

**ALGORITHMS FOR INTERFERENCE IMMUNITY AND EFFICIENT
RADIO RESOURCE UTILIZATION IN WIRELESS COMMUNICATIONS
SYSTEMS**

A DISSERTATION SUBMITTED TO
THE GRADUATE SCHOOL OF
ENGINEERING AND NATURAL SCIENCES
OF ISTANBUL MEDIPOL UNIVERSITY
IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR
THE DEGREE OF
DOCTOR OF PHILOSOPHY
IN
ELECTRICAL, ELECTRONICS ENGINEERING AND CYBER SYSTEMS

By
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July, 2023

ALGORITHMS FOR INTERFERENCE IMMUNITY AND EFFICIENT RADIO
RESOURCE UTILIZATION IN WIRELESS COMMUNICATIONS SYSTEMS

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July 2023

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in scope and in quality, as a dissertation for the degree of Doctor of Philosophy.

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ACKNOWLEDGEMENT

In the name of Allah (God), the Gracious, the Merciful.

I would like to express my deepest gratitude to my esteemed Ph.D. supervisor, Prof. Hüseyin Arslan, for his exceptional support and invaluable contributions. Without his unwavering patience and countless hours dedicated to guiding me, this thesis would never have come to fruition. His vast knowledge, extensive experience, continuous encouragement, and devoted guidance have been the driving force behind my passion for research and my aspirations for a successful academic career in the future. It has been an honor to be mentored by him, and I am grateful for all he has done for me. Sincerely, I extend my humble appreciation to the esteemed members of my Ph.D. defense committee: Prof. Dr. Ali Emre Pusane, Prof. Dr. Selim Akyokuş, Assoc. Prof. Dr. Ertuğrul Başar, and Prof. Dr. Mehmet Kemal Özdemir. Their constructive comments and invaluable suggestions on my research and thesis have been instrumental in shaping the quality and depth of my work. Furthermore, I would like to thank Prof. Dr. Yasemin Yüksel Durmaz and the administrative staff of the School of Engineering and Natural Sciences at Istanbul Medipol University, with special recognition for Mr. Barış Yasin Yoldaş, Mr. Burak Gokay. Their unwavering efforts have greatly facilitated and eased the entire process for me, making this journey much smoother and more manageable.

Being a member of the Communications, Signal Processing, and Networking Center (CoSiNC) has been a delightful and honorable experience. My heartfelt gratitude extends to all my friends in CoSiNC family for their limitless support, both as friends and colleagues. The experiences and knowledge gained within the CoSiNC group have shaped me as a researcher and individual, leaving a profound impact on my academic journey. Now, as I step into this new professional endeavor, I carry with me the lessons learned and the cherished memories from my time with the CoSiNC group. I want to express my special thanks to Ahmet Yazar, Abu Bakari Kihero, Ebubekir Memisoğlu, Halise Türkmen, and Muhammad Sohaib Solaija for their friendship, sincerity, and unforgettable coffee times. Furthermore, I want to thank Abdyl Nëntor Musliu, Muhamed Halid Eryılmaz, Riad Domazeti, and Olti Guni for taking me out of the routine and offering memorable times. Dear fellows, your unhesitating support and friendship have made my time studying and living in the majestic city of Istanbul, Turkey, truly wonderful. I am eager to explore uncharted territories, expand my horizons, and I am enthusiastic about the possibilities that lie ahead, ready to embrace the challenges and achievements that await me in this exciting new phase of my career.

Throughout my life, I have held deep gratitude for my parents, Fahri Tusha and Bukuriye Tusha, my sister Aurela Metvelaj and her husband Gentian Metvelaj, my sister Almeda Tusha, and my grandmother Zyke Tusha for always being by my side. Their unwavering presence and unconditional support have been instrumental in helping me navigate through these challenging years and complete my studies successfully. Furthermore, I extend my heartfelt thanks to my parents-in-law, Recep Doğan and Zubeyde Doğan, my brother-in-law Ertuğrul H. Doğan, sister-in-law Nida Çiftçi and her husband Mehmet A. Çiftçi, for their support, help, and prayers throughout this journey. To my beloved family, your presence has been a constant source of comfort and strength for me, lasting for a lifetime.

Finally, I want to extend my warm gratitude to my dear wife, Seda Tusha, for her selfless care and endless support. She has always been there to comfort me during moments of helplessness, to bring a smile to my face when I feel down, and to offer immense help

and encourage during the ups and downs of this long and challenging journey. Thank you for being not just my best friend and colleague, but also my life companion, enriching my life throughout the years.

Armed Tusha

July, 2023



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LIST OF SYMBOLS

λ	: wavelength
ζ	: symbol span
η	: spectral efficiency
η_f	: frequency domain efficiency
η_t	: time domain efficiency
θ	: phase imbalance
ϵ	: gain imbalance
h	: channel coefficient
Δh	: error of channel coefficient
$\mathbf{H}$	: channel matrix
$\mathbf{H}_e$	: channel error matrix
$\mathbf{H}_{\text{eff}}$	: effective channel matrix
$\mathbf{H}_i$	: integer delay and Doppler channel matrix
$\mathbf{H}_w$	: windowed channel matrix
l_i	: delay shift of i-th channel tap
k_i	: Doppler shift of i-th channel tap
Δf	: frequency subcarrier spacing
σ	: noise standard deviation
n	: AWGN
N_0	: noise power density
$(\cdot)^H$	: Hermitian (conjugate transpose)
$(\cdot)^T$	: transpose
$(\cdot)^*$	: conjugate
$\text{tr}(\cdot)$	: trace of matrix $\mathbf{A}$
$ \cdot $	: absolute value
$\mathbf{I}_N$	: $N \times N$ identity matrix
κ	: fractional Doppler shift index
L	: number of channel taps
β	: power allocation coefficient
B	: bandwidth
$M \Delta f$	: bandwidth of OTFS frame
T_f	: time duration of OTFS frame
R	: data rate
P_e	: probability of error
$\text{Pr}\{\cdot\}$	: pairwise error probability
$\mathbf{F}_N$	: N-point DFT
γ	: signal-to-interference and noise ratio
$v_{\max}$	: maximum speed

ABBREVIATIONS

3GPP	: 3rd Generation Partnership Project
4G	: fourth generation
5G	: fifth generation
AWGN	: additive white Gaussian noise
BER	: bit error rate
CDM	: code division multiplexing
CDMA	: code division multiple accessing
CFO	: carrier frequency offset
CP	: cyclic prefix
CSI	: channel state information
DCT	: direct conversion transmitter
DFT	: discrete Fourier transform
DNN	: deep neural network
EDGE	: enhanced data rates for global evolution
eMBB	: enhanced mobile broadband
FDMA	: frequency division multiple accessing
FFT	: fast Fourier transform
FIR	: finite impulse response
GFDM	: generalized frequency division multiplexing
GPRS	: general packet radio service
HSPA	: high speed packet access
ICI	: inter-carrier interference
IDI	: inter-Doppler interference
IDFT	: inverse discrete Fourier transform
IM	: index modulation
i.i.d.	: independent and identically distributed
IFFT	: inverse fast Fourier transform
IQ	: in-phase and quadrature
ISFT	: inverse symplectic Fourier transform
ISI	: inter-symbol interference
LO	: local oscillator
LTE	: Long-Term Evolution
mMTC	: massive machine-type communications
mmWave	: millimeter wave
MRC	: maximum ratio combining
MMSE	: minimum mean square error
MDI	: mirror Doppler interference
MIMO	: multiple-input multiple-output
NR	: new radio
NOMA	: non-orthogonal multiple accessing
NMSE	: normalized mean square error
OFDM	: orthogonal frequency division multiplexing
OFDM-IM	: orthogonal frequency division multiplexing with index modulation
OFDM-GIM	: orthogonal frequency division multiplexing with generalized IM
OFDM-SNM	: OFDM with subcarrier number modulation
OTFS	: orthogonal time-frequency space
OOBE	: out-of-band emission
PN	: phase noise

SC	: single carrier
SHT	: super-heterodyne transmitter
SIC	: successive interference cancellation
SINR	: signal-to-interference and noise ratio
SNR	: signal-to-noise ratio
PAPR	: peak-to-average power ratio
PDF	: probability distribution function
PHY	: physical
PSD	: power spectral density
PSK	: phase shift keying
QPSK	: Quadrature Phase Shift Keying
QAM	: quadrature amplitude modulation
RAT	: radio access technologies
RB	: resource block
RF	: radio frequency
RTT	: round trip time
Rx	: receiver
SINR	: signal-to-interference-plus-noise ratio
SISO	: single-input single-output
SNR	: signal-to-noise ratio
TDD	: time-division duplexing
TDM	: time-division multiplexing
TDMA	: time division multiple accessing
Tx	: transmitter
UE	: user equipment
uRLLC	: ultra-reliable low latency communication
WLAN	: wireless local area network
ZF	: zero-forcing

KABLOSUZ İLETİŞİM SİSTEMLERİNDE GİRİŞİM DİRENCİ VE VERİMLİ RADYO KAYNAĞI KULLANIMI İÇİN ALGORİTMALAR

ÖZET

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Temmuz, 2023

Girişim, kablosuz iletişim için en yaygın engellerden birini temsil eder. Yeni nesil haberleşme sistemleri, sistem karmaşıklığını ve gecikmesini kontrol ederek yüksek veri hızı, yüksek haberleşme güvenilirliği ile beraber herkesin ve her şeyin sürekli olarak bağlı olduğu bir haberleşme ortamı hedefler. Bu ortamda, girişim, kablosuz iletişim sistemlerinin ilkel doğasını ortaya çıkarır; iletişim bağlantısını bozan istenmeyen fiziksel bir sinyal her zaman mevcuttur. Bu durum, iletim sinyalinin fiziksel katman mimarisinden, kablosuz kanal ile etkileşiminden, kullanılan radyo kaynaklarından (RR) ve özellikle verici/alıcı mimarisinden kaynaklanır.

Geniş uygulama ve kullanım durumları arasında, ultra-güvenilir ve düşük gecikmeli (URLLC) olarak nitelendirilen talepler, katı gereksinimlere sahiptir. Bu çalışmada, tek taşıyıcılı (SC) iletim, URLLC sistemlerin gereksinimlerini karşılamak için umut vadeden bir çözüm olarak ele alınmıştır. İlk olarak, güvenilirlik ve gecikme üzerinde etkili olan RR-tabanlı parametreler ayrıntılı bir şekilde açıklanmıştır. Ek olarak, bu parametreler zaman ve frekans alanında optimize edilerek URLLC sistemlerinin gereksinimlerine ulaşılması amaçlanmıştır. Daha sonra, SC blok içinde adaptif sönmüleme faktör kullanımı, girişim düzeyini kontrol ederek daha verimli kaynak kullanımı elde etmek için önerilmiştir. Optimize edilmiş parametreler ile SC sistemlerin önemli gecikme ve güvenilirlik iyileştirmesi elde edilebileceği gösterilmiştir.

İndeks modülasyon (IM) ile çok taşıyıcılı iletim, yüksek veri hızları, yüksek güvenilirlik ve düşük gecikme ile uygulamaları karşılamak için esnek yapıları sayesinde umut verici bir aday olarak kabul edilmiştir. Genel olarak, mevcut IM şemaları yalnızca ideal iletişim senaryoları için değerlendirilir. Bu çalışmada, üç frekans alanı IM şemasının hata performansı araştırılmış ve taşıyıcı frekans ofseti (CFO) ve faz içi ve dördün (IQ) dengesizlik girişimi varlığında karşılaştırılmıştır. IM ile ortogonal frekans bölme (OFDM-IM), genelleştirilmiş IM ile OFDM (OFDM-GIM) ve alt taşıyıcı sayı modülasyonu ile OFDM (OFDM-SNM) değerlendirilir. OFDM-GIM ve OFDM-SNM'in OFDM-IM'e göre daha yüksek spektral verim sağlamasına rağmen, RF bozulmalarına karşı daha duyarlı oldukları bilgisayar tabanlı benzetimlerle gösterilmiştir. Bunlar arasında OFDM-IM, CFO ve IQ dengesizliği altında en iyi hata performansını sağlar.

Ortogonal frekans bölmeli çoğullama (OFDM) ve OFDM-IM teknolojilerinin üst üste binmesi yoluyla hibrit güç alanı non-ortogonal çoklu erişim (NOMA) şeması olan IM-NOMA sunulmuştur. Bilgisayar tabanlı simülasyonlar ve matematiksel analizler ile gösterilmiştir ki, IM-NOMA, toplam güç kısıtlaması altında bit hata oranı (BER) ve ulaşılabilir toplam veri oranı açısından klasik OFDM-NOMA'yı geride bırakmaktadır. IM-NOMA'nın sistem performansı, sadece üst üste binen müdahale eden kullanıcılar arasındaki güç farkına bağlı değil, aynı zamanda OFDM-IM sinyalinin RR kullanımının özelliklerine de bağlıdır. Bu nedenle, bu şema, birbirine benzer güç atamalarında olası kritik hat Ortogonal zaman frekans uzayı (OTFS) modülasyonu, güçlü gecikme-Doppler dayanıklılığı sayesinde bağlantı güvenilirliği konusunda önemli avantajlar sunan yeni bir teknolojidir. Klasik hücreli teknolojiye, genel sistem performansı, belirli bir kullanıcı ekipmanının (UE) radyo kaynakları üzerinde deneyimlediği zayıflama ve girişimle sınırlıdır. Ayrıca, OTFS performansı, UE'lerin kanal çeşitliliğine bağlıdır. Bu nedenle, bu çalışma pratik senaryoları göz önünde bulundurarak OTFS sistemleri için çoklu kullanıcı çeşitliliğinin potansiyelini inceler. Özellikle, mevcut OTFS çerçevesinin RR'nın en fazla sayıda kanal dokunuşuna sahip UE'ye atanmasını öneren yeni bir zamanlama algoritması tanıtılmaktadır, bu da artan kanal çeşitliliğine yol açar. Hem matematiksel analiz hem de simülasyon sonuçları aracılığıyla, önerilen zamanlama tekniğinin, maksimum olasılık (ML) alıcı ile sistem BER açısından gecikme-Doppler RR'a rastgele erişim şemasını geride bıraktığı gösterilmektedir. BER sonuçları, çeşitli sistem yapılandırmalarını dikkate alan MMSE (minimum ortalama kare hatası) dedektörü ile OTFS iletimi için önerilen tekniklerin doğruluğunu doğrulamaktadır. a performansına karşı dayanıklıdır.

Ortogonal zaman frekans uzayı (OTFS) modülasyonu, güçlü gecikme-Doppler dayanıklılığı sayesinde bağlantı güvenilirliği konusunda önemli avantajlar sunan yeni bir teknolojidir. Klasik hücreli teknolojiye, genel sistem performansı, belirli bir kullanıcı ekipmanının (UE) radyo kaynakları üzerinde deneyimlediği sönümleme ve girişimle sınırlıdır. Ek olarak, OTFS performansı, UE'lerin kanal çeşitliliğine bağlıdır. Bu nedenle, bu çalışma pratik senaryoları göz önünde bulundurarak OTFS sistemleri için çoklu kullanıcı çeşitliliğinin potansiyelini inceler. Özellikle, mevcut OTFS çerçevesinin RR'nın en fazla sayıda kanal dokunuşuna sahip UE'ye atanmasını öneren yeni bir zamanlama algoritması tanıtılmaktadır, bu da artan kanal çeşitliliğine yol açar. Hem matematiksel analiz hem de simülasyon sonuçları aracılığıyla, önerilen zamanlama tekniğinin, maksimum olasılık (ML) alıcı ile sistem BER açısından gecikme-Doppler RR'a rastgele erişim şemasını geride bıraktığı gösterilmektedir. BER sonuçları, çeşitli sistem yapılandırmalarını dikkate alan minimum ortalama kare hatası (MMSE) dedektörü ile OTFS iletimi için önerilen tekniklerin doğruluğunu kanıtlar.

Bu tezde, IQ dengesizliğinden kaynaklanan girişimin OTFS tabanlı iletimin performansı üzerindeki etkisi araştırılmıştır. OTFS, zamanla değişen kablolu kanalın zaman ve frekans seçiciliğine karşı umut verici bir dayanıklılığa sahip yeni bir iki boyutlu iletim şemasıdır ve IQ bozulması, verici/alıcı tasarımındaki evre uyumsuzluklarından kaynaklanır. Bu çalışmada, OTFS'de evre uyumsuzluğuna sahip vericinin fiziksel etkisini göstermek için matematiksel bir model sunulmuştur, bu da sadece Doppler ekseninde meydana gelen yeni bir girişim türü olan ayna-Doppler girişimi (MDI) ile sonuçlanır. Ayrıca, sabit bir zarfa sahip M-ary sembolünü ve ML alıcı dikkate alarak OTFS'in ortalama BER'i için kapalı formda matematiksel ifadesi türetilmiştir. Önemli bir şekilde, bulgularımız, IQ dengesiz vericinin, OTFS iletiminin çeşitlilik kazancında önemli bir azalmaya neden olduğunu, ancak bir boyutlu mevcut iletişim sistemlerindeki BER performansında bir hata düzeyine neden olmadığını kanıtlamaktadır. Monte Carlo tabanlı

simülasyonlar, sabit zarf M-ary sembolleri altında farklı sistem yapılandırmaları için elde edilen matematiksel formüllerin doğruluğunu kanıtlamak için sunulmuştur.

5G ve ötesi (B5G) kablosuz sistemlerde düşük hesaplama karmaşıklığı, düşük güç tüketimi ve uygun maliyet sağlamak için basit verici/alıcı mimarilerinin tasarımı kaçınılmazdır, ancak bu, iletimin performans güvenilirliğini önemli ölçüde düşüren donanım bozulmalarına yol açar. Bu nedenle, bu istenmeyen donanım bozulmalarının etkisini telafi etmek için ek donanım veya sinyal işleme gereklidir, bu da spektral verimliliği azaltırken enerji tüketimini ve işleme süresini artırır. Gecikme-Doppler alanında IQ dengesizliğinin etkisini azaltmak için, sistem tasarımında çekici bir RR (radyo kaynakları) esnekliği sağlayan IM (indeks modülasyon) kavramının yardımıyla enerji ve spektral verimli bir IQ dengesizliği telafi şeması öneriyoruz. Klasik kablosuz teknolojilerde kullanılan iteratif ve pilot tabanlı çözümlerin aksine, önerilen şema, IQ dengesizliği etkisinin tahmin ve telafisi sırasında ek enerji tüketimini ve önemli spektral verim kaybını önler. Elde edilen BER sonuçları, pratik senaryolarda mükemmel/kusurlu kanal durumu bilgisini dikkate alan farklı OTFS sistem yapılandırmaları için önerilen telafi edici mekanizmanın doğruluğunu doğrulamaktadır.

OTFS, çift yönlü dağılımlı kablosuz ortamın hem zaman hem de frekans seçiciliğinden faydalanır, bu, klasik kablosuz iletişim teknolojilerinde sınırlayıcı faktörlerden biri olarak kabul edilir. Öte yandan, OTFS tabanlı kablosuz sistemlerin performansı, gecikme-Doppler düzleminde her zaman tamsayı Doppler değerleri olmamasının kaçınılmaz ve doğal özelliklerinden kaynaklanan kesirli Doppler nedeniyle ciddi şekilde etkilenir. Bu durum, ciddi aralık-Doppler girişimi (IDI) ve gecikme-Doppler alanında düşük bir kanal seyrekliği ile sonuçlanır. Bu bağlamda, bu çalışma, OTFS tabanlı iletimin hata performansı üzerinde kesirli Doppler'in etkisine karşı mücadele etmek için transceiver'da pratik darbe şekillendirme filtreleri ile birlikte yeni bir alıcı pencereleme tekniği sunmaktadır. Özellikle, IDI'nin etkisini azaltmak için alıcıda global çevrim önekli (CP) tabanlı yükselen kosinüs (RC) pencereleme önerilmiştir. Bu amaçla, önerilen şema, IDI düzeyini ve etkili kanalın seyrekliğini tanımlayan DD alanındaki OTFS sinyalinin darbe şeklini değiştirir. Ayrıca, önerilen alıcı pencerelemeli OTFS (RW-OTFS) şemasının çeşitli pratik iletişim senaryolarında performansını değerlendiriyoruz. Elde edilen sayısal ve simülasyon sonuçları, önerilen IDI önleyicinin yüksek hareketlilik senaryolarında OTFS sistemleri için doğruluğunu kanıtlar.

Anahtar sözcükler: Girişim, darbe şekillendirme, kanal bozulmaları, RF bozulmaları, ortak kaynak kullanımı, SC, OFDM, IM, OTFS, NOMA.

ALGORITHMS FOR INTERFERENCE IMMUNITY AND EFFICIENT RADIO RESOURCE UTILIZATION IN WIRELESS COMMUNICATIONS SYSTEMS

ABSTRACT

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July, 2023

Interference represents one of the most common barriers for the wireless communications society to bring the fully connected world to life, where everybody and everything is connected at any time, aiming to support a wide range of services and applications with increasing demand in terms of data rate with a higher degree of reliability and security, while keeping an affordable overall system capacity, complexity, and latency. Essentially, interference clearly explains the primitive nature of the wireless communications systems, where there is always an unwanted physical signal that disrupts the communication link, occurring from the physical layer architecture of transmission signal, its interaction with the wireless channel, utilized radio resources (RR) and transceiver architecture in particular.

Among the wide set of applications and use-cases, demands characterized by ultra-reliable and low-latency (URLLC) have stringent requirements. In this work, single carrier (SC) transmission is considered as a promising solution to meet demands of systems with URLLC. Firstly, RR-based parameters affecting reliability and latency are comprehensively explained. Additionally, these parameters are optimized in time and frequency domain to achieve URLLC. Later, adaptive roll-off factor usage within a SC block is proposed to provide more efficient resource utilization under a fixed intended interference level. It is shown via computer simulations that significant latency and reliability improvement can be achieved for SC systems with the aid of optimized parameters.

Multicarrier transmission through index modulation (IM) has been considered as a promising candidate owing to its flexible structures in order to satisfy the applications with high data rates, high reliability, and low latency. In general, existing IM schemes are only evaluated for ideal communication scenarios. In this work, the error performance of three frequency domain IM schemes is investigated and compared in the presence of carrier frequency offset (CFO) and in-phase and quadrature (IQ) imbalance interference. Orthogonal frequency division multiplexing with IM (OFDM-IM), OFDM with generalized IM (OFDM-GIM), and OFDM with subcarrier number modulation (OFDM-SNM) are assessed. It is shown via computer-based simulations that although OFDM-GIM and OFDM-SNM provide higher spectral efficiency compared with OFDM-IM, they are more sensitive against RF impairments. Among them, OFDM-IM provides the best error performance under CFO and IQ imbalance.

A hybrid power domain non-orthogonal multiple accessing (NOMA) scheme by the superposition of OFDM and index modulated OFDM (OFDM-IM) technologies is presented and named IM-NOMA. It is shown via both computer-based simulations and mathematical analysis that IM-NOMA outperforms the classical OFDM-NOMA in terms of bit error rate (BER) under a total power constraint and achievable sum rate. The system performance of IM-NOMA not only depends on the power difference between the overlapped interfering users but also on the features of RR utilization for the OFDM-IM signal. Hence, this scheme is robust against possible catastrophic error performance in case similar power is assigned to the coexisting users.

Orthogonal time frequency space (OTFS) modulation is a recent technology that offers a significant advantage on link reliability due to its strong delay-Doppler resilience. In classical cellular technology, overall system performance is limited by the fading and interference experienced by particular user equipment (UE) over its radio resources RR. Besides, OTFS performance depends on the channel diversity of the UEs. Therefore, this work examines the potential of multi-user diversity for OTFS systems considering practical scenarios. Specifically, we introduce a novel scheduling algorithm that proposes the assignment of the available RR of OTFS frame to the UE with the largest number of channel taps, resulting in enhanced channel diversity. It is shown via both mathematical analysis and simulation results that the proposed scheduling technique outperforms the scheme with random access to delay-Doppler RR in terms of system BER with maximum likelihood (ML) receiver. The BER results validate the accuracy of the proposed technique for OTFS transmission with minimum mean square error (MMSE) detector considering various system configurations.

The influence of interference due to IQ imbalance on the performance of OTFS-based transmission is investigated in this thesis. OTFS is a recent two-dimensional transmission scheme with promising robustness against the time and frequency selectivity of the time-varying wireless channel, and IQ impairment arises from inevitable imperfections between in-phase and quadrature branches in the transceiver design. In this work, the mathematical model is presented to illustrate the physical effect of the IQ mismatched transmitter in OTFS, which results in a new type of interference named mirror-Doppler interference (MDI), occurring only along the Doppler axis. Moreover, a closed-form expression for the average BER of OTFS with IQ imbalanced transmitter is derived taking into account M-ary symbol with a constant envelope and maximum likelihood (ML) detection principle. Importantly, our findings prove that IQ imbalanced transmitter causes a significant reduction in the diversity gain of OTFS transmission, but does not lead to an error floor in the BER performance as in one-dimensional current communication systems, i.e., OFDM-based techniques. Monte-Carlo-based simulations are provided to validate the derived mathematical formulas for OTFS transmission with constant envelope M-ary symbols under different system configurations.

Design of simple transceiver architectures is inevitable in order to provide low computational complexity, low power consumption and affordable cost in beyond 5G (B5G) wireless systems, but it results in hardware impairments that significantly degrade the performance reliability of transmission. Therefore, extra hardware or signal processing is needed to compensate for the effect of these undesired hardware impairments at the cost of reducing spectral efficiency, increasing energy consumption and processing time. To mitigate the effect of IQ imbalance in delay-Doppler domain, we propose an energy and spectral efficient IQ imbalance compensation scheme with the aid of IM concept, which provides an attractive RR flexibility in the system design. In

contrast to conventional solutions used in classical wireless technologies, such as iterative and pilot-based techniques, the proposed scheme avoids additional energy consumption and significant spectral efficient loss during the estimation and compensation of the IQ imbalance effect. The obtained BER results validate the accuracy of the proposed compensator for different OTFS system configurations considering perfect/imperfect channel state information in practical scenarios.

On the one hand, OTFS takes advantage of both time and frequency selectivity of the doubly dispersive wireless environment, which has been considered as one of the limiting factors in classical wireless communication technologies. On the other hand, the performance of OTFS-based wireless systems is severely affected by fractional Doppler that arises due to inevitable and inherent characteristics of the wireless channel taps of not being always integer Doppler values in the delay-Doppler plane. Thus, this results in severe inter-Doppler interference (IDI) and a low channel sparsity in delay-Doppler domain. To this end, this work introduces a novel receiver windowing technique to combat the impact of fractional Doppler on error performance of the OTFS-based transmission with practical pulse-shaping filters at the transceiver. Specifically, we introduce a global cyclic prefix (CP) based raised cosine (RC) windowing at Rx in order to alleviate the effect of IDI on the received signal. To do so, the proposed scheme changes the pulse shape of the OTFS signal in DD domain that defines the IDI level and sparsity of the effective channel. Furthermore, we assess the performance of the proposed Rx windowed OTFS (RW-OTFS) scheme under various practical communication scenarios. The obtained numerical and simulation results validate the accuracy of the proposed IDI mitigator for OTFS systems in practical high-mobility scenarios.

