO correction, and a mean EVM of -14.62 dB and a standard deviation of 5.79 dB were for SFO estimation done with the method by Tsai et al. [TKC05] with residual SFO correction. In all cases, no transmission errors were experienced due to the chosen LDPC channel code [MKJ⁺24] of rate $2/3$ and the parity check matrix for the case with 64800 bits in an LDPC code block from [ETSI14].

While the EVM performance remains virtually unaffected by the choice between the three considered SFO estimation and correction methods, the bistatic sensing performance is significantly influenced by residual SFO that lead to range

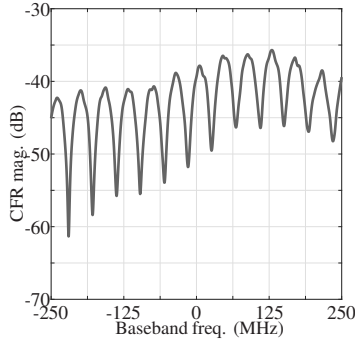


Figure 4.68: Measured magnitude response of the experienced communications CFR including the effects of the RF AFEs. © 2025 IEEE. Reprinted, with permission, from [22].

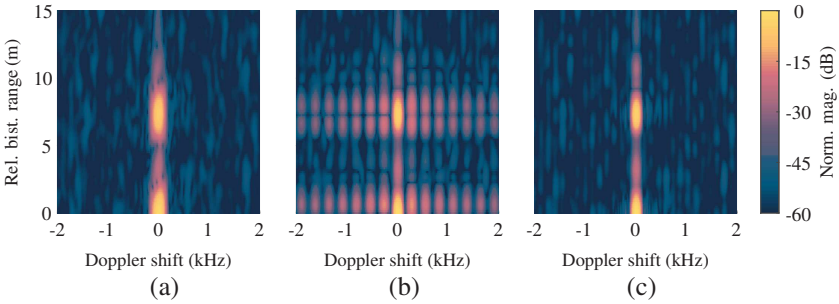


Figure 4.69: Measured bistatic radar images assuming SFO estimation and correction based on (a) the method by Tsai et al. without residual SFO correction, (b) the method by Tsai et al. with residual SFO correction, and (c) the TITO method. © 2025 IEEE. Reprinted, with permission, from [22].

and Doppler shift migrations that can be derived from delay and frequency shift migrations in (3.29) and (3.31), respectively. Figure 4.69 shows the measured bistatic radar images ultimately obtained when using the aforementioned SFO estimation and correction approaches. During both range and Doppler shift processing, Chebyshev windowing with 100 dB sidelobe suppression was performed. As in Section 3.2.2 and [6, 50], the required full transmit frame for the bistatic radar processing was reconstructed based on the knowledge of the pilot subcarriers as well as the receive payload QPSK symbols from the constellations in Figure 4.67. Furthermore, biases to the relative bistatic ranges

and Doppler shifts in the radar image are removed by shifting the radar image so that the reference path, which is the stronger peak in the radar image, appears at a relative bistatic range of 0 m and Doppler shift of 0 Hz. It is observed that the bistatic radar image obtained after SFO estimation with the TITO method presents a well-defined radar target peak at a relative bistatic range of 7.19 m. In the radar image obtained after synchronization with the method by Tsai et al. without residual SFO correction, the radar target peak appears at the same range as in the previous case and a negligible SNR loss of 0.38 dB is observed. This only happens due to the compensation of range migration via discrete-time domain shifting of the OFDM symbols in the frame. However, this comes at the cost of extra computational complexity and a significant spread in the Doppler shift direction, with multiple high sidelobes that could not be successfully suppressed by windowing. As for the case where residual SFO correction was performed after the estimation with the Tsai algorithm and resampling, a significant degradation of the range resolution and a slight degradation of the Doppler shift resolution are observed compared with the bistatic radar image obtained when the TITO method is used. In this case, the radar target peak is at a relative bistatic range of 7.34 m and a SNR loss of 2.65 dB is experienced w.r.t. the TITO case.

4.4 Chapter conclusion

This chapter presented a performance evaluation of the bistatic OFDM-based ISAC system concept previously presented in Chapter 3. First, a numerical analysis of performance degradation caused by ISI, ICI, and SFO was presented. The first two impairments can be caused either by inaccurate time and frequency synchronization, as well as due to the delays and Doppler shifts associated with the ranges and velocities of targets. Since these impairments have been widely investigated in the literature, brief analyses with a focus on sensing performance degradation were presented. As for SFO, a detailed analysis including achievable accuracy and computational complexity associated with its estimation was presented to demonstrate the superiority of the proposed TITO method over conventional SFO estimation approaches originally proposed for OFDM-based communications systems. Next, the induced constellation distortion and sensing performance degradation by relevant hardware impairments, namely amplifier non-linear distortion, mixer I/Q imbalance, PN, and SJ, were

numerically analyzed. In these analyses, both tolerable impairment level and the effectiveness of possible countermeasures were assessed. Finally, a measurement setup based on a proposed RIS-assisted bistatic ISAC system concept was presented, and both the system design considerations and measurement results discussed to verify the proposed system concept, signal processing chain, and synchronization procedure in practice.

Focusing on the full-frame bistatic radar signal processing approach, in which estimation of the transmit OFDM frame via communications processing is required, the obtained simulation and measurement results validated the feasibility of the proposed bistatic OFDM-based ISAC system. It was shown that any static path can be used regardless of being an LoS or NLoS one. The only conditions this path must fulfill are having constant range, Doppler shift, and angle, besides having sufficient SNR to contribute to synchronization and communications, the latter in the full-frame processing case. It was also observed that enough synchronization accuracy can be achieved either with a two-stage procedure as in the case of STO and CFO, where coarse and fine estimates are respectively obtained via preamble and pilots, or with dedicated algorithms, as is the case with the proposed TITO method for robust and accurate SFO estimation. In addition, it was shown that hardware impairments can be either avoided in the system design phase, or remedied afterwards either via signal-processing-based compensation or solely relying on the radar processing gain to suppress the impairment-induced interference. Based on the previously carried out discussion in Sections 3.3 and 3.4, and on the presented results in this chapter, the considered synchronization offsets, their resulting impairments, as well as possible countermeasures are summarized in Table 4.9. Similarly, Table 4.10 presents a summary of the remaining investigated hardware impairments and identified countermeasures. Together with detailed link budget calculations, signal-level simulations, and the considerations on how to jointly consider the effects of different impairments in Section 4.2.5, these summaries can serve as guidelines to efficiently design bistatic OFDM-based ISAC systems. Since many of the considered impairments are also present in monostatic architectures, the observations can also be reused or adapted when designing such systems.

Table 4.9: Summary of synchronization offsets, resulting impairments, and possible countermeasures. Adapted from [52], © 2026 IEEE.

Synchronization offset/impairment	Countermeasure
STO/ISI	Correction for reference path via synchronization possible. Ensure enough CP length to avoid relevant peak power loss and interference level increase for radar targets.
CFO/ICI	Correction for reference path via synchronization possible. Ensure enough subcarrier spacing length to avoid relevant peak power loss and interference level increase for radar targets.
SFO	Correction via signal processing feasible but highly accurate estimate of around 0.01 ppm required for coherent processing of long OFDM frames, which is possible with TITO for realistic SFO levels. Using conventional SFO estimation methods may result in severe radar image distortion.

Table 4.10: Summary of hardware impairments and countermeasures. Adapted from [52], © 2026 IEEE.

Hardware impairment	Countermeasure
Antenna element coupling	Minimize in the array design phase.
Beam squint	Compensation via signal processing possible if digital beamforming or hybrid beamforming with high RF chain count is performed.
Amplifier/mixer non-linear distortion	Ensure comparable IBO to OFDM signal PAPR.
Mixer I/Q imbalance	Compensation via signal processing possible for realistic amplitude and phase imbalances.
PN	Ensure low combined transmit and receive PN level. Ideally ≤ -45 dBc or, realistically, ≤ -30 dBc (3GPP PN model).
Quantization noise	Ensure enough SQNR by adopting appropriate number of DAC bits $N_{\text{bit}}^{\text{DAC}}$ and ADC bits $N_{\text{bit}}^{\text{ADC}}$ of around. For instance, 12 bits results in SQNR = 74 dB even without oversampling.
DAC/ADC clipping	Ideally avoid, otherwise similar effects to amplifier hard clipping in saturation may be expected. Further investigation required.
Image spectra and aliasing	Ensure sufficient oversampling and analog LP filtering.
Sampling jitter	Ensure smaller RMS SJ than 10^{-11} s. Feasible with modern converters with jitter cleaning.

5 Bistatic MIMO-OFDM-based ISAC

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In Chapters 3 and 4, the effects and countermeasures to synchronization offsets and hardware impairments were described and investigated for bistatic SISO-OFDM-based ISAC. Focusing on synchronization offsets, it was concluded that their estimation in ISAC systems require much higher accuracy than in regular communications systems to ensure unbiased radar sensing. When scaling up to a MIMO architecture, additional challenges are imposed on the ISAC system. Focusing on practical deployments, these include, e.g., considerations of the hardware architecture and its impact on synchronization as well as the following communications and radar sensing processing chains. While the offset estimation and correction techniques from Chapter 3 can be applied when scaling up to a MIMO architecture, prior considerations of how the aforementioned offsets manifest across different receive channels must be made, and corresponding adaptations applied. More specifically, different challenges arise depending on how LOs and sampling clocks are distributed across the multiple transmit and receive channels, as well as on how the synchronization offset estimations are combined across each receive channel to ensure accurate sufficient coherence among the channels and therefore avoid communications and sensing performance degradation. Furthermore, aspects such as calibration, particularly when relying on OTA synchronization, remain open challenges that must be addressed to enable effective sensing in MIMO architectures where DoA estimation requires phase coherence among the receive channels.

In this context, this chapter introduces a bistatic MIMO-OFDM-based ISAC system concept. It features a multichannel transmitter with beamforming capability that generates both a reference path, besides additional beams to enable detecting radar targets. As in the concept for the SISO case from Chapter 3 depicted in Figure 3.1, the aforementioned reference path is used for synchroniza-

tion, communications and sensing reference, and it is henceforth assumed to be a LoS path without loss of generality. In addition, a multichannel receiver with digital beamforming is considered. For that receiver, a distributed synchronization concept is proposed. Finally, further processing steps for communications, i.e., channel estimation, equalization, and diversity combining, and for sensing, i.e., calibration to ensure coherence among the receive channels and generation of radar images with range, Doppler shift, and DoA information, are described.

The remainder of this chapter is organized as follows. Section 5.1 and Section 5.2 present the system model and the processing chain, respectively, for the proposed bistatic MIMO-OFDM-based ISAC system concept. Next, Section 5.3 and Section 5.4 present simulation and measurement results, respectively, besides discussing the achievable communications and radar sensing performances in the considered bistatic MIMO-OFDM-based ISAC system. Finally, concluding remarks are presented in Section 5.5.

5.1 System model

In the considered bistatic MIMO-OFDM-based ISAC system, it is assumed that both transmitting and receiving stations are static and that both their positions are known to each other. This is, e.g., the case for gNBs as previously considered in the SISO case from Chapter 3 and depicted in Figure 3.1. In addition, it is w.l.o.g. assumed that a LoS reference path is present between the stations and that both are equipped with uniform linear arrays (ULAs) stretching along the x -axis. Although these ULAs can only be used to steer beams or estimate DoA in azimuth, it is assumed that the LoS and any other relevant paths or radar targets are covered by the elevation beamwidth of both transmit and receive ULAs. Detailed descriptions of the antenna array geometry and transmit signal generation and propagation are provided in Sections 5.1.1 and 5.1.2, respectively.

5.1.1 Antenna array geometry and coordinate system

The arrangement and coordinate system of the adopted ULA for the transmitting and receiving ISAC stations in this article are as depicted in Fig-

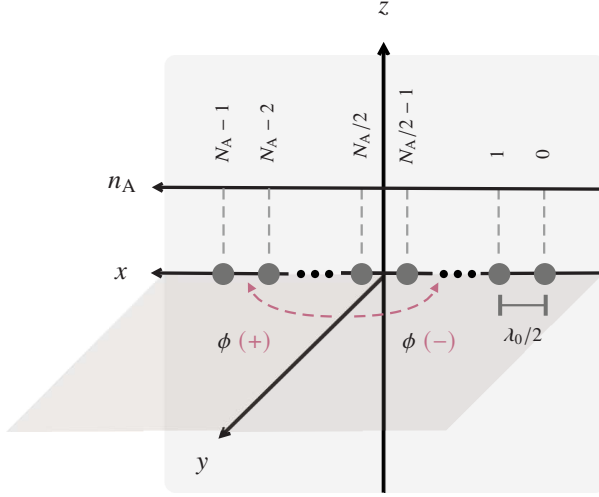
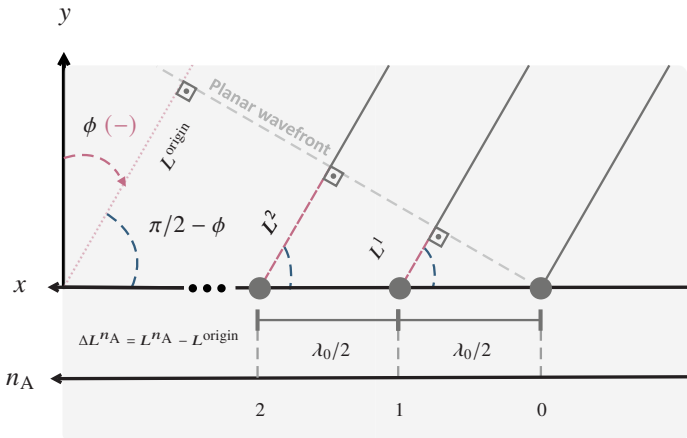


Figure 5.1. Adopted ULA arrangement and coordinate system. © 2026 IEEE. Reprinted, with permission, from [24].