Keywords: Interference, pulse shaping, channel impairments, RF impairments, coexistence, SC, OFDM, IM, OTFS, NOMA.

CHAPTER 1

1. INTRODUCTION

Over the past decade, the world of wireless technology has been experiencing revolutionary advances mainly with the emerge of a broad range of applications, which will continue increasing at a rapid pace in the fifth generation (5G) and beyond (5G-B) [1]–[3]. As a result, substantial effort by academia and industry is imperative to pave the way for a better future for our world.

1.1. Background

Early generations of wireless systems have been designed considering the needs of human users where system throughput has been the primary concern [4], [5]. Besides human-centric networks, emerge of machine users has brought unprecedented demands on data rate, system capacity, security, transmission latency, coverage area, and reliability, in 5G New Radio (NR) [6], [7]. According to International Telecommunication Union (ITU), these requirements have been categorized into three main services; enhanced Mobile Broadband (eMBB), massive Machine-Type Communications (mMTC), and Ultra-Reliable Low-Latency Communications (URLLC) [8]–[10]. High data rate services, such as broadband internet access, high-definition video streaming, and virtual reality, mainly determine the eMBB use-cases, which can be considered as an extension of the conventional 4th generation (4G) Long Term Evolution (LTE) technology. Applications with mMTC are expected to support a massive number of machine UEs with sporadic transmission, low power consumption, wide coverage area, and moderate data payloads. Differently, URLLC services represent the most challenging ones due to conflicting requirements of ultra-reliability and low-latency. In addition to the above demands, new application families, such as New Radio (NR)-Lite (NR-Lite) and secure URLLC (SURLLC),

have been envisioned in beyond 5G [11], [12]. This unprecedented global trend, where everything and everyone is expected to be connected at any time on our planet with real-time data exchange from a wide variety of services, requires efficient use of radio resources (RR) by wireless operators and adaptation of new communication techniques. Hence, future radio access technologies (RAT) need to support user equipment (UE) while performing transmission when traveling in public transportation, in crowded environments, and in unexpected traffic jams [4]. Basically, the ultimate goal is to offer a society where people, vehicles, drones and sensors will be able to connect with each other through terrestrial and non-terrestrial deployments in order to support innovative services such as smart cities, self-driving cars, remote surgery, and smart homes [13].

1.2. Scope Of The Thesis Dissertation

With the urge to be always connected to the network along with data-hungry applications, the evolution of traditional wireless deployments toward the coexistence of a wide range of services and use-case has become a necessity; starting from the early time of wireless technology histories such as General Packet Radio Service (GPRS), Enhanced Data rates for Global Evolution (EDGE), High Speed Packet Access (HSPA) and ending up with the most prominent today's wireless technologies such as Wireless Fidelity (Wi-Fi), Long Term Evolution (LTE), LTE-Advanced, and 5G NR [14]. Moreover, an intelligent integration of these technologies with heterogeneous networks (HetNet), mmWave-based schemes, multiple-input-multiple-output (MIMO) signaling, and potential future wireless systems that recently have been attracting attention of both academia and industry such as re-configurable intelligent surfaces (RIS), rate-splitting and multiple accessing (RSMA), unmanned aerial vehicle (UAV), index modulation (IM), orthogonal time frequency space (OTFS) modulation and machine learning (ML) are expected to carry the society of wireless communications to the stages of 5G-B, as illustrated in **Figure 1.1** [15]–[18]. Consequently, the presence of various wireless systems in the network, high conjunction of using the same RRs per square meter, and broadcast nature of the medium are highly expected to expose the transmission from the desired UE to external interference leaking from other sources (i.e., UEs and wireless networks) and internal interference coming from its own communication link. In line with this, interference between UEs indeed dominates the

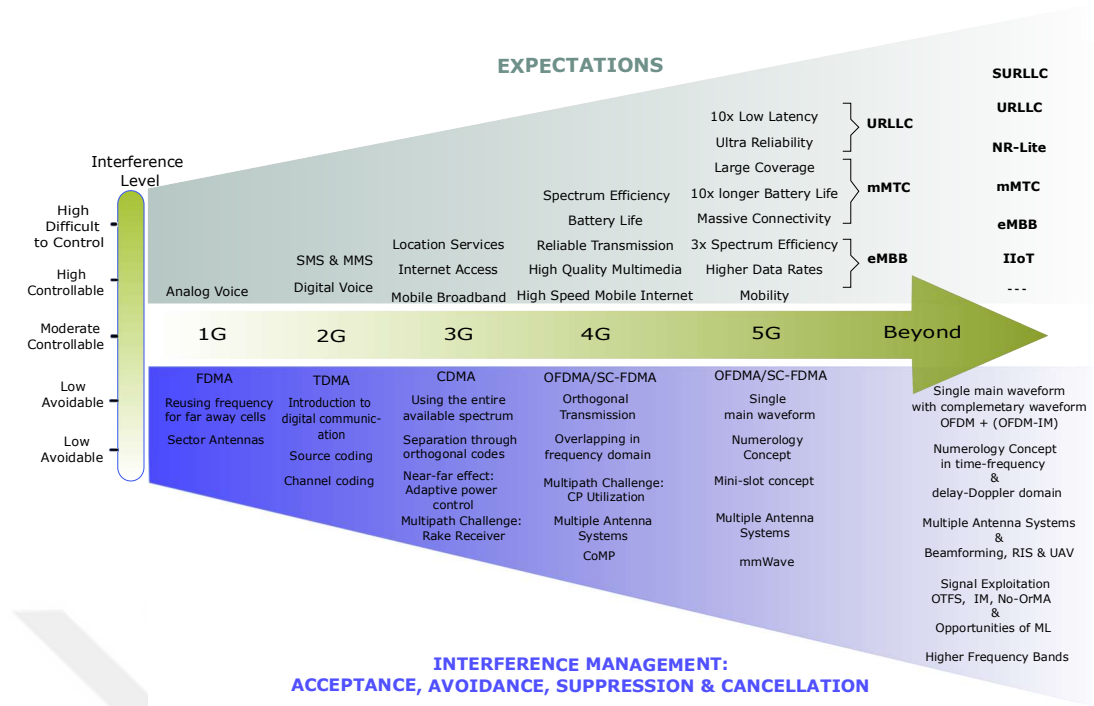


Figure 1.1: An overview of wireless communications transition from 1G to beyond 5G.

noise power and results in interference-limited wireless systems, which represents a critical challenge towards accomplishment of ambitious UE requirements in future wireless systems [19]. Importantly, new interference types and models have been revealed with the emerge of multi-functional applications, multi-tier networks, utilization of unlicensed bands, different derivatives of waveforms, and novel waveforms in order to serve various services and use-cases [20]. It is worth mentioning that the future of interference will not be the same with what we have seen in the past last decades. Interference is becoming a major area that will shape the future of wireless systems by disclosing research opportunities and future direction.

With this motivation, there is a need more than ever to understand, model, avoid, control, and generate interference, since it is becoming and will be essential to the performance of wireless communications realm for the better. In this line, the scope of this thesis is to provide a holistic perspective indeed which considers potential interference sources and offers physical (PHY) layer-based novel management techniques, i.e., interference control, mitigation, and cancellation schemes, towards 5G-B wireless communications systems. In particular, this thesis study focuses on flexible frame design for single UE and multiple UEs communications, the coexistence of multiple UEs with diverse demands via hybrid NOMA, and the exploitation of the delay-Doppler domain with a focus on the

OTFS technique, aiming to facilitate a wide array of services in future communication systems.

1.3. Motivation And Thesis Contributions

This thesis centers around research studies that concentrate on interference control and efficient utilization of radio resources (RRs) in next-generation communication. The main contributions of this research are given as follows.

1.3.1. Single carrier transmission for URLLC with adaptive radio resource utilization

In literature, several solutions have been proposed to meet the requirements of services with URLLC by both academia and industry. Design of new frame structures and various transmission protocols have been considered for URLLC in NR [21]. In legacy Long Term Evolution (LTE), one subframe consists of 14 orthogonal frequency division multiplexing (OFDM) symbols with 15 kHz subcarrier spacing, and it defines the smallest resource unit with 1 ms transmission latency [22]. In order to satisfy the requirements of URLLC, OFDM frame structure has been redesigned in NR, and new concept of 'mini-slot' is introduced where down to 2-OFDM symbols can be used per transmission. However, establishment of the transmission with latency below 1 ms and reliability up to 10^{-9} BER still remains a significant challenge. In [23], downlink and uplink latency for LTE Rel.15 are evaluated about 0.64 ms and 1.64 ms with mini-slot, respectively. In Rel.15, large subcarrier spacings (SCS) up to 120 kHz for below-6 GHz, and up to 240 kHz for above-6 GHz have been introduced to reduce transmission latency of applications with URLLC [24]. Transmission utilizing SCS equal or above 60 kHz can meet the latency requirement even for use cases in factory automation. However, since length of cyclic prefix (CP) is a linear ratio of OFDM symbol duration, it linearly decreases with the increase of SCS. Hence, system's reliability decreases due to inter-symbol interference (ISI) caused by reduced CP length. Moreover, grant-free transmission has been proposed in ongoing standardization meetings to avoid latency caused by handshaking between users and base station (BS) for scheduling [25], [26]. Lack of coordination between users and BS leads to asynchronous transmission. But, achieving a reliable system with OFDM requires strict

Table 1.1: Transmission requirements for different URLLC applications and use cases

Use case	Latency (ms)	Reliability
Factory automation	0.25 – 10	10^{-9}
Smart grids	3 – 20	10^{-6}
Professional audio	2	10^{-6}
Intel. Trans. System	10 – 100	$10^{-3} - 10^{-5}$

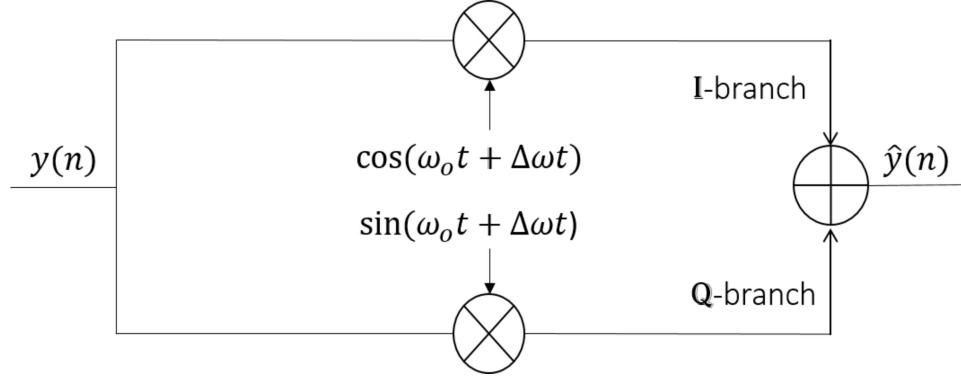
synchronization. Otherwise, OFDM based systems suffer from inter-carrier interference (ICI) [27]. Furthermore, higher level of peak-to-average-power ratio (PAPR) and out of band emission (OOBE) severely degrade the performance of OFDM systems compared with single carrier (SC) systems [28].

Inspired from the aforementioned reasons, SC transmission can be considered as a possible candidate to support applications and use cases with URLLC. Raised Cosine (RC) pulses are used for SC communication systems since they are Nyquist pulses that have the capability to eliminate ISI. RC pulse is localized in the frequency domain, but it requires infinite time to generate. Thus, system's reliability decays due to occurrence of ISI since the pulse is truncated in time domain [29]. Minimizing ISI caused by truncation depends on two parameters, which are span (ζ) and roll-off factor (α). ζ decides number of symbols for truncation of RC filter. Higher ζ value improves system reliability by reducing ISI, but it increases transmission latency. In SC communication, latency is determined by both data size (S) and ζ . Moreover, an increase of ζ value reduces the amount of OOBE, but it increases PAPR. α specifies both power of the filter's side lobes in time domain and occupied bandwidth in frequency domain. Increasing α factor results in a more localized signal in time domain and wider occupied bandwidth in frequency domain, and vice versa. Moreover, high α factor improves system's reliability by reducing both PAPR and OOBE [30]. On the other hand, unnecessary increasing of the α factor causes loss of spectral efficiency. Hence, the goal of this work is to present the significance of ζ , α and S optimization to achieve URLLC with SC transmission, while considering system resource efficiency (η), PAPR and OOBE in the system. Moreover, adaptive α usage for SC transmission is proposed in order to further enhance the system efficiency.

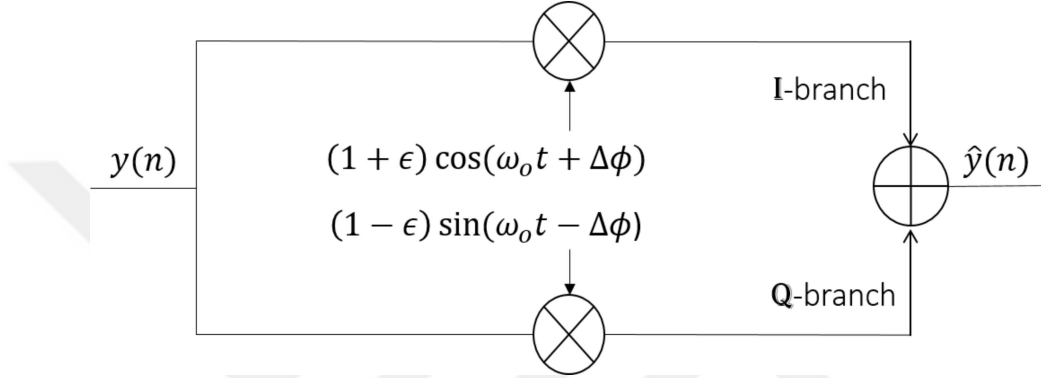
1.3.2. Performance analysis of frequency domain IM schemes under CFO and IQ imbalance

At first, the concept of IM has been introduced as a spatial modulation (SM) technique to convey additional information through active antenna indices [31]. Later, in [32] orthogonal frequency division multiplexing (OFDM) is combined with IM, and it is named as OFDM-IM. It performs transmission of data bits not only by M -ary modulated subcarriers but also by the indices of active subcarriers. Therefore, OFDM-IM can exceed the spectral efficiency of classical OFDM for low modulation order. It is shown that OFDM-IM provides better bit error rate (BER) performance compared with conventional OFDM under frequency selective channels. Partial subcarrier activation of IM allows designing an adaptable system by considering various demands [33]. In [34], OFDM-IM is proposed as an eligible solution to alleviate the impact of inter-carrier interference due to sporadic transmission of applications with mMTC. In [35], coexistence of OFDM and OFDM-IM over shared resources is evaluated for URLLC. In [36], OFDM with general index modulation (OFDM-GIM) has been proposed to further improve spectral efficiency by activating different numbers of subcarriers in each subblock. In [37], instead of using the active subcarrier indices to convey extra data bits, the number of utilized subcarriers has been considered as an information source in OFDM with subcarrier number modulation (OFDM-SNM). It further improves system's spectral efficiency while containing similar error performance compared with OFDM-IM.

OFDM-based system owing to the attractive properties of robustness against frequency fading channels, single tap equalization, and high spectral efficiency, it has been the most common multicarrier communication technology of fourth generation (4G) standards [38]. Unfortunately, OFDM suffers performance degradation due to non-idealities in the RF transceiver, which can cause carrier frequency offset (CFO) and in-phase and quadrature (IQ) imbalance [39]. CFO corresponds to mismatch of local oscillators between transmitter (Tx) and receiver (Rx), while non-identical amplitudes and not fixed $\frac{\pi}{2}$ phase difference between I-branch and Q-branch lead to IQ imbalance [40], as shown in Figure 1.2. In [41], the error probability of OFDM-IM is evaluated in the presence of CFO. It is shown that OFDM-IM is superior to OFDM under CFO impact due to fractional subcarrier activation. In [42], effect of IQ imbalance is investigated for OFDM-IM, and a simple non-iterative Rx is proposed to mitigate this impairment. Although recent IM techniques are evaluated



(a): Carrier frequency offset.



(b): IQ imbalance.

Figure 1.2: Baseband model of the received signal with RF impairments.

and compared from the perspective of spectral efficiency, error performances of these IM techniques under RF impairments is still unknown. Therefore, the aim of this paper is to analyze and compare the performance of various IM techniques including OFDM-IM, OFDM-GIM and OFDM-SNM under the impact of CFO and IQ imbalance.

1.3.3. A hybrid downlink NOMA with OFDM and OFDM-IM for beyond 5G wireless networks

The rapid growth of data traffic with emerging applications and use-cases increases the need for new technologies in beyond 5G wireless multiple accessing (MA) networks [43]. Orthogonal frequency division MA (OFDMA) is the most commonly used technique in downlink transmission. It provides high system throughput for near user equipment (UE) due to assignment of distinct resources and high signal-to-noise ratio (SNR). Unfortunately, throughput of edge UEs significantly reduces with the decay of SNR due to path-loss. Hence, unequal achievable rate is provided to UEs which significantly reduces system fairness.

Non-orthogonal multiple accessing (NOMA) via superposition of different UEs over the same resources (i.e., time and frequency) is actively investigated to enhance achievable rate while considering UE fairness [44]. In NOMA, power domain multiplexing is utilized at Tx, and successive interference cancellation (SIC) is used to separate the superposed signals at Rx [45]. However, this scheme needs signals that possess significant power difference to have a successful multi-user detection (MUD). In case of insufficient power difference between UEs, SIC can not eliminate inter-user interference (IUI), and thus multiplexing of UEs by means of power becomes infeasible. Consequently, signal separability appears as a very challenging task due to low signal-to-interference plus noise ratio (SINR) stemming from high IUI. In the literature, plenty of work attempts to maximize the achievable sum rate of OFDM-NOMA systems without compromising reliability. In [46], an appropriate power domain user pairing is suggested to enhance the sum rate of system for a given user service quality. Differently, in [47], multi-numerology multiplexing along with power domain multiplexing is proposed for both increasing sum rate and mitigating IUI. Recently, the use of OFDM and index-modulated OFDM is introduced for the coexistence of grant-free ultra-reliable low-latency communication (URLLC) and grant-based latency-tolerant service, respectively [35]. Available resources are shared among the services when URLLC occurs. K -repetitions transmission and maximum ratio combining (MRC) Rx are adopted to provide URLLC. Moreover, NOMA concept is incorporated with space-domain IM in [48], [49].

In this paper, a new hybrid NOMA scheme exploring power domain multiplexing of OFDM and OFDM-IM is introduced and termed IM-NOMA. Inherent power difference between OFDM and OFDM-IM subcarriers exists due to partial subcarrier activation in OFDM-IM. Hence, the performance of MUD in downlink IM-NOMA depends not only on the UEs' power differences but also on the subcarrier activation ratio of OFDM-IM. It is demonstrated that IM-NOMA is superior to the conventional OFDM-NOMA in terms of achievable sum rate and system reliability while guaranteeing target rate of far UE. Additionally, a novel Rx design is provided to enhance MUD performance when the UEs have similar power.

1.3.4. Exploiting user diversity in OTFS transmission for beyond 5G wireless systems

Basically, OTFS-based communication schemes convey the data information in delay-Doppler domain that offers a more comprehensive description of the environmental motion, surrounding reflectors, and number of arrival arrays, compared to the scenarios with either time or frequency channel state information (CSI) [50]. Since OTFS transmission is able to benefit from the full diversity of wireless channels, it has attracted the attention of academia and industry [51]. In [52], Raviteja *et al.* have exploited delay-Doppler domain representation of the wireless channel in order to perform target range and velocity sensing of the surrounding obstacles in the environment. Moreover, in [53], Rakib *et al.* have assessed the concept of OTFS not only as a potential sensing and modulation technique but also as a multiple accessing scheme, where different users are allocated at distinct radio resources in the delay-Doppler domain. In addition, Ding *et al.* have combined OTFS with non-orthogonal multiple accessing (NOMA), in [54]. Moreover, authors in [55] and [56] have discussed the integration of OTFS signals in multiple-input multiple-output technology and the design of proper pulse shaped signals, respectively, improving the link reliability, and enhancing the overall spectral efficiency of the system accordingly. On the other hand, several works in the literature have investigated the performance of OTFS transmission under radio frequency (RF) imperfections, such as peak-to-average power ratio (PAPR), phase noise (PN) and in-phase and quadrature (IQ) imbalance for practical wireless communication scenarios [57]–[59].

Despite the aforementioned opportunities and applications of OTFS signaling, there is still limited research activities by both academia and industry to take advantage of the channel independence of different user equipment (UE) under practical scenarios with multi-user systems [60]–[63]. Different from classical time/frequency-based multi-carrier schemes, in [64], Yuan *et al.* explicitly show that the diversity of OTFS-based wireless technologies becomes dominant with the increase of the number of resolvable incoming paths. To the best of the author's knowledge, the influence of the multi-user diversity with the focus on varying number of channel taps has not yet been investigated in the literature, and the impact of diversity of each UE with OTFS transmission on the error performance of the overall system is unknown. Hence, the main motivation of this work is to close this gap and to provide useful insights for both academia and industry. In this regard, this study

proposes a novel scheduling algorithm for multi-user OTFS-based wireless technology in which the BS attempts to maximize the average link reliability of the overall system via utilization of the diverse number of channel taps of each UE in downlink transmission.

1.3.5. Performance analysis of OTFS under in-phase and quadrature imbalance at T_x

The exponential growth of user devices with new requirements and use cases shows that inexperienced wireless techniques are required in 6G and beyond wireless systems, especially for vehicular communications [12], [65]–[67]. The most widely used wireless technology standards rely on time-division multiplexing (TDM) and orthogonal frequency division multiplexing (OFDM) that define the two most prominent technologies of wireless communication systems, respectively. However, designing suitable transceivers becomes a challenge in controlling inter-symbol interference (ISI) caused by the frequency selectivity of the radio channel and inter-carrier interference (ICI) due to the time selectivity of the radio channel in TDM-based and OFDM-based wireless communication technologies, respectively [68], [69]. Even though TDM and OFDM-based schemes represent the main driving wireless technologies, none of these concepts are robust against doubly dispersive channels. Therefore, a new two-dimensional (2-D) modulation scheme named orthogonal time frequency space (OTFS) is proposed in [17] both to relax and to manage the effect of selective wireless channel on the communication link.

The main idea of OTFS transmission is to convey information via delay-Doppler domain, which provides a more comprehensive description of the wireless channel, i.e., presence of surrounding reflectors, environmental motion, and angle of arrival, compared to either time or frequency channel information [50], [70]–[72]. Thus, OTFS transmission offers better performance compared with classical transmission technologies under doubly dispersive channel conditions, since it is capable to fully explore the diversity of wireless channel due to the high resolvability of any path differing in delay or Doppler shift [51], [73]–[75]. In [52], Raviteja *et al.* utilize the channel state information (CSI) of delay-Doppler domain for sensing target range and velocity of obstacles in the surrounding medium. Differently, [53] introduces OTFS transmission not only as a modulation concept but also as a multiple accessing waveform with multiple users allocated at different delay-Doppler radio resources. Moreover, authors propose the use of guard band in order

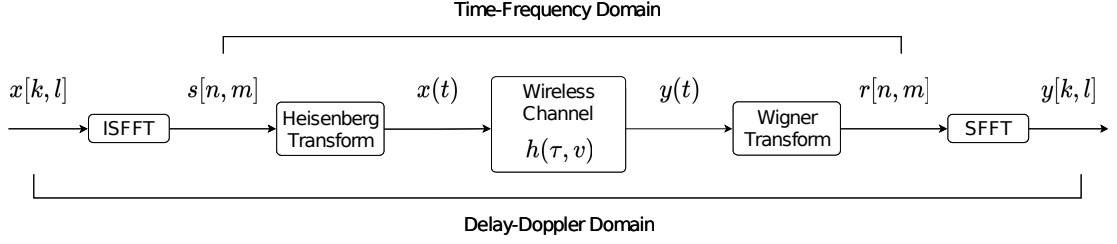


Figure 1.3: System model of the OTFS transmission in doubly dispersive channel.

to reduce interference between users due to inter-Doppler interference (IDI). Since the OTFS transmission with practical pulse shapes suffers from interference, its performance is compared with non-orthogonal waveform of generalized frequency division multiplexing (GFDM) in [76], and it is combined with non-orthogonal multiple accessing (NOMA) scheme in [54]. Farhang *et al.* propose OFDM-based OTFS to simplify the implementation complexity of the system at the cost of lower spectral efficiency [77]. As a potential solution, [56] and [55] discuss the design of various pulse shaped waveforms for OTFS transmission and its combination with multiple-input multiple-output (MIMO) technology, respectively, in order to not only increase spectral efficiency but also enhance system performance under realistic scenarios.

Despite the promising benefits and use-cases of OTFS transmission outlined above, still limited research by both academia and industry exists to assess the opportunities and challenges of the OTFS transmission under practical scenarios including hardware and channel impairments. For this reason, its performance under practical implementation scenarios requires a comprehensive assessment. In [78] and [79], authors elaborate the performance of the OTFS transmission under non-linear power amplifiers and phase noise (PN) effect, respectively. Moreover, [57] clearly shows that OTFS signalling outperforms conventional OFDM transmission in the presence of PN. In general, super-heterodyne Tx (SHT) and direct conversion Tx (DCT) represent the main designs used in modern wireless communication industry. SHT is superior in terms of link reliability compared with DCT due to the use of both image rejection filters and intermediate frequency while converting the baseband signal to high frequencies and vice versa. However, its hardware complexity limits the energy efficiency and communication latency that represent stringent demands for enhanced mobile broadband (eMBB) and ultra-reliable and low-latency (uRLLC) applications in wireless technologies. Different from SHT, DCT provides low

complexity and low power consumption since neither image rejection filter nor intermediate frequency is used [80]. Moreover, its architecture offers a prospective solution for easier integration with massive devices and significantly reducing the overall cost compared with SHT due to the simple structure [81]. Despite the aforementioned favorites of DCT design, its implementation represents one of the most critical challenges in practical scenarios, since the performance of the transmitted signal is heavily affected by the presence of hardware non-idealities due to in-phase and quadrature (IQ) imperfection between modulator branches, limiting the signal-to-interference plus noise (SINR) in conventional wireless systems [82]. In [83], only the experimental results for OTFS under joint non-linear power amplifier and IQ imbalance are discussed considering a 60 GHz testbed in a laboratory environment. However, to the best of author's knowledge, the IQ imbalance impact in delay-Doppler domain is still unknown, and the influence of the IQ imbalance on OTFS transmission has not yet been investigated and theoretically analyzed in the literature. Thus, the aim of this work is to close this gap and to provide useful knowledge for both academia and industry. In particular, we list the main contributions of the work as follows.

- We derive the mathematical model to understand the physical effect of IQ imbalance in delay-Doppler domain for OTFS transmission.
- We prove that IQ imbalanced Tx in OTFS transmission leads to mirror interference occurring only across the Doppler axis.
- We show that different from conventional single-domain transmission techniques, i.e., time-based TDM and frequency-based OFDM, the error performance of OTFS transmission does not saturate under IQ imbalanced Tx.
- Importantly, IQ imbalance significantly deteriorates the diversity gain of OTFS transmission.
- We provide the closed-form BER expression of OTFS with constant envelope M-ary symbols under IQ imbalanced Tx for the validation and assessment of the system performance considering ideal and realistic wireless channel scenarios.

1.3.6. Index modulation-aided IQ imbalance compensator for OTFS communications systems

Recently, a new modulation scheme named OTFS is proposed to control the effect of doubly dispersive wireless channels on the communication link [17]. The basic idea of OTFS modulation is to carry data symbols via a two-dimensional plane named delay-Doppler (DD) domain, which provides a more comprehensive description of the wireless medium, i.e., the gain of each channel tap, time of arrival and angle of arrival, in comparison to either time or frequency channel representation. Therefore, the OTFS scheme outperforms the traditional transmission technologies (i.e., time division multiplexing (TDM) and OFDM under time-varying channel conditions, as it is able to explore the diversity of wireless channels due to the high resolvability of each path differing in delay or Doppler shift [74]. With this motivation, in [52], authors propose the use of the channel state information (CSI) of the DD domain for sensing target range and velocity of obstacles in the surrounding medium. Despite the promising favorites of OTFS modulation, its performance is affected due to hardware imperfections [57], [78]. Among these impairments, in-phase and quadrature (IQ) imbalance is subject to this paper. IQ imbalance occurs due to the mismatch on phase and amplitude response between I and Q branches of the local oscillators (LO) [84] that leads to interference. In particular, the impact of the IQ imbalance limits the diversity gain of OTFS and becomes severe with the increase in modulation order [58], [85]. Thus, IQ imbalance parameters need to be estimated and compensated in order to maintain the reliability of the communication link [82].

Mainly, blind and pilot-aided compensation approaches are proposed in the literature to mitigate the effect of IQ imbalance on conventional communication systems [86], [87]. Specifically, in [88], Naikoti *et. al.* propose the combination of OTFS transmission with deep neural network (DNN) architecture in order to perform both compensations of IQ imbalance and channel training in the DD domain. However, blind and pilot-aided compensation schemes need a high number of iterations and training pilots leading to high system complexity and spectral efficiency loss. In contrast to the above solutions, in this work, we propose a novel and index modulation (IM)-aided IQ imbalance compensation technique independent from iterative processing and training/pilot sequences within the OTFS frame. The concept of IM conveys data bits not only by the modulated carriers, but also by the partial activation of carrier indices that enhances the system diversity [89],

[90]. Recently, OTFS is combined with IM to improve its error performance and spectral efficiency [91], [92]. Importantly, the sparse utilization of radio resources via IM enables conventional communication schemes (i.e., TDM and OFDM) to alleviate and control the impact of interference as discussed in [93]. Hence, in this work, we utilize the sparsity of IM to mitigate the impact of IQ imbalance and simultaneously perform reliable communication via active carriers in the DD domain.

1.3.7. Low complex inter-Doppler interference mitigation for OTFS systems via global Rx windowing

The concept of OTFS has brought new research perspectives owing to its two-dimensional (2-D) lattice structure while interacting with the wireless channel. Basically, OTFS signaling conveys information jointly via delay-Doppler (DD) domains, which represent the frequency and time variation of the wireless channel, respectively [94]. Hence, it provides a more comprehensive description regarding the environmental motion, presence of surrounding reflectors, and angle of arrival, compared to either time or frequency channel state information (CSI), existing between Tx and Rx [50], [72]. As a result, OTFS signaling offers better performance compared with conventional wireless technologies under time varying channel conditions due to its high resolvability in DD domain at the cost of higher system complexity [51], [75]. For instance, DD domain CSI is exploited for sensing target range and velocity of obstacles in the surrounding medium in [52]. Moreover, in [55], authors have introduced the combination of OTFS signaling with the conventional multiple-input multiple-output (MIMO) technology, and thus, they offer a simplified channel estimation method in order to relax the overall computational complexity of the system. In [95], an approximation of Message Passing (MP) algorithm is introduced in order to reduce Rx complexity. In line with this, to ensure the compatibility of OTFS concept with the prominent OFDM-based technology, in [77], Farhang *et al.* have introduced OFDM-based OTFS at the cost of lower spectral efficiency. In [56], Raviteja *et al.* have discussed the performance of the OTFS signaling with a focus on practical pulse shaping, where reduced cyclic prefix (CP) OTFS has been proposed, which has been differently known as common CP in 5G wireless technologies. Technically speaking, authors utilize common CP only to control the impact of time dispersion of wireless channel (i.e., ISI) on OTFS frame. In [96], Zegrar *et al.* have studied the effect of various prefix/suffix

configurations on OTFS transmission with rectangular pulse shaping. Furthermore, Li *et al.* took the next step towards the error performance exploitation of coded OTFS system under practical scenarios, in [64].

Differently from the waveform perspectives of OTFS signal, in [53], Rakib *et al.* have discussed OTFS signaling as a potential multiple accessing option, where multiple UEs are allocated at distinct DD radio resources. Moreover, an intelligent OTFS based multiple accessing is proposed in [97], where each UE is assigned at equally spaced intervals in the DD domain. In [98], Mary *et al.* have suggested multiplexing of UEs in a non-contiguous and interleaved manner in accordance with various demands. In [54], [99], [100], OTFS signaling is considered as a potential solution not only for orthogonal multiple accessing but also for non-orthogonal multiple accessing (NOMA). In addition, the performance of OTFS under practical RF impairments has attracted the attention of literature and industry. In [57], Surabhi *et al.* have illustrated that OTFS outperforms OFDM signaling under phase noise (PN) in mmWave bands. Moreover, in [59], [101] and [78], authors have elaborated the performance of the OTFS transmission under in-phase and quadrature (IQ) imbalance and non-linear power amplifiers effect, respectively.

As a new waveform, the OTFS scheme has introduced new challenges and research opportunities, in addition to its promising benefits. In particular, the effect of wireless channel needs to be equalized via advanced Rx, which become overwhelming in the presence of fractional Doppler effect [102]. It is worth noting that majority of the research in the literature considers only the integer Doppler indices, which maximizes the sparsity of the wireless channel, leading to a simplified Rx design. However, fractional Doppler destroys the channel sparsity since it is spread along all the Doppler axis, leading to severe inter-Doppler interference (IDI). In line with this, Li *et al.* have proposed a novel algorithm, performing cross-domain iterative detection and estimation to improve the error performance of OTFS scheme under fractional Doppler at the cost of additional Rx complexity in [103]. Moreover, Wei *et al.* have discussed the impact of Tx windowing and Rx windowing for OTFS scheme in the time-frequency (TF) domain by utilizing either mercury/water-filling power allocation scheme or Dolph-Chebyshev (DC) window at the cost of higher complexity and emerge of non-orthogonal pulses in DD domain, respectively [104]. The performance of this scheme relies on the quality of channel state information (CSI) in TF domain while performing a bisection search. Moreover, in [105], Qu

et al. have introduced a Rx design for OTFS transmission that leverages the sparsity of the channel matrix in order to perform both symbol detection and IDI mitigation in an iterative manner under the impact of fractional Doppler. However, the power inefficiency and round trip time (RTT) latency become a major concern with increased system complexity due to fractional Doppler in OTFS-based wireless technologies. Therefore, the design of lightweight approaches that provide easier integration with the current technologies and massive devices has become an imperative need to alleviate the effect of IDI in OTFS systems.

Given the discussions above, this work introduces a novel Rx windowing (RW) approach for OTFS waveform named as RW-OTFS. The proposed RW-OTFS design considers the relation between the orthogonality constraints and the pulse shape of the OTFS carrier, i.e., the Dirichlet function in order to mitigate IDI. In contrast to the conventional Rxs that use iterative exchange of the extrinsic information between disjoint components, i.e., encoder and decoder, we utilize pulse localization in Doppler domain with the aid of proper CP padding in time domain [104], [105]. To the best of the authors' knowledge, the impact of raised cosine (RC) based Rx windowing in DD domain is still unknown, and its influence on the OTFS transmission has not yet been modeled and theoretically investigated in the literature. In this regard, the main contributions of this work are listed as follows.

- This is the first work that proposes time domain-based RC windowing for OTFS transmission in order to mitigate and control the impact of side lobes for the OTFS carriers in DD domain.
- In particular, the proposed scheme performs global CP-based RC windowing to mitigate IDI since the entire OTFS frame duration is needed to observe the Doppler characteristics of the transmission.
- We develop a mathematical model in order to analyze the physical effects of RC pulse shaping in the DD domain for OTFS transmission.
- Importantly, we show that different from the iterative-based interference mitigation techniques, i.e., water filling power allocation and cross domain CSI exploitation, the channel sparsity of the OTFS waveform can be significantly enhanced with the aid of the proposed RC windowing at the Rx side.

- We provide the closed-form of upper bound BER expression for the proposed RW-OTFS scheme under fractional Doppler impairment in order to validate and assess the overall system performance, while considering realistic wireless environment scenarios and system configurations.



CHAPTER 2

2. THEORETICAL PART

This chapter describes the utilization of RRs by prominent wireless communications technologies to accomplish the demands of UEs in the wireless realm.

2.1. Transmission Domains

Basically, we revise the basic concepts of these techniques from a transmission domain perspective, aiming to provide an essential background for the remainder of the work.

2.1.1. Frequency domain

2.1.1.1. Frequency division multiplexing

It splits the entire frequency spectrum into multiple RF carriers, where each carrier is assigned to a particular UE. Moreover, a frequency guard band is utilized in between to avoid signal emission between adjacent channels and maintain the orthogonality. FDM is one of the most common analog technologies of 1G for mobile phones known as frequency division multiple accessing (FDMA). FDM is characterized by a large symbol duration (T_s) compared to the delay spread (τ_{max}) of the wireless channel. It reduces the impact of ISI at the cost of a lower data rate. Unfortunately, the insertion of the guard band leads to a waste of spectral efficiency. Moreover, the hardware impairments become a critical factor due to the implication of narrow-band filter in the transceiver architecture.

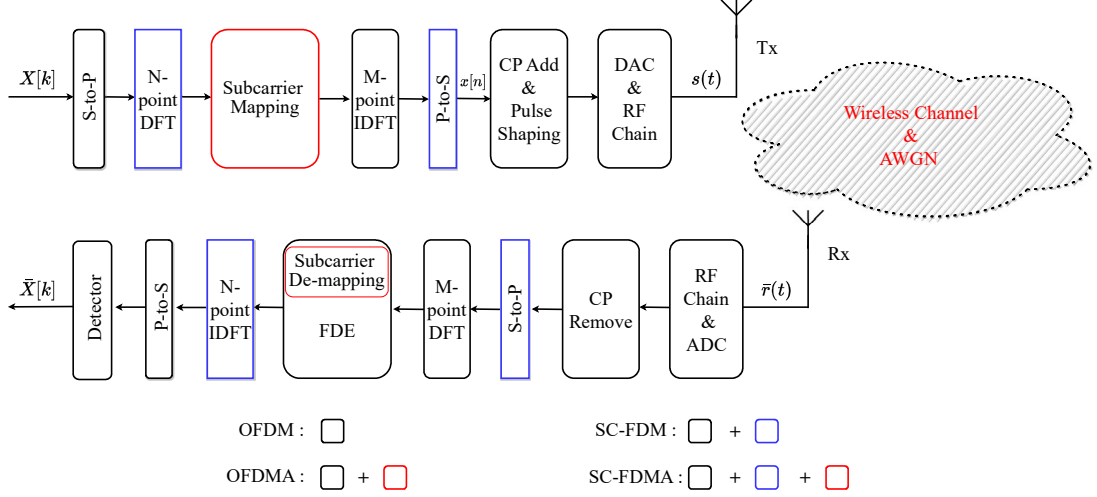


Figure 2.1: Transceiver model for conventional wireless technologies utilizing FDE methodology.

2.1.1.2. Orthogonal frequency division multiplexing

OFDM represents the most popular multi-carrier technology due to its low complexity, efficient spectrum utilization, and easy compatibility, and thus, it has been adopted in wireless standards, such as IEEE 802.16, wireless local area networks (WLANs), 4G LTE, and Wi-Fi [106]. In particular, this technology takes a wide-band fading channel and splits it into several narrow-band subchannels. Occasionally, interference on the desired channel can be bandwidth-limited. Consequently, this will avoid impacting all subchannels, hence not affecting all the data. As a result, it is resilient against CCI interference. OFDM can avoid ISI which is associated with the frequency selectivity of wireless channels and the presence of multiple narrow-band subchannels. As shown in **Figure 2.1**, data information $X[k]$ (i.e., M -ary quadrature amplitude modulation (QAM)/phase shift keying (PSK) symbols) are defined in the frequency domain. Later, the inverse-discrete Fourier Transform (IDFT) process converts the information signal to time domain $x[n]$, leading to an extended symbol duration T_s compared with time-domain based transmission schemes. Technically speaking, to prevent ISI in $x[n]$, a cyclic prefix (CP) extension at the front is needed, and thus, a simple one-tap FDE is performed at the Rx side. Unfortunately, OFDM signal is highly sensitive to ICI occurring among sub-channels and caused by the strict frequency synchronization demands. Hence, it is very sensitive to phase noise, asynchronous transmission, and corrupt RF-front-end.

2.1.2. Time domain

2.1.2.1. Time division multiplexing

In contrary to FDM, in TDM, the entire bandwidth has been used for transmission or shared among the UEs i.e., which is the case of time division multiple accessing (TDMA), while utilizing predetermined time slots or dynamic time slots. Due to fast time slots the data rate is higher than FDM, and thus, the transmission bandwidth is much higher. In this line, this scheme allows much more flexible rates. Importantly, due to shared bandwidth access, no guard band is needed. Thus, this removes the necessity for narrow-band filter processing and implementation. On the other hand, wireless systems performing broadband communication require advanced equalization techniques, since T_s gets smaller. Consequently, complex signal processing techniques are required to overcome ISI and relevant synchronization problems within the TDM/A frame.

2.1.2.2. Single carrier - Frequency division multiplexing

Inspired by the simple structure of OFDM, single carrier-frequency division multiplexing (SC-FDM) is introduced in the literature [107], as shown in **Figure 2.1**. It is characterized by a similar structure as classical OFDM. Thus, it is currently adopted as the uplink transmission named single carrier-frequency division multiple access (SC-FDMA) for the 3GPP LTE technology due to its lower level of envelope fluctuation compared to the OFDM-based scheme. Basically, different from the OFDM transmission, an additional discrete Fourier transform (DFT) and IDFT processes is implemented at Tx and Rx, respectively, in order to enable the FDE [108]. In other words, a sequential transmission of the subchannels is achieved instead of parallel transmission as in OFDM at the cost of a moderate system complexity enhancement in order to achieve lower PAPR. SC-FDM can be considered as an OFDM system with an additional DFT mapper, and thus, it is also called DFT spread OFDM (DFT-s-OFDM) in the literature. Obviously, this shows that in SC-FDMA, the information of any symbol is carried by each subchannel, while OFDM based subchannel conveys only a single data symbol information. Therefore, SC-FDM reduces the sensitivity to ICI while losing immunity against multi-path fading, i.e., limiting form the frequency selectivity.

2.1.3. Code domain

Code Division Multiplexing: CDM has been adopted in 3G wireless communication networks, where orthogonal codes are assigned to the desired signal $x[n]$, spreading it over the entire spectrum [109]. In other words, different CDM based signals share the radio resources in both time and frequency dimensions but are orthogonal in the code domain. Hence, in the case of code division multiple accessing (CDMA), each UE has a unique de-spreading sequence to recuperate the information at the Rx side. CDM technique provides satisfactory performance under the conditions of low SNR by means of achieved processing gain via spread-spectrum technology. Therefore, CDM/A heavily has been used in military applications for confidentiality. CDMA offers better multi-path resolution due to the wide-band transmission. Moreover, RR sharing among multiple UEs allows a high spectral efficiency alongside a relaxed CCI handoff problem that has been a major problem in cellular systems. On the other hand, the near-far power issue emerges while utilizing CDMA techniques. Even though different UEs are allowed to coexist over the same RR, the number of orthogonal codes and their corresponding lengths are governed by orthogonality and latency constraints, making code selection a tedious process.

2.1.4. Index domain

The idea behind IM is to convey additional data bits via different transmit entities in time, frequency, space, and code domains in order to provide reliability and energy efficiency without compromising SE [110]. At first, IM concept is combined with MIMO systems, i.e., spatial modulation (SM), and OFDM systems, i.e., OFDM-IM, by transmitting additional information bits through antenna and subcarrier indices, respectively [111], [112]. For instance, OFDM-IM partially activates subcarriers to convey the data bits by both activated subcarriers and M -ary symbols [112]. In this line, SC transmission with IM (SC-IM), generalized frequency division multiplexing GFDM with IM (GFDM-IM), and code-domain IM (CD-IM) through direct-sequence spread spectrum (DSSS) are introduced in the literature [113]–[115]. Incoming information bits are divided into subblocks and each subblock is split into two parts, where the first part decides the status of information-bearing entities, i.e., active or inactive, and the second part is conventionally modulated.

Consequently, the IM concept introduced the utilization of RR in a sparser manner, diverging from classical communication schemes, while maintaining a significant level of SE. To emphasize, our main focus is on elaborating the potential advantages of IM schemes in managing and mitigating various types of interference.

2.2. SC Transmission

In SC transmission, data information bits are firstly mapped to a continuous sequence of M -ary complex symbols. Later, data symbols are filtered to generate baseband signal $x(t)$, which is expressed as

$$x(t) = \sum_{n=-(\frac{S-1}{2})}^{\frac{S-1}{2}} s_m g(t - mT), \quad (2.1)$$

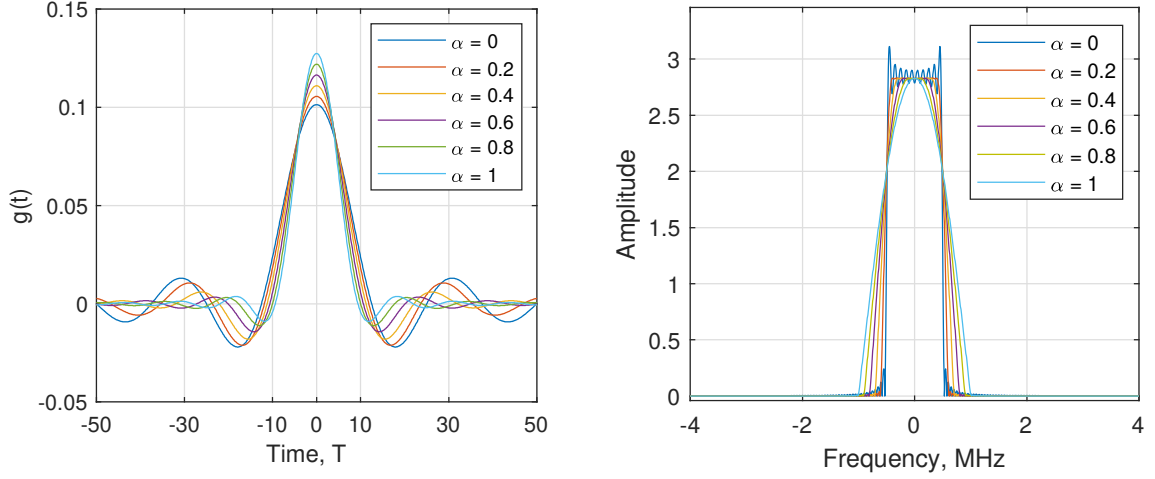
where s_m denotes the m -th data symbol, T is symbol duration and $g(n)$ represents the Tx filter. In order to achieve Nyquist pulse, Root-Raised-Cosine (RRC) pulse is used at both Tx and Rx in this study. Time domain representation of RRC filter $g(t)$ is given as

$$g(t) = \frac{2\alpha}{\pi\sqrt{T}} \frac{\cos\left[(1+\alpha)\frac{\pi t}{T}\right] + \frac{\cos\left[(1-\alpha)\frac{\pi t}{T}\right]}{\frac{4\alpha t}{T}}}{1 - \left(\frac{4\alpha t}{T}\right)^2}. \quad (2.2)$$

$g(t)$ with various α values is illustrated in Figure 2.2(a). Filter's side lobes decrease with an increase of α . Therefore, low time duration is needed to generate RRC filter with high α factor. Fourier transform of $g(t)$ is frequency domain representation of RRC filter $G(f)$, which is written as

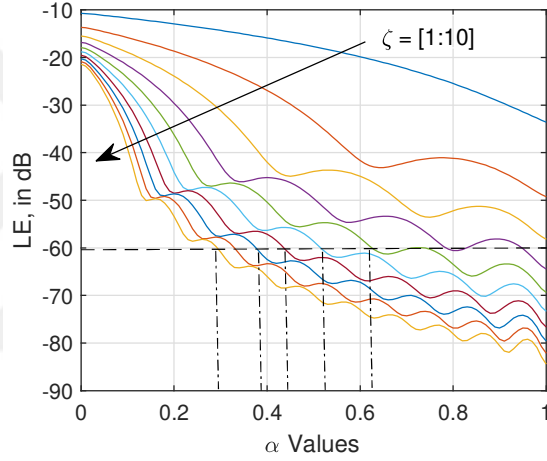
$$G(f) = \begin{cases} \frac{1}{\sqrt{2W}}, & 0 \leq |f| < (1-\alpha)W \\ \sqrt{\frac{1}{4W}} \left\{ 1 + \Upsilon \right\}, & W(1-\alpha) \leq |f| < W(1+\alpha) \\ 0, & |f| \geq W(1+\alpha) \end{cases} \quad (2.3)$$

where $\Upsilon = \cos\left[\frac{\pi}{2W\alpha}(|f| - (1-\alpha)W)\right]$, and W denotes Nyquist bandwidth associated with $R = \frac{R_b}{\log_2(M_{ary})} = 2W$ Nyquist rate. Figure 2.2(b) illustrates occupied bandwidth in frequency domain, and OOB caused by different α values. In SC with RRC pulse shaping, occupied bandwidth equals to $OB = R(1+\alpha)$. Hence, use of high α values increases occupied bandwidth, and it gives rise to loss of spectral efficiency in the system. For instance, OB increases twice for $\alpha = 1$ compared with $\alpha = 0$ case for the same data



(a): Time domain representation of RRC filter $g(t)$ for different α factors.

(b): Frequency domain representation of RRC filter $g(t)$ for different α factors.



(c): Amount of leakage energy for different α and ζ pairs.

Figure 2.2: Impact of α and ζ on filter's side lobes, occupied bandwidth, and ISI due to truncation.

rate. Therefore, the value of α should be minimized to increase the spectral efficiency of the system. On the other hand, reducing α value increases OOB, as seen in Figure 2.2(b). High OOB severely degrades overall system reliability. Clearly, a conflict of interest exists between spectral efficiency and OOB due to α factor. In order to provide further insights on the design of RRC filter, leakage energy (LE) regarding to different α and ζ parameters is shown in Figure 2.2(c). For a given α , LE represents the difference between energy of perfect RRC pulse and energy of truncated RRC pulse with span ζ . In other words, LE implies the amount of ISI due to truncation. It is obvious that a fixed amount of LE can be achieved by various α and ζ pairs. In this regard, SC communication

provides opportunities to fulfill different service requirements such as latency, reliability, and spectral efficiency by intelligent utilization of α and ζ .

2.3. OTFS Transmission

This section revisits and discusses the basic concept of OTFS transmission under fractional Doppler from a pulse-shaping perspective in order to provide an essential background for the remainder of the study. Without loss of generality, this study considers an OTFS architecture, where both Tx and Rx are equipped by a single antenna, as illustrated in **Figure 1.3**. Specifically, we consider an OTFS frame that consists of N number of symbols and M number of subcarriers with T symbol duration and Δf subcarrier spacing, respectively. Hence, the OTFS frame occupies a total bandwidth of $B = M\Delta f$ over the time duration of $T_f = TN$. In this line, the input binary bits stream is first mapped to one of the uncoded digital modulation schemes, such as Phase-Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM) symbols, and then rearranged in the DD domain as $X \in \mathbb{C}^{M \times N}$, where $x[l, k] \in \mathbb{A}$ refers to the element of X in the l -th delay and k -th Doppler indices. The DD lattice is represented as follows

$$\Psi = \left\{ \left(\frac{l}{M\Delta f}, \frac{k}{NT} \right), l=0, \dots, M-1, k=0, \dots, N-1 \right\}, \quad (2.4)$$

where NT and $M\Delta f$ define the Doppler and delay resolution, respectively. Consequently, $1/NT$ denotes the step size of Doppler axis and $1/M\Delta f$ is the step size of delay, while $T\Delta f = 1$. Since the data transmission is performed in time domain, we utilize the Inverse Symplectic Fast Fourier Transform (ISFFT) in order to first convert the DD domain signal X to TF domain [70], as follows

$$x_{\text{TF}}[n, m] = \frac{1}{\sqrt{NM}} \sum_{k=0}^{N-1} \sum_{l=0}^{M-1} x[k, l] e^{j2\pi \left(\frac{nk}{N} - \frac{ml}{M} \right)}, \quad (2.5)$$

where $E\{x[k, l]x[k', l']\} = \sigma_s^2 \delta(k - k', l - l')$, with $\delta(\cdot)$ denoting the Dirac delta function and σ_s^2 is the signal power. Later, as shown in **Figure 1.3**, we utilize Heisenberg transform in order to obtain the time domain signal as in [56], given as

$$s(t) = \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} x_{\text{TF}}[n, m] g_{\text{tx}}(t - nT) e^{j2\pi m \Delta f (t - nT)}, \quad (2.6)$$

where $g_{\text{tx}}(t)$ is the Tx pulse shaping with the pulse duration of T . After the OTFS frame $s(t)$ leaves the Tx side, it passes over the doubly dispersive wireless channel as

$$h(\tau, \nu) = \sum_{i=0}^{L-1} h_i \delta(\tau - \tau_i) \delta(\nu - \bar{\nu}_i), \quad (2.7)$$

where h_i , τ_i , and $\bar{\nu}_i$ are the path gain, delay shift and Doppler shift for the i -th channel path, respectively, with $i \in \mathbb{N}[0, L-1]$. L denotes the number of channel paths. Regarding (3.55), the delay and Doppler indices for the i -th channel component equal to

$$\tau_i = \frac{l_i + \iota_i}{M\Delta f}, \quad \text{and} \quad \bar{\nu}_i = \frac{k_i + \kappa_i}{NT}, \quad (2.8)$$

where ι_i and κ_i denote the fractional delay and fractional Doppler, respectively. The impact of fractional delay components becomes insignificant since they are resolvable in current wideband communication systems [104]. Hence, this study focus on the effect of fractional Doppler κ_i that is inevitable in beyond 5G mobile networks, and κ_i can take a value between $-0.5 < \kappa_i \leq 0.5$.