Figure 5.1. This figure is composed by $N_A \in \mathbb{N}_{>0}$ antenna elements labeled as $n_A \in \{0, 1, \dots, N_A - 1\}$. All antenna elements are placed along the x axis with spacing $\lambda_0/2$ among consecutive elements. In addition, it is assumed that they face the positive y direction. The x , y , and z coordinates of the n_A th antenna element are denoted by the position vector $\mathbf{p}^{n_A} \in \mathbb{R}^{3 \times 1} | \mathbf{p}^{n_A} = [\rho_x^{n_A}, \rho_y^{n_A}, \rho_z^{n_A}]^T$ such that $\rho_x^{n_A} = (\lambda_0/2)[(-N_A + 1)/2 + n_A]$ and $\rho_y^{n_A} = \rho_z^{n_A} = 0 \forall n_A$. An azimuth angle $\phi \in [-\pi/2, \pi/2)$, which can either be a direction-of-departure (DoD) in the case of a transmit ULA or DoA in the case of a receive ULA, is also defined in Figure 5.1. Positive ϕ angles, i.e., $\phi \in [0, \pi/2)$, are measured if they correspond to directions between positive x and y axes, while negative ϕ angles, i.e., $\phi \in [-\pi/2, 0)$, are measured for directions between negative x and positive y axes. If signals are either beamformed towards or received from an angle ϕ , a different electromagnetic wave propagation length will be traveled by the signals associated with each of the N_A antennas. Based on Figure 5.2,



wavefront. © 2026 IEEE. Reprinted, with permission, from [24].

the length difference ΔL^{n_A} for the n_A th antenna with respect to the origin of the array at $(x, y, z) = (0, 0, 0)$ is given by

$$\begin{aligned}\Delta L^{n_A} &= -(\lambda_0/2) [(-N_A + 1)/2 + n_A] \cos(\pi/2 - \phi) \\ &= -(\lambda_0/2) [(-N_A + 1)/2 + n_A] \sin(\phi).\end{aligned}\quad (5.1)$$

5.1.2 Transmit signal generation and propagation

At the transmitter side, a baseband transmit signal $x(t) \in \mathbb{C}$ that occupies a bandwidth $B \leq F_s$, where F_s is the sampling frequency associated with sampling period $T_s = 1/F_s$, is generated. Next, $x(t)$ is up-converted to the carrier frequency $f_c \gg B$ with an I/Q mixer, and beamforming at an array with $N_{\text{Tx}} \in \mathbb{N}_{>0}$ transmit antennas is performed on the resulting signal towards $P \in \mathbb{N}_{>0}$ steering directions. Neglecting the beam sidelobes for simplicity, this ultimately results in P propagation paths labeled as $p \in \{0, 1, \dots, P-1\}$ towards each out of the $N_{\text{Rx}} \in \mathbb{N}_{\geq 0}$ receive antennas also arranged as depicted in Figure 5.1.

The first, $p = 0$, is the LoS direction towards the receiver. The other $P - 1$ beams, i.e., $p = 1$ to $p = P - 1$, are steered in strategic directions to perform sensing and possibly serve other communications receivers. Every p th path is associated with an attenuation factor $\alpha_p^{n_{\text{Tx}}, n_{\text{Rx}}}$, a delay $\tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}$, an a Doppler shift $f_{\text{D},p}^{n_{\text{Tx}}, n_{\text{Rx}}}$, with $n_{\text{Tx}} \in \{0, 1, \dots, N_{\text{Tx}} - 1$ and $n_{\text{Rx}} \in \{0, 1, \dots, N_{\text{Rx}} - 1\}$ denoting the transmit and receive antenna indexes, respectively. Furthermore, it is assumed that the p th path is associated with an ideal azimuth DoD $\phi_{\text{Tx}-\text{T},p}$, which corresponds to the true transmitter-referred direction of the receiver for $p = 0$ and the true direction of radar targets for $p \in \{1, \dots, P - 1\}$.

The described transmit beamforming is achieved by transmitting weighted copies of the same up-converted version of $x(t)$ through each of the N_{Tx} transmit channels, which are assumed to be associated to their own antenna. The weight at each n_{Tx} th channel corresponds to the sum of the n_{Tx} th element of the transmit beamsteering vectors $\mathbf{b}_{\text{Tx},p} \in \mathbb{C}^{N_{\text{Tx}} \times 1}$ associated with steering directions of index p over all P steering directions. The n_{Tx} th element of $\mathbf{b}_{\text{Tx},p}$ is given by

$$b_{\text{Tx},p}^{n_{\text{Tx}}} = e^{-j2\pi(\rho_{\text{Tx},x}^{n_{\text{Tx}}}/\lambda_0)\sin(\widehat{\phi}_{\text{Tx}-\text{T},p})}. \quad (5.2)$$

In this equation, $\rho_{\text{Tx},x}^{n_{\text{Tx}}}$ is the x coordinate of the n_{Tx} th transmit antenna, whose position is represented by the vector $\boldsymbol{\rho}_{\text{Tx},n_{\text{Tx}}} \in \mathbb{R}^{3 \times 1}$ that is denoted as

$$\boldsymbol{\rho}_{\text{Tx},n_{\text{Tx}}} = \left[\rho_{\text{Tx},x}^{n_{\text{Tx}}}, \rho_{\text{Tx},y}^{n_{\text{Tx}}}, \rho_{\text{Tx},z}^{n_{\text{Tx}}} \right]^T. \quad (5.3)$$

As the model from Figure 5.1 is considered, it holds that

$$\rho_{\text{Tx},x}^{n_{\text{Tx}}} = (\lambda_0/2)[(-N_{\text{Tx}} + 1)/2 + n_{\text{Tx}}] \quad (5.4)$$

as well as $\rho_{\text{Tx},y}^{n_{\text{Tx}}} = \rho_{\text{Tx},z}^{n_{\text{Tx}}} = 0 \forall n_{\text{Tx}}$. In addition, $\widehat{\phi}_{\text{Tx}-\text{T},p}$ is the estimate of $\phi_{\text{Tx}-\text{T},p}$ that is used as the azimuth DoD of the p th steering direction.

After propagation through the aforementioned P paths, attenuated, delayed and Doppler-shifted versions of the up-converted and beamformed version of $x(t)$ are captured by the n_{Rx} th receive antenna. This constitutes the BB signal at the

n_{Rx} th receive channel, which after down-conversion into the baseband with an I/Q mixer is denoted as $y^{n_{\text{Rx}}}(t) \in \mathbb{C}$ and expressed as

$$\begin{aligned}
 y^{n_{\text{Rx}}}(t) = & \left[\sum_{n_{\text{Tx}}=0}^{N_{\text{Tx}}-1} \left(\alpha_0^{n_{\text{Tx}}, n_{\text{Rx}}} b_{\text{Tx},0}^{n_{\text{Tx}}} x(t - \tau_0^{n_{\text{Tx}}, n_{\text{Rx}}} - \tau_{\Delta}) e^{-j2\pi f_c \tau_0^{n_{\text{Tx}}, n_{\text{Rx}}}} e^{j2\pi f_{\Delta} t + \psi_{\Delta}} \right. \right. \\
 & + \left. \sum_{p=1}^{P-1} \alpha_p^{n_{\text{Tx}}, n_{\text{Rx}}} b_{\text{Tx},p}^{n_{\text{Tx}}} x(t - \tau_p^{n_{\text{Tx}}, n_{\text{Rx}}} - \tau_{\Delta}) e^{-j2\pi f_c \tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}} e^{j2\pi f_{\text{D},p}^{n_{\text{Tx}}, n_{\text{Rx}}} t} e^{j2\pi f_{\Delta} t + \psi_{\Delta}} \right) \Bigg] \\
 & * h_{\text{ABE}}^{n_{\text{Rx}}}(t).
 \end{aligned} \tag{5.5}$$

In this equation, $\alpha_p^{n_{\text{Tx}}, n_{\text{Rx}}}$, $\tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}$, and $f_{\text{D},p}^{n_{\text{Tx}}, n_{\text{Rx}}}$ are the attenuation factor, delay, and Doppler shift associated with the p th path between the n_{Tx} th transmit and the n_{Rx} th receive antennas. Note that $f_{\text{D},0}^{n_{\text{Tx}}, n_{\text{Rx}}} = 0$ Hz since transmitting and receiving stations are assumed to be static and there is therefore no Doppler shift for the LoS path between them labeled as $p = 0$. The propagation delays $\tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}$ also result in phase shifts, which are denoted by $e^{-j2\pi f_c \tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}}$. Time and frequency offsets between transmitter and receiver are also considered in (5.5). These are namely the STO τ_{Δ} , as well as the CFO f_{Δ} and its resulting phase rotation ψ_{Δ} . Finally, $h_{\text{ABE}}^{n_{\text{Rx}}}(t) \in \mathbb{C}$ is the ABE CIR of the n_{Rx} th receive channel associated with a corresponding CFR $H_{\text{ABE}}^{n_{\text{Rx}}}(f) \in \mathbb{C}$. They account for hardware non-idealities that may lead to slightly different delays and therefore phases among the N_{Rx} receive channels. It is henceforth assumed that the aforementioned CIR has a dominant path with delay $\tau_{\text{ABE}}^{n_{\text{Rx}}}$.

5.2 Transmit and receive signal processing

Based on the system model outlined in Section 5.1, the transmit and receive signal processing steps in the proposed bistatic MIMO-OFDM-based ISAC system are discussed in this section. More specifically, Section 5.2.1 discusses the choice of the transmit beamforming weights, while Section 5.2.2 discusses the receive signal processing chain for both communications and bistatic radar sensing.

5.2.1 Choice of transmit beamforming weights

Based on (5.1) and the discussion on it in Section 5.1.1, it is known that a electromagnetic wave propagation length difference $\Delta L_p^{n_{\text{Tx}}}$ for the n_{Tx} th antenna with respect to the origin of the transmit array at $(x_{\text{Tx}}, y_{\text{Tx}}, z_{\text{Tx}}) = (0, 0, 0)$ given by

$$\Delta L_p^{n_{\text{Tx}}} = -(\lambda_0/2)[(-N_{\text{Tx}} + 1)/2 + n_{\text{Tx}}] \sin(\phi_{\text{Tx}-\text{T},p}) \quad (5.6)$$

will be experienced. Converting this length difference into delay yields $\tau_{\Delta,p}^{n_{\text{Tx}}} = \Delta L_p^{n_{\text{Tx}}}/c_0$, which can be alternatively expressed as

$$\tau_{\Delta,p}^{n_{\text{Tx}}} = -(2f_c)^{-1} [(-N_{\text{Tx}} + 1)/2 + n_{\text{Tx}}] \sin(\phi_{\text{Tx}-\text{T},p}). \quad (5.7)$$

Next, the propagation delays $\tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}$ are denoted as

$$\tau_p^{n_{\text{Tx}}, n_{\text{Rx}}} = \tau_{\text{Tx}-\text{T},p}^{n_{\text{Tx}}} + \tau_{\text{T}-\text{Rx},p}^{n_{\text{Rx}}}, \quad (5.8)$$

where, for $p \in \{1, \dots, P-1\}$, $\tau_{\text{Tx}-\text{T},p}^{n_{\text{Tx}}}$ is the propagation delay between the n_{Tx} th transmit antenna and the radar target and $\tau_{\text{T}-\text{Rx},p}^{n_{\text{Rx}}}$ is the propagation delay between the radar target and the n_{Rx} th receive antenna, both for the p th path. For the LoS path, i.e., $p = 0$, it is assumed that $\tau_{\text{Tx}-\text{T},0}^{n_{\text{Tx}}} = \tau_0^{n_{\text{Tx}}, n_{\text{Rx}}}$ and $\tau_{\text{T}-\text{Rx},0}^{n_{\text{Rx}}} = 0$ s. Using the result from (5.7), the propagation delays $\tau_{\text{Tx}-\text{T},p}^{n_{\text{Tx}}}$ are further expanded as

$$\tau_{\text{Tx}-\text{T},p}^{n_{\text{Tx}}} = \tau_{\text{Tx}-\text{T},p} + \tau_{\Delta,p}^{n_{\text{Tx}}}, \quad (5.9)$$

where

$$\tau_{\text{Tx}-\text{T},p} = R_{\text{Tx}-\text{T},p}/c_0 \quad (5.10)$$

and $R_{\text{Tx}-\text{T},p}$ denotes the range of the p th target with respect to the origin of the transmit array at $(x_{\text{Tx}}, y_{\text{Tx}}, z_{\text{Tx}}) = (0, 0, 0)$. The results from equations (5.6) to (5.9) allow expanding the expression of the delay-induced phase shifts $e^{-j2\pi f_c \tau_p^{n_{\text{Tx}}, n_{\text{Rx}}}}$ as

$$\left(e^{-j2\pi f_c \tau_{\text{Tx}-\text{T},p}} e^{-j2\pi f_c \tau_{\text{T}-\text{Rx},p}^{n_{\text{Rx}}}} \right) e^{-j2\pi f_c \tau_{\Delta,p}^{n_{\text{Tx}}}}, \quad (5.11)$$

or further as

$$\left(e^{-j2\pi f_c \tau_{\text{Tx}-\text{T},p}} e^{-j2\pi f_c \tau_{\text{T}-\text{Rx},p}^{n_{\text{Rx}}}} \right) e^{j\pi [(-N_{\text{Tx}}+1)/2 + n_{\text{Tx}}] \sin(\phi_{\text{Tx}-\text{T},p})}. \quad (5.12)$$

To ensure maximum transmit beamforming gain, the weights $b_{\text{T}_x, p}^{n_{\text{T}_x}}$ must be chosen such that $\widehat{\phi}_{\text{T}_x-\text{T}, p} = \phi_{\text{T}_x-\text{T}, p}$. This can be achieved by setting the arguments of the exponential function defining the beamforming weights in (5.2), equal to the negative of the arguments of the rightmost exponential function in (5.12), i.e.,

$$j2\pi \left(\rho_{\text{T}_x, x}^{n_{\text{T}_x}} / \lambda_0 \right) \sin \left(\widehat{\phi}_{\text{T}_x-\text{T}, p} \right) = j\pi [(-N_{\text{T}_x} + 1)/2 + n_{\text{T}_x}] \sin \left(\phi_{\text{T}_x-\text{T}, p} \right). \quad (5.13)$$