At the Rx side, the time domain received signal is given in [102], as follows

$$r(t) = \int \int h(\tau, \nu) s(t - \tau) e^{j2\pi\nu(t-\tau)} d\tau d\nu + n(t), \quad (2.9)$$

where $n(t)$ is the additive white Gaussian noise (AWGN) in the time domain. In this line, we perform Wigner transform in order to obtain the TF domain signal given as

$$R(t, f) = \int r(t) g_{\text{rx}}^*(t' - t) e^{-j2\pi m \Delta f (t' - t)} dt', \quad (2.10)$$

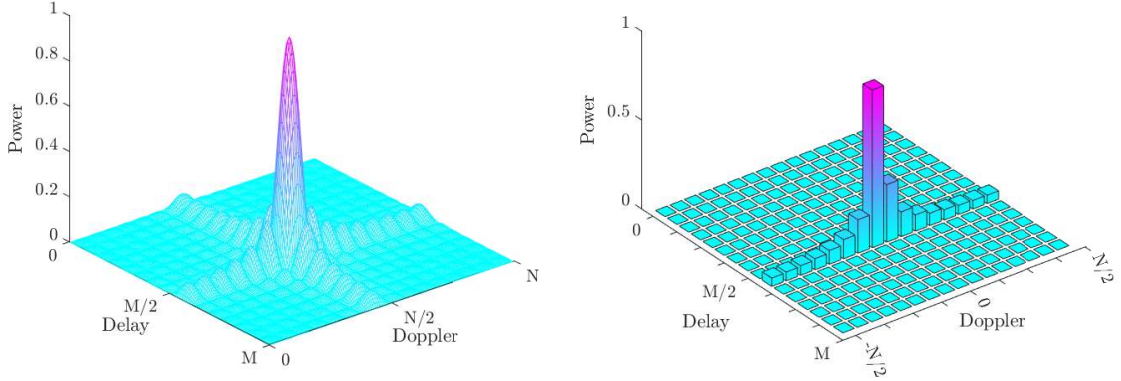
where $g_{\text{rx}}(t)$ is the Rx pulse shaping that needs to be designed considering the Tx pulse shape in order to maximize the signal-to-noise ratio (SNR). The obtained discrete samples $r_{TF}[n, m]$ of $R(t, f)$ given in (2.10) need to obey the bi-orthogonality condition of ideal pulse shaping, which is given as

$$\begin{aligned} & \int g_{\text{tx}}(t - nT) e^{j2\pi m \Delta f (t - nT)} g_{\text{rx}}^*(t - n'T) \\ & \times e^{-j2\pi m' \Delta f (t - n'T)} dt = \delta[n - n'] \delta[m - m'], \end{aligned} \quad (2.11)$$

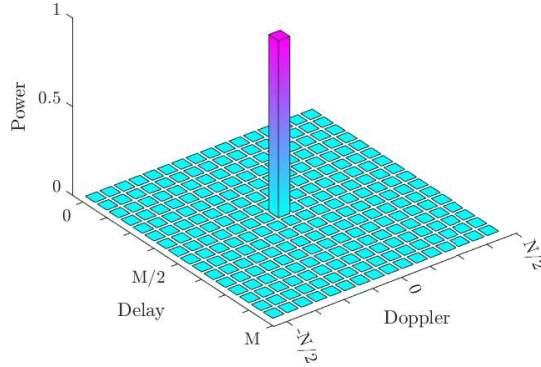
where $n' \in \{0, \dots, M-1\}$ and $m' \in \{0, \dots, N-1\}$ denote the time and frequency indices, respectively. Lastly, as shown in **Figure 1.3**, the DD domain samples of the OTFS frame

are recovered as

$$y[k, l] = \frac{1}{\sqrt{NM}} \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} r_{\text{TF}}[n, m] e^{-j2\pi(\frac{nk}{N} - \frac{ml}{M})}, \quad (2.12)$$



(a): Conventional 3-D OTFS carrier in DD **(b):** Received OTFS carrier when $\kappa = 0.3$. domain.



(c): Received OTFS carrier when $\kappa = 0$.

Figure 2.3: The effect of IDI in OTFS carrier under fractional Doppler.

where $r_{\text{TF}}[n, m]$ represents the discrete samples of TF domain received signal $R(t, f)$. After substituting (2.9) into (2.10) and taking into consideration the ambiguity function in (2.11), it is shown that not only the channel parameters, i.e., L , τ_i , and ν_i , but also the DD characteristics of the pulse design affect the received signal, which is expressed as

$$y[k, l] = \sum_{k'=0}^{N-1} \sum_{l'=0}^{M-1} x[k', l'] h_w[k - k', l - l'] + \hat{n}[k, l], \quad (2.13)$$

where $\hat{n}[k, l] = \sum_{k'=0}^{N-1} \sum_{l'=0}^{M-1} n[k', l']$ denotes the Gaussian noise. $h_w[k - k', l - l']$ is

the effective channel that is composed of wireless channel parameters and pulse shaping characteristics as in [102]

$$h_w[k, l] = \sum_{i=0}^{L-1} h_i w(k - k_i - \kappa_i, l - l_i) e^{-j2\pi \frac{(k_i + \kappa_i)l_i}{NM}}, \quad (2.14)$$

where $w(k - k_i - \kappa_i, l - l_i)$ is the filtering shape of OTFS carrier in DD domain with ideal pulse shaping. This pulse shaping is given in [104] as

$$w(k, l) = \frac{1}{\sqrt{NM}} \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} e^{-j2\pi n \frac{k}{N}} e^{-j2\pi m \frac{l}{M}} \quad (2.15a)$$

$$= \mathcal{G}_N^{Ideal}(k) \mathcal{F}_M^{Ideal}(l), \quad (2.15b)$$

where, $\mathcal{G}_N^{Ideal}(k)$ and $\mathcal{F}_M^{Ideal}(l)$ are given as follows

$$\mathcal{G}_N^{Ideal}(k) = e^{j(N-1)\phi_k} \mathcal{D}_N(N\phi_k) \quad (2.16a)$$

$$\mathcal{F}_M^{Ideal}(l) = e^{j(M-1)\phi_l} \mathcal{D}_M(M\phi_l), \text{ and} \quad (2.16b)$$

$$\mathcal{D}_N(\phi) = \frac{\sin(N\phi)}{N\sin(\phi)}, \quad (2.16c)$$

where $\phi_k = -(\pi/N)(k - k' - k_i - \kappa_i)$ and $\phi_l = -(\pi/M)(l - l' - l_i)$. $\mathcal{D}_N(\cdot)$ is the Dirichlet function, representing the impact of pulse shaping of OTFS transmission in DD domain. In this line, after substituting (2.14), (2.15), and (2.16) into (2.13), the mathematical model of received signal with ideal pulse shaping in DD domain is given as

$$y[k, l] = \sum_{i=0}^{L-1} h_i e^{-j2\pi \frac{(k_i + \kappa_i)l_i}{NM}} \left[\frac{1}{N} \sum_{q=-N/2}^{N/2-1} \mathcal{D}_N(\phi'_q) e^{j\phi'_q(N-1)} \times x[[k - k_i + q]_N, [l - l_i]_M] \right] + \hat{n}[k, l], \quad (2.17)$$

where $\phi'_q = -(\pi/N)(-q - \kappa_i)$ and $q = [k_i + k' - k]$.

This paper considers practical rectangular pulses as potential signaling for OTFS-based transmission where their power representation is visualized in **Figure 2.3(a)**. This power depends on the multiplication of two Dirichlet functions with respect to delay and Doppler

axes as in [25]. In this regard, the received signal in DD domain can be expressed as

$$y[k, l] = \sum_{i=0}^{L-1} h_i e^{j2\pi(\frac{l-l_i}{M})(\frac{k_i+\kappa_i}{N})} \left[\sum_{q=-N/2}^{N/2-1} \alpha_i(k, l, q) \times x[[k - k_i + q]_N, [l - l_i]_M] \right] + \hat{n}[k, l], \quad (2.18)$$

where

$$\alpha_i(k, l, q) = \begin{cases} \frac{1}{N}(\mathcal{D}_N(\phi'_q) - 1)e^{-j2\pi\frac{[k-k_i+q]_N}{N}}, & 0 \leq l < l_i, \\ \frac{1}{N}\mathcal{D}_N(\phi'_q), & l_i \leq l < M, \end{cases} \quad (2.19)$$

which is shown in **Figure 2.3(b)**. The amount of IDI caused by fractional Doppler given in (2.19), it is spread over the all Doppler axes (k) as in the case of ideal pulse shaping given in (2.17) due to the loss of orthogonality in (2.16c). Therefore, control of the energy introduced by the Dirichlet pulse shaping is imperative to alleviate the effect of IDI in (2.19). In particular, the maximum sparsity (L) in DD domain occurs in case of zero IDI, i.e., channel taps with only integer delay and Doppler shifts, as illustrated in **Figure 2.3(c)**. Thus, $y[k, l]$ with $\kappa_i=0$ is given as

$$y[k, l] = \sum_{i=0}^{L-1} h_i e^{j2\pi\frac{k_i}{N}(\frac{l-l_i}{M})} \alpha_i(k, l) \times x[[k - k_i]_N, [l - l_i]_M] + \hat{n}[k, l], \quad (2.20)$$

where

$$\alpha_i(k, l) = \begin{cases} \frac{N-1}{N}e^{-j2\pi\frac{[k-k_i]_N}{N}}, & 0 \leq l < l_i, \\ 1, & l_i \leq l < M. \end{cases} \quad (2.21)$$

Unlike to (2.18), in (2.20) the impact of the Dirichlet shaping $\mathcal{D}_N(\phi'_q)$ does not appear in all the Doppler indices. Hence, $y[k, l]$ maximizes the sparsity features.

Remark 1. Regarding (2.20), we see that the received signal in DD domain $y[k, l]$ depends not only on the environment parameters, i.e., h_i , additional phase component $e^{-j2\pi\frac{(k_i+\kappa_i)l_i}{NM}}$, and L , but also on the $\mathcal{D}_N(\phi'_q)$. Thus, the desired signal $y[k, l]$ is composed of not only L replicas, coming from L reflectors in the environment but also all other Doppler indices $q \neq 0$ contribute to the desired data symbol. As a result, $y[k, l]$ suffers from severe IDI, which depends on the characteristics of the $\mathcal{D}_N(\phi'_q)$.

CHAPTER 3

3. EXPERIMENTAL PART

In this section, we offer the necessary experimental setups employed for the studies conducted throughout this Ph.D. thesis.

3.1. Single Carrier For Beyond 5G Wireless Systems

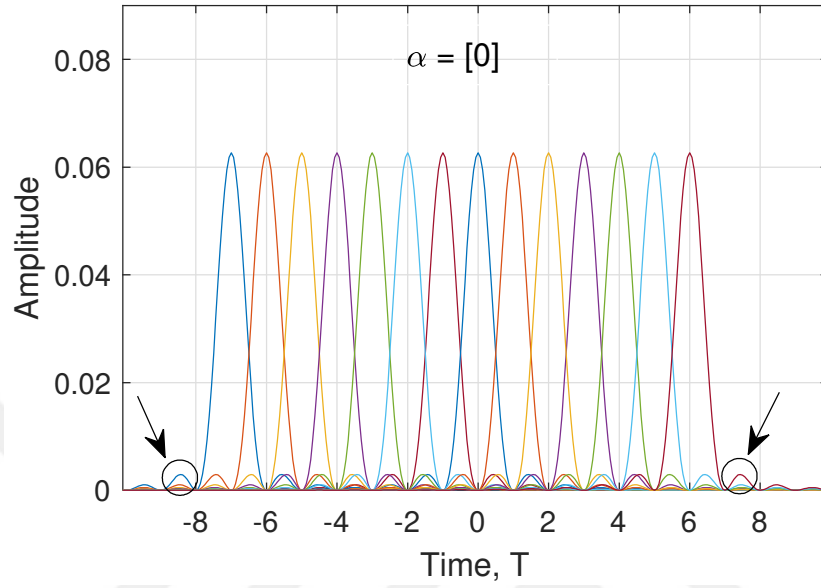
3.1.1. Proposed single carrier frame design

In conventional SC transmission, RRC pulses with the same α factor are used within a block as illustrated in **Figure 3.1(a)**. In this paper, adaptive α usage within a block is proposed in order to achieve higher spectral efficiency, while maintaining low OOB and high reliability. In the proposed SC system, RRC pulses with low α values for inner symbols, and RRC pulses with higher α values for edge symbols are considered as shown in **Figure 3.1(b)**. The side lobes of edge RRC pulses easily fade owing to high α factor.

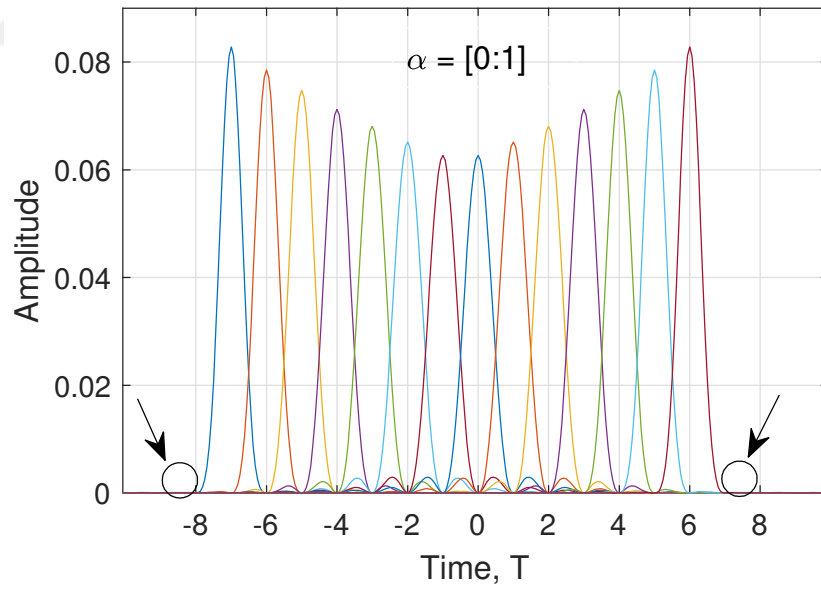
3.1.2. Optimization of SC system parameters

As previously described, system reliability, transmission latency, spectral efficiency, OOB and PAPR relies on utilized α and ζ pair. In order to achieve desired reliability and latency requirements for different URLLC applications, resource optimization in both time and frequency domain is necessary by α and ζ pair, which avoids loss of system efficiency.

Overall system efficiency (η) of SC transmission is proportional to the multiplication of time efficiency and frequency efficiency, and it is expressed as



(a): Conventional SC transmission using fixed α factor.



(b): Proposed SC transmission using adaptive α factor.

Figure 3.1: Pulse shaping for conventional SC transmission and the proposed SC transmission.

$$(\zeta, \alpha) = \underset{\zeta, \alpha}{\operatorname{argmax}}(\eta_t \times \eta_f), \quad (3.1)$$

where η_t and η_f represent efficiency in time and frequency domain, respectively. Time domain efficiency is the ratio of number of data symbols to block size as follows

$$\eta_t = \frac{S}{S + \zeta + 1}, \quad (3.2)$$

while frequency domain efficiency depends on the use of α factor as

$$\eta_f = \frac{1}{1 + \alpha}. \quad (3.3)$$

3.2. Investigation Of Frequency Domain IM

In this section, general system model of frequency domain IM is first revised. Later, mathematical models of the received signal under CFO and IQ imbalance are provided in particular.

3.2.1. System model of potential frequency domain IM for beyond 5G

In this work, it is considered a classical OFDM block with N subcarriers, which is operating on a frequency-selective Rayleigh fading channel. Number of conveyed information bits is $m = N \cdot \log_2(M)$, where M denotes modulation order. After information bits are mapped to subcarriers, frequency domain signal X passes over serial to parallel converter process. Later, Inverse Fast Fourier Transform (IFFT) is applied over X to obtain time domain samples of OFDM block as follows

$$x(n) = \sum_{k=0}^{N-1} X(k) e^{j \frac{2\pi n k}{N}}, \quad 0 \leq n \leq N-1. \quad (3.4)$$

The transmitted signal is sent through an independent and identically distributed (i.i.d) Rayleigh channel $h(n) = \sum_{l=0}^{L-1} h(l) \delta(n-l)$ as follows

$$y(n) = \sum_{l=0}^{L-1} h(l) x(n-l) + w(n), \quad (3.5)$$

where $w(n)$ represents complex Gaussian noise with zero mean and variance σ_w^2 . L denotes the number of channel taps, and path gains $h(l)$ are Gaussian random variables with

$CN(0, \frac{1}{L})$. The frequency domain input-output relationship for the k -th subcarrier is

$$Y(k) = X(k)H(k) + W(k), \quad (3.6)$$

where $W(k)$ and $H(k)$ denote the frequency response of noise $w(n)$ and channel $h(n)$, respectively. Moreover, a cycle prefix (CP) with length L is appended on $x(n)$ to prevent inter-symbol interference.

Different from classical OFDM, OFDM-IM carries data information not only by subcarriers but also by indices of active subcarriers. N size OFDM block is equally split in G subblock with $n = \frac{N}{G}$ subcarriers per subblock. Number of active subcarriers is fixed to k out of n in each subblock. $p_1 = \lfloor \log_2 C(n, k) \rfloor$ bits are carried by indices of active subcarriers where $\lfloor . \rfloor$ and $C(n, k)$ denote floor function and binomial coefficient, respectively. $p_2 = k \cdot \log_2(M)$ bit stream is converted to M -ary symbols s_i and carried by active subcarriers, with $s_i \in [s_1, s_2, \dots, s_M]$. Hence, total number of data bits conveyed by one OFDM-IM block is

$$m = gp = g(p_1 + p_2), \quad (3.7)$$

where $p = p_1 + p_2$ is the number of bits per OFDM subblock.

In OFDM-GIM, different number of active subcarriers can be utilized in a subblock. The set of all allowed number of active subcarriers is $\mathbf{a} = \{k(1), k(2), \dots, k(b)\}$, where $\xi \in [1, 2, \dots, n]$ and total number of transmitted bits per subblock is

$$p = \lfloor \log_2 \left(\sum_{\xi=1}^b M^{k(\xi)} \cdot C(n, k(\xi)) \right) \rfloor. \quad (3.8)$$

Note that OFDM-IM and OFDM-GIM transmits information bits through indices of active subcarrier. On the other hand, number of active subcarriers is mapped to information bits in OFDM-SNM. Therefore, $p_1 = \lfloor \log_2 N \rfloor$, and $k \in [1, \dots, n]$ is decided according to coming p_1 bits. Having decided k , $p_2 = k \cdot \log_2(M)$. For further details about mapping of information bits to either subcarrier indices in OFDM-GIM or subcarrier number in OFDM-SNM refer to [36] and [37].

In this study, a log-likelihood ratio (LLR) detector is used at the Rx to identify indices and the number of active subcarriers for a given subblock. The Rx calculates logarithmic ratio $\lambda(k)$ of *a posteriori* probability for active subcarrier and inactive subcarrier within

a subblock as

$$\lambda(k) = \ln\left(\frac{k}{n-k}\right) + \frac{|Y(k)|^2}{N_0} + \ln\left(\sum_{i=1}^M \exp\left(-\frac{|Y(k) - H(k)s_i|^2}{N_0}\right)\right), \quad (3.9)$$

where k subcarriers with maximum λ values represent indices of the utilized subcarriers that are mapped to p_1 bits. Later, M-ary symbols s_i carried by detected active subcarriers are demodulated to get p_2 data bits.

In OFDM-GIM, a two-stage Rx is used. Firstly, LLR Rx is applied for every $k(\xi) \in \mathbf{a}$, and corresponding $\lambda(\xi)$ values are evaluated. Secondly, indices of active subcarrier (I) and M-ary symbols (s_a) for $k(\xi)$ which provide minimum distance between received signal and reconstructed signal are selected as

$$\begin{aligned} (I, s_a) = \operatorname{argmax}_{I(\xi), s} & \left(\sum_{k=1}^n |Y(k)|^2 \right. \\ & \left. + \sum_{\gamma=1}^{k(\xi)} |Y(I(\gamma)) - H(I(\gamma))s(\gamma)|^2 \right). \end{aligned} \quad (3.10)$$

Then, I and s_a are mapped to p_1 and p_2 bit sequences, respectively.

3.2.2. Carrier frequency offset model

The CFO occurs because of mismatch between Tx carrier frequency (f_{tx}) and Rx carrier frequency (f_{rx}), as shown in **Figure 1.2(a)**. In this work, it is assumed that CFO does not change during one OFDM symbol. The received time domain OFDM block with CFO equals [116]

$$\hat{y}(n) = y(n)e^{\frac{j2\pi n\Delta\omega}{N}}, \quad (3.11)$$

where $\Delta\omega = \frac{f_{tx} - f_{rx}}{\Delta f}$ denotes normalized CFO, and Δf is subcarrier spacing. The frequency response of $\hat{y}(n)$ equals

$$\hat{Y}(k) = c_0 X(k) H(k) + \underbrace{\sum_{i=0, i \neq k}^{N-1} c_{i-k} X(i) H(i)}_{\text{interference}} + D(k), \quad (3.12)$$

where c_{i-k} represents the inter-carrier interference component

$$c_{i-k} = \frac{\sin(\pi(i-k+\Delta\omega))}{N \sin(\frac{\pi}{N}(i-k+\Delta\omega))}, \quad (3.13)$$

with $i, k \in [0, 1, \dots, N-1]$. $D(k)$ represents colored noise. In (3.12), c_0 clearly shows amplitude distortion and phase rotation that desired signal $X(k)$ experiences, where

$$c_0 = \frac{\sin(\pi\Delta\omega)}{N\sin(\frac{\pi}{N}\Delta\omega)}, \quad (3.14)$$

with $i = k$. Moreover, besides amplitude and phase impairment, all the other $N-1$ remaining subcarriers cause interference in k -th subcarrier with $k \neq i$.

3.2.3. IQ imbalance model at Rx in frequency domain

The effect of IQ imbalance occurs due to the mismatch in amplitude ($\epsilon \neq 0$) and phase ($\Delta\phi \neq 0$) between I and Q branches, as illustrated in **Figure 1.2(b)**. The received baseband signal with IQ imbalance $\hat{y}(n)$ can be represented as

$$\hat{y}(n) = \alpha y(n) + \beta y^*(n), \quad (3.15)$$

where $(.)^*$ denotes complex conjugate. α and β denote the distortion parameters due to IQ imbalance, which are modeled as [39]

$$\alpha = \cos(\Delta\phi) + j\epsilon\sin(\Delta\phi), \quad (3.16)$$

$$\beta = \epsilon\cos(\Delta\phi) - j\sin(\Delta\phi). \quad (3.17)$$

By substituting $Y(k)$ from (3.6) into (3.15), frequency domain signal with IQ imbalance $\hat{Y}(k)$ is expressed as

$$\hat{Y}(k) = \alpha X(k)H(k) + \underbrace{\beta X^*(-k)H^*(-k)}_{\text{interference}} + Z(k), \quad (3.18)$$

where $Z(k) = \alpha W(k) + \beta W^*(-k)$, and it has a Gaussian distribution with $Z(k) \sim \mathcal{N}(0, N'_0)$ and $N'_0 = (1 + \epsilon^2)N_0$.

As seen from (3.18), IQ imbalance degrades the power of $X(k)$ and causes interference coming from $X(-k)$. If $X(-k) = 0$, $X(k)$ is only exposed to self-degradation. Furthermore, there is no degradation and zero interference for $\alpha = 1$ and $\beta = 0$, which correspond to $\epsilon = 0$ and $\Delta\phi = 0^\circ$.

3.3. Proposed Transceiver Design For Hybrid NOMA

The system model of proposed IM-NOMA is illustrated in **Figure 3.2**, where near UE (N_{UE}) and far UE (F_{UE}) perform OFDM and OFDM-IM transmission, respectively. A single antenna is used at Tx and Rx, while UEs share N subcarriers of an OFDM block via superposition coding. Total power P is fractionally split among UEs considering UEs' channel gains and target rate of F_{UE} ($R_{F_{UE}}$). Corresponding power allocation factors for N_{UE} and F_{UE} are β_1 and β_2 , respectively, with $P = \sum_{u=1}^2 \beta_u P$. After superposition, the resultant signal is expressed as follows

$$\mathbf{x}_f = \sum_u \sqrt{\beta_u P} \mathbf{x}_{f,u}, \quad u \in [1, 2], \quad (3.19)$$

where $\mathbf{x}_{f,u} = [x_{f,u}(1) \ x_{f,u}(2) \ \dots \ x_{f,u}(N)]^T$ denotes the frequency domain signal of u -th UE. N_{UE} with OFDM ($\mathbf{x}_{f,1}$) modulates all subcarriers to convey data bits and thus $x_{f,1}(k) \in S$, where S is the set of M -ary symbols $S = \{s_0 \ s_1 \ \dots \ s_{M-1}\}$. Differently, F_{UE} with OFDM-IM ($\mathbf{x}_{f,2}$) carries information not only by subcarriers but also by indices of active ones with $x_{f,2}(k) \in \{0, S\}$ [32]. In OFDM-IM, OFDM block is equally split into G subblocks of $b = \frac{N}{G}$ subcarriers each, where only a out of b subcarriers are utilized. A given subblock realization of selected subcarrier indices is given as $\eta_g = \{i_{g,1}, \dots, i_{g,b}\}$ with $i_{g,l} \in [1, \dots, b]$, $g = 1, \dots, G$, and $l = 1, \dots, a$. This partial resource activation of $\frac{a}{b}$ increases the power of active subcarriers to $\frac{b}{a} \beta_2 P$. The total bit number conveyed by F_{UE} equals $m = G([\log_2 C(b, a)] + a \log_2(M)) = Gp$. $\mathbf{x}_{f,2}$ is generated after combination of all G subblocks.

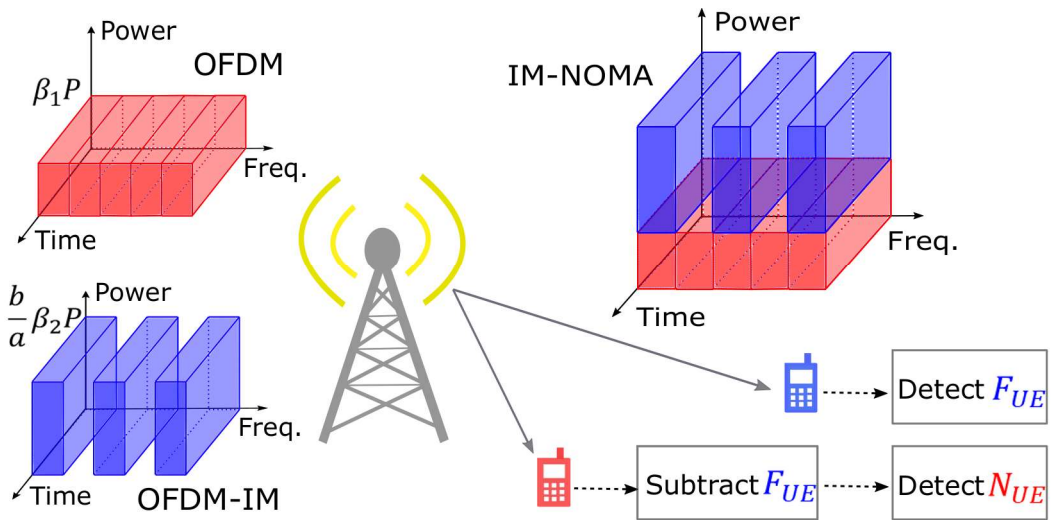


Figure 3.2: Downlink IM-NOMA by superposition of OFDM and OFDM-IM.