Rearranging the expression in (5.13) and knowing from (5.4) that $n_{\text{T}_x} = \rho_{\text{T}_x, x}^{n_{\text{T}_x}} / (\lambda_0/2) - (-N_{\text{T}_x} + 1)/2$, $\widehat{\phi}_{\text{T}_x-\text{T}, p} = \phi_{\text{T}_x-\text{T}, p}$ is obtained, which implies that the maximum transmit beamforming gains are obtained if the steering angles $\widehat{\phi}_{\text{T}_x-\text{T}, p}$ are set equal to the angles $\phi_{\text{T}_x-\text{T}, p}$ corresponding to the LoS path ($p = 0$) or the radar targets ($p \in 1, \dots, P - 1$). This ultimately results in

$$\sum_{n_{\text{T}_x}=0}^{N_{\text{T}_x}-1} b_{\text{T}_x, p}^{n_{\text{T}_x}} e^{-j2\pi f_c \tau_p^{n_{\text{T}_x}, n_{\text{R}_x}}} = N_{\text{T}_x} \left(e^{-j2\pi f_c \tau_{\text{T}_x-\text{T}, p}} e^{-j2\pi f_c \tau_{\text{T}-\text{R}_x, p}^{n_{\text{R}_x}}} \right). \quad (5.14)$$

It is then considered that the delays for every p th path are approximately the same for every combination of transmit and receive channels, i.e.,

$$\tau_p^{n_{\text{T}_x}, n_{\text{R}_x}} \approx \tau_p \quad \forall n_{\text{T}_x}, n_{\text{R}_x}, p. \quad (5.15)$$

This assumption is reasonable since the phase rotations induced by the slight difference in the delays are still considered. Based on (5.14) and (5.15), (5.5) can be rewritten as

$$\begin{aligned} y^{n_{\text{R}_x}}(t) = & \left(\alpha_0^{n_{\text{T}_x}, n_{\text{R}_x}} N_{\text{T}_x} x(t - \tau_0 - \tau_\Delta) e^{-j2\pi f_c \tau_{\text{T}_x-\text{T}, 0}} e^{-j2\pi f_c \tau_{\text{T}-\text{R}_x, 0}^{n_{\text{R}_x}}} e^{j2\pi f_\Delta t + \psi_\Delta} \right. \\ & + \sum_{p=1}^{P-1} \alpha_p^{n_{\text{T}_x}, n_{\text{R}_x}} N_{\text{T}_x} x(t - \tau_p - \tau_\Delta) e^{-j2\pi f_c \tau_{\text{T}_x-\text{T}, p}} e^{-j2\pi f_c \tau_{\text{T}-\text{R}_x, p}^{n_{\text{R}_x}}} e^{j2\pi f_{\text{D}, p}^{n_{\text{T}_x}, n_{\text{R}_x}} t} e^{j2\pi f_\Delta t + \psi_\Delta} \Big) \\ & * h_{\text{ABE}}^{n_{\text{R}_x}}(t). \end{aligned} \quad (5.16)$$

It is worth highlighting that, in practical deployments, the N_{T_x} transmit channels of the AFEs will likely have different CFRs $H_{\text{AFE}}^{n_{\text{T}_x}}(f) \in \mathbb{C}$. These must be

estimated beforehand with calibration measurements and compensated via DPD to ensure that transmit beamforming is correctly performed.

5.2.2 Receive signal processing chain

After being sampled at each n_{Rx} th receive channel, the discrete-time domain equivalent sequences to the baseband signals $y^{n_{\text{Rx}}}(t)$ must undergo synchronization followed by communications processing to allow estimating the transmit OFDM frame which will then be used for bistatic radar signal processing as described in [6, 50]. The three aforementioned processing steps are described in further detail as follows.

5.2.2.1 Synchronization

It is assumed that the STO and CFO are individually estimated for each receive channel based on what is described for the SISO case in Sections 3.3.1 and 3.3.2 as well as in [6, 50]. More specifically, a first, coarse estimation of the CFO is obtained at each receive channel with the S&C algorithm. After correcting the CFO, an STO estimate for each receive channel is obtained via cross-correlation with a preamble OFDM symbol to yield a more accurate frame start point estimate than what can be achieved with the S&C algorithm. Next, the SFO is also estimated for each receive channel with the TITO algorithm proposed in Section 3.3.3 and [22]. After resampling based on the aforementioned SFO estimate, a fine-tuning of the STO and CFO estimates is performed based on pilot subcarriers.

Since a MIMO architecture is assumed, the CFO and SFO estimates are averaged to yield improved accuracy before these offsets can be corrected in the proposed bistatic MIMO-OFDM-based ISAC system. This can be done since there are two groups of LOs and sampling clocks, one shared by all transmit channels and another one shared by all receive channels. Consequently, the same CFO and SFO must be experienced by all N_{Rx} receive channels. The CFO estimate $\hat{f}_{\Delta}^{\text{MIMO}}$ and the estimate $\hat{\delta}^{\text{MIMO}}$ of the normalized SFO by the sampling frequency T_s

[22] in the considered bistatic MIMO-OFDM-based ISAC system are therefore obtained as

$$\hat{f}_{\Delta}^{\text{MIMO}} = \frac{1}{N_{\text{Rx}}} \sum_{n_{\text{Rx}}=0}^{N_{\text{Rx}}-1} \hat{f}_{\Delta}^{n_{\text{Rx}}} \quad (5.17)$$

and

$$\hat{\delta}^{\text{MIMO}} = \frac{1}{N_{\text{Rx}}} \sum_{n_{\text{Rx}}=0}^{N_{\text{Rx}}-1} \hat{\delta}^{n_{\text{Rx}}}, \quad (5.18)$$

where $\hat{f}_{\Delta}^{n_{\text{Rx}}}$ and $\hat{\delta}^{n_{\text{Rx}}}$ are the CFO and normalized SFO estimates for the n_{Rx} th receive channel, respectively.

As for the STO, the same cannot be done since different delays are experienced as each n_{Rx} th receive channel is associated with its own ABE CIR $h_{\text{ABE}}^{n_{\text{Rx}}}(t)$. Although $h_{\text{ABE}}^{n_{\text{Rx}}}(t)$ tends to have a dominant tap as the connections between the ABE elements are matched, the experienced delays may differ among the N_{Rx} even if only by a fraction of the sampling period T_s . This is, however, enough to break the phase coherence among receive channels and impair the DoA estimation processing performed as later explained in Section 5.2.2.3. In this sense, the earliest estimated OFDM frame start point among all N_{Rx} receive channels is taken as a common start point for all channels to ensure that none of them is impaired by ISI. The STO estimate $\hat{\tau}_{\Delta}^{\text{MIMO}}$ in the considered bistatic MIMO-OFDM-based ISAC system is therefore calculated as

$$\hat{\tau}_{\Delta}^{\text{MIMO}} = \min_{n_{\text{Rx}} \in \{0, \dots, N_{\text{Rx}}-1\}} \hat{\tau}_{\Delta}^{n_{\text{Rx}}}, \quad (5.19)$$

where $\hat{\tau}_{\Delta}^{n_{\text{Rx}}}$ is the STO estimate for the n_{Rx} th receive channel. The procedure to obtain the global STO, CFO, and SFO estimates, i.e., $\hat{\tau}_{\Delta}^{\text{MIMO}}$, $\hat{f}_{\Delta}^{\text{MIMO}}$, and $\hat{\delta}^{\text{MIMO}}$, is illustrated in Figure 5.3. It is worth highlighting that the STO estimate will not only include the STO itself, but also the delay of the LoS reference path as they are not distinguishable at the receiver side. Disregarding estimation bias for the sake of simplicity, the STO estimate $\hat{\tau}_{\Delta}^{n_{\text{Rx}}}$ at the n_{Rx} th receive channel will ideally be given by

$$\hat{\tau}_{\Delta}^{n_{\text{Rx}}} = \tau_0 + \tau_{\Delta} + \tau_{\text{ABE}}^{n_{\text{Rx}}}. \quad (5.20)$$

In other words, $\hat{\tau}_{\Delta}^{n_{\text{Rx}}}$ will be ideally the result of the sum of the delay associated with the LoS reference path, the actual STO between transmitter and receiver,

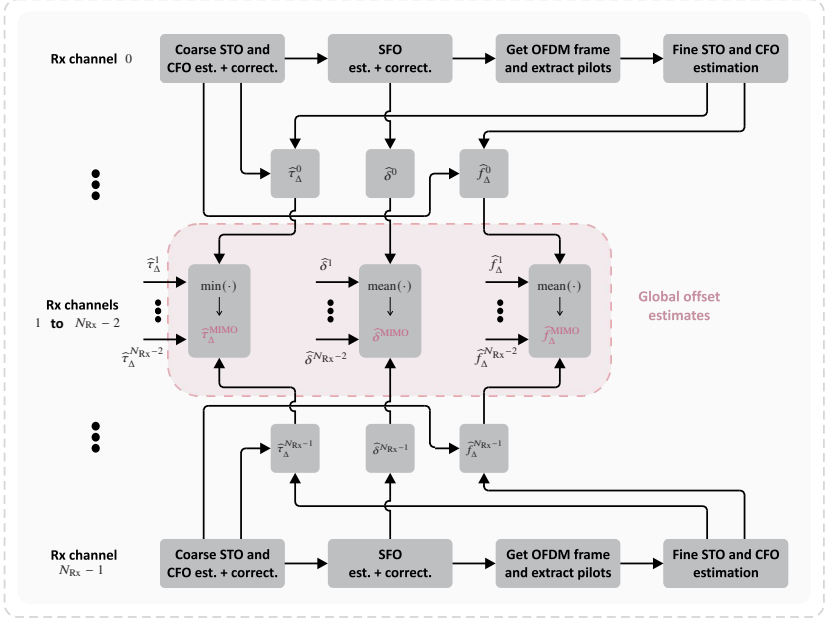


Figure 5.3. Representation of procedure to obtain global STO $\hat{\tau}_\Delta^{MIMO}$, CFO $\hat{f}_\Delta^{MIMO}$ and SFO $\hat{\delta}^{MIMO}$ estimates at the receiver. Adapted from [24], © 2026 IEEE.

and the delay of the dominant path of the ABE CIR at the n_{Rx} th receive channel. Consequently, (5.19) can be rewritten as

$$\hat{\tau}_\Delta^{MIMO} = \tau_0 + \tau_\Delta + \min_{n_{Rx} \in \{0, \dots, N_{Rx}-1\}} \tau_{ABE}^{n_{Rx}} \quad (5.21)$$

if the assumption of the absence of estimation bias is kept.

Once the start point of the OFDM frame has been defined based on the STO estimate $\hat{\tau}_\Delta^{MIMO}$, the normalized SFO estimate $\hat{\delta}^{MIMO}$ is used to perform re-sampling of the receive signal, followed by a CFO correction based on $\hat{f}_\Delta^{MIMO}$. This correction happens at each receive channel and is illustrated in Figure 5.4, along with the remaining processing steps at the receiver of the bistatic MIMO-OFDM-based ISAC system, including communications and bistatic radar signal processing later described in Sections 5.2.2.2 and 5.2.2.3. Since the STO was

corrected only at the sample level and the ABE CIR delays $\tau_{\text{ABE}}^{n_{\text{Rx}}}$ are different among the N_{Rx} receive channels, a fine tuning is required. Available pilot subcarriers in the OFDM frame are then used to estimate the residual STO and CFO associated with the LoS reference path, consequently allowing their correction and aligning the LoS reference path at a delay of 0 s and a Doppler shift of 0 Hz throughout all N_{Rx} receive channels.

The described synchronization for the proposed bistatic MIMO-OFDM-based ISAC system results in a discrete-time domain sequence $y_s^{n_{\text{Rx}}} \in \mathbb{C}$ for every n_{Rx} th receive channel, which is expressed as

$$\begin{aligned}
 y_s^{n_{\text{Rx}}} \approx & \left[\left(\alpha_0^{n_{\text{Tx}}, n_{\text{Rx}}} N_{\text{Tx}} x(t) e^{-j2\pi f_c \tau_{\text{Tx}-\text{T},0}} e^{-j2\pi f_c \tau_{\text{T}-\text{Rx},0}^{n_{\text{Rx}}}} e^{j\psi_\Delta} \right. \right. \\
 & + \sum_{p=1}^{P-1} \alpha_p^{n_{\text{Tx}}, n_{\text{Rx}}} N_{\text{Tx}} x(t - (\tau_p - \tau_0)) e^{-j2\pi f_c \tau_{\text{Tx}-\text{T},p}} e^{-j2\pi f_c \tau_{\text{T}-\text{Rx},p}^{n_{\text{Rx}}}} e^{j2\pi f_{\text{D},p}^{n_{\text{Tx}}, n_{\text{Rx}}} t} e^{j\psi_\Delta} \Big) \\
 & \left. \left. * h_{\text{ABE}}^{n_{\text{Rx}}} \left(t - [\hat{\tau}_\Delta^{\text{MIMO}} - (\tau_0 + \tau_\Delta)] \right) \right] \right]_{t=sT_s}.
 \end{aligned} \tag{5.22}$$

At the n_{Rx} th receive channel, this sequence undergoes S/P conversion to yield a discrete-time domain OFDM frame. After CP removal, the OFDM symbols in the aforementioned frame undergo DFT resulting in the discrete-frequency domain OFDM frame $\mathbf{Y}^{n_{\text{Rx}}} \in \mathbb{C}^{N \times M}$ with $N \in \mathbb{N}_{\geq 0}$ subcarriers and $M \in \mathbb{N}_{\geq 0}$ OFDM symbols. From this frame, pilots can be extracted to perform the aforementioned residual STO and CFO corrections before further communications and bistatic radar signal processing take place.