Later, inverse Fast Fourier Transform (IFFT) is applied to obtain time domain samples of $\mathbf{x}_f$. A cyclic prefix (CP) is added to eliminate the influence of Rayleigh channel $\mathbf{h}_{t,u} = [h_{t,u}(1) \ h_{t,u}(2) \ \dots \ h_{t,u}(L)]^T$ with L Gaussian random variable taps following $\mathcal{CN}(0, \frac{1}{L})$ distribution. After channel propagation, the received frequency domain signal is expressed as

$$\mathbf{y}_{f,u} = \mathbf{X}_f \mathbf{h}_{f,u} + \mathbf{w}_{f,u}, \quad (3.20)$$

where $\mathbf{X}_f = \text{diag}\{\mathbf{x}_f\}$. $\mathbf{w}_{f,u}$ and $\mathbf{h}_{f,u}$ denote additive noise $w_{f,u} \sim \mathcal{CN}(0, \sigma_{o,u}^2)$ and channel $h_{f,u} \sim \mathcal{CN}(0, \sigma_{f,u}^2)$ for u -th UE, respectively.

3.3.1. Multiuser detection for different power multiplexing

In case of $\beta_1 < \beta_2$ to meet R_{FUE} considering the channel gains of UEs, MUD of $\mathbf{y}_{f,u}$ is performed via SIC by utilizing the power difference between UEs under the assumption of known channel state information (CSI) at BS and Rx. The received signals at N_{UE} and F_{UE} are given as

$$\mathbf{y}_{f,1} = \left(\sqrt{\beta_1 P} \mathbf{X}_{f,1} + \sqrt{\beta_2 P} \mathbf{X}_{f,2} \right) \mathbf{h}_{f,1} + \mathbf{w}_{f,1}, \quad (3.21a)$$

$$\mathbf{y}_{f,2} = \left(\sqrt{\beta_1 P} \mathbf{X}_{f,1} + \sqrt{\beta_2 P} \mathbf{X}_{f,2} \right) \mathbf{h}_{f,2} + \mathbf{w}_{f,2}. \quad (3.21b)$$

In IM-NOMA, communication reliability lays in the detection of F_{UE} carrying additional IM data. Thus, maximum likelihood (ML) decoder is applied in a subblock-wise manner, and it can jointly estimate data bits carried not only by the M -ary symbols but also by the active indices. ML criterion for g -th subblock is given as

$$\hat{\mathbf{x}}_{f,2,1}^g = \arg \min_{\mathbf{x}_{f,2}^g} \|\mathbf{y}_{f,1}^g - \mathbf{X}_{f,2}^g \mathbf{h}_{f,1}^g\|^2, \quad (3.22a)$$

$$\hat{\mathbf{x}}_{f,2,2}^g = \arg \min_{\mathbf{x}_{f,2}^g} \|\mathbf{y}_{f,2}^g - \mathbf{X}_{f,2}^g \mathbf{h}_{f,2}^g\|^2. \quad (3.22b)$$

In order to get $\mathbf{x}_{f,1}$, N_u needs to detect $\mathbf{x}_{f,2}$ and remove from $\mathbf{y}_{f,1}$. In case of correct detection $\hat{\mathbf{x}}_{f,2,1}^g \cong \mathbf{x}_{f,2}$, the signal of N_{UE} (3.21a) with SIC can be written as follows

$$\mathbf{y}_{f,1} = \sqrt{\beta_1 P} \mathbf{X}_{f,1} \mathbf{h}_{f,1} + \mathbf{w}_{f,1}. \quad (3.23)$$

N_{UE} applies classical M -ary demodulation after removing $\hat{\mathbf{x}}_{f,2,1}$ by SIC and channel effect via zero-forcing technique.

3.3.2. Multiuser detection for equal power multiplexing

In OFDM-NOMA, SIC leads to a catastrophic error performance in case comparable $\beta_1 \cong \beta_2$ are assigned to UEs. Due to partial subcarrier activation of F_{UE} , $y_{f,1}(k)$ for $x_{f,2}(k) \neq 0$ mathematically can be rewritten as

$$y_{f,1}(k) = h_{f,1}(k) \left(\sqrt{\beta_1 P} x_{f,1}(k) + \sqrt{\frac{b}{a}} \beta_2 P x_{f,2}(k) \right) + w_{f,1}(k), \quad (3.24)$$

where both users provide transmission over the same radio resources, otherwise ($x_{f,2}(k) = 0$)

$$y_{f,1}(k) = h_{f,1} \sqrt{\beta_1 P} x_{f,1}(k) + w_{f,1}(k), \quad (3.25)$$

where N_{UE} performs orthogonal transmission. In this way, the performance of proposed IM-NOMA not only depends on the power difference but also on the portion of non-orthogonal transmission between UEs. It is important to emphasize that F_{UE} with OFDM-IM has higher power than N_{UE} only due to partial subcarrier activation, as seen in (3.24). However, this power difference is not higher than the power of N_{UE} on the subcarriers, where ($x_{f,2}(k) = 0$). This leads to wrong detection of active subcarriers and M -ary data symbols for F_{UE} when the conventional SIC is utilized. Motivated from this, a new Rx design is proposed for IM-NOMA to sustain the communication when power multiplexing is inapplicable. It is illustrated in **Figure 3.3** and explained below.

1. Since N_{UE} experiences partial interference, $\tilde{\mathbf{x}}_{f,1}$ is detected and removed from $\mathbf{y}_{f,u}$ while treating F_{UE} as noise. In this way, $\tilde{\mathbf{x}}_{f,2}$ signal format is similar to $\mathbf{x}_{f,2}$.
2. ML detector is applied by F_{UE} to estimate $\hat{\mathbf{x}}_{f,2} \cong \mathbf{x}_{f,2}$.
3. Lastly, N_{UE} removes $\hat{\mathbf{x}}_{f,2}$ from $\mathbf{y}_{f,1}$ to get $\hat{\mathbf{x}}_{f,1} \cong \mathbf{x}_{f,1}$.

Due to ML Rx, the total complexity of F_{UE} and N_{UE} equal $O(2^{p_1} M^a)$ and $O(M(1 + 2^{p_1} M^{a-1}))$ while $\beta_1 < \beta_2$, and an additional complexity $O(M)$ is introduced due hard decision process while $\beta_1 \cong \beta_2$.

3.3.3. Performance analysis for hybrid NOMA

In order to provide beneficial insights for the proposed IM-NOMA, theoretical analysis of achievable sum rate and ABEP are evaluated for both N_{UE} and F_{UE} .

3.3.3.1. Achievable Rate Analysis

According to (3.23), under the assumption of successful decoding of F_{UE} 's signal, mathematical expression for N_{UE} 's achievable rate is written as

$$R_{N_{UE}} = \log_2 \left(1 + \frac{|h_{f,1}(k)|^2 |x_{f,1}(k)|^2 \beta_1 P}{\sigma_{o,1}^2} \right). \quad (3.26)$$

On the other hand, the rate calculation of F_{UE} is not straightforward since OFDM-IM signal performs a subblock-based modulation. Hence, achievable rate of F_{UE} is given as [117]

$$\begin{aligned} R_{F_{UE}} &= \frac{1}{b} \mathbf{I}(\mathbf{x}_f^g; \mathbf{y}_{f,2}^g | \mathbf{h}_{f,2}^g) \\ &= \frac{1}{b} (p - E_h[H(\mathbf{x}_f^g | \mathbf{y}_{f,2}^g, \mathbf{h}_{f,2}^g)]) \end{aligned} \quad (3.27)$$

considering the mutual information between $\mathbf{x}_f^g$ and $\mathbf{y}_{f,2}^g$. In order to drive $H(\mathbf{x}_f^g | \mathbf{y}_{f,2}^g, \mathbf{h}_{f,2}^g)$, probability density functions of $f(\mathbf{y}_{f,2}^g | \mathbf{x}_f^g, \mathbf{h}_{f,2}^g)$ and $f(\mathbf{y}_{f,2}^g | \mathbf{h}_{f,2}^g)$ are needed and given as

$$\begin{aligned} f(\mathbf{y}_{f,2}^g | \mathbf{x}_f^g, \mathbf{h}_{f,2}^g) &= \frac{1}{(\pi \sigma_{o,2}^2)^b} \exp(-\|\mathbf{y}_{f,2}^g - \mathbf{X}_f^g \mathbf{h}_{f,2}^g\|^2) \\ f(\mathbf{y}_{f,2}^g | \mathbf{h}_{f,2}^g) &= \frac{1}{2^p} \sum_{i=k}^{2^p} f(\mathbf{y}_{f,2}^g | \mathbf{x}_f^g(i), \mathbf{h}_{f,2}^g), \end{aligned} \quad (3.28)$$

where $\mathbf{x}_f^g(i)$ represents the i -th observation of $\mathbf{x}_f^g$ realization. Since the utilization of (3.28) into (3.27) can not give a closed form for R_{UE} , mathematical evaluation of lower bound achievable rate is used as

$$\begin{aligned} R_{F_{UE}} &\triangleq \frac{p}{b} - \log_2(e) - 1 \\ &\quad - \frac{2^{-p}}{b} \sum_{i=1}^{2^p} \left(\log_2 \sum_{j=1}^{2^p} 1 + \frac{1}{\det(\mathbf{I}_b + \vartheta_{i,j})} \right), \end{aligned} \quad (3.29)$$

with $\vartheta_{i,j} = \frac{1}{2\sigma_{o,2}^2} \text{diag}\{\mathbf{x}_f^g(i) - \mathbf{x}_f^g(j)\}^H \text{diag}\{\mathbf{x}_f^g(i) - \mathbf{x}_f^g(j)\}$. Total lower bound achievable rate equals $R = R_{N_{UE}} + R_{F_{UE}}$.

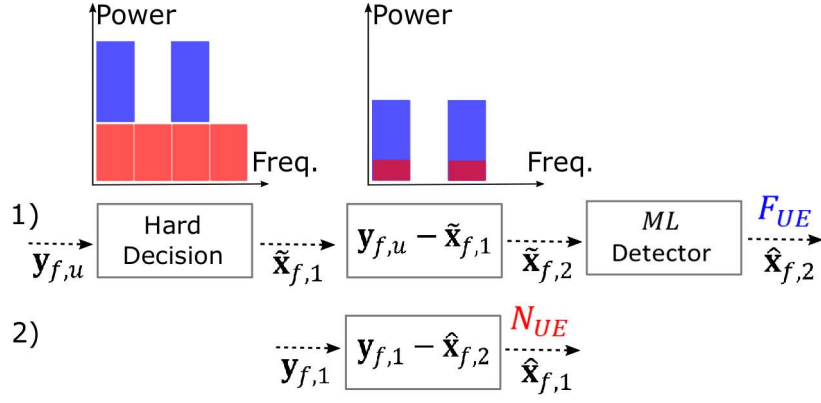


Figure 3.3: Proposed Rx for IM-NOMA with symmetric channel gains.

3.3.3.2. Average bit error probability analysis

Sharing of radio resources between multiple UEs gives rise to IUI, which significantly decays transmission reliability, in NOMA. Hence, in order to determine the error probability of a given UE, the energy of the received signal should be considered according to superposition coding principles.

3.3.3.2.1 Far user In order to derive error performance of F_{UE} , the impact of interference coming from N_{UE} must be considered. First, conditional pairwise error probability (CPEP) is evaluated to determine ABEP. F_{UE} makes decision error in case $\mathbf{X}_f$ is sent and erroneously detected as $\hat{\mathbf{X}}_f$, which can be a decision error on M -ary symbols or indices of active subcarriers. Since data information carried by IM-NOMA is independent from one block to another, a single subblock is considered for CPEP, which is given as

$$P\left(\mathbf{X}_f^g \rightarrow \hat{\mathbf{X}}_f^g | \mathbf{h}_{f,2}^g\right) = Q\left(\sqrt{\frac{\|(\mathbf{X}_f^g - \hat{\mathbf{X}}_f^g)\mathbf{h}_{f,2}^g\|^2}{2\sigma_{o,2}^2}}\right), \quad (3.30)$$

where $\|(\mathbf{X}_f^g - \hat{\mathbf{X}}_f^g)\mathbf{h}_{f,2}^g\|^2 = (\mathbf{h}_{f,2}^g)^H \mathbf{R} \mathbf{h}_{f,2}^g$, and $\mathbf{R} = (\mathbf{X}_f^g - \hat{\mathbf{X}}_f^g)^H (\mathbf{X}_f^g - \hat{\mathbf{X}}_f^g)$. A well-known approximation of Q-function ($Q(v) = (1/\sqrt{2\pi}) \int_v^\infty \exp(-(v^2/2)) dv$) is adopted in this work as $Q(v) \cong \frac{1}{12}e^{-\frac{v^2}{2}} + \frac{1}{4}e^{-\frac{2v^2}{3}}$ [118]. It is important to note that this approximation provides a very tight result in case of the conventional OFDM-IM is used, and thus Craig's formula-based analysis can be performed for the advanced IM scenarios [119]. Thus, unconditional pairwise error probability (UPEP) is obtained by averaging over instantaneous

channel states as

$$P\left(\mathbf{X}_f^g \rightarrow \hat{\mathbf{X}}_f^g\right) \cong \mathbb{E}_{\mathbf{h}_{f,2}^g} \left\{ \frac{e^{-\frac{E_s \delta}{4\sigma_{o,2}^2}}}{12} + \frac{e^{-\frac{E_s \delta}{3\sigma_{o,2}^2}}}{4} \right\} \quad (3.31)$$

$$\cong \frac{1/12}{\det(\mathbf{I}_b + q_1 \mathbf{C}_b \mathbf{R})} + \frac{1/4}{\det(\mathbf{I}_b + q_2 \mathbf{C}_b \mathbf{R})},$$

where $\mathbf{C}_b = \mathbb{E}\left\{\mathbf{h}_{f,2}^g(\mathbf{h}_{f,2}^g)^H\right\}$, $q_1 = \frac{1}{4\sigma_{o,2}^2}$, and $q_2 = \frac{1}{3\sigma_{o,2}^2}$.

Finally, upper bound ABEP of F_{UE} using ML detection is calculated as

$$P_e \leq \frac{1}{p\xi} \sum \sum P\left(\mathbf{X}_f^g \rightarrow \hat{\mathbf{X}}_f^g\right) \varepsilon(\mathbf{X}_f^g, \hat{\mathbf{X}}_f^g), \quad (3.32)$$

where $\varepsilon(\mathbf{X}_f^g, \hat{\mathbf{X}}_f^g)$ is the bit errors that $\mathbf{X}_f^g$ differs from $\hat{\mathbf{X}}_f^g$ and ξ is the number of possible realizations considering both UEs.

3.3.3.2.2 Near user By applying closed-form expressions of BER performances to $\mathbf{y}_{f,1}$ in (3.23), lower bound ABEP of N_{UE} for binary phase shift keying (BPSK) is obtained as [120]

$$P_e = \frac{1}{2} \left(1 - \sqrt{\frac{\gamma_1}{1 + \gamma_1}} \right), \quad (3.33)$$

where $\gamma_1 = \frac{|h_{f,1}(k)|^2 |x_{f,1}(k)|^2 \beta_1 P}{\sigma_{o,1}^2}$ is SNR of N_{UE} after correct detection of F_{UE} . For M -QAM, readers are referred to [120].

3.4. Proposed Frame Design For Multiuser OTFS Systems

3.4.1. System model of multiuser OTFS frame

In this work, we consider a single cell with N_u active users that are served by one base station in downlink transmission, as illustrated in **Figure 3.4**.

The BS selects the UE for transmission based on the corresponding CSI of each UE. Basically, in downlink systems with multi-user transmission, the ultimate goal is to maximize the achievable system throughput, and thus, the BS in classical wireless technologies chooses the UE with the best SNR for the next transmission time. Since the impact of channel fading reduces with the increase of the number of channel taps in OTFS transmission [64], we propose to allocate the radio resources (RR) of the OTFS frame to the UE with the largest number of channel taps, improving the error performance of the overall

system that is formulated as follow

$$P_u \triangleq \max\{P_i\}$$

$$s.t. \ P_{min} \leq P_i \leq P_{max}, \text{ for } i \in [0, 1, \dots, N_u - 1], \quad (3.34)$$

where P_u denotes the path number of the selected UE for OTFS transmission. It is worth mentioning that the number of channel taps for each UE is known at the BS. In this study, we consider a system model with single antennas at the Tx and Rx. The system frame consists of N number of symbols and M number of subcarriers with T symbol duration and Δf subcarrier spacing (SCS), respectively. Therefore, the OTFS frame occupies a total bandwidth of $B = M\Delta f$ with a frame duration $T_f = TN$, while $T\Delta f = 1$. Consequently, the OTFS signal conveys a block of $N \times M$ uncoded Q -ary data symbols $x[k, l] \in \mathcal{A}$, where $\mathcal{A} = \{a_0, a_1, \dots, a_{Q-1}\}$ is the modulation alphabet. $\mathbf{X}_u \in \mathbb{C}^{M \times N}$ represents data symbols in the delay-Doppler grid for the selected user u with $x_u[k, l]$ being an element of $\mathbf{X}_u$. After we apply Inverse Symplectic Fast Fourier Transform (ISFFT), we convert the data symbols to time-frequency domain. Accordingly, we apply N -point inverse Fast Fourier Transform (IFFT) and M -point FFT operation over the rows and columns of $\mathbf{X}_u$, respectively. To generate the time domain OTFS signal, we perform Heisenberg Transform by using M -point IFFT over time-frequency data symbols. Consequently, time domain OTFS frame can be written as $\mathbf{S}_u = \mathbf{G}_{tx} \mathbf{F}_M^H (\mathbf{F}_M \mathbf{X}_u \mathbf{F}_N^H)$, where $\mathbf{G}_{tx}$ denotes a diagonal matrix that has the samples of the pulse shape in time domain, and $\mathbf{F}_N$ denotes the DFT matrix given as

$$[\mathbf{F}_N]_{k,l} = \frac{1}{\sqrt{N}} \exp\left(\frac{-j2\pi kl}{N}\right), \quad k, l \in \{0, 1, \dots, N-1\}. \quad (3.35)$$

The Hermitian of N -point DFT is $\mathbf{F}_N^H$. It is worth mentioning that $\mathbf{G}_{tx}$ corresponds to the identity matrix in case of rectangular pulse shaping, where considering the column-wise vectorization of $\mathbf{S}_u$, we can write it as

$$\mathbf{s}_u = (\mathbf{F}_N^H \otimes \mathbf{G}_{tx}) \mathbf{x}_u, \quad (3.36)$$

where $\otimes$ denotes the Hadamard product and $\mathbf{x}_u = \text{vec}(\mathbf{X}_u)$. We add a cyclic prefix (CP) of L length in order to mitigate the inter-symbol interference between the OTFS signals, and thus, L is larger than the number of channel paths (P_u). After the transmission, the OTFS signal $\mathbf{s}_u$ undergoes the time-varying wireless channel given as $h_u(\tau, \nu) = \sum_{i=0}^{P_u-1} h_u^i \delta(\tau - \tau_u^i) \delta(\nu - \nu_u^i)$, where h_u^i , τ_u^i , ν_u^i , denote the path gain, delay shift and Doppler shift of the

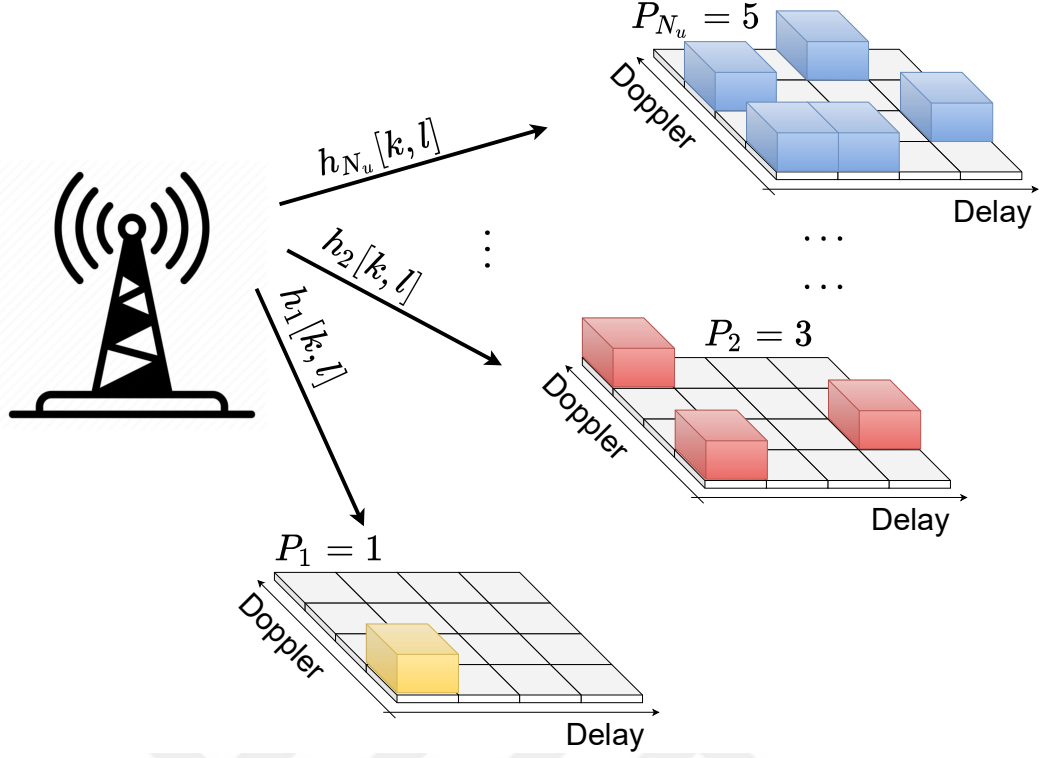


Figure 3.4: Channel model for a multi-user scenario in downlink transmission.

i -th channel path, respectively, and $\delta(\cdot)$ is the Dirac delta function. Moreover, the delay and Doppler bins for the i -th channel tap equal to $\tau_u^i = l_u^i / (M\Delta f)$ and $\nu_u^i = (k_u^i + \kappa) / (NT)$, respectively, where κ denotes the fractional Doppler. Moreover, we neglect the impact of fractional delay, since the typical value of the sampling time $1/(M\Delta f)$ is sufficiently small, especially in wideband wireless technologies.

At the Rx side, after discarding the additional CP, the received signal is given as follows

$$\mathbf{r}_u = \mathbf{H}_u \mathbf{s}_u + \mathbf{w}, \quad (3.37)$$

where $\mathbf{w} \in \mathbb{C}^{MN \times 1}$ denotes the additive white Gaussian noise (AWGN) random variable with $\mathcal{CN}(0, N_0 \mathbf{I})$. $\mathbf{H}_u \in \mathbb{C}^{MN \times MN}$ is the channel matrix given by $\mathbf{H}_u = \sum_{i=0}^{P_u-1} h_u^i \mathbf{\Pi}^{l_i} \mathbf{\Delta}^{k_i}$, where $\mathbf{\Pi} = \text{circ}[0, 1, 0, \dots, 0]_{MN \times MN}$ is a permutation matrix based on forward cyclic shift, and $\mathbf{\Delta} = \text{diag}[z^0, z^1, \dots, z^{MN-1}] \in \mathbb{C}^{MN \times MN}$ with $z = \exp(j2\pi/MN)$. In order to get the delay-Doppler domain symbols, we first perform Wigner transform as $\mathbf{F}_M \mathbf{G}_{rx} \mathbf{R}_u$, given that $\mathbf{R}_u = \text{vec}^{-1}(\mathbf{r}_u)$ and $\mathbf{G}_{rx}$ is a diagonal matrix containing samples of the Rx pulse shape. Following that, we perform SFFT operation to get the received signal in the delay-Doppler

domain as $\mathbf{Y}_u = \mathbf{F}_M^H (\mathbf{F}_M \mathbf{G}_{rx} \mathbf{R}_u) \mathbf{F}_N$. After vectorization, the received signal is written as

$$\mathbf{y}_u = \mathbf{H}_{u,eff} \mathbf{x}_u + \bar{\mathbf{w}}, \quad (3.38)$$

where $\mathbf{H}_{u,eff} = (\mathbf{F}_N \otimes \mathbf{G}_{rx}) \mathbf{H}_u (\mathbf{F}_N^H \otimes \mathbf{G}_{tx})$ is the effective channel and $\bar{\mathbf{w}}$ is the noise in the delay-Doppler domain. In case of using a rectangular pulse shape, i.e., $\mathbf{G}_{rx} = \mathbf{I}_M$, $\bar{\mathbf{w}}$ preserves the same statistical features of $\mathbf{w}$. Regarding (3.125), we use either maximum likelihood (ML) or minimum mean square error (MMSE) detector to estimate the data symbols as follows

$$\hat{\mathbf{x}}_{u,ML} = \arg \min_{\mathbf{x} \in \mathcal{A}^{NM}} \|\mathbf{y} - \mathbf{H}_{u,eff} \mathbf{x}_u\|^2, \quad (3.39a)$$

$$\hat{\mathbf{x}}_{u,MMSE} = \mathbf{H}_{u,eff}^H (\mathbf{H}_{u,eff} \mathbf{H}_{u,eff}^H + N_0 \mathbf{I})^{-1} \mathbf{y}_u, \quad (3.39b)$$

where $\hat{\mathbf{x}}_u = \text{vec}(\hat{\mathbf{X}}_u)$ and we demodulate $\hat{\mathbf{x}}_u \in \mathbb{C}^{MN \times 1}$ symbols considering the modulation alphabet ($\mathcal{A}$) in order to get the information bits.