5.2.2.2 Communications signal processing

The communications signal processing for the n_{Rx} th receive channel starts with channel estimation based on pilot subcarriers. As in Section 3.1 and [22], they are assumed to be distributed in the OFDM frame with a regular spacing of $\Delta N_{\text{pil}} \in \mathbb{N}_{>0}$ subcarriers and $\Delta M_{\text{pil}} \in \mathbb{N}_{>0}$ OFDM symbols, resulting in a total of $M_{\text{pil}} = M / \Delta M_{\text{pil}}$ pilot OFDM symbols, each with $N_{\text{pil}} = N / \Delta N_{\text{pil}}$ pilot subcarriers. The estimated communications CFRs at the positions of the

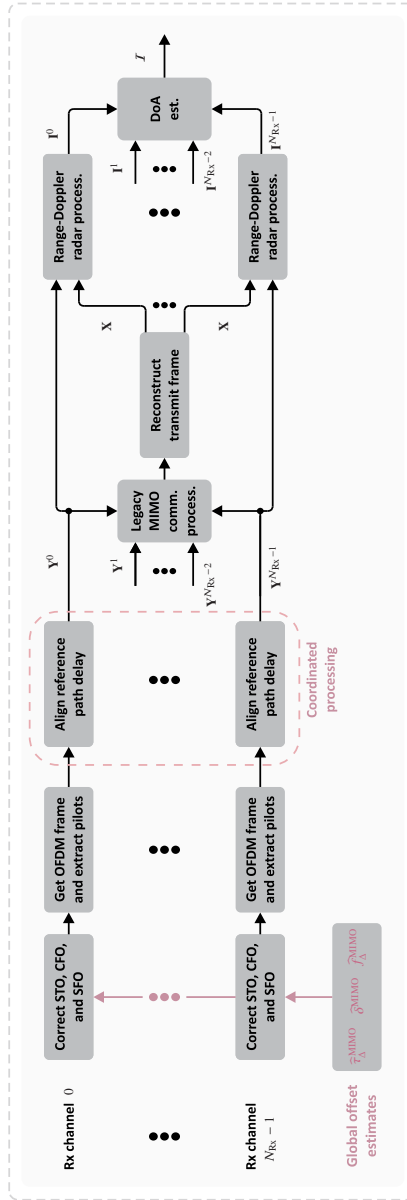


Figure 5.4. Representation of remaining ISAC processing chain after estimation of global synchronization offset estimates at the receiver. Adapted from [24], © 2026 IEEE.

mentioned pilot subcarriers then undergo two-dimensional interpolation to yield the communications CFR matrix $\mathbf{H}_{\text{CFR}}^{n_{\text{Rx}}} \in \mathbb{C}^{N \times M}$.

After channel estimation, the discrete-frequency domain OFDM frames $\mathbf{Y}^{n_{\text{Rx}}}$ at all N_{Rx} receive channels undergo maximal-ratio combining (MRC) [Kah54, Bre59] as an additional step to what is performed in the SISO case in Section 3.2.2 and [6, 50]. This results in a single discrete-frequency domain OFDM frame $\mathbf{Y}^{\text{MRC}} \in \mathbb{C}^{N \times M}$, whose element at its n th row, $n \in \{0, 1, \dots, N - 1\}$, and m th column, $m \in \{0, 1, \dots, M - 1\}$, is given by

$$Y_{n,m}^{\text{MRC}} = \frac{\sum_{n_{\text{Rx}}=0}^{N_{\text{Rx}}-1} \left[Y_{n,m}^{n_{\text{Rx}}} \left(H_{\text{CFR},n,m}^{n_{\text{Rx}}} \right)^* \right]}{\sum_{n_{\text{Rx}}=0}^{N_{\text{Rx}}-1} \left| H_{\text{CFR},n,m}^{n_{\text{Rx}}} \right|^2}. \quad (5.23)$$

In this equation, $Y_{n,m}^{n_{\text{Rx}}} \in \mathbb{C}$ and $H_{\text{CFR},n,m}^{n_{\text{Rx}}} \in \mathbb{C}$ are the elements at the n th row and m th column of $\mathbf{Y}^{n_{\text{Rx}}}$ and $\mathbf{H}_{\text{CFR}}^{n_{\text{Rx}}}$, respectively. The estimated transmit modulation symbols can be directly extracted from $\mathbf{Y}^{\text{MRC}}$ to form an estimate of the transmit frame $\mathbf{X} \in \mathbb{C}^{N \times M}$ that will be later also used for bistatic radar signal processing. If channel coding was applied, $\mathbf{Y}^{\text{MRC}}$ must first undergo decoding to obtain the transmit bit stream estimate, which will then be re-encoded to form an estimate of $\mathbf{X}$ as discussed in Section 3.2.2 and [6].

5.2.2.3 Bistatic radar signal processing

Having estimated the transmit frame $\mathbf{X}$ as described in Section 5.2.2.2, the bistatic signal processing starts by estimating the radar CFR matrix $\mathbf{D}^{n_{\text{Rx}}} \in \mathbb{C}^{N \times M}$ for every n_{Rx} th receive channel. Since the ABE-induced delay has already been considered during STO correction as described in Section 5.2.2.1, only the magnitude of the ABE CFR must be calibrated to avoid distorting the ultimately obtained radar images. In this sense, the element at the n th row and m th column of $\mathbf{D}^{n_{\text{Rx}}}$ is calculated as

$$D_{n,m}^{n_{\text{Rx}}} = Y_{n,m}^{n_{\text{Rx}}} \left/ \left(\left| H_{\text{ABE},n,m}^{n_{\text{Rx}}} \right| X_{n,m} \right) \right., \quad (5.24)$$

where $H_{\text{ABE},n,m}^{n_{\text{Rx}}} \in \mathbb{C}$ is the element at the corresponding subcarrier and OFDM symbol position of the ABE CFR matrix $\mathbf{H}_{\text{ABE}}^{n_{\text{Rx}}} \in \mathbb{C}^{N \times M}$. The aforementioned ABE CFR matrix is associated with $h_{\text{ABE}}^{n_{\text{Rx}}}(t)$ and can be estimated in advance

via calibration measurements, e.g., over-the-air [VRD⁺20] or feeding a known broadband signal to all receive channels and estimating their response. Since the ABE CFR is time-invariant, $H_{\text{ABE},n,m}^{n_{\text{Rx}}}$ only varies along with subcarrier index n , assuming the same complex values for different OFDM symbol indexes m . It is also worth highlighting that only the absolute value of $H_{\text{ABE},n,m}^{n_{\text{Rx}}}$ is taken to avoid distortion in the sidelobe profile of radar targets in the ultimately obtained range-Doppler shift radar images. This is done since the phase of the aforementioned CFR is associated with delay experienced through convolution with the corresponding CIR. Since this delay has already been compensated during synchronization, it does not have to be considered again when performing bistatic radar signal processing.

By performing the same bistatic radar signal processing steps as for the SISO case with full-frame processing from Section 3.2.2 and [6, 50] on $\mathbf{D}^{n_{\text{Rx}}}$, a range-Doppler shift radar image $\mathbf{I}^{n_{\text{Rx}}} \in \mathbb{C}^{N \times M}$ can be obtained for the n_{Rx} th receive channel. In this radar image, the targets reflections associated with paths $p \in \{1, \dots, P-1\}$ will have image SNR defined as

$$\text{SNR}_{\text{image},p}^{n_{\text{Rx}}} = \frac{P_{\text{Tx}} (N_{\text{Tx}} G_{\text{Tx}}) G_{\text{Rx}} \sigma_{\text{RCS},p} \lambda_0^2 G_p}{(4\pi)^3 R_{\text{Tx}-\text{T},p}^2 R_{\text{T}-\text{Rx},p}^2 k_B B T_{\text{therm}} \text{NF}}. \quad (5.25)$$

In this equation, G_{Tx} and G_{Rx} are the single-element transmit and receive antenna gains, respectively, previously defined in Section 2.1. $\sigma_{\text{RCS},p}$ is the RCS of the p th target as previously defined in Section 3.1, which is assumed to be the same for every transmit and receive antenna pair for simplicity. $G_p = NM$ is the range-Doppler shift radar processing gain assuming the full-frame processing discussed in Section 3.2.2. In addition,

$$R_{\text{Tx}-\text{T},p} = c_0 \tau_{\text{Tx}-\text{T},p}, \quad (5.26)$$

which is a rearranged form of (5.10), and

$$R_{\text{T}-\text{Rx},p} = c_0 \tau_{\text{T}-\text{Rx},p} \quad (5.27)$$

are the ranges from transmitter to the p th target and the p th target to the receiver, respectively. Similarly to (5.15), they are defined assuming

$$\tau_{\text{Tx}-\text{T},p}^{n_{\text{Tx}}} \approx \tau_{\text{Tx}-\text{T},p} \quad \forall n_{\text{Tx}}, p \quad (5.28)$$

and

$$\tau_{\text{T-Rx},p}^{n_{\text{Rx}}} \approx \tau_{\text{T-Rx},p} \quad \forall n_{\text{Rx}}, p. \quad (5.29)$$

Finally, the term $k_B B T_{\text{therm}} \text{NF}$ in the denominator accounts for the AWGN power. It is defined by the Boltzmann constant k_B , the previously defined OFDM signal bandwidth B , the standard room temperature in Kelvin T_{therm} , and the overall receiver noise figure NF.

Next, Fourier beamforming [NSF⁺17] is performed for DoA estimation, which allows obtaining a three-dimensional radar image represented by the matrix $\mathcal{I} \in \mathbb{C}^{N \times M \times N_{\text{Rx}}^S}$, where N_{Rx}^S is the number of evaluated DoAs. Specifically, the depths of $\mathcal{I}$ are associated with range, Doppler shift, and azimuth DoA, respectively.

Assuming a steering vector $\mathbf{b}_{\text{Rx},n_{\text{Rx}}^S} \in \mathbb{C}^{1 \times N_{\text{Rx}}}$ for the n_{Rx}^S th evaluated DoA $\phi_{\text{T-Rx},n_{\text{Rx}}^S}$, whose n_{Rx}^S th element is given by

$$b_{\text{Rx},n_{\text{Rx}}^S}^{n_{\text{Rx}}} = e^{j2\pi(\rho_{\text{Rx},x}^{n_{\text{Rx}}}/\lambda_0) \sin(\phi_{\text{T-Rx},n_{\text{Rx}}^S})}, \quad (5.30)$$

the element at the n th row, m th column, and n_{Rx}^S th depth of $\mathcal{I}$ can be expressed as

$$\mathcal{I}_{n,m,n_{\text{Rx}}^S} = \sum_{n_{\text{Rx}}^S=0}^{N_{\text{Rx}}^S-1} I_{n,m}^{n_{\text{Rx}}} b_{\text{Rx},n_{\text{Rx}}^S}^{n_{\text{Rx}}} = \sum_{n_{\text{Rx}}^S=0}^{N_{\text{Rx}}^S-1} I_{n,m}^{n_{\text{Rx}}} e^{j2\pi(\rho_{\text{Rx},x}^{n_{\text{Rx}}}/\lambda_0) \sin(\phi_{\text{T-Rx},n_{\text{Rx}}^S})}. \quad (5.31)$$

In the previous equations, $\rho_{\text{Rx},x}^{n_{\text{Rx}}}$ is the x coordinate of the n_{Rx} th receive antenna in the coordinate system $(x_{\text{Rx}}, y_{\text{Rx}}, z_{\text{Rx}})$ of the receive array. Extending the reasoning used for transmit beamforming to the Fourier beamforming-based DoA estimation, the phase shifts $e^{-j2\pi f_c \tau_{\text{T-Rx},p}^{n_{\text{Rx}}}}$ in (5.22) will lead to target peaks at for the same ranges and Doppler shifts as in the range-Doppler shift radar images $\mathbf{I}^{n_{\text{Rx}}}$, but now at every n_{Rx}^S th depth of $\mathcal{I}$ that is associated with DoA $\phi_{\text{T-Rx},n_{\text{Rx}}^S}$ such that $\phi_{\text{T-Rx},n_{\text{Rx}}^S} = \phi_{\text{T-Rx},p}$, where $\phi_{\text{T-Rx},p}$ is the actual DoA with respect to the receiver associated with the p th path. Target reflections in

this radar image associated with paths $p \in \{1, \dots, P-1\}$ have an image SNR given by

$$\text{SNR}_{\text{image},p}^{\text{MIMO}} = \frac{P_{\text{Tx}} (N_{\text{Tx}} G_{\text{Tx}}) (N_{\text{Rx}} G_{\text{Rx}}) \sigma_{\text{RCS},p} \lambda_0^2 G_p}{(4\pi)^3 R_{\text{Tx}-\text{T},p}^2 R_{\text{T}-\text{Rx},p}^2 k_B B T_{\text{therm}} \text{NF}}, \quad (5.32)$$

where a gain of factor N_{Rx} with respect to (5.25) is observed due to the coherent processing of N_{Rx} range-Doppler shift radar images $\mathbf{I}^{n_{\text{Rx}}}$ for DoA estimation.

5.3 Numerical analysis

To analyze the performance of the proposed bistatic MIMO-OFDM-based ISAC system concept, $N_{\text{Tx}} = 4$ transmit and $N_{\text{Rx}} = 8$ receive channels are considered, with both transmit and receive arrays assumed to be ULAs with a $\lambda_0/2$ element spacing. Furthermore, it is assumed that BP sampling at a digital IF of $f_{\text{IF}} = 3.68$ GHz is performed. This specific digital IF is used in the described testbed in [NKN⁺24] that is later used for measurement-based validation in Section 5.4.

The adopted OFDM signal parameters including the ones discussed so far, besides subcarrier spacing $\Delta f = B/N$ and CP length $N_{\text{CP}} \in \mathbb{N}_{\geq 0}$, are listed in Table 5.1. In addition, QPSK modulation for all subcarriers and an LDPC channel code [MKJ⁺24] of rate 2/3 and the parity check matrix for the case with 64800 bits in an LDPC code block from [ETS14] were assumed. This resulted in the ISAC performance parameters calculated according to the discussion in Section 3.2.2 and [VRD⁺20, 50], which are also listed in the same table. These are namely communications data rate at 100% duty cycle, $\mathcal{R}_{\text{comm}}$, processing gain G_p considering the DoA estimation gain, range resolution ΔR , maximum unambiguous range $R_{\text{max,ua}}$, maximum ISI-free range $R_{\text{max,ISI}}$, Doppler shift resolution Δf_{D} , maximum unambiguous Doppler shift $f_{\text{D,max,ua}}$, maximum ICI-free Doppler shift $f_{\text{D,max,ICI}}$, Rx azimuth resolution $\Delta \phi_{\text{Rx}}$, and maximum unambiguous Rx azimuth $\phi_{\text{Rx,max,ua}}$.

Most relevant aspects of bistatic OFDM-based ISAC system performance have been analyzed in Chapters 3 and 4, as well as in studies such as [6] for the SISO case. In the adopted MIMO architecture, the main difference is the MRC diversity gain obtained according to Section 5.2.2.2 for communications, the

Table 5.1: MIMO-OFDM signal and ISAC performance parameters. Adapted from [24], © 2026 IEEE.

OFDM signal parameters		ISAC performance parameters	
f_c	27.5 GHz	$\mathcal{R}_{\text{comm}}$	0.39 Gbit/s
B	491.52 MHz	G_p	60.22 dB
N	2048	ΔR	0.61 m
Δf	240 kHz	$R_{\text{max,ua}}$	1249.14 m
N_{CP}	512	$R_{\text{max,ISI}}$	312.28 m
M	512	Δf_D	375 Hz
ΔN_{pil}	2	$f_{D,\text{max,ua}}$	± 96 kHz
ΔM_{pil}	2	$f_{D,\text{max,ICI}}$	± 24 kHz
N_{pil}	1024	$\Delta \phi_{\text{Rx}}$	14.32°
M_{pil}	256	$\phi_{\text{Rx,max,ua}}$	$\pm 90^\circ$

DoA estimation capability for radar as described in Section 5.2.2.3, besides the distributed synchronization discussed in Section 5.2.2.1. In this section, the focus is placed on the latter. Since all $N_{\text{Tx}} = 4$ transmit channels and $N_{\text{Rx}} = 8$ receive channels share the same LO reference and sampling clocks locally at the transmitter and receiver, respectively, the same CFO and SFO will be experienced at all receive channels as discussed in Section 5.2.2.1. Therefore, no further analysis of these impairments besides the one for the SISO case in Chapters 3 and 4 as well as in [6, 22] is performed in this section. An additional hardware impairment that may degrade the performance of the considered bistatic MIMO-OFDM-based ISAC system is PN. The effect of PN on bistatic OFDM-based ISAC has been investigated assuming PLL-based oscillators in a SISO architecture in Sections 3.4.4.1 and 4.2.3 as well as in [21], showing that it leads to ICI and CPE, which degrade the performance of range and Doppler shift estimations, respectively. In the MIMO case considered in this chapter, PN may additionally degrade both the transmit beamforming and the DoA estimation at the receiver side. The extent of the degradation will depend on whether independent LOs are adopted for each receive channel or whether a single LO is distributed to the mixers of all N_{Rx} receive channels [PLN⁺16]. In the first case, the uncorrelated PN of the different channels tends to average out

at the beamforming or DoA directions [HR10]. As for the latter case, although the different delays $\tau_{\text{ABE}}^{n_{\text{Rx}}}$ experienced at each receive channel lead to reduction of the correlation of their PN contributions, performance degradation may still occur if no low-PN LO source is adopted [CNF25]. This, however, is left as an open topic for a future study.

Regarding the STO, however, an analysis of the robustness of the proposed system concept and processing chain to mismatches among the delays $\tau_{\text{ABE}}^{n_{\text{Rx}}}$ associated with the N_{Rx} receive channels of the ABE becomes necessary as they may not be perfectly aligned even after synchronization as described in Section 5.2.2.1. For that purpose, it is henceforth assumed that CFO and SFO are absent and that perfect time synchronization is performed for the receive channel $n_{\text{Rx}} = 0$. For all other $N_{\text{Rx}} = 7$ channels, i.e., $n_{\text{Rx}} \in \{1, \dots, 7\}$, random delays defined according to a Rayleigh distribution with standard deviation σ_τ considered. Due to the use of a digital IF, a corresponding angular standard deviation σ_θ will be experienced among the channels. The choice of the Rayleigh distribution is made to ensure that no ISI immediately occurs for receive channels having smaller delay than $n_{\text{Rx}} = 0$. Based on the previous assumptions, the communications and radar sensing performances of the proposed bistatic MIMO-OFDM-based ISAC system are analyzed in Sections 5.3.1 and 5.3.2. In these analyses, the critical sampling period $T_s = 1/B$ is used as a reference to which the delay standard deviation is compared.

5.3.1 Communications performance

Figure 5.5 shows the EVM and BER simulated as functions of the normalized delay standard deviation by the sampling period, i.e., σ_τ/T_s . The obtained results assume MRC among the $N_{\text{Rx}} = 8$ receive channels according to (5.23). Figure 5.5(a) shows that the EVM increases linearly until around $\sigma_\tau = 10^{-1} T_s$. This corresponds to 10^{-1} discrete samples or 203.45 ps and is associated with a phase standard deviation of $\sigma_\theta = 139.03^\circ$ at the digital IF of $f_{\text{IF}} = 3.68$ GHz. Although a large deviation between the phases of the channels is experienced, this delay standard deviation does not result in ISI and therefore allows the MRC processing to perform sufficiently well, which is supported by the low EVM values that are equal to or lower than -47.84 dB in the aforementioned region. Between $\sigma_\tau = 10^{-1} T_s$ and $10^2 T_s$, the EVM trend changes and its values decrease until as low as -74.94 dB. For $\sigma_\tau > 10^2 T_s$, the EVM starts increasing

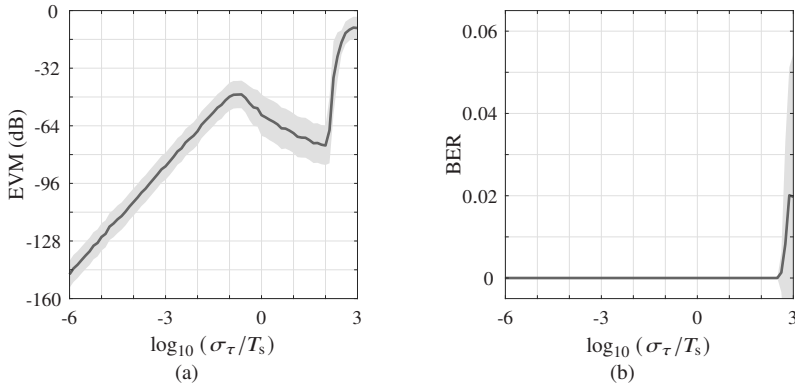


Figure 5.5. Simulated EVM (a) and BER (b) as functions of the normalized delay standard deviation σ_τ by the sampling period T_s . Both parameters were calculated for an ideal, noiseless channel. The continuous lines represent the mean value of the calculated parameters and the shading in the background represents the standard deviation. © 2026 IEEE. Reprinted, with permission, from [24].

again as such high delay standard deviations may result in ISI for some of the receive channels. Although distortions of the QPSK constellation were observed, it can be seen in Figure 5.5(b) that the adopted LDPC channel code can ensure null BER until around $\sigma_\tau = 10^{2.5} T_s$. Afterwards the experienced ISI at some of the receive channels can no longer be handled by the channel code and the BER increases up to 0.02 at $\sigma_\tau = 10^3 T_s$.

5.3.2 Radar sensing performance

To solely analyze the influence of STO mismatch among the receive channels, *genie-aided* decoding is assumed to allow perfect reconstruction of the transmit frame needed for the full-frame bistatic radar signal processing described in Section 3.2.2 and [6, 50]. In addition, a single, static radar target is considered and AWGN is disregarded to allow solely focusing on the STO effect. Figure 5.6 shows the PPLR, range PSLR and ISLR, as well as azimuth PSLR and ISLR calculated for the considered target in a range-azimuth cut of the radar image calculated as in (5.31) at zero Doppler shift as a function of the normalized delay standard deviation among receive channels $n_{R_X} \in \{1, \dots, 7\}$ by the sampling

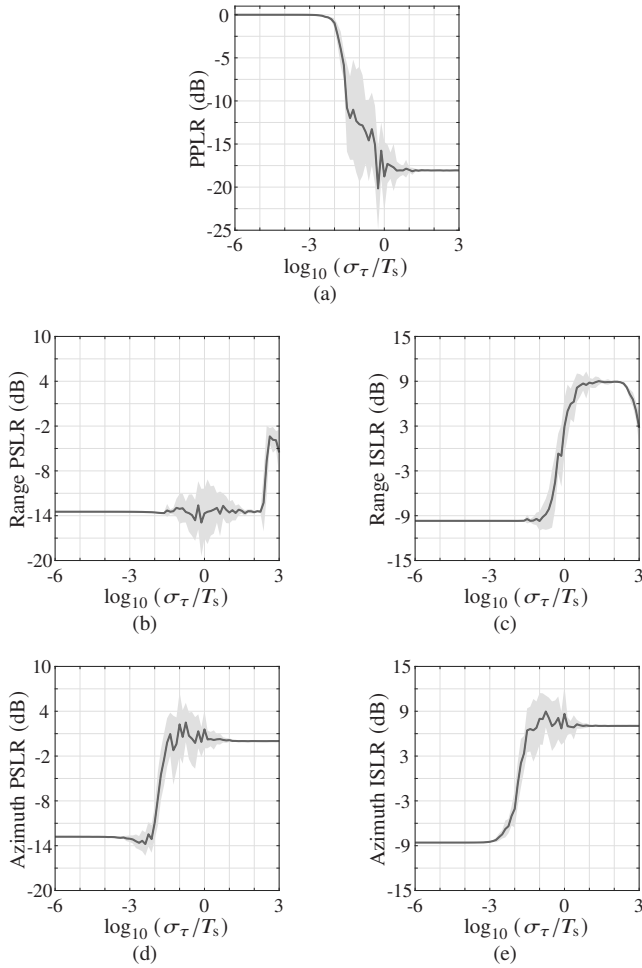


Figure 5.6. Simulated PPLR (a), range PSLR (b) and ISLR (c), and Doppler shift PSLR (e) and ISLR (d) as functions of the normalized delay standard deviation σ_τ by the sampling period T_s . The continuous lines represent the mean value of the calculated parameters and the shading in the background represents the standard deviation. © 2026 IEEE. Reprinted, with permission, from [24].

period T_s . For the calculation of these parameters, rectangular windowing was used for both range and azimuth processing to avoid window-specific sidelobe suppression. The obtained results show that, until around $\sigma_\theta = 10^{-3} T_s$, which corresponds to 10^{-3} discrete samples or 20.35 ps and is associated with a phase standard deviation of $\sigma_\theta = 2.70^\circ$ at the digital IF of $f_{IF} = 3.68$ GHz, no significant mainlobe or sidelobe degradation is observed. Afterwards, azimuth PSLR and ISLR are the first to degrade, with visible changes in sidelobes observable before $\sigma_\tau = 10^{-2} T_s$, which is associated with $\sigma_\theta = 26.96^\circ$. After this point, PPLR degradation is no longer negligible, with 3 dB peak power loss being reached at a delay standard deviation of around $\sigma_\tau = 6 \cdot 10^{-2} T_s$ and $\sigma_\theta = 100^\circ$. Both range PSLR and ISLR only start to degrade at the same aforementioned σ_τ value. However, while the ISLR continuously increases, which shows that the overall sidelobe level becomes higher, the range PSLR oscillates and only starts to continuously increase at $\sigma_\tau = 177.83 T_s$, which corresponds to $\sigma_\theta = 219.13^\circ$. At this high delay standard deviation, ISI may not be entirely ruled out for all receive channels as CP of length $N_{CP} = 512$ was adopted.

Combined with the earlier degradation of azimuth PSLR and ISLR, the PPLR and range PSLR and ISLR results show that the azimuth estimation is more severely affected by the mismatch of the experienced STOs among the $N_{RX} = 8$ receive channels due to the experienced phase mismatch among the receive channels. While the achieved results seem to indicate that tolerable performance degradation is attained even for high phase rotations, as is the case for $\sigma_\tau = 6 \cdot 10^{-2} T_s$ and $\sigma_\theta = 100^\circ$, one must consider that the use of rectangular windowing results in high sinc-shaped sidelobes that may mask the overall radar sensing performance degradation. In this sense, a further performance parameter, namely the image SIR is analyzed as follows.

Figure 5.7 shows the mean image SIR as a function of σ_τ/T_s . The results were obtained from simulations with the considered single, static target and assuming the use of Chebyshev windowing with 100 dB sidelobe suppression for both range and azimuth processing to ensure that only distortion in the cut of the radar image at a Doppler shift of 0 Hz, and not range or azimuth sidelobes are considered. The mean image SIR is calculated as the ratio between the peak power of the target at 0 m range and 0° azimuth, which is associated with the receive channel $n_{RX} = 0$ that has unbiased synchronization, and the average power at the rest of the radar image cut. The obtained results show that

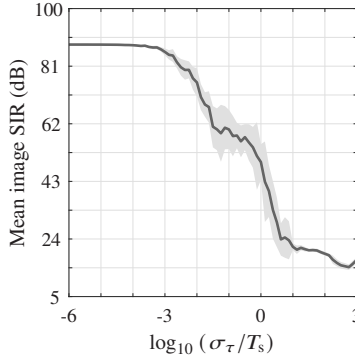


Figure 5.7. Simulated mean image SIR as a function of the normalized delay standard deviation σ_τ by the sampling period T_s . The continuous lines represent the mean value of the calculated parameters and the shading in the background represents the standard deviation. © 2026 IEEE. Reprinted, with permission, from [24].

only negligible mean image SIR degradation is observed for $\sigma_\tau \leq 10^{-3.25} T_s$, which corresponds to $\sigma_\theta \leq 1.52^\circ$. Since the aforementioned values are low, they do not lead to either significant range offsets nor phase rotation among the receive channels, which explains the nearly constant mean image SIR. For $\sigma_\tau > 10^{-2} T_s$, the mean image SIR starts to decrease rapidly, with a 10 dB reduction with respect to the highest achievable mean image SIR being already experienced at $\sigma_\tau \leq 10^{-2} T_s$. Since, however, the overall performance degradation in the aforementioned region is moderate and mainly due to phase mismatches among the receive channels, it can be considered that STO can be sufficiently compensated afterwards at the pilot-based fine tuning discussed in Section 5.2.2.1 since it may achieve sub-sample accuracy, e.g., by using CZT as discussed in Section 4.1.3.2 and [22].