3.4.2. Performance analysis of multiuser scheduling in delay-Doppler domain

In this section, we outline the theoretical analysis in order to investigate the performance of the proposed multi-user OTFS scheme. In particular, we derive the closed-form expression of the average bit error rate (ABER). To analyze the overall error performance, we rewrite the input-output relation (3.38) as follows

$$\mathbf{y}_u = \Phi_{\mathbf{v}_u, \tau_u}(\mathbf{x}_u) \mathbf{h}_u + \mathbf{w}, \quad (3.40)$$

where $\mathbf{h}_u = [h_u^0, h_u^1, \dots, h_u^{P_u-1}]^T$, and each element of $\mathbf{h}_u$ represents an independent and identically distributed complex Gaussian random variable as $h_u^i \sim \mathcal{CN}(\mu_u, 1/P_u)$. $\Phi_{\mathbf{v}_u, \tau_u}(\mathbf{x}_u)$ is a concatenated matrix with size $MN \times P_u$ that involves the data vector $\mathbf{x}_u$ given as

$$\Phi_{\mathbf{v}_u, \tau_u}(\mathbf{x}_u) = [\Psi_0 \mathbf{x}_u \mid \Psi_1 \mathbf{x}_u \mid \dots \mid \Psi_{P_u-1} \mathbf{x}_u], \quad (3.41)$$

where $\Psi_i = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{\Pi}^{l_i} \Delta^{k_i} (\mathbf{F}_N^H \otimes \mathbf{I}_M)$. In this work, $\mathbf{v}_u$ and τ_u represent the vectors of Doppler values and delay values for the user of interest, respectively, where $\mathbf{v}_u = [k_u^0, k_u^1, \dots, k_u^{P_u-1}]$ and $\tau_u = [l_u^0, l_u^1, \dots, l_u^{P_u-1}]$. Moreover, these indices follow a discrete uniform distribution. To calculate the pairwise error probability (PEP), we assume ML estimation and ideal CSI at the Rx.

In particular, for this work, we consider the number of channel taps of each user as a

sequence $Z_0, Z_1, \dots, Z_{N_u-1}$ of i.i.d random variables with support on $[P_{\min}, P_{\max}]$ and following a probability density function (PDF) and cumulative distribution function (CDF) given by $f_Z(z)$ and $F_Z(z)$, respectively. In this line, we use the random variable P_u to model the user performing transmission with the maximum number of channel taps, given as $P_u = \max_{0 \leq i \leq N_u-1} Z_i$, where the PDF of P_u is derived in the followings. Considering the fact that $P_u < z$ if and only if all elements of the sample are less than z , and Z_i 's are i.i.d random variables, we can deduce that

$$\Pr\{P_u \leq z\} = \Pr\{Z_0 \leq z, \dots, Z_{N_u-1} \leq z\}, \quad (3.42a)$$

$$= \prod_{i=0}^{N_u-1} \Pr(Z_i \leq z) = [F_Z(z)]^{N_u}, \quad (3.42b)$$

where $[F_Z(z)]^{N_u}$ represents the CDF of P_u . Hence, after some mathematical manipulations, the PDF of P_u is given as

$$f_{P_u}(z) = N_u [F_Z(z)]^{N_u-1} f_Z(z). \quad (3.43)$$

In case Z_i is uniformly distributed on $[P_{\min}, P_{\max}]$, the PDF of P_u is given as follows

$$f_{P_u}(z) = N_u \frac{(z - P_{\min})^{N_u-1}}{(P_{\max} - P_{\min})^{N_u}}. \quad (3.44)$$

To assess the error performance of the proposed OTFS technique, we utilize the union bound error performance, which is given as

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u\} \leq \int_{P_u} \Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u | P_u\} f_{P_u}(P_u) dP_u, \quad (3.45)$$

where $\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u | P_u\}$ denotes the PEP in case the transmitted symbol block $\mathbf{x}_u$ is erroneously detected as $\hat{\mathbf{x}}_u$ for the desired UE with P_u channel spreading. In line with the principle of ML Rx, we give the conditional PEP as follows

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u | \mathbf{h}_u, \mathbf{v}_u, \tau_u, P_u\} = \Pr\{||\mathbf{y}_u - \Phi_{\mathbf{v}_u, \tau_u}(\mathbf{x})_u \mathbf{h}_u||^2 > ||\mathbf{y}_u - \Phi_{\mathbf{v}_u, \tau_u}(\hat{\mathbf{x}})_u \mathbf{h}_u||^2\}, \quad (3.46)$$

where, after performing some mathematical manipulations and substituting (3.40) in (3.46), the upper bound of conditional PEP is given as in [121]

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u | \mathbf{h}_u, \mathbf{v}_u, \tau_u, P_u\} \leq \exp\left(-\frac{E\{s\}}{4N_0} \sum_{l=1}^r \lambda_l^2 |\bar{h}_u^l|^2\right), \quad (3.47)$$

where $E\{s\}$ is the symbol energy, λ_l 's are the diagonal elements of $\mathbf{\Gamma}$, representing the

singular values of $\mathbf{D} = \Phi(\hat{\mathbf{x}}_u) - \Phi(\mathbf{x}_u)$. Moreover, $\mathbf{D}^H \mathbf{D} = \mathbf{U} \mathbf{\Gamma} \mathbf{U}^H$ and $\bar{\mathbf{h}}_u = \mathbf{U} \mathbf{h}_u^H$, while $\mathbf{U} \mathbf{U}^H = \mathbf{I}$. In this regard, r represents the rank of the matrix $\mathbf{\Gamma}$, and $\lambda_l \in \text{diag}\{\mathbf{\Gamma}\}$ with $l \in [1, 2, \dots, r]$. After averaging over (3.47) with the aid of the moment generating function (MGF) [122] and considering each $|\bar{h}_u^l|^2$, we derive the upper bound of the unconditioned PEP under a given delay and Doppler realizations [123] as follows

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u \mid \mathbf{v}_u, \tau_u, P_u\} \leq \left(\prod_{l=1}^r \frac{\lambda_l^2}{P_u} \right)^{-1} \left(\frac{\gamma}{4} \right)^{-r}, \quad (3.48)$$

where $\gamma = E\{s\}/N_0$. Moreover, referring to [64], an approximation of the upper bound for the unconditional PEP can be given as

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u \mid P_u\} \lesssim \Theta^{-r} \left(\frac{\gamma}{4P_u} \right)^{-r}, \quad (3.49)$$

where $\Theta = \mathbf{e}^H \mathbf{e}$ represents the squared Euclidean distance between any pair of $\hat{\mathbf{x}}$ and $\mathbf{x}$ in particular, whilst $\mathbf{e} = \mathbf{x}_u - \hat{\mathbf{x}}_u$. Consequently, to obtain the closed form expression of the unconditional PEP for the proposed OTFS-based scheduling scheme we need to combine (3.44) and (3.49), and substitute into (3.45). Hence, we rewrite (3.45) with the aid of the binomial theorem as

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u\} \leq \zeta \sum_{s=0}^{N_u-1} \frac{(-P_{\min})^{N_u-1-s}}{s!(N_u-1-s)!} \Upsilon, \quad (3.50)$$

where $\zeta = N_u! \Theta^{-r} (P_{\max} - P_{\min})^{-N_u}$, and Υ equals

$$\Upsilon = \int_{P_u} P_u^{s+r} \left(\frac{4}{\gamma} \right)^r dP_u. \quad (3.51)$$

We need to integrate (3.51) in order to have a closed form solution for the unconditional PEP in (3.50). Therefore, we perform some mathematical manipulations in order to simplify the integral part of (3.50) as follows

$$\Upsilon = \left(\frac{\gamma}{4} \right)^{s+1} \int_{\varkappa} \varkappa^{(s+r)} d\varkappa, \quad (3.52)$$

where $\varkappa = \frac{4P_u}{\gamma}$. After combining (3.52) with (3.50), the upper bound of the unconditional PEP reduces to

$$\Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u\} \leq \zeta \sum_{s=0}^{N_u-1} \left(\frac{\gamma}{4} \right)^{s+1} \frac{(-P_{\min})^{N_u-1-s}}{s!(N_u-1-s)!} \frac{\varkappa^{s+r+1}}{s+r+1} \Big|_{\varkappa_n}^{\varkappa_x}, \quad (3.53)$$

where $\varkappa_n = \frac{4P_{\min}}{\gamma}$ and $\varkappa_x = \frac{4P_{\max}}{\gamma}$. Then, the ABEP of the proposed system model can be evaluated by

$$\Pr\{e\} = \sum_{\mathbf{x}} \sum_{\hat{\mathbf{x}}} \frac{B^{-1}}{NM \log_2 Q} \Pr\{\mathbf{x}_u \rightarrow \hat{\mathbf{x}}_u\} \varepsilon(\mathbf{x}_u, \hat{\mathbf{x}}_u), \quad (3.54)$$

where $\varepsilon(\mathbf{x}_u, \hat{\mathbf{x}}_u)$ represents the bit-based Hamming distance among particular $\mathbf{x}_u$ and $\hat{\mathbf{x}}_u$ pair, $NM\log_2 Q$ is the total number of bits transmitted per OTFS transmission, and $B = Q^{NM}$.

3.5. The Impact Of IQ Imbalance In Delay-Doppler Domain

This section revises the basic concept of OTFS transmission and IQ imbalance to provide an essential background for the remainder of the work, as shown in **Figure 3.5** and **Figure 3.6**, respectively.

3.5.1. Ideal OTFS transceiver model

This work considers a single antenna at both the Tx and the Rx, while the OTFS Tx emits out a block of $N \times M$ uncoded M -ary data symbols. The OTFS frame consists of N number of symbols and M number of subcarriers with T symbol duration and Δf subcarrier spacing, respectively, occupying a total bandwidth of $B = M\Delta f$ and a total frame duration of $T_f = TN$. In this work, we represent the delay-Doppler signal by $x[k, l]$, conveying M -ary symbols with $x[k, l] \in \mathcal{A}$, where $\mathcal{A} = \{a_0, a_1, \dots, a_{M-1}\}$ is the modulation alphabet. Moreover, the delay-Doppler grid is given as

$$\Phi = \left\{ \left(\frac{k}{NT}, \frac{l}{M\Delta f} \right), k=0, \dots, N-1, l=0, \dots, M-1 \right\}, \quad (3.55)$$

where $1/(M\Delta f)$ and $1/(NT)$ denote the step size of delay and Doppler axis, respectively, while $T\Delta f = 1$. The Inverse Symplectic Fast Fourier Transform (ISFFT) converts $x[k, l]$ delay-Doppler signal to time-frequency domain as follows [70]

$$s[n, m] = \frac{1}{\sqrt{NM}} \sum_{k=0}^{N-1} \sum_{l=0}^{M-1} x[k, l] e^{j2\pi \left(\frac{nk}{N} - \frac{ml}{M} \right)}, \quad (3.56)$$

where n and m denote time and frequency indices, respectively. After applying Heisenberg Transform as given in [70], the time domain signal is obtained as

$$x(t) = \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} s[n, m] g_t(t - nT) e^{j2\pi \Delta f m(t - nT)}, \quad (3.57)$$

where $g_t(t)$ denotes the transmit periodic pulse of duration T , and Ψ represents the time-frequency grid given as

$$\Psi = \{(nT, m\Delta f), n=0, \dots, N-1, m=0, \dots, M-1\}. \quad (3.58)$$

After the OTFS frame $x(t)$ leaves the Tx, it undergoes the doubly dispersive channel that is given as follows

$$h(\tau, \nu) = \sum_{i=0}^{L-1} h_i \delta(\tau - \tau_i) \delta(\nu - \nu_i), \quad (3.59)$$

where h_i , τ_i , and ν_i denote the path gain, delay shift and Doppler shift regarding the channel path i , respectively, while L is the total number of channel paths and $\delta(\cdot)$ is Dirac delta function. Consequently, considering (3.55), the delay and Doppler bins for the channel path i equal to $\tau_i = l_i / (M\Delta f)$ and $\nu_i = (k_i + \kappa) / (NT)$, respectively, where κ denotes the fractional Doppler that is neglected ($\kappa = 0$) to assess the physical effect of IQ imbalance in this paper.

At the Rx, the received time domain OTFS frame is

$$y(t) = \int_{\tau} \int_{\nu} h(\tau, \nu) x(t - \tau) e^{j2\pi\nu(t-\tau)} d\tau d\nu + z(t), \quad (3.60)$$

where $z(t)$ denotes additive white Gaussian noise (AWGN). Further, by the application of Wigner transform [70], the received signal $y(t)$ is converted from time domain to time-frequency domain as follows

$$r[n, m] \triangleq \int_{t'} y(t') g_r(t' - t) e^{-j2\pi f(t'-t)} dt' |_{t, f}, \quad (3.61)$$

where $t = nT$, $f = m\Delta f$, and $g_r(t)$ denotes the Rx pulse obeying the criteria of match filtering. From the perspective of time-frequency data, the received signal in (3.161) is rewritten as follows

$$r[n, m] = \sum_{n'=0}^{N-1} \sum_{m'=0}^{M-1} H_{n,m}[n', m'] s[n', m'], \quad (3.62)$$

where $H_{n,m}[n', m']$ is given by [70]

$$H_{n,m}[n', m'] = \int_{\tau} \int_{\nu} h(\tau, \nu) \times A_g((n - n')T - \tau, (m - m')\Delta f - \nu) e^{j2\pi(\nu + m'\Delta f)((n - n')T - \tau)} e^{j2\pi\nu n'T} d\tau d\nu, \quad (3.63)$$

where $A_g(t, f)$ denotes the cross-ambiguity function for perfectly localized pulse shapes in both time and frequency domains and is readily decomposed as

$$A_g(t, f) = \int_{t'} g_r(t' - t) g_t(t') e^{j2\pi f(t'-t)} dt' = \delta(t) \delta(f). \quad (3.64)$$

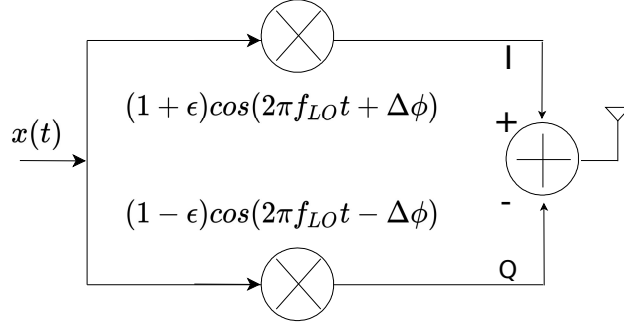


Figure 3.5: The system model of baseband IQ imbalance at Rx side.

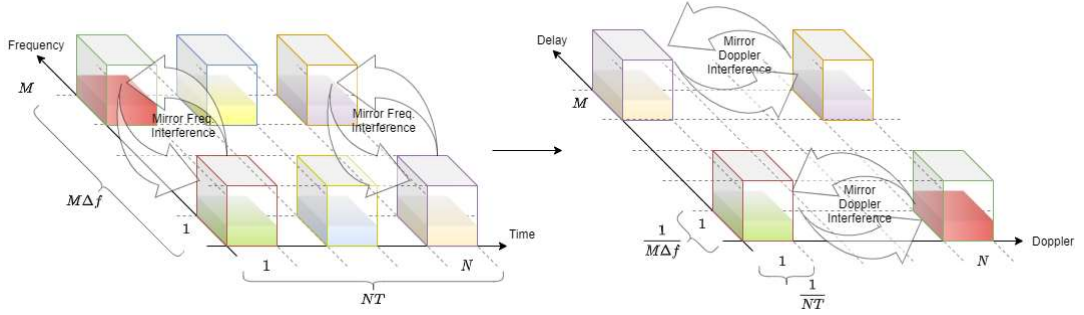


Figure 3.6: The physical effect of Tx IQ imbalance in time-frequency and delay-Doppler domains.

In addition, this decomposition $A_g(t, f) = \delta(t)\delta(f)$ represents the OTFS transmission free of interference, since $H_{n,m}[n', m']$ is different from zero only when $n' = n$ and $m' = m$ that corresponds to $r[n, m] = H_{n,m}[n, m]X[n, m]$ [102]. Hence, (3.63) simplifies to

$$H_{n,m}[n, m] = \int_{\tau} \int_{\nu} h(\tau, \nu) e^{j2\pi\nu nT} e^{-j2\pi(\nu + m\Delta f)\tau} d\tau d\nu. \quad (3.65)$$

For further insights about $H_{n,m}[n, m]$, we suggest the reader to visit [17]. The result of SFFT operation, which converts time-frequency signal $r[n, m]$ to delay-Doppler domain, is

$$y[k, l] = \frac{1}{NM} \sum_{\hat{k}=0}^{N-1} \sum_{\hat{l}=0}^{M-1} x[k', l'] h_w\left(\frac{k-k'}{NT}, \frac{l-l'}{M\Delta f}\right), \quad (3.66)$$

where $h_w(\tau, \nu) = \int_{\tau'} \int_{\nu'} h(\tau', \nu') w(\nu - \nu', \tau - \tau') e^{-j2\pi\nu\nu'\tau'} d\tau' d\nu'$ and the windowing function $w(\tau, \nu)$ is given as

$$w(\tau, \nu) = \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} e^{-j2\pi(\nu nT - \tau m\Delta f)}, \quad (3.67)$$

while both Tx and Rx pulses have a rectangular shape in the time-frequency domain. Consequently, considering the sparse wireless channel in delay-Doppler domain, the mathematical expression for the input-output relation of OTFS transmission in (3.66) is rewritten as follows [102]

$$y[k, l] = \sum_{i=0}^{L-1} h_i e^{-j2\pi v_i \tau_i} x[(k - v_i)_N, (l - \tau_i)_M] + z[k, l], \quad (3.68)$$

whose vector representation, in accordance with [50], is

$$\mathbf{Y} = \mathbf{H}\mathbf{X} + \mathbf{Z}, \quad (3.69)$$

where $\mathbf{Y}, \mathbf{Z} \in \mathbb{C}^{1 \times NM}$, $\mathbf{X} \in \mathbb{C}^{L \times NM}$, and $\mathbf{H} \in \mathbb{C}^{1 \times L}$ whose i th entity is equal to $h_i e^{-j2\pi v_i \tau_i}$. The elements of $\mathbf{Z}$ follow additive white Gaussian noise (AWGN) random variable with $\mathcal{CN}(0, N_0)$ while $h_i \sim \mathcal{CN}(0, 1/L)$.

The n -th column of $\mathbf{X}$ is given as

$$\mathbf{X}(n) = \begin{bmatrix} u((k - v_0)_N + N(l - \tau_0)_M) \\ \vdots \\ u((k - v_{L-1})_N + N(l - \tau_{L-1})_M) \end{bmatrix}, \quad (3.70)$$

where $u(k + Nl) = x[k, l]$ with $k = 0, 1, \dots, N-1$, $l = 0, 1, \dots, M-1$. In case of perfect CSI at the Rx, we can show the detection of the transmitted data by utilizing maximum likelihood (ML) detection principle as follows

$$\hat{\mathbf{x}} = \arg \min_{\mathbf{x} \in \mathcal{A}^{NM}} \|\mathbf{Y} - \mathbf{H}\mathbf{x}\|^2. \quad (3.71)$$

3.5.2. IQ Imbalance model at Tx

IQ imbalance demonstrates an effect of the mismatches on amplitude and phase response of the local oscillators (LO) between I and Q branches at either Tx or Rx side [84], as shown in **Figure 3.5**. On the other words, IQ imbalance appears as a consequence of any gain ($\epsilon \neq 0$) and phase ($\Delta\phi \neq 0$) imperfection between sine and cosine carriers that limits the orthogonality among each other, and thus becomes severe in multicarrier communication systems. An ideal design consists of I and Q branches with the same signal energy level and a $\frac{\pi}{2}$ phase difference. Otherwise, IQ imbalance leads to in-band self degradation and in-band mirror interference in classical OFDM-based signals [124]. In the literature, the mathematical expression for the effect of IQ imbalance on any transmitted

signal $m(t)$ is given in [84] as

$$m^{tx}(t) = \alpha m(t) + \beta m^*(t), \quad (3.72)$$

where α and β denote the IQ imbalance parameters that equal

$$\alpha = \cos(\Delta\phi) + j\epsilon \sin(\Delta\phi) \quad (3.73)$$

$$\beta = \epsilon \cos(\Delta\phi) - j \sin(\Delta\phi), \quad (3.74)$$

respectively. The notation $(\cdot)^*$ denotes the complex conjugate operator. Note that we investigate the impact of IQ imbalanced Tx in OTFS transmission in this study.

3.5.3. Analysis the effect of IQ imbalanced Tx in delay-Doppler domain

In this section, we model the physical effect of the IQ imbalanced Tx in OTFS transmission. With the aid of (3.57) and (3.72), the time domain distorted OTFS transmission due to IQ imbalance is given as

$$x^{tx}(t) = \alpha x(t) + \beta x^*(t), \quad (3.75)$$

which leads to a new interference type that we name mirror Doppler interference (MDI) as given in the following theorem.

Theorem 1 (Mirror Doppler Interference). *The physical effect of IQ imbalanced Tx in delay-Doppler domain results in interference occurring only with respect to Doppler axis that is given by*

$$x^{tx}[k, l] = \alpha x[k, l] + \beta x^*[-k, l], \quad (3.76)$$

where $-k$ Doppler bin denotes the mirror of k Doppler bin.

Proof. After performing Wigner operation as in (3.161), we obtain $x^{tx}(t)$ in time-frequency domain as follows

$$s^{tx}[n, m] = \int_{t'} x^{tx}(t') g_t^*(t' - t) e^{-j2\pi f(t' - t)} dt' |_{t, f}, \quad (3.77a)$$

$$\begin{aligned} &= \alpha \int_{t'} x(t') g_t^*(t' - t) e^{-j2\pi f(t' - t)} dt' |_{t, f} + \\ &\quad \beta \int_{t'} x^*(t') g_t(t' - t) e^{j2\pi f(t' - t)} dt' |_{t, f}, \end{aligned} \quad (3.77b)$$

$$= \alpha s[n, m] + \beta s^*[n, -m] \quad (3.77c)$$

where $t = nT$, $f = m\Delta f$ and $s^*[n, -m]$ is the complex conjugate of $s[n, m]$. In time-frequency domain, IQ imbalance at Tx causes mirror interference only with respect to frequency components of OTFS transmission. To observe the impact of IQ imbalanced Tx in delay-Doppler domain, we apply SFFT operation as

$$x^{tx}[k, l] = \alpha \sum_{k=0}^{N-1} \sum_{l=0}^{M-1} s[n, m] e^{-j2\pi(\frac{nk}{N} - \frac{ml}{M})} + \quad (3.78a)$$

$$\begin{aligned} & \beta \sum_{k=0}^{N-1} \sum_{l=0}^{M-1} s^*[n, -m] e^{-j2\pi(\frac{nk}{N} - \frac{ml}{M})} \\ & = \alpha x[k, l] + \beta x^*[-k, l], \end{aligned} \quad (3.78b)$$

where $-k$ represents mirror index with respect to k in Doppler axis, and vice versa. Importantly, according to (3.72), α represents the self degradation of signal power on index k , while β shows the MDI on k -th index coming from its corresponding mirror $-k$. In case I and Q branches perfectly match $\alpha = 1$ and $\beta = 0$, there is no signal power loss on index k and also zero MDI leaks from its imaginary carrier $-k$. In **Figure 3.6**, we illustrated the physical effect of the IQ imbalanced Tx on the OTFS transmission. For instance, the green index is the mirror of the red index and vice versa. It is worth mentioning that OFDM is a single-dimensional based transmission associated with mirror interference occurring in the frequency domain due to IQ imbalance. On the other hand, even though OTFS is a two-dimensional scheme, interference only occurs with respect to a single dimension (i.e., Doppler domain).

Note that we utilize the conjugate property of *Fourier transform* ($\mathcal{F}$) to establish the relation between the transmitted OTFS signal and IQ imbalance, where $\mathcal{F}\{\omega^*(\varrho)\} = \Omega^*(-\rho)$ with $\omega(\varrho) \xleftrightarrow{\mathcal{F}} \Omega(\rho)$. $\square$

Considering the Theorem 1, the mathematical expression of the received signal equals

$$y[k, l] = \sum_{i=0}^{L-1} h_i e^{-j2\pi v_i \tau_i} x^{tx}[(k - v_i)_N, (l - \tau_i)_M] + z[k, l], \quad (3.79)$$

where both desired signal and MDI experience the same channel. Regarding (3.79), for the case the symbol block i is transmitted, we can write the vectorized form of the received signal as

$$\mathbf{Y} = \mathbf{H} \mathbf{X}^{tx} + \mathbf{Z}, \quad (3.80a)$$

$$= \mathbf{H}(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^*) + \mathbf{Z}, \quad (3.80b)$$

where $\mathbf{X}^{tx}$ denotes the transmitted symbol block which is given as $\mathbf{X}^{tx} = \alpha \mathbf{X} + \beta \mathbf{X}^\pi$, and $\mathbf{X}^\pi$ is such a permuted version of $\mathbf{X}$ that we derive its n -th column as

$$\mathbf{X}^\pi(n) = \begin{bmatrix} u^*((-k - v_0)_N + N(l - \tau_0)_M) \\ \vdots \\ u^*((-k - v_{L-1})_N + N(l - \tau_{L-1})_M) \end{bmatrix}, \quad (3.81)$$

where $u^*(-k - v_{L-1})_N$ denotes the complex conjugate of $u(-k - v_{L-1})_N$.

3.5.4. Performance analysis of OTFS under Tx mirror-Doppler interference

In this section, we derive the theoretical analysis of the bit error probability (BEP) in order to provide beneficial insights about the performance of the IQ imbalanced Tx in OTFS transmission over the wireless channel with ideal and imperfect CSI. Thus, we consider the pairwise error probability (PEP) to calculate the bit error rate (BER) performance of OTFS Tx under IQ imbalance, assuming ML estimation at the Rx side in all the scenarios.

3.5.4.1. Bit Error Probability Analysis of OTFS with Tx-IQ

We drive the BER performance of the IQ imbalanced Tx in OTFS transmission by taking into account the union bound concept as follows

$$\Pr\{e\} = \frac{1}{(MN2^{MN})} \sum_{\mathbf{X}_i} \sum_{\mathbf{X}_j} \Pr(\mathbf{X}_i \rightarrow \mathbf{X}_j) \varepsilon(\mathbf{X}_i, \mathbf{X}_j), \quad (3.82)$$

where $\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j\}$ defines the pairwise error probability (PEP) of the decision when the transmitted symbol block $\mathbf{X}_i$ is erroneously detected as $\mathbf{X}_j$ and their corresponding bit-based pairwise error is denoted by $\varepsilon(\mathbf{X}_i, \mathbf{X}_j)$. Furthermore, we can write the PEP in terms of averaging with respect to the joint PDF of $\mathbf{H}$, that is

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j\} = \int \Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}\} f_{\mathbf{H}}(\mathbf{H}) d\mathbf{H} \quad (3.83)$$

where $f_{\mathbf{H}}(\mathbf{H})$ is the joint probability density function (PDF) of $\mathbf{H}$, and $\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}\}$ denotes the conditional PEP, i.e., the PEP conditioned on $\mathbf{H}$. In accordance with (3.80), using ML decoding, we express the conditional PEP as follows

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}\} = \Pr\{\|\mathbf{Y} - \mathbf{H}\mathbf{X}_i\|^2 > \|\mathbf{Y} - \mathbf{H}\mathbf{X}_j\|^2\}, \quad (3.84)$$

which we obtain in the closed-form in the following theorems.