5.4 Measurement-based verification

To validate the proposed bistatic MIMO-OFDM-based ISAC system concept, the setup shown in Figure 5.8 was adopted. It has the same considered transmit and receive arrays with $N_{Tx} = 4$ and $N_{Rx} = 8$ antenna elements, respectively, as assumed in Section 5.3. The aforementioned elements are microstrip patch antennas that were selected from ULAs designed with an element spacing of

$\lambda_0/2$ at 28 GHz, i.e., 5.35 mm. More specifically, two serially-fed microstrip patch antennas were connected vertically to increase the antenna gain in elevation at the cost of reduced half-power beamwidth (HPBW). The resulting HPBWs are 40.8° for elevation and 86° for azimuth, and the measured antenna element gain is of 4.1 dB including losses. The used AFEs were designed to support RFs in the range from 26.5 GHz to 29.5 GHz, having a maximum output power of 10 dBm. As for the ABEs, the same frequency range is covered, and a noise figure of 2.1 dB was calculated, although the actual value is expected to be higher. Further details on the adopted ULAs and their associated AFEs and ABEs can be found in [Eis21]. For signal generation and sampling at the receiver side, the testbed described in [NKN⁺24], with specific components further analyzed in [KNN⁺24a, KNN⁺24b, SWR⁺24], was adopted. As in Section 5.3, BP sampling with a digital IF of $f_{\text{IF}} = 3.68$ GHz was performed, and the same OFDM signal parameters from Table 5.1 were adopted. In the adopted measurement scenario, the transmit ULA was placed at approximately 55 cm and an approximate DoA of 3° with respect to the receive ULA. In addition, a reflector with an approximate bistatic range of 70 cm and an approximate DoA of -20° was used as a static radar target. At the transmitter side, the AFEs associated with each transmit channel were calibrated and beamforming was performed towards the DoDs of 0° and 30° , which approximately correspond to the DoDs of the LoS reference path towards the receiver and the reflector, respectively.

Before conducting the actual bistatic ISAC measurements, the CFRs of the $N_{\text{Tx}} = 4$ transmit channels were estimated via calibration measurements with a reference receiver. Figure 5.9 shows the normalized magnitude of the AFE CFRs $H_{\text{ABE}}^{n_{\text{Rx}}}(f)$ as a function of the BB frequency for all transmit channels. Their discrete-frequency domain equivalents were smoothed with a moving average filter and used to perform DPD, ensuring beamforming towards the previously mentioned DoDs. In addition, the CFRs and the corresponding CIRs of all $N_{\text{Rx}} = 8$ receive channels of the ABE were estimated via calibration measurements with a reference transmitter. Figure 5.10 shows the normalized magnitude of the ABE CFRs $H_{\text{ABE}}^{n_{\text{Rx}}}(f)$ as a function of the BB frequency only for the receive channels $n_{\text{Rx}} \in \{0, 3, 7\}$ for better visualization. Their equivalent discrete-frequency domain representations $H_{\text{ABE},n,m}^{n_{\text{Rx}}}$ are later used for bistatic radar signal processing as described by (5.24). In addition, the normalized magnitude of the corresponding ABE CIRs $h_{\text{ABE}}^{n_{\text{Rx}}}(t)$ are also shown. The delays of their dominant paths, i.e., $\tau_{\text{ABE}}^{n_{\text{Rx}}}$, ultimately influence both the local

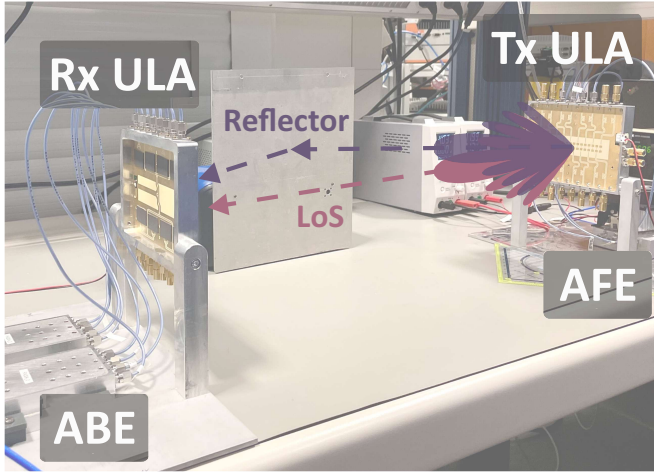


Figure 5.8. Bistatic MIMO-OFDM-based ISAC measurement setup. © 2026 IEEE. Reprinted, with permission, from [24].

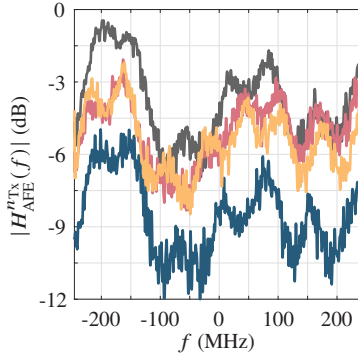


Figure 5.9. Measured normalized magnitude of the AFE CFR $H_{AFE}^{n_{Tx}}(f)$ versus BB frequency f for transmit channels $n_{Tx} = 0$ (—), $n_{Tx} = 1$ (—), $n_{Tx} = 2$ (—), and $n_{Tx} = 3$ (—). Adapted from [24], © 2026 IEEE.

STO estimates at each channel as described in (5.20) as well as the global STO estimate calculated as in (5.21).

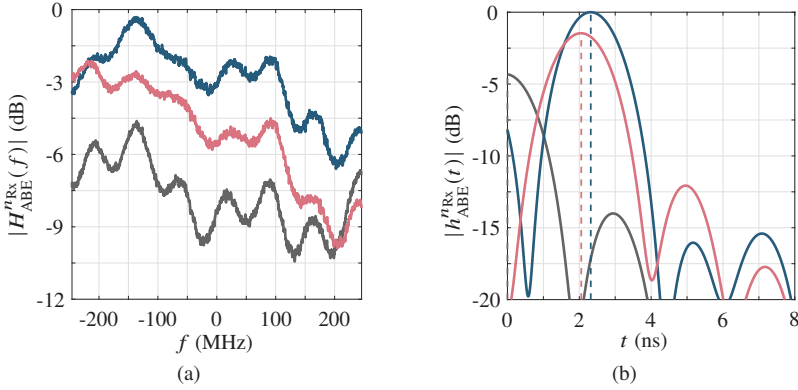


Figure 5.10. Measured normalized magnitude of the ABE CFR $H_{\text{ABE}}^{n_{\text{Rx}}}(f)$ versus BB frequency f (a) and normalized magnitude of the ABE CIR $h_{\text{ABE}}^{n_{\text{Rx}}}(t)$ versus time t (b) for receive channels $n_{\text{Rx}} = 0$ (—), $n_{\text{Rx}} = 3$ (—), and $n_{\text{Rx}} = 7$ (—). Their delays $\tau_{\text{ABE}}^{n_{\text{Rx}}}$ associated with the peaks of the three shown CIRs are $\tau_{\text{ABE}}^0 \approx 0$ ns (—), $\tau_{\text{ABE}}^3 \approx 2.32$ ns (—), and $\tau_{\text{ABE}}^7 \approx 2.06$ ns (—). Adapted from [24], © 2026 IEEE.

Table 5.2: Measured local and global synchronization offset estimates. © 2026 IEEE. Reprinted, with permission, from [24].

Receive channel	Residual STO	CFO	SFO
$n_{\text{Rx}} = 0$	0.0235 ns	15.4077 kHz	−4.1615 ppm
$n_{\text{Rx}} = 1$	0.4069 ns	15.3095 kHz	−4.1604 ppm
$n_{\text{Rx}} = 2$	1.4242 ns	15.3760 kHz	−4.1606 ppm
$n_{\text{Rx}} = 3$	1.6276 ns	15.5673 kHz	−4.1613 ppm
$n_{\text{Rx}} = 4$	1.0173 ns	15.6186 kHz	−4.1609 ppm
$n_{\text{Rx}} = 5$	1.2207 ns	15.4008 kHz	−4.1609 ppm
$n_{\text{Rx}} = 6$	1.4242 ns	15.5489 kHz	−4.1586 ppm
$n_{\text{Rx}} = 7$	2.0345 ns	15.5884 kHz	−4.1607 ppm
Global estimate	N/A	15.4772 kHz	−4.1606 ppm

To enable both communications and radar sensing processing, STO, CFO and SFO estimation and compensation were individually performed at each receive

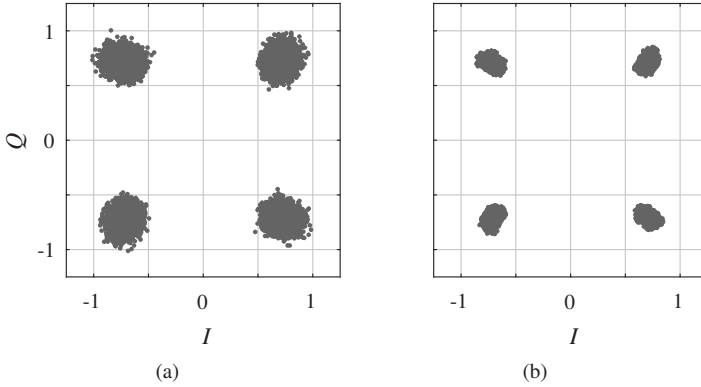


Figure 5.11. Measured normalized QPSK constellations: (a) single channel with ZF equalization and (b) MRC of all $N_{\text{Rx}} = 8$ receive channels. © 2026 IEEE. Reprinted, with permission, from [24].

channel as described in Section 5.2.2.1. The obtained local synchronization offset estimates, as well as the resulting global estimates are shown in Table 5.2. Since the earliest estimated OFDM frame start point among all N_{Rx} receive channels, which in the measurements was the one for receive channels $n_{\text{Rx}} \in \{0, 1\}$, is taken as a common start point for all channels, only the residual STO later estimated via pilots is shown. The obtained residual STO values show that larger compensation was needed for the remaining receive channels, i.e., $n_{\text{Rx}} = 2$ through $n_{\text{Rx}} = 7$ as expected. Overall, a mean value of 1.17 ns and standard deviation of 0.11 ns was observed among the local residual STO estimates, which respectively correspond to 57.50% and 5.38% of the critical sampling period $T_s = 1/B$. The CFO estimates, in turn, include both the first estimation via the S&C algorithm and the fine tuning with pilots. A standard deviation of 116.18 Hz was observed among the CFO estimates, which corresponds to only 0.05% of the subcarrier spacing of $\Delta f = 240$ kHz. This shows that the local estimates were already accurate and therefore converged to an even more accurate global CFO estimate. As for SFO, a standard deviation of only 48.45×10^{-3} ppm was observed among the local estimates, which was the case due to the use of the TITO algorithm introduced in Section 3.3.3 and [22] that allows performing highly accurate SFO estimation.

Afterwards channel estimation and equalization were performed and the constellations shown in Figure 5.11 were obtained. Specifically, Figure 5.11(a) shows the receive QPSK constellation for the channel $n_{\text{Rx}} = 0$ after ZF FDE, which has EVM with a mean value of -28.05 dB and 5.73 dB standard deviation. After MRC following (5.23) of all $N_{\text{Rx}} = 8$ receive channels, the constellation shown in Figure 5.11(b) was obtained. This constellation is associated with a mean EVM of -28.98 dB and 5.34 dB standard deviation. After performing demodulation on the constellation from Figure 5.11(b) and LDPC decoding, followed by re-encoding and re-modulation, an estimate of the transmit OFDM frame $\mathbf{X}$ was obtained and used for bistatic radar signal processing as described in (5.24). First, range-Doppler radar images $\mathbf{I}^{\text{rx}}$ were obtained for all receive channels, with an example for the channel $n_{\text{Rx}} = 0$ shown in Figure 5.12(a). In this image, it can be seen that a single reflection is seen at a relative bistatic range of 0 m with respect to the LoS reference path. This happens since the LoS path and the reflector are unresolved, which is due to the fact that the difference of 15.23 cm between their bistatic ranges is smaller than the range resolution $\Delta R = 0.61$ m. The second reflection seen at a relative bistatic range of 4.12 m is due to reflections in the room where the measurement was performed. After DoA estimation via Fourier beamforming was performed as described in (5.31), a three-dimensional radar image $\mathcal{I}$ was obtained. A cut of this image at a Doppler shift of 0 Hz is shown in Figure 5.12(b). In this image, it can be seen that the LoS path and reflector are resolved, which confirms that the necessary phase coherence for DoA estimation was achieved with the presented synchronization, communications and radar signal processing scheme for the proposed bistatic MIMO-OFDM-based ISAC system in Section 5.1. The deviations from the ground truth DoAs provided at the beginning of this section can be explained by limited accuracy and possible systematic errors in the ground truth measurements.

5.5 Chapter conclusion

This chapter introduced a system concept for bistatic MIMO-OFDM-based ISAC. After a system model description, the additional processing steps with respect to a bistatic SISO-OFDM-based ISAC system were described. These include transmit beamforming, a proposed distributed synchronization concept

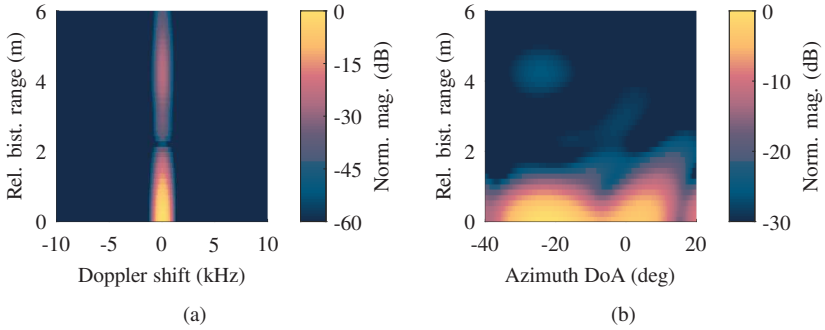


Figure 5.12. Measured bistatic radar images: (a) range-Doppler shift radar image showing unresolved reflector and LoS paths at 0 m relative bistatic range, and (b) range-azimuth DoA radar image with beamforming at the transmitter for null Doppler shift showing resolved reflector and LoS paths at -22.5° and 3° , respectively. Adapted from [24], © 2026 IEEE.

including STO, CFO, and SFO estimation and correction, as well as bistatic radar signal processing to estimate range, Doppler shift, and DoA.

In the proposed synchronization concept, it was shown that local offset estimates obtained at each receive channel can be averaged to obtain a single global estimate that is then used for synchronization at all channels in the CFO and SFO cases. As for STO, different CIRs and consequently delays are experienced at each ABE receive channel due to hardware non-idealities. Consequently, the earliest locally estimated OFDM frame start point among all receive channels is set as the global start point estimate to avoid ISI. The residual STOs can be later individually corrected at each receive channel based on estimates obtained via pilot subcarriers.