After performing algebraic manipulations using the norm expansion $\|Y - \mathbf{H}\mathbf{X}\|^2 = \|Y\|^2 + \|\mathbf{H}\mathbf{X}\|^2 - 2\Re\{Y\mathbf{X}^H\mathbf{H}^H\}$, where $\Re\{\cdot\}$ denotes the real-part operator, we readily show that conditional PEP, given in (3.174), can be rewritten as

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = \Pr\{\|\mathbf{H}\mathbf{X}_i\|^2 - \|\mathbf{H}\mathbf{X}_j\|^2 > 2\Re\{Y(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H\}\}, \quad (3.85)$$

where we notice that

$$\|\mathbf{H}\mathbf{X}_i\|^2 - \|\mathbf{H}\mathbf{X}_j\|^2 = \mathbf{H}(\mathbf{X}_i\mathbf{X}_i^H - \mathbf{X}_j\mathbf{X}_j^H)\mathbf{H}^H \quad (3.86a)$$

$$= \mathbf{H}\mathbf{P}_{ij}\mathbf{H}^H, \quad (3.86b)$$

where $\mathbf{P}_{ij} = \mathbf{X}_i\mathbf{X}_i^H - \mathbf{X}_j\mathbf{X}_j^H$. Moreover, for any symbol block index k , we also notice

$$Y\mathbf{X}_k^H\mathbf{H}^H = (\mathbf{H}(\alpha\mathbf{X}_i + \beta\mathbf{X}_i^\pi) + \mathbf{Z})\mathbf{X}_k^H\mathbf{H}^H, \quad (3.87a)$$

$$= \alpha\mathbf{H}\mathbf{X}_i\mathbf{X}_k^H\mathbf{H}^H + \beta\mathbf{H}\mathbf{X}_i^\pi\mathbf{X}_k^H\mathbf{H}^H + \mathbf{Z}\mathbf{X}_k^H\mathbf{H}^H, \quad (3.87b)$$

where, as being conditioned on both $\mathbf{X}_k$ and $\mathbf{H}$, $\mathbf{Z}\mathbf{X}_k^H\mathbf{H}^H$ follows a complex Gaussian distribution. Then, with the aid of (3.87b), we rewrite the term $Y(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H$ in (3.85) as

$$\begin{aligned} Y(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H &= \alpha\mathbf{H}\mathbf{X}_i(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H + \\ &\quad \beta\mathbf{H}\mathbf{X}_i^\pi(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H + \\ &\quad \mathbf{Z}(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H. \end{aligned} \quad (3.88)$$

In accordance, after we perform some mathematical manipulations we reduce (3.85) to

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = \Pr\{\mathbf{H}\mathbf{P}_{ij}\mathbf{H}^H > 2\Re\{\mathbf{H}\mathbf{C}_{i,j}\mathbf{H}^H\} + 2\Re\{\mathbf{Z}(\mathbf{X}_i - \mathbf{X}_j)^H\mathbf{H}^H\}\}, \quad (3.89)$$

where $\mathbf{C}_{ij}$ is given as

$$\mathbf{C}_{ij} = (\alpha\mathbf{X}_i + \beta\mathbf{X}_i^\pi)(\mathbf{X}_i - \mathbf{X}_j)^H, \quad (3.90)$$

where we can easily see the parameters of IQ imbalance α and β defined in (3.73) and

(3.74), respectively. Since C_{ij} includes the correlated X_i and X_j^π matrices, we define the correlation matrix due to IQ imbalance in the following theorem.

Theorem 2 (Correlation difference matrix R_{ij}). *We express $2\Re\{HC_{ij}H^H\}$ as a correlation matrix and conditioned only on H as*

$$2\Re\{HC_{ij}H^H\} = HR_{ij}H^H, \quad (3.91)$$

where the correlation difference matrix is given as

$$R_{ij} = C_{ij} + C_{ij}^H. \quad (3.92)$$

Proof. With the aid of norm expansion, we rewrite (3.91) as follows

$$2\Re\{HC_{ij}H^H\} = \|H\|^2 + \|HC_{ij}^H\|^2 - \|H - HC_{ij}^H\|^2, \quad (3.93a)$$

$$= HH^H + HC_{ij}^H C_{ij} H^H - \|H(I - C_{ij}^H)\|^2, \quad (3.93b)$$

$$= HH^H + HC_{ij}^H C_{ij} H^H - \quad (3.93c)$$

$$H(I - C_{ij}^H)(I - C_{ij})H^H,$$

$$= H(I + C_{ij}^H C_{ij} - (I - C_{ij}^H)(I - C_{ij}))H^H, \quad (3.93d)$$

$$= H(C_{ij} + C_{ij}^H)H^H. \quad (3.93e)$$

□

Regarding (3.89), the statistical characteristics of the noise term are given as

$$2\Re\{ZX_k^H H^H\} \sim \mathcal{CN}(0, 2N_0 H D_{ij} H^H), \quad (3.94)$$

where N_0 denotes the spectral density of additive white Gaussian noise, and D_{ij} is the squared distance matrix between X_i and X_j defined as

$$D_{ij} = (X_i - X_j)(X_i - X_j)^H. \quad (3.95)$$

In accordance with (3.89) and the aforementioned theorems, we obtain the closed-form expression of conditioned PEP as shown in the following theorem.

Theorem 3 (Conditional PEP). *For OTFS transmission with IQ imbalanced Tx, the conditioned PEP of the decision when the transmitted symbol block X_i is erroneously detected*

as the symbol block $\mathbf{X}_j$ is given by

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = Q\left(\frac{\mathbf{H}(\mathbf{R}_{ij} - \mathbf{P}_{ij})\mathbf{H}^H}{\sqrt{2N_0 \mathbf{H}\mathbf{D}_{ij}\mathbf{H}^H}}\right). \quad (3.96)$$

Proof. After mathematical manipulations, we rewrite (3.89) as

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = \Pr\{\mathbf{H}(\mathbf{P}_{ij} - \mathbf{R}_{ij})\mathbf{H}^H > + 2\Re\{\mathbf{Z}(\mathbf{X}_i - \mathbf{X}_j)^H \mathbf{H}^H\}\}, \quad (3.97)$$

which completes the proof of Theorem 3. $\square$

Theorem 4 (Conditional PEP). *After applying singular value decomposition (SVD), we express the conditional PEP given in (3.96) as follows*

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = Q\left(\frac{\sum_{l=1}^L t_l |\bar{h}_l|^2}{\sqrt{2N_0 \sum_{l=1}^L g_l |\tilde{h}_l|^2}}\right). \quad (3.98)$$

Proof. Considering the fact that $\mathbf{R}_{ij}$, $\mathbf{P}_{ij}$ and $\mathbf{D}_{ij}$ are Hermitian matrices, we diagonalize $\mathbf{R}_{ij} - \mathbf{P}_{ij}$ and $\mathbf{D}_{ij}$ by applying the unitary transformation as follows

$$\mathbf{R}_{ij} - \mathbf{P}_{ij} = \mathbf{U}\mathbf{T}_{ij}\mathbf{U}^H, \quad (3.99a)$$

$$\mathbf{D}_{ij} = \mathbf{V}\mathbf{G}_{ij}\mathbf{V}^H, \quad (3.99b)$$

where $\mathbf{U}$ and $\mathbf{V}$ are unitary matrices and obey the rule of $\mathbf{U}\mathbf{U}^H = \mathbf{I}$ and $\mathbf{V}\mathbf{V}^H = \mathbf{I}$, respectively. Importantly, $\mathbf{T}_{ij}$ and $\mathbf{G}_{ij}$ are diagonal matrices with the singular values of $\mathbf{R}_{ij} - \mathbf{P}_{ij}$ and $\mathbf{D}_{ij}$ on its diagonal, respectively. We define $\bar{\mathbf{H}} = \mathbf{H}\mathbf{U}$ and $\tilde{\mathbf{H}} = \mathbf{H}\mathbf{V}$, and after substituting (3.181) in (3.96), we write (3.96) as follows

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j \mid \mathbf{H}\} = Q\left(\frac{\bar{\mathbf{H}}\mathbf{T}_{ij}\bar{\mathbf{H}}^H}{\sqrt{2N_0 \tilde{\mathbf{H}}\mathbf{G}_{ij}\tilde{\mathbf{H}}^H}}\right), \quad (3.100)$$

where $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ have the same distribution characteristics with $\mathbf{H}$ since $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ are related to $\mathbf{H}$ via a unitary matrix. Thus, we rewrite the numerator and denominator of (3.182) as follows

$$\mathbf{H}(\mathbf{R}_{ij} - \mathbf{P}_{ij})\mathbf{H}^H = \bar{\mathbf{H}}\mathbf{T}_{ij}\bar{\mathbf{H}}^H = \sum_{l=1}^L t_l |\bar{h}_l|^2, \quad (3.101a)$$

$$\mathbf{H}\mathbf{D}_{ij}\mathbf{H}^H = \tilde{\mathbf{H}}\mathbf{G}_{ij}\tilde{\mathbf{H}}^H = \sum_{l=1}^L g_l |\tilde{h}_l|^2. \quad (3.101b)$$

For further simplification of (3.98), we next express the denominator in terms of numerator in the following theorem.

Theorem 5 (Conditional PEP). *We express the conditional PEP as the ratio of independent and identically distributed (i.i.d) exponential random variables, given by*

$$\Pr\{X_i \rightarrow X_j | \mathbf{H}\} = Q\left(\sqrt{\frac{1}{2N_0} \frac{(\sum_{l=1}^L \gamma_l)^2}{\sum_{l=1}^L b_l \gamma_l}}\right). \quad (3.102)$$

Proof. We simplify (3.101a) with the aid of Cholesky decomposition as

$$\bar{\mathbf{H}}\bar{\mathbf{T}}_{ij}\bar{\mathbf{H}}^H = \bar{\mathbf{H}}\mathbf{F}_{ij}^H\mathbf{F}_{ij}\bar{\mathbf{H}}^H, \quad (3.103a)$$

$$= \mathbf{K}_{ij}\mathbf{K}_{ij}^H = \sum_{l=1}^L \gamma_l, \quad (3.103b)$$

where $\bar{\mathbf{T}}_{ij} = \mathbf{T}_{ij} + \varkappa^2 \mathbf{I}$ with $\varkappa \lll 1$, and $\varkappa^2$ is larger than the smallest singular value of $\mathbf{T}_{ij}$ that equals zero. $\mathbf{F}_{ij}$ is Cholesky factor of $\bar{\mathbf{T}}_{ij}$ and $\mathbf{F}_{ij}^H\mathbf{F}_{ij}$ is a diagonal matrix, and $\mathbf{K}_{ij} = \bar{\mathbf{H}}\mathbf{F}_{ij}^H$. Thus, we define $\bar{\mathbf{H}} = \mathbf{K}_{ij}\text{inv}(\mathbf{F}_{ij}^H)$, $\gamma_l = \bar{t}_l|\bar{h}_l|^2$. In this sense, we rewrite (3.101b) as

$$\tilde{\mathbf{H}}\mathbf{G}_{ij}\tilde{\mathbf{H}}^H = \mathbf{K}_{ij}\mathbf{B}_{ij}\mathbf{K}_{ij}^H = \sum_{l=1}^L b_l \gamma_l, \quad (3.104)$$

where $\mathbf{B}_{ij} = \text{inv}(\mathbf{U}\mathbf{F}_{ij}^H)\mathbf{V}\mathbf{G}_{ij}\mathbf{V}^H\text{inv}(\mathbf{F}_{ij}\mathbf{U}^H)$ is a diagonal matrix. $\square$

The alternative formula of Q -function is expressed to simplify (3.102) as follows

$$P(X_i \rightarrow X_j | \mathbf{H}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \exp\left(-\frac{\xi}{2N_0 \sin^2(\theta)}\right) d\theta, \quad (3.105)$$

where the random variable (RV) $\xi = (\sum_{l=1}^L \gamma_l)^2 / \sum_{l=1}^L b_l \gamma_l$. The unconditional PEP is obtained by averaging (3.105) over the RV ξ given as

$$\Pr\{X_i \rightarrow X_j\} = \int_{-\infty}^{\infty} \Pr\{X_i \rightarrow X_j | \mathbf{H}\} f_{\xi}(\chi) d\chi, \quad (3.106)$$

where $f_{\xi}(\chi)$ denotes the PDF of ξ RV. Importantly, all γ_l represent L i.i.d RVs and all b_l values can take either zero or positive values. Now, consider that out of the L values there are $r \leq L$ nonzero values. Out of the r values, there exist $K \leq r$ distinctive values. With

the aid of [125, Eq. (16)], we obtain the PDF of $f_\xi(\chi)$ with K -distinct values of b_l as

$$f_\xi(\chi) = \sum_{i=1}^K \sum_{j=1}^{n_i} \frac{\mu(i, j)}{b_i^j \Gamma(j) \Gamma(r-j)} \sum_{k=0}^{r-j-1} \binom{r-j-1}{k} \times \frac{(-1)^k \Gamma(r+j+k)}{b_i^k \chi^{k+j+1}} \left(1 - e^{-b_i \chi}\right) \sum_{l=0}^{r+j+k-1} \frac{b_i^l \chi^l}{l!} \quad (3.107)$$

where n_i denotes the occurrence of b_i and $\mu(i, j)$ is given as

$$\begin{aligned} \mu(i, j) &= \frac{1}{b_i^{n_i-j} (n_i-j)!} \times \sum_{l_1 \geq 0} \cdots \sum_{l_{i-1} \geq 0} \sum_{l_{i+1} \geq 0} \cdots \sum_{l_K \geq 0} \\ &\times \binom{n_i-j}{l_1, \dots, l_{i-1}, l_{i+1}, \dots, l_K} \prod_{j=1, j \neq i}^N (-1)^{l_j} \\ &\times \frac{\Gamma(n_j + l_j)}{\Gamma(n_j)} \left(\frac{b_i}{b_i - b_j}\right)^{n_j + l_j} b_j^{l_j}. \end{aligned} \quad (3.108)$$

Although $f_\xi(\chi)$ can be calculated as in (3.107), it is not easy to find and calculate the average PEP in (3.106). As the number of channel taps (L) increases, the calculation of $\mu(i, j)$ values becomes difficult, computational cost increases and numerical evaluation becomes infeasible. For this reason, we further simplify (3.102) as we show in the following theorem.

Theorem 6 (Conditional PEP). *With the aid of Cauchy-Schwarz inequality, we express the conditional PEP in (3.102) as the sum of exponential RVs, given by*

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}\} \leq Q\left(\sqrt{\frac{1}{2N_0} \sum_{l=1}^r b_l^{-1} \gamma_l}\right). \quad (3.109)$$

Proof. We rewrite ξ as follows

$$\xi = \frac{(\gamma_1 + \gamma_2 + \cdots + \gamma_r)^2}{b_1 \gamma_1 + b_2 \gamma_2 + \cdots + b_r \gamma_r}, \quad (3.110)$$

where we utilize Cauchy-Schwarz inequality for the numerator as given below

$$\left(\sum_{l=1}^r \frac{\gamma_l \sqrt{b_l \gamma_l}}{\sqrt{b_l \gamma_l}}\right)^2 \leq \left(\sum_{l=1}^r \frac{\gamma_l}{b_l}\right) \left(\sum_{l=1}^r b_l \gamma_l\right). \quad (3.111)$$

Then, substituting (3.111) in (3.110) yields (3.109), which proves Theorem 3. $\square$

After considering $\gamma_l = \bar{t}_l |\bar{h}_l|^2$ and averaging over (3.109) using the moment generating function (MGF) [122], we give the upper bound of unconditioned PEP [123] as follows

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j\} \leq \prod_{l=1}^r \left(1 + \frac{1}{4b_l N_0 L}\right)^{-1}, \quad (3.112)$$

where r denotes the rank of $\mathbf{B}_{ij}$. Since $\mathbf{B}_{ij} = \text{inv}(\mathbf{U}\mathbf{F}_{ij}^H)\mathbf{D}_{ij}\text{inv}(\mathbf{F}_{ij}\mathbf{U}^H)$ and $\mathbf{F}_{ij}$ is the Cholesky factor of $\bar{\mathbf{T}}_{ij}$ that depends on $\mathbf{R}_{ij} - \mathbf{P}_{ij}$, b_l values are dependent both on $\mathbf{R}_{ij} - \mathbf{P}_{ij}$ and on $\mathbf{D}_{ij}$. It is worth noting that $\mathbf{R}_{ij} - \mathbf{P}_{ij}$ in the derivations contains the effect of IQ imbalance in OTFS system. However, in case of zero IQ imbalance (i.e., $\alpha = 1$ and $\beta = 0$) $\mathbf{R}_{ij} - \mathbf{P}_{ij}$ converts to $\mathbf{D}_{ij}$, and thus, b_l only relies on $\mathbf{D}_{ij}$. In accordance with (3.112), it is worth noting that the impact of IQ imbalance on the Tx of OTFS transmission importantly limits the diversity gain of the communication, and IQ imbalance does not cause saturation of the SNR as in conventional only time-based or frequency-based wireless systems.

3.5.4.2. OTFS with Tx-IQ under imperfect channel estimation

In this subsection, we derive the effects of imperfect CSI on the PEP performance of OTFS transmission with IQ imbalanced Tx via union bound, as given in the following theorem.

Theorem 7 (PEP under Imperfect CSI). *The PEP of OTFS transmission with IQ imbalanced Tx considering imperfect CSI while the transmitted symbol block $\mathbf{X}_i$ is erroneously detected as $\mathbf{X}_j$ is given by*

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j\} \leq \prod_{l=1}^r \left(1 + \frac{1}{4b_l(N_0 + \sigma_i^2)L}\right)^{-1}, \quad (3.113)$$

where $\sigma_i^2 = \sigma_\epsilon^2 \|\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi\|^2$ denotes the impact of IQ imbalance and imperfect CSI, σ_ϵ^2 is variance of the channel estimation error $\Delta \mathbf{H} \sim \mathcal{CN}(0, eN_0)$ with e denoting the error amount of perfect CSI. We define the erroneous channel as follows

$$\mathbf{H}_e = \mathbf{H} + \boldsymbol{\epsilon}, \quad (3.114)$$

where $\boldsymbol{\epsilon}$ is the vector error of channel with $\mathbb{E}\{\boldsymbol{\epsilon}^H \boldsymbol{\epsilon}\} = \sigma_\epsilon^2 \mathbf{I}$.

Proof. Considering the effect of not only IQ imbalanced Rx but also the error of imperfect CSI, we rewrite (3.80b) as follows

$$\mathbf{Y} = \mathbf{H}_e(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi) + \mathbf{Z}_e, \quad (3.115)$$

where $\mathbf{Z}_e = -\boldsymbol{\epsilon}(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi) + \mathbf{Z}$. After performing mathematical changes, we give (3.89) as

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}_e\} = \Pr\{\mathbf{H}_e(\mathbf{P}_{ij} - \mathbf{R}_{i,j})\mathbf{H}_e^H > 2\Re\{\mathbf{Z}_e(\mathbf{X}_i - \mathbf{X}_j)^H \mathbf{H}_e^H\}\}, \quad (3.116)$$

where the variance of the signal introduced due to imperfect CSI is $2\sigma_e^2 \|(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi)(\mathbf{X}_i - \mathbf{X}_j)^H \mathbf{H}_e^H\|^2$. We further simplify the variance by using the inequality given as follows

$$\|(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi)(\mathbf{X}_i - \mathbf{X}_j)^H \mathbf{H}_e^H\|^2 \leq \|(\alpha \mathbf{X}_i + \beta \mathbf{X}_i^\pi)\|^2 \|(\mathbf{X}_i - \mathbf{X}_j)^H \mathbf{H}_e^H\|^2, \quad (3.117)$$

where after performing norm expansion, we can write (3.96) as follows

$$\Pr\{\mathbf{X}_i \rightarrow \mathbf{X}_j | \mathbf{H}_e\} \leq Q \left(\frac{\mathbf{H}_e(\mathbf{R}_{ij} - \mathbf{P}_{ij})\mathbf{H}_e^H}{\sqrt{2(N_0 + \sigma_i^2)\mathbf{H}_e \mathbf{D}_{ij} \mathbf{H}_e^H}} \right). \quad (3.118)$$

Consequently, substituting the new noise variance in (3.112) proves the Theorem 7. $\square$

3.6. Proposed IQ Imbalance Model And Mitigation In OTFS Transmission

3.6.1. System model of OTFS with IM

In this section, we briefly revise the basic concept of OTFS system and IM implementation in OTFS to provide an explicit representation of the work for the reader.

3.6.1.1. OTFS Transceiver Model

Figure 1.3 shows the block diagram of OTFS systems, where we consider a single antenna at Tx and Rx. The OTFS frame consists of M number of subcarriers and N number of symbols with Δf subcarrier spacing (SCS) and T symbol duration, respectively. Accordingly, the frame duration of OTFS signal is $T_f = TN$ and occupies a total bandwidth of $B = M\Delta f$, while $T\Delta f = 1$. Consequently, the OTFS signal transmits a block of $N \times M$ uncoded Q -ary data symbols $x[k, l] \in \mathcal{A}$, where $\mathcal{A} = \{a_0, a_1, \dots, a_{Q-1}\}$ is the modulation alphabet. In this work, $\mathbf{X} \in \mathbb{C}^{N \times M}$ denotes information symbols in DD grid with $x[k, l]$ being element of $\mathbf{X}$. We apply Inverse Symplectic Fast Fourier Transform (ISFFT) to convert data information from DD domain to time-frequency domain. In other words, we perform M -point Fast Fourier Transform (FFT) and N -point inverse-FFT (IFFT) operation over the columns and rows of $\mathbf{X}$, respectively. Later, we use Heisenberg Transform to generate the time domain OTFS signal via performing M -point IFFT over time-frequency data information. Therefore, time domain OTFS signal is given as follows

$$\mathbf{S} = \mathbf{G}_{tx} \mathbf{F}_M^H (\mathbf{F}_M \mathbf{X} \mathbf{F}_N^H), \quad (3.119)$$

where $\mathbf{G}_{tx}$ is a diagonal matrix that has the samples of pulse shape in time domain and $\mathbf{S} = \mathbf{X}\mathbf{F}_N^H$. We denote N -point FFT matrix as $\mathbf{F}_N$ give as

$$[\mathbf{F}_N]_{k,l} = \frac{1}{\sqrt{N}} \exp\left(\frac{-j2\pi kl}{N}\right), \quad k, l \in \{0, 1, \dots, N-1\} \quad (3.120)$$

where N -point IFFT is $\mathbf{F}_N^H$. Importantly, $\mathbf{G}_{tx}$ converts to the identity matrix in case of rectangular waveforms. Considering the column-wise vectorization of $\mathbf{S}$ matrix, we can write (3.119) as

$$\mathbf{s} = (\mathbf{F}_N^H \otimes \mathbf{G}_{tx})\mathbf{x}, \quad (3.121)$$

where $\otimes$ denotes the Hadamard product and $\mathbf{x} = \text{vec}(\mathbf{X})$. In order to control inter-symbol interference between OTFS frames, we pad a cyclic prefix of $L-1$ length, where L is the number of wireless channel taps. Later, the OTFS signal $\mathbf{s}$ interacts with the time-varying wireless channel that is given as

$$h(\tau, \nu) = \sum_{i=0}^{L-1} h_i \delta(\tau - \tau_i) \delta(\nu - \nu_i), \quad (3.122)$$

where h_i , τ_i , and ν_i denote the path gain, delay shift and Doppler shift regarding the channel path i , respectively, and $\delta(\cdot)$ is Dirac delta function. Moreover, the delay and Doppler bins for the i -th channel tap equal to $\tau_i = l_i / (M\Delta f)$ and $\nu_i = (k_i + \kappa) / (NT)$, respectively, where κ denotes the fractional Doppler that is neglected ($\kappa = 0$) in this work.

At the Rx side, after discarding the additional CP, the received signal is given as follows

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \mathbf{w}, \quad (3.123)$$

where $\mathbf{w} \in \mathbb{C}^{MN \times 1}$ denotes the additive white Gaussian noise (AWGN) random variable with $\mathcal{CN}(0, N_0 \mathbf{I})$. $\mathbf{H} \in \mathbb{C}^{MN \times MN}$ is the channel matrix given by

$$\mathbf{H} = \sum_{i=1}^P h_i \mathbf{\Pi}^{l_i} \mathbf{\Delta}^{k_i}, \quad (3.124)$$

where $\mathbf{\Pi} = \text{circ}[0, 1, 0, \dots, 0]_{MN \times MN}$ is a permutation matrix based on forward cyclic shift, and $\mathbf{\Delta} = \text{diag}[z^0, z^1, \dots, z^{MN-1}] \in \mathbb{C}^{MN \times MN}$ with $z = \exp(j2\pi/MN)$. In order to get the DD domain symbols, we first perform Wigner transform as $\mathbf{F}_M \mathbf{G}_{rx} \mathbf{R}$, given that $\mathbf{R} = \text{vec}^{-1}(\mathbf{r})$ and $\mathbf{G}_{rx}$ is a diagonal matrix containing samples of Rx pulse shape $g_{rx}(t)$. Later, we perform SFFT operation to get the received signal in DD domain as

$\mathbf{Y} = \mathbf{F}_M^H (\mathbf{F}_M \mathbf{G}_{rx} \mathbf{R}) \mathbf{F}_N$. After vectorization, the received signal is written as

$$\mathbf{y} = \mathbf{H}_{eff} \mathbf{x} + \bar{\mathbf{w}}, \quad (3.125)$$

where $\mathbf{H}_{eff} = (\mathbf{F}_N \otimes \mathbf{G}_{rx}) \mathbf{H} (\mathbf{F}_N^H \otimes \mathbf{G}_{tx})$ is the effective channel and $\bar{\mathbf{w}}$ is the noise in DD domain. In case of rectangular pulse shape, i.e., $\mathbf{G}_{rx} = \mathbf{I}_M$, $\bar{\mathbf{w}}$ preserves the same statistical features of $\mathbf{w}$. Regarding (3.125), we use minimum mean square error (MMSE) detector to estimate the data symbols as follows

$$\hat{\mathbf{x}} = \mathbf{H}_{eff}^H (\mathbf{H}_{eff} \mathbf{H}_{eff}^H + N_0 \mathbf{I})^{-1} \mathbf{y}, \quad (3.126)$$

where $\hat{\mathbf{x}} = \text{vec}(\hat{\mathbf{X}})$ and we demodulate $\hat{\mathbf{x}} \in \mathbb{C}^{MN \times 1}$ to get the information bits.

3.6.1.2. OTFS with IM

Different from conventional OTFS, OTFS-IM conveys data symbols not only by carriers but also by indices of the active carriers [91]. We split the radio resources of NM size OTFS block into G subblocks with $n = NM/G$ carriers per subblock. To perform IM, we activate u out of n carriers in each subblock. Consequently, $p_1 = \lfloor \log_2 C(n, u) \rfloor$ is the number of bits used to select the indices of active carriers, where $\lfloor \cdot \rfloor$ and $C(\cdot, \cdot)$ denote floor function and binomial coefficient, respectively. The indices of the active carriers within subblock b are given as

$$\Gamma_b = [i_{b,1}, \dots, i_{b,u}], \quad (3.127)$$

where $i_{b,q} \in \{1, \dots, n\}$ with $1 \leq b \leq G$ and $1 \leq q \leq u$. The number of possible subblock combinations equals $2^{\lfloor \log_2 C(n, u) \rfloor}$. Later, with the modulation of the elements of Γ_b , subblock b is given as

$$\mathbf{x}_b = [a_{b,1}, \dots, a_{b,u}], \quad (3.128)$$

which transmits $p_2 = u \log_2(Q)$ bits through modulated symbols. Therefore, the total number of data bits carried by OTFS-IM equals $p = G(p_1 + p_2)$. We follow the same procedure for each subblock. After combining all the subblocks, we generate $\mathbf{X}$ in DD domain. Later, we apply the same procedures as those of classical OTFS.