Focusing on STO mismatch among the receive channels levels, simulation results showed that negligible communications performance degradation despite visibly increased EVM as well as negligible peak and sidelobe distortion in radar sensing were observed for an STO standard deviation of 1% the critical sampling period among the receive channels. At this point, however, an image SIR drop of 10 dB compared to the case without STO mismatch among the receive channels is experienced, which although moderate is not negligible. This level of imbalance among the STOs can, however, be corrected during STO fine tun-

ing based on pilots as sub-sample accuracy can be achieved. Finally, the claims and simulation results were confirmed by verification measurements with a 4×8 MIMO setup at 27.5 GHz. The results confirmed the similar CFO and SFO experienced by each individual receive channel, and also showed the mismatch in the STOs. All the aforementioned offsets could, however be estimated and corrected with the proposed synchronization approach, and communications and radar sensing capabilities in the proposed bistatic MIMO-OFDM-based ISAC system concept were successfully demonstrated.

6 Conclusion

Parts of this chapter are adapted from [22] © 2025 IEEE.

With the increasing demand for more complex hardware and spectral resources, e.g., to support wide bandwidths and high carrier frequencies needed for high-speed communications and high resolution sensing, ISAC has significantly gained popularity in recent years. An ISAC system was reported for the first time by Mealey in [Mea63], where a missile range instrumentation radar transmits a group of time- or on-and-off-modulated pulses to jointly detect and communicate with space vehicles equipped with beacon receivers. Later in 2009, an OFDM-based ISAC system concept was first proposed by Sturm et al. in [SZW09]. Since then, significant progress has been made in the field.

Especially in the communications-centric context, ISAC became one of the most anticipated technologies to be enabled by future cellular networks in 6G and beyond [DBB⁺21, WBV21, LCM⁺22, CBMD23, WSN⁺24]. Indeed, 3GPP Release 19 and future 6G releases already include studies on ISAC. In this context, ISAC will come with the potential of turning cellular networks into the so-called perceptive mobile networks [ZRH⁺21], enabling applications such as monitoring of areas with campus networks in intralogistics scenarios with humans and AGVs as well as environment sensing for critical infrastructure or safe urban mobility [KZPH23, SNS⁺24]. In addition to the mentioned ones, several other potential use cases ISAC have been recently discussed by European Telecommunications Standards Institute (ETSI) [ETS25].

To facilitate the advancement of future practical ISAC deployments, a number of challenges must be analyzed. These include the choice for suitable modulation schemes for ISAC, the understanding of the effects of well-known hardware impairments in communications systems on the radar sensing performance, besides the development of synchronization and radar signal processing strategies to enable distributed sensing, starting with bistatic ISAC deployments. In this dissertation, research efforts towards advancing the theoretical foundation of

ISAC in all of the aforementioned fields are presented, discussed and supported by detailed analyses corroborated by simulation and measurement results. The major contributions of this dissertation are the following:

1. Improved processing chain for PMCW-based ISAC.
2. Introduction of C-OCDM- and SM-OCDM-based ISAC system concepts.
3. A robust synchronization procedure for OFDM-based bistatic ISAC covering all relevant offsets for the first time.
4. Radar signal processing approaches for bistatic OFDM-based ISAC relying on pilots only or on transmit frame reconstruction via communications processing.
5. Analysis of OFDM-based ISAC under major hardware impairments, with straightforward system design guidelines to ensure system performance.
6. System concept for bistatic MIMO-OFDM-based ISAC, with practical synchronization aspects being considered for the first time.
7. Verification of contributions via simulations and measurements.

Overall, the contributions of this dissertation are summarized as follows:

- **The development of a communications processing chain for PMCW-based ISAC.** In this chain, pulse compression and channel equalization are performed in the discrete-frequency domain. Although simple, the introduced concept offers a more computationally efficient and better performing approach to the state of the art.
- **The introduction of C-OCDM-based ISAC.** In the first modulation scheme, the radar signal processing is based on matched filtering. While more resilient, e.g., than OFDM to frequency-selective channels due to its spread spectrum nature, C-OCDM may present limited radar sensing performance due to transmit-data-dependent range sidelobes. Out of this reason, C-OCDM is better suited for communications-centric ISAC.
- **The development of SM-OCDM-based ISAC to overcome the radar sensing performance limitations of C-OCDM with respect to range sidelobes.** In this system, a strategic subchirp allocation is performed for radar sensing and communications. The radar processing relies on the

DFnT for pulse compression, which compresses the radar sensing sub-chirp into the autocorrelation function of the root Zadoff-Chu sequence. While virtually ideal for most cases, this autocorrelation may suffer from delay-Doppler coupling for high Doppler shifts. This effect can, however, be avoided with appropriate waveform parameterization. As for communications, the robustness of C-OCDM is kept, but data rate is reduced due to required guard regions during subchirp allocation. Consequently, SM-OCDM becomes more suitable for radar-centric applications.

- **Detailed discussions and thorough comparative performance analysis of ISAC systems based on the PMCW, OFDM, C-OCDM, and SM-OCDM modulation schemes.** Based on simulation and measurement results, it was concluded that PMCW and SM-OCDM are better suited for radar-centric ISAC as they present robust radar sensing performance and lower PAPR, while having low-data-rate but robust communications capability. As for OFDM and C-OCDM, they tend to better suit communications-centric ISAC systems. The first due to its high PAPR that may affect radar link budget, and the second additionally due to its increased, data-dependent range sidelobe level.
- **The proposal of a bistatic OFDM-based ISAC system concept and radar signal processing strategies based on either the sole use of pilot subcarriers or transmit frame reconstruction via communications processing.** While the first enables straightforward sensing and simpler processing chain, the latter allows profiting from the full achievable unambiguous range and Doppler shift, besides full radar processing gain.
- **The formulation of a synchronization procedure for bistatic OFDM-based ISAC to allow accurate estimation and compensation of offsets and, consequently, coherent radar signal processing of the receive signal to avoid biases and artifacts in the ultimately obtained radar image.** While STO and CFO estimation and correction approaches rely on existing techniques, adaptations of them to the ISAC context and integration into the full synchronization procedure were made. In addition, a novel SFO estimation algorithm, TITO, was proposed. TITO estimates the SFO by first estimating its induced delay migration estimation. Unlike existing methods in the literature, TITO benefits from a processing gain similar to the radar range processing gain during the aforementioned delay estimation. Consequently, a higher effective SNR is experienced

during SFO estimation, yielding improved estimation accuracy. In addition, TITO avoids bias due to SFO-induced ISI by adaptively choosing the number of OFDM symbols containing pilot subcarriers to be used for delay migration and subsequent SFO estimation. Consequently, TITO allows for SFO estimation with RMSE as low as 0.01 ppm for realistic conditions, while traditional approaches such as the method by Tsai et al. [TKC05] yield RMSEs of 1 ppm or higher.

- **Thorough discussions and numerical analyses of key hardware impairments in bistatic OFDM-based ISAC, which can also be extended to the monostatic case.** In this context, considerations were made on beam squint and element coupling in antenna arrays, besides quantization noise, aliasing, and clipping in DACs and ADCs. In addition, brief numerical analyses were performed for amplifier and mixer non-linear distortion and I/Q imbalance. Notably, extensive analyses were performed to assess the influence of PN at the mixing stage and SJ at the sampling stage in OFDM-based ISAC. It was concluded that beam squint may lead to offsets to the intended the beam direction, potentially resulting in bias during angular estimation and interference among beams in the multicast case. As for quantization noise, its effect can be minimized both by ensuring a sufficient number of bits and relying on the radar processing gain. As for aliasing, it can be avoided with sufficient oversampling and analog filtering, whereas clipping is expected to result in similar effects as amplifier and mixer non-linear distortion. The latter, in turn, result in noise-like interference that can be tolerated for IBOs equal to or higher than the OFDM signal PAPR. Next, it was discussed that although relevant, the effects of realistic I/Q imbalance levels can be fully compensated via signal processing. Regarding PN, it was shown that its induced ICI can be suppressed by the radar processing gain to a certain extent, and that CPE must ideally be estimated and corrected to avoid artifacts along the Doppler shift direction of radar images. Finally, it was shown that colored SJ arising from the use of PLL-based oscillators to derive sampling clocks leads to OOB emission at the DAC, besides ICI. In the specific case of BP sampling, SJ also causes similar effects to PN to the digital IF. In modern DACs ADCs, however, which can ensure femtosecond RMS SJ, these effects become tolerable or negligible.

- **The introduction of a signal processing chain and a distributed synchronization concept for bistatic MIMO-OFDM-based ISAC.** It is assumed that beamformed transmission is performed at the transmitter side, and combination of signals of distinct receive channels is later performed via MRC for communications and digital beamforming for DoA estimation in radar sensing. To ensure the necessary coherence for the aforementioned processing, a strategy to estimate STOs, CFOs and SFOs locally at each receive channel and combine them into a global synchronization offset estimates for all receive channels based on assumptions on the hardware architecture was proposed. Since it is assumed that all transmit channels share the same sampling clocks and LO for the up-conversion, and that a similar setting is observed at the receiver side, the CFO and SFO estimates at different receive channels can be simply averaged to obtain global estimates. As for STO, the earliest estimated OFDM start sample among all receive channels is taken as a global estimate to account for differences in their CIRs and avoid ISI in all channels. A subsequent residual STO correction is then performed to align all channels and ensure the required coherence for DoA estimation in the radar processing chain. Based on the obtained results, it was shown that the aforementioned coherence can be achieved even under realistic constraints such as considerable differences in the CIRs of distinct receive channels and limited residual STO estimation accuracy.
- **The assembly of measurement setups and verification measurements for both SISO and MIMO bistatic OFDM-based ISAC system, signal processing, and synchronization concepts.** For the SISO case, a RIS-assisted bistatic system concept was proposed, and measurements at 26.2 GHz with a bandwidth of 500 MHz were performed. As for the MIMO case, measurements at 27.5 GHz with a bandwidth of 491.52 MHz were performed using ULAs for both transmitter and receiver, with a reference LoS between them. The obtained measurements successfully demonstrated the capabilities of the proposed system concepts and signal processing techniques under realistic conditions.

In spite of the listed contributions and further research efforts in the literature, several open questions remain to be addressed, mainly as to how developed techniques for ISAC should be integrated into radar-centric systems, adding communications capability, e.g., to automotive radar sensors, and to

communications-centric systems, turning cellular networks into PMNs. The contributions of this dissertation combined, however, address and provide solutions for a multitude of relevant challenges, therefore serving as a reference to guide researchers and practitioners in the system-level design and development of signal processing algorithms towards enabling future ISAC systems.

A Appendix

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A.1 Effects of frequency and time shifts in the discrete-Fresnel domain

In this section, the effects of frequency and time shifts on OCDM signals after DFnT transformation into the discrete-Fresnel domain are analyzed. Specifically, Sections A.1.1 and A.1.2 discuss the effects of frequency and time shifts, respectively, when individually applied to OCDM signals. Based on the findings of these two sections, Section A.1.3 discusses the effects of jointly applied frequency and time shifts to OCDM signals. Finally, concluding remarks are presented in Section A.1.4

A.1.1 Frequency shifts

If a frequency shift f_Δ is the only experienced effect during propagation of the transmit OCDM signal through a channel, then the element at the n th row and m th column of the discrete-time domain frame $\mathbf{y}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ obtained after CP removal at the receiver side can be expressed as

$$\begin{aligned} y_{n,m}^{\text{OCDM}} &= x_{n,m}^{\text{OCDM}} e^{j2\pi f_\Delta [m(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}}) + N_{\text{CP}}^{\text{OCDM}} + n] / B^{\text{OCDM}}} \\ &= x_{n,m}^{\text{OCDM}} e^{j2\pi f_\Delta n / B} e^{j2\pi f_\Delta [m(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}}) + N_{\text{CP}}^{\text{OCDM}}] / B^{\text{OCDM}}} . \end{aligned} \quad (\text{A.1})$$

The indices n and m are related to the sampling instant t of $y(t)$ with sampling frequency F_s^{OCDM} after analog conditioning and down-conversion, such that

$t = mT_{\text{block}}^{\text{OCDM}} + (N_{\text{CP}}^{\text{OCDM}} + n)T_s^{\text{OCDM}}$. This holds for the further definitions of $y_{n,m}^{\text{OCDM}}$ in this section and will therefore not be mentioned again for the sake of conciseness.

For the sake of simplicity, the normalized frequency shift $k_{\Delta} \in \mathbb{R}$ and the phase $\phi_m \in \mathbb{R}$ are defined as

$$k_{\Delta} \triangleq f_{\Delta} / \Delta f^{\text{OCDM}} \quad (\text{A.4})$$

and

$$\phi_m \triangleq 2\pi k_{\Delta} \left[m \left(N_{\text{subch}}^{\text{OCDM}} + N_{\text{CP}}^{\text{OCDM}} \right) + N_{\text{CP}}^{\text{OCDM}} \right] / N_{\text{subch}}^{\text{OCDM}}, \quad (\text{A.5})$$

respectively. Next, performing column-wise DFNTs on $\mathbf{y}^{\text{OCDM}}$ yields the matrix $\dot{\mathbf{Y}}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$, whose element at the k th row and m th column is expressed as

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= e^{-j\frac{\pi}{4}} \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \left(x_{n,m}^{\text{OCDM}} e^{j2\pi k_{\Delta} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_m} \right) e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(n-k)^2} \\ &= e^{j\phi_m} e^{-j\frac{\pi}{4}} \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} x_{n,m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(n^2 - 2n(k-k_{\Delta}) + k^2)}. \end{aligned} \quad (\text{A.6})$$

Knowing that

$$n^2 - 2nk + k^2 + 2nk_{\Delta} = [n - (k - k_{\Delta})]^2 + 2kk_{\Delta} - k_{\Delta}^2, \quad (\text{A.7})$$

it is possible to rewrite (A.6) as

$$\dot{Y}_{k,m}^{\text{OCDM}} = e^{j\phi_m} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(2kk_{\Delta} - k_{\Delta}^2)} e^{-j\frac{\pi}{4}} \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} x_{n,m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}[n - (k - k_{\Delta})]^2}. \quad (\text{A.8})$$

Expressing $x_{n,m}^{\text{OCDM}}$ as the IDFnT of $\dot{X}_{k,m}^{\text{OCDM}}$ in (A.8) and rearranging the resulting expression yields

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= \frac{e^{j\phi_m}}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{\kappa,m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(k^2-\kappa^2)} \\ &\quad \times \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}n(-2k+2k_{\Delta}+2\kappa)}, \end{aligned} \quad (\text{A.9})$$

for $\kappa \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} - 1\}$. The rightmost sum in (A.9) is a finite geometric series, which, according to the result from the Appendix, can be evaluated to yield

$$\dot{Y}_{k,m}^{\text{OCDM}} = \frac{e^{j\phi_m}}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{\kappa,m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(k^2-\kappa^2)} \left(\frac{e^{j2\pi(k_{\Delta}-k+\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(k_{\Delta}-k+\kappa)} - 1} \right). \quad (\text{A.10})$$

Since $k \in \mathbb{N}$ and $\kappa \in \mathbb{N}$, it holds for the specific case where $k_{\Delta} \in \mathbb{Z}$, which is when the frequency shift f_{Δ} is an integer multiple of Δf^{OCDM} , that

$$\left. \frac{e^{j2\pi(k_{\Delta}-k+\kappa)} - 1}{e^{j\frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}}(k_{\Delta}-k+\kappa)} - 1} \right|_{k_{\Delta} \in \mathbb{Z}} = \delta \left[\langle \kappa - (k - k_{\Delta}) \rangle_{N_{\text{subch}}^{\text{OCDM}}} \right]. \quad (\text{A.11})$$

The result from (A.11) finally allows rewriting (A.10) as

$$\dot{Y}_{k,m}^{\text{OCDM}} = \frac{e^{j\phi_m}}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{\kappa,m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(k^2-\kappa^2)} \delta \left[\langle \kappa - (k - k_{\Delta}) \rangle_{N_{\text{subch}}^{\text{OCDM}}} \right], \quad (\text{A.12})$$

which after further manipulation becomes

$$\dot{Y}_{k,m}^{\text{OCDM}} = e^{j\phi_m} \dot{X}_{\langle k-k_{\Delta} \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m}^{\text{OCDM}} e^{j\frac{\pi}{N_{\text{subch}}^{\text{OCDM}}}(2kk_{\Delta}-k_{\Delta}^2)}. \quad (\text{A.13})$$

In summary, the results from (A.10) and (A.13) reveal that a frequency shift f_{Δ} during the propagation of the transmit OCDM signal through a channel results

in a circular shift by $k_\Delta = f_\Delta / \Delta f^{\text{OCDM}}$ samples and a multiplication by the complex exponential $e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (2k k_\Delta - k_\Delta^2)}$ that rotates the phases of the elements within the columns of $\mathbf{\dot{Y}}^{\text{OCDM}}$, which represent discrete-Fresnel domain receive OCDM symbols. Additionally, the introduced frequency shift rotates the phase of each m th OCDM symbol with respect to the previous one, which is expressed as the phase ϕ_m of the complex exponential $e^{j\phi_m}$ that multiplies all elements of the m th column of $\mathbf{\dot{Y}}^{\text{OCDM}}$.

A.1.2 Time shifts

Let an OCDM signal be transmitted through a channel with a CIR represented in the discrete-time domain by $\mathbf{h} \in \mathbb{C}^{N_h \times 1} | \mathbf{h} = [h_0, h_1, \dots, h_{N_h-1}]^T$, with $N_h \in \mathbb{N}_{>0} | N_h < N_{\text{CP}}^{\text{OCDM}}$ and $(\cdot)^T$ denoting the transpose operator. In the absence of frequency shifts, the columns of the discrete-time domain receive OCDM frame after CP removal $\mathbf{y}^{\text{OCDM}}$ are the result of the circular convolution of length $N_{\text{subch}}^{\text{OCDM}}$ between the columns of $\mathbf{x}^{\text{OCDM}}$ and $\mathbf{h}$. In other words, the element at the n th row and m th column of $\mathbf{y}^{\text{OCDM}}$ is expressed as $y_{n,m}^{\text{OCDM}} = x_{n,m}^{\text{OCDM}} \circledast_{N_{\text{subch}}^{\text{OCDM}}} h_n$, where $\circledast_{N_{\text{subch}}^{\text{OCDM}}}$ is the $N_{\text{subch}}^{\text{OCDM}}$ -length circular convolution operator. Alternatively, one can write

$$y_{n,m}^{\text{OCDM}} = \sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_\nu^{\text{ZP}} x_{\langle n-\nu \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m}^{\text{OCDM}}, \quad (\text{A.14})$$

for $\nu \in \{0, 1, \dots, N_{\text{subch}}^{\text{OCDM}} - 1\}$. In this equation, h_ν^{ZP} represents the ν th element of the zero-padded version of $\mathbf{h}$ to the length $N_{\text{subch}}^{\text{OCDM}}$, which is denoted by $\mathbf{h}^{\text{ZP}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times 1} | \mathbf{h}^{\text{ZP}} = [h_0^{\text{ZP}}, h_1^{\text{ZP}}, \dots, h_{N_h-1}^{\text{ZP}}, 0, \dots, 0]^T$.

By performing column-wise DFHTs on $\mathbf{y}^{\text{OCDM}}$, the matrix representation $\mathbf{\dot{Y}}^{\text{OCDM}}$ of the discrete-Fresnel domain frame is yielded. Its element at the k th row and m th column is expressed as

$$\dot{Y}_{k,m}^{\text{OCDM}} = e^{-j\frac{\pi}{4}} \sum_{n=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \left(\sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_\nu^{\text{ZP}} x_{\langle n-\nu \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m}^{\text{OCDM}} \right) e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (n-k)^2}. \quad (\text{A.15})$$

Substituting $l = \langle n - \nu \rangle_{N_{\text{subch}}^{\text{OCDM}}}$ in (A.15) and rearranging the resulting expression yields

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= \sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (\nu^2 - 2\nu k)} e^{-j \frac{\pi}{4}} \\ &\quad \times \sum_{l=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \left(x_{l,m}^{\text{OCDM}} e^{j 2\pi \nu l / N} \right) e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (l-k)^2}. \end{aligned} \quad (\text{A.16})$$

Based on the obtained results in Section A.1.1, more specifically in (A.10), (A.16) can be rewritten as

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= \sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (\nu^2 - 2\nu k)} \\ &\quad \times \left[\frac{1}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{\kappa,m}^{\text{OCDM}} e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (k^2 - \kappa^2)} \right. \\ &\quad \left. \times \left(\frac{e^{j 2\pi (\nu - k + \kappa)} - 1}{e^{j \frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}} (\nu - k + \kappa)} - 1} \right) \right]. \end{aligned} \quad (\text{A.17})$$

Since $k \in \mathbb{N}$, $\kappa \in \mathbb{N}$, and $\nu \in \mathbb{N}$, the same principle from (A.11) can be applied to the rightmost term in (A.17), which then becomes

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= \sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (\nu^2 - 2\nu k)} \\ &\quad \times \left(\frac{1}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \dot{X}_{\kappa,m}^{\text{OCDM}} e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (k^2 - \kappa^2)} \right. \\ &\quad \left. \times \delta \left(\langle \kappa - (k - \nu) \rangle_{N_{\text{subch}}^{\text{OCDM}}} \right) \right). \end{aligned} \quad (\text{A.18})$$

After further manipulation, (A.18) can be finally rewritten as

$$\dot{\mathbf{y}}_{k,m}^{\text{OCDM}} = \sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} \dot{X}_{\langle k-\nu \rangle_{N_{\text{subch}}^{\text{OCDM}},m}}^{\text{OCDM}}. \quad (\text{A.19})$$

The relationship between $\dot{\mathbf{y}}_{k,m}^{\text{OCDM}}$ and $\dot{\mathbf{x}}_{k,m}^{\text{OCDM}}$ as in (A.19) reveals that the same circular shifts and amplitude weightings suffered by the columns of $\mathbf{x}^{\text{OCDM}}$ are observed in the columns of both $\mathbf{y}^{\text{OCDM}}$ and its discrete-Fresnel domain representation $\dot{\mathbf{Y}}^{\text{OCDM}}$. In other words, $y_{n,m}^{\text{OCDM}} = x_{n,m}^{\text{OCDM}} \otimes_{N_{\text{subch}}^{\text{OCDM}}} h_n$ results in $\dot{\mathbf{y}}_{k,m}^{\text{OCDM}} = \dot{\mathbf{x}}_{k,m}^{\text{OCDM}} \otimes_{N_{\text{subch}}^{\text{OCDM}}} h_k$, which is also known as the convolution theorem of the DFNT [OAG⁺15, OZ16a].

A.1.3 Simultaneous time and frequency shifts

Should the propagation of the OCDM transmit signal through a channel results not only in time shifts represented by the CIR $\mathbf{h}$, but also in a normalized frequency shift k_{Δ} , the element at the n th row and m th column of the discrete-time domain receive frame $\mathbf{y}^{\text{OCDM}}$ obtained after CP removal at the receiver side can be expressed as

$$y_{n,m}^{\text{OCDM}} = \left(x_{n,m}^{\text{OCDM}} \otimes_{N_{\text{subch}}^{\text{OCDM}}} h_n \right) e^{j2\pi k_{\Delta} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_m}. \quad (\text{A.20})$$

For the sake of simplicity, the auxiliary matrix $\mathbf{r}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ is defined, being the element $r_{n,m}^{\text{OCDM}} \in \mathbb{C}$ located at its n th row and m th column given by

$$r_{n,m}^{\text{OCDM}} \triangleq x_{n,m}^{\text{OCDM}} \otimes_{N_{\text{subch}}^{\text{OCDM}}} h_n. \quad (\text{A.21})$$

Consequently, (A.20) can be alternatively expressed as

$$y_{n,m}^{\text{OCDM}} = r_{n,m}^{\text{OCDM}} e^{j2\pi k_{\Delta} n / N_{\text{subch}}^{\text{OCDM}}} e^{j\phi_m}. \quad (\text{A.22})$$

Defining $\dot{\mathbf{R}}^{\text{OCDM}} \in \mathbb{C}^{N_{\text{subch}}^{\text{OCDM}} \times M^{\text{OCDM}}}$ as the output of column-wise DFNTs on $\mathbf{r}^{\text{OCDM}}$, the relationship between the element $\dot{R}_{k,m}^{\text{OCDM}} \in \mathbb{C}$ at the k th row and m th column of $\dot{\mathbf{R}}^{\text{OCDM}}$ and the element $\dot{X}_{k,m}^{\text{OCDM}}$ at the same position of $\dot{\mathbf{X}}^{\text{OCDM}}$

can be defined based on the result from (A.18). Additionally, the element $\dot{Y}_{k,m}^{\text{OCDM}}$ at the k th row and m th column of $\dot{\mathbf{Y}}^{\text{OCDM}}$ can be derived from $\dot{R}_{k,m}^{\text{OCDM}}$ based on the result from (A.10). Consequently, $\dot{Y}_{k,m}^{\text{OCDM}}$ can be expressed as a function of $\dot{X}_{k,m}^{\text{OCDM}}$ as

$$\begin{aligned} \dot{Y}_{k,m}^{\text{OCDM}} &= \frac{e^{j\phi_m}}{N_{\text{subch}}^{\text{OCDM}}} \sum_{\kappa=0}^{N_{\text{subch}}^{\text{OCDM}}-1} \left(\sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} \dot{X}_{\langle \kappa-\nu \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m}^{\text{OCDM}} \right) e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (k^2 - \kappa^2)} \\ &\quad \times \left(\frac{e^{j2\pi(k_{\Delta} - k + \kappa)} - 1}{e^{j \frac{2\pi}{N_{\text{subch}}^{\text{OCDM}}} (k_{\Delta} - k + \kappa)} - 1} \right). \end{aligned} \quad (\text{A.23})$$

In the specific case where $k_{\Delta} \in \mathbb{Z}$, the property used in (A.11) can be finally applied to (A.23), yielding

$$\dot{Y}_{k,m}^{\text{OCDM}} = e^{j\phi_m} \left(\sum_{\nu=0}^{N_{\text{subch}}^{\text{OCDM}}-1} h_{\nu}^{\text{ZP}} \dot{X}_{\langle (k-k_{\Delta})-\nu \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m}^{\text{OCDM}} \right) e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (2kk_{\Delta} - k_{\Delta}^2)}, \quad (\text{A.24})$$

$$\text{i.e., } \dot{Y}_{k,m}^{\text{OCDM}} = e^{j\phi_m} (\dot{X}_{\langle k-k_{\Delta} \rangle_{N_{\text{subch}}^{\text{OCDM}}}, m} \otimes_{N_{\text{subch}}^{\text{OCDM}}} h_k) e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (2kk_{\Delta} - k_{\Delta}^2)}.$$

The results from (A.23) and (A.24) indicate that the convolution with the CIR followed by a frequency shift f_{Δ} during the propagation of the transmit OCDM signal through a channel results in a circular shift by $k_{\Delta} = f_{\Delta}/\Delta f^{\text{OCDM}}$ samples on top of the circular shifts caused by the convolution with the multiple taps of the discrete-time domain CIR representation $\mathbf{h}$. This effect is similar to known range- or delay-Doppler coupling in chirp-based radar systems [Win07]. In addition, a multiplication by the complex exponential $e^{j \frac{\pi}{N_{\text{subch}}^{\text{OCDM}}} (2kk_{\Delta} - k_{\Delta}^2)}$ of the shifted versions of $\dot{\mathbf{X}}^{\text{OCDM}}$ takes place as observed in Section A.1.1, increasingly rotating the phases of the $N_{\text{subch}}^{\text{OCDM}}$ elements of each column of $\dot{\mathbf{Y}}^{\text{OCDM}}$. As in the case of Section A.1.1, every m th column of $\dot{\mathbf{Y}}^{\text{OCDM}}$ is additionally multiplied by the complex exponential $e^{j\phi_m}$, which rotates the phase of all elements of the m th discrete-Fresnel domain OCDM symbol. The described effects ultimately distort the transmit subchirps and cause interchirp interference (IChI) in the discrete-Fresnel domain, which is somewhat similar to the ICI in the discrete-frequency domain in OFDM-based systems.

A.1.4 Concluding remarks

This section discussed the effects of time and frequency shifts on OCDM symbols in the discrete-Fresnel domain, both when experienced individually or jointly. The achieved results showed that frequency shifts result in both a shift and subchirp-dependent phase rotation in the discrete-Fresnel domain, while time shifts simply result in a shift in the discrete-Fresnel domain, which is due to the fact that the DF_nT itself is a convolution as described in Section 2.4.1. When time and frequency shifts are simultaneously experienced, both aforementioned effects are combined, and a delay-Doppler coupling is experienced due to the fact that both effects result in shifts in the discrete-Fresnel domain.

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