Different from conventional OTFS Rx, in OTFS-IM, we need to detect not only the data bits conveyed via Q -ary symbols but also the information carried via active indices. Hence, after MMSE equalization as in (3.126), we use the possible subblock realizations

to detect the transmitted data information shown as

$$\{\hat{\Gamma}_b, \hat{\mathbf{a}}_b\} = \arg \min_{\Gamma_b \in \mathbf{\Gamma}, \mathbf{a}_{b,j} \in \mathcal{A}} \sum_{j=1}^k |\hat{x}_{b,j} - a_{b,j}|^2, \quad (3.129)$$

where $\hat{\Gamma}_b$ and $\hat{\mathbf{a}}_b$ are the estimated index realization and data symbols for subblock b , respectively.

3.6.2. IQ imbalance mitigation in OTFS with IM

In this section, firstly, we show the effect of IQ imbalance in OTFS transmission at Tx and Rx sides. Then, we introduce the proposed IQ imbalance compensation scheme for OTFS.

3.6.2.1. Mathematical model of IQ imbalance in OTFS with IM

The effect of IQ imbalance occurs due to the misalignment on phase and amplitude response between I and Q branches of the LOs at Tx and Rx sides [84]. In the literature, the impact of IQ imbalance on any baseband signal $m(t)$ is modeled via the mathematical model as follows

$$m_{iq}(t) = \alpha m(t) + \beta m^*(t), \quad (3.130)$$

where α and β represent the IQ imbalance parameters given as

$$\alpha = \cos(\theta) + j\epsilon \sin(\theta) \quad (3.131a)$$

$$\beta = \epsilon \cos(\theta) - j \sin(\theta), \quad (3.131b)$$

respectively. The cases of $(\epsilon \neq 0)$ and $(\theta \neq 0)$ stands for gain and phase imperfection between sine and cosine carriers, respectively. The I and Q branches of a perfect handset have a $\frac{\pi}{2}$ phase difference and the same signal energy level considering either Tx or Rx side. Therefore, in case of IQ imperfection at Tx side, after substituting s from (3.121) into (3.130), time domain signal equals

$$s_{iq} = \alpha s + \beta s^*, \quad (3.132)$$

where $s^* = \{(\mathbf{F}_N^H \otimes \mathbf{G}_{tx})\mathbf{x}\}^*$ that leads to

$$\mathbf{S}_{iq} = \alpha \mathbf{S} + \beta \mathbf{S}_m^*, \quad (3.133)$$

where $\mathbf{S}_m$ denotes the mirror-form of $\mathbf{S}$. For example, if t is time axes and f is frequency axes of $S[t, f] \in \mathbf{S}$, its mirror is $S_m[t, f] = S[t, (M - f + 2)_M]$ with $0 \leq f \leq M - 1$ and $S_m[t, f] \in \mathbf{S}_m$. After performing SFFT process, the signal under Tx IQ imbalance in DD domain equals

$$\mathbf{X}_{iq} = \underbrace{\alpha \mathbf{X}}_{\text{SD}} + \underbrace{\beta \mathbf{X}_m^*}_{\text{MDI}}, \quad (3.134)$$

where $\mathbf{X}_m$ is the mirror-form of $\mathbf{X}$, and the mirror of $X[k, l] \in \mathbf{X}$ is $X_m[k, l] = X[(N - k + 2)_N, l]$ with $0 \leq k \leq N - 1$ and $X_m[k, l] \in \mathbf{X}_m$. Therefore, the received signal under Tx IQ imbalance is

$$\mathbf{y}_{iq}^t = \underbrace{\alpha \mathbf{H}_{eff} \mathbf{x}}_{\text{SD}} + \underbrace{\beta \mathbf{H}_{eff} \mathbf{x}_m^*}_{\text{MDI}} + \mathbf{w}, \quad (3.135)$$

where SD and MDI shows in-band self degradation and in-band mirror Doppler interference of the desired signal, respectively, as discussed in [58].

Different from Tx IQ imperfection, the received signal in DD domain under Rx IQ imbalance is given as

$$\mathbf{y}_{iq}^r = \underbrace{\alpha \mathbf{H}_{eff} \mathbf{x}}_{\text{SD}} + \underbrace{\beta \mathbf{H}_{eff,m}^* \mathbf{x}_m^*}_{\text{MDI}} + \tilde{\mathbf{w}}, \quad (3.136)$$

where $\mathbf{H}_{eff,m}$ is the channel impairment, hence the desired signal and interference are exposed to different channels. Moreover, $\tilde{\mathbf{w}} = \alpha \mathbf{w} + \beta \mathbf{w}_m^*$ is noise under the Rx IQ imbalance. In particular, the variance of $\tilde{\mathbf{w}}$ is as follows

$$\tilde{N}_0 = N_0(|\alpha|^2 + |\beta|^2) = N_0(1 + \epsilon^2). \quad (3.137)$$

Equation (3.137) shows that statistical features of $\tilde{\mathbf{w}}$ depends only on the gain imbalance (ϵ). Thus, $\tilde{N}_0 \simeq N_0$ whilst the gain mismatch is very small.

3.6.2.2. Proposed IQ imbalance Compensator

In this section, we propose a simple energy and spectral efficient IM-based compensator to estimate and mitigate the effect of IQ imbalance at Tx or Rx for OTFS-based wireless communications. In order to provide a simple understanding of the proposed schemes, we visualize the physical effect of IQ imbalance in the delay Doppler domain, as illustrated in **Figure 3.7**. For instance, only SD occurs in Doppler bin k_2 since no information symbol is transmitted via its mirror bin $-k_2$, leading to zero MDI due to IQ imperfection. However,

Algorithm 3.1 Estimation of α and β for Tx IQ imbalance

```

0: procedure EstimatorTx( $\alpha, \beta$ )
1:  $\{\hat{\Gamma}, \hat{\mathbf{a}}\} \leftarrow$  Eqs.(3.126), (3.129)  {Hard Decision}
2:  $\bar{\mathbf{X}} \leftarrow$  Eq.(3.128)
3:  $d_{ij} \in \hat{\Gamma}_{\bar{\mathbf{X}}}$   { $d_{ij}=1$  is active,  $d_{ij}=0$  is silent}
4:  $i = 0$   { $i$  denotes delay access}
5: while  $i \leq M$  do
6:    $\mathbf{c}_i \leftarrow \hat{\Gamma}$   { $\mathbf{c}_i$  shows unused Doppler indices}
7:    $\mathbf{c}_m \leftarrow \mathbf{c}_i$   { $\mathbf{c}_m$  shows mirror indices ( $m$ ) of  $\mathbf{c}_i$ }
8:   for  $j = 1 : N(k/n)$  do
9:     if  $\mathbf{c}_m(j) \notin \mathbf{c}_i$  then
10:       $\alpha \leftarrow \hat{X}(\mathbf{c}_m(j), i) / \bar{X}(\mathbf{c}_m(j), i)$ 
11:       $\beta \leftarrow \hat{X}(\mathbf{c}_i(j), i) / \bar{X}(\mathbf{c}_m(j), i)$ 
12:    end if
13:  end for
14:   $i = i + 1$ 
15: end while
16: end procedure
  
```

this is not the case in conventional OTFS transmission, where each Doppler bin confronts both SD and MDI. Therefore, regarding (3.134) and (3.135) where the Tx IQ imbalance takes place, the baseband received signal in DD domain for the carrier k with inactive mirror ($X_m^* = 0$) can be written as

$$\hat{X} = \alpha X + W \quad (3.138)$$

that makes α independent of β parameter, and $W = \text{vec}^{-1}(\mathbf{w})$. Moreover, we can write the received signal at the mirror carrier $-k$ as

$$\hat{X} = \beta X_m^* + W, \quad (3.139)$$

that makes β independent of α parameter. However, we need to know the indices of active carriers and data symbols carried by X in order to benefit from (3.138) and (3.139), resulting in the estimation of the IQ imbalance parameters. In line with this, we perform hard decision on the received signal given in (3.135) via ((3.126)) and ((3.129)) to know X and its mirror X_m that we represent with $\bar{X}$ and $\bar{X}_m$, respectively. As a result, we are able to estimate α and β parameter of Tx IQ imbalanced as follows

$$\alpha = \hat{X} / \bar{X} + W, \quad (3.140a)$$

$$\beta = \hat{X} / \bar{X}_m^* + W, \quad (3.140b)$$

respectively. In this way, we compensate the Tx IQ imbalance as follows

$$\hat{x}_r = (\hat{x} - \beta \bar{x}) / \alpha, \quad (3.141)$$

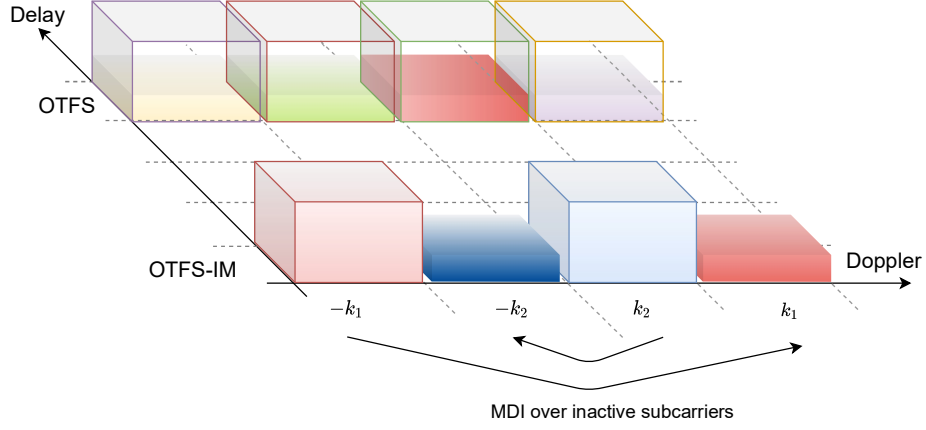


Figure 3.7: The visualization of mirror Doppler interference in OTFS and OTFS-IM caused by IQ imbalance.

where $\hat{\mathbf{x}}_r$ is the recovered data, and $\bar{\mathbf{x}} = \text{vec}(\bar{\mathbf{X}})$. In Algorithm (3.1), we summarize the proposed method for estimating the parameters of Tx IQ imbalance.

Algorithm 3.2 Estimation of α and β for Rx IQ imbalance

- 0: **procedure** EstimatorRx(α, β)
 - 1: $\{\hat{\Gamma}, \hat{\mathbf{a}}\} \leftarrow \text{Eqs. (3.126), (3.129)} \quad \{\text{Hard Decision}\}$
 - 2: $\bar{\mathbf{X}} \leftarrow \text{Eq. (3.128)}$
 - 3: $\bar{\mathbf{Y}} \leftarrow \text{Eq. (3.135)}$
 - 4: $d_{ij} \in \hat{\Gamma}_{\bar{\mathbf{Y}}} \quad \{d_{ij}=1 \text{ is active, } d_{ij}=0 \text{ is silent}\}$
 - 5: Follow the same steps as in Algorithm 1 regarding $\bar{\mathbf{Y}}$
 - 5: **end procedure**=0
-

Regarding (3.136), it is obvious that we can not directly explore the sparsity of $\mathbf{X}$ to estimate and compensate the impact of Rx IQ imbalance in OTFS transmission. Different from Tx IQ imbalance (3.135), Rx IQ imbalance not only effects the transmitted data $\mathbf{X}$ but also the effective channel $\mathbf{H}_{eff}$. Hence, we utilize the sparsity of $\mathbf{Y}_{iq}^r$ where the IQ impact is obvious, with $\mathbf{Y}_{iq}^r = \text{vec}^{-1}(\mathbf{y}_{iq}^r)$. In case of only SD with zero MDI and vise versa, we write the received signal in (3.136) as

$$\mathbf{Y}_{iq}^r = \alpha \mathbf{Y} + \tilde{\mathbf{W}}, \quad (3.142a)$$

$$\mathbf{Y}_{iq}^r = \beta \mathbf{Y}_m^* + \tilde{\mathbf{W}}, \quad (3.142b)$$

respectively, and $\tilde{\mathbf{W}} = \text{vec}^{-1}(\tilde{\mathbf{w}})$. Similar to the Tx IQ imbalance, we perform hard decision over (3.136) to know $\mathbf{Y}$ and its mirror $\mathbf{Y}_m$ that we show with $\bar{\mathbf{Y}}$ and $\bar{\mathbf{Y}}_m$, respectively. Algorithm (3.2) briefly shows the mathematical process how to estimate the parameters of Rx IQ imbalance. In particular, we find $\bar{\mathbf{Y}}$ in line 5 and follow the same steps as in

Algorithm (3.1). Later, we compensate the Rx IQ imbalance as follows

$$\hat{\mathbf{y}}_r = (\mathbf{y}_{iq}^r - \beta \bar{\mathbf{y}}) / \alpha. \quad (3.143)$$

After performing MMSE equalization over $\hat{\mathbf{y}}_r$, we use ML Rx to detect the data bits transmitted via both Q -ary symbols and indices.

To provide more insights on the performance and efficiency of the proposed method, we offer the computational complexity for the estimation of IQ imbalance parameters in terms of multiplications and considering (3.140) as follows

$$C = N_F G \min(k, n - u), \quad (3.144)$$

where $\min(k, n - u)$ represents the number of maximum used subcarriers within a subblock while satisfying the condition of $(X \neq 0 \ \& \ X_m = 0)$ or $(X_m = 0 \ \& \ X_m \neq 0)$, and N_F is the number of OTFS transmission frames. It is important to note that the same computational complexity is needed to estimate α and β parameters. Consequently, the overall system complexity introduced by the proposed scheme is $C_t = 2N_F G \min(u, n - u)$.

3.7. Proposed Transceiver Design For OTFS Against Inter-Doppler Interference

3.7.1. System model of proposed Rx windowing OTFS

This section introduces the system model of the proposed RW-OTFS for beyond 5G wireless communication systems. For the sake of clarity, we present the vectorial representation of the input-output relationship of the proposed RW-OTFS scheme. Hence, according to [51], the TF domain version of $\mathbf{X}$ is obtained as

$$\mathbf{X}_{TF} = \mathbf{F}_M \mathbf{X} \mathbf{F}_N^H, \quad (3.145)$$

where $\mathbf{F}_M \in \mathbb{C}^{M \times M}$ and $\mathbf{F}_M^H \in \mathbb{C}^{M \times M}$ represent M -point discrete Fourier transform (DFT) matrix and its Hermitian matrix, respectively. Then, considering practical rectangular pulse shaping for the Tx, the time domain symbol matrix is given as

$$\mathbf{S} = \mathbf{I}_M \mathbf{F}_M^H \mathbf{X}_{TF} = \mathbf{X} \mathbf{F}_N^H, \quad (3.146)$$

where the Heisenberg transform is performed column wise on $\mathbf{X}_{\text{TF}}$. Moreover, the vectorial form of $\mathbf{S}$ can be expressed as

$$\mathbf{s} = (\mathbf{F}_N^H \otimes \mathbf{I}_M) \mathbf{x}, \quad (3.147)$$

where $\mathbf{x} = \text{vec}\{\mathbf{X}\}$ and $\mathbf{s} = \text{vec}\{\mathbf{S}\}$. Later, a cyclic prefix (CP) with the length of G is utilized in order not only to combat the effect of multipath channel [126] but also to mitigate IDI via the proposed Rx windowing. The proposed scheme is shown in **Figure 3.8**. Basically, the idea of guard interval is to replicate part of the OTFS time domain frame from back to the front, as illustrated in **Figure 3.9(a)**.

Remark 2. *The length of the CP denoted by G must be at least $L_m - 1$ long, where L_m denotes the maximum excess delay, in order to avoid ISI due to the delay spread of wireless channel for conventional OTFS based transmission. Differently, in this work, we exploit non-utilized $G - L_m$ CP samples to mitigate IDI due to fractional Doppler. Since Doppler characteristics of OTFS signal appear in a global manner, we introduce a novel technique that utilizes global CP extension in order to manipulate the pulse shaping of OTFS carriers in DD domain. Global CP extension offers better spectral efficiency than local CP extension, where CP is added between each M sample.*

In line with **Remark 2** and (3.147), the OTFS time domain frame is extended from MN samples to $MN + G$ as

$$\mathbf{s}_G = \mathbf{A} \mathbf{s}, \quad (3.148)$$

where $\mathbf{A}$ refers to the CP addition matrix given by

$$\mathbf{A} = \begin{bmatrix} \mathbf{0}_{G \times (MN-G)} & \mathbf{I}_G \\ & \mathbf{I}_{MN} \end{bmatrix}.$$

After the global CP added OTFS frame undergoes doubly dispersive wireless channel (2.7), the received signal in the discrete-time domain is written as

$$r[n] = \sum_{i=0}^{L-1} h_i e^{j \frac{2\pi}{(MN+W)} (k_i + \kappa_i)(n - l_i)} s_G[(n - l_i)_{(MN+G)}], \quad (3.149)$$

where $W = G - L_m$, with L_m denoting the number of samples affected by ISI.

At the Rx side, after discarding the L_m samples, time domain baseband signal is given as

$$\mathbf{r} = \mathbf{B} \mathbf{H}_T \mathbf{s}_G + \mathbf{n}, \quad (3.150)$$

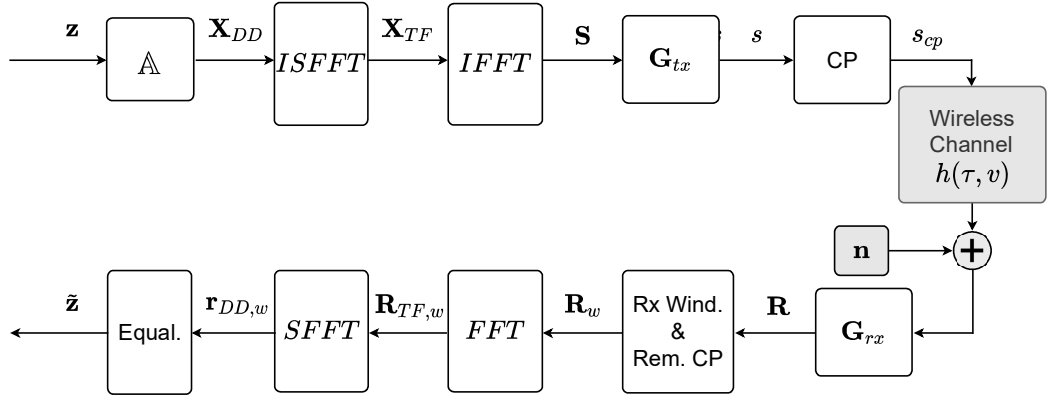


Figure 3.8: Block diagram of the proposed RW-OTFS system design.

with $\mathbf{B} = [0_{(MN+W) \times L_m} \mathbf{I}_{MN+W}] \in \mathbb{C}^{(MN+W) \times (MN+G)}$, referring to the CP removal matrix for the impact of time dispersion of wireless channel, which is illustrated via the red color part in **Figure 3.9(b)**. Moreover, the desired signal $\mathbf{r} \in \mathbb{C}^{(MN+W) \times 1}$ utilizes the remaining $W = G - L_m$ samples for the Rx windowing process. $\mathbf{n} \in \mathbb{C}^{(MN+W) \times 1}$ denotes the AWGN vector with distribution $\mathcal{CN}(0, \sigma_o^2 \mathbf{I})$. Moreover, $\mathbf{H}_T$ is the wireless channel matrix for (3.149) given as follows

$$\mathbf{H}_T = \sum_{i=0}^{L-1} h_i \mathbf{\Pi}^{l_i} \Delta^{(k_i + \kappa_i)}, \quad (3.151)$$

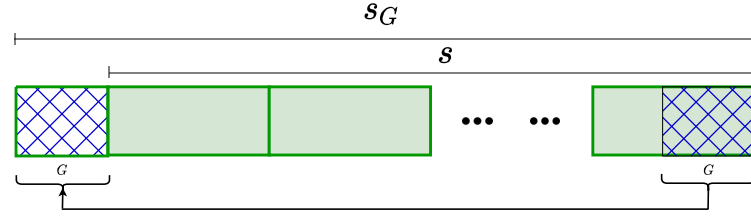
where Δ is a diagonal matrix given as

$$\Delta^{(k_i + \kappa_i)} = \text{diag}\{z^{n(k_i + \kappa_i)/N}\}, \quad (3.152)$$

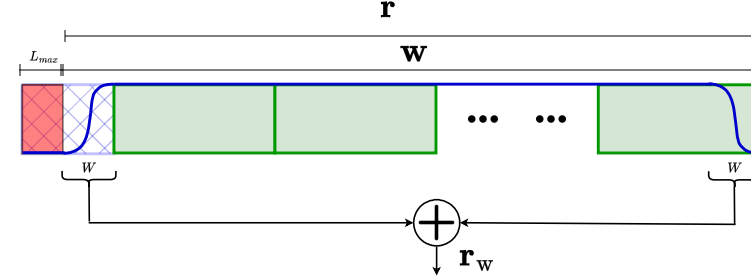
with $n \in [-G, -G + 1, \dots, 0, \dots, MN - 1]$ and $z = e^{j2\pi/(MN+G)}$, while $\mathbf{\Pi} \in \mathbb{C}^{(MN+G) \times (MN+G)}$ denotes a permutation matrix based on forward cyclic shift, given as

$$\mathbf{\Pi} = \begin{bmatrix} 0 & \cdots & 0 & 1 \\ 1 & \ddots & 0 & 0 \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \cdots & 1 & 0 \end{bmatrix}. \quad (3.153)$$

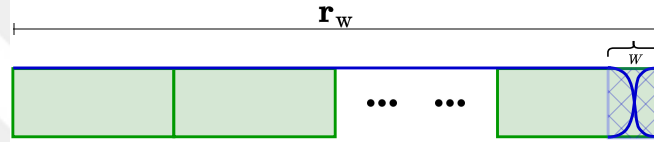
At the Rx side, the windowing function is represented by the bold blue line in **Figure 3.9(b)**. In conventional multicarrier systems, the windowing functions have been studied and utilized in order to boost side lobe suppression [127]. With this motivation, this work introduces a simple Rx windowing for OTFS based systems in order to mitigate the effect of IDI caused by fractional Doppler. Specifically, this work adopts RC windowing that is



(a): Cycle prefix extension G at Tx side.



(b): CP removal and windowing process at Rx side.



(c): Rx windowed OTFS frame.

Figure 3.9: Illustration of the signal processing at Tx and Rx for the proposed RW-OTFS scheme.

modeled as follows

$$w[n] = \begin{cases} \frac{1}{2}(1 + \cos(\pi + \frac{n\pi}{W})), & n=0, 1, \dots, W-1, \\ 1, & n=W, \dots, MN-1, \\ \frac{1}{2}(1 + \cos(\pi + \frac{n\pi}{W})), & n=MN, \dots, MN+W-1, \end{cases} \quad (3.154)$$

where the values of $w[n]$ represent the non-zero entities of the diagonal matrix $\mathbf{W} \in \mathbb{C}^{(MN+W) \times (MN+W)}$ in particular. After discarding the L_m -samples affected by ISI, a smooth transition is established by point-by-point multiplication of the windowing function $\mathbf{W}$ and the received OTFS frame $\mathbf{r}$ in (3.150). Later, W samples of G are discarded from the beginning of $\mathbf{r}$ and added to the end, as shown in **Figure 3.9(b)** and **Figure 3.9(c)**. As a result, RW-OTFS signal in matrix form is obtained as

$$\mathbf{r}_w = \mathbf{V} \mathbf{W} \mathbf{r}, \quad (3.155)$$

where the matrix $\mathbf{V}$ is given as

$$\mathbf{V} = \begin{bmatrix} 0_{(MN-W) \times W} & \\ & \mathbf{I}_{MN} \\ \mathbf{I}_W & \end{bmatrix}.$$

In accordance with (3.150) and (3.155), we obtain the closed-form expression of the received RW-OTFS frame in DD domain as detailed in the following proposition.

Remark 3. *As illustrated in Figure 3.9(b), this paper converts the conventional rectangular pulse-shaped OTFS frame to pulses with soft edges with the aid of time domain windowing. With the smooth edge transition, we ensure that the proposed RX windowed OTFS scheme offers a better spectral localization than that of the sharp-edged OTFS-based communications systems in DD domain.*

Proposition 1 (Proposed RW-OTFS frame). *In DD domain, the received signal considering the Rx windowed OTFS transmission with practical pulse shaping is represented as*

$$\mathbf{y} = \mathbf{H}_w \mathbf{x} + \mathbf{n}_w, \quad (3.156)$$

where $\mathbf{H}_w$ is given as

$$\mathbf{H}_w = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{H}_{w,T} (\mathbf{F}_N^H \otimes \mathbf{I}_M), \quad (3.157)$$

and

$$\mathbf{H}_{w,T} = \mathbf{V} \mathbf{W} \mathbf{B} \mathbf{H}_T \mathbf{A}. \quad (3.158)$$

Moreover, $\mathbf{n}_w$ denotes the noise after the Rx windowing in DD domain, and is given by

$$\mathbf{R}_w \mathbf{n}_w = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{V} \mathbf{W} \mathbf{n}. \quad (3.159)$$

Proof. After completing the Rx windowing process (3.155), the OTFS demodulation is equivalent to the inverse operation of the modulation, which consists of a TF demodulator and a post-processing module in order to receive the DD domain data symbols. The received 2-D signal in time domain after the serial-to-parallel conversion is given as

$$\mathbf{R}_w = \text{vec}^{-1}(\mathbf{r}_w), \quad (3.160)$$

and $\mathbf{R}_w \in \mathbb{C}^{M \times N}$. In this line, with the aid of Wigner transform [70], the TF domain signal

$\mathbf{R}_{w,TF}$ is obtained as

$$\mathbf{R}_{w,TF} = \mathbf{F}_M \mathbf{R}_w, \quad (3.161a)$$

$$\mathbf{r}_{w,TF} = (\mathbf{I}_N \otimes \mathbf{F}_M) \mathbf{r}_w, \quad (3.161b)$$

where $\mathbf{r}_{w,TF} = \text{vec}(\mathbf{R}_{w,TF})$. Later, we utilize symplectic finite Fourier transform (SFFT) in order to convert the data information from TF domain to DD domain as

$$\mathbf{y} = (\mathbf{F}_N \otimes \mathbf{F}_M^H) \mathbf{r}_{w,TF}, \quad (3.162)$$

and $\mathbf{Y} = \text{vec}^{-1}(\mathbf{y})$. In this line, after substituting (3.150), (3.155), and (3.161b) into (3.162) we complete the proof of Proposition 1. $\square$

Remark 4. It is worth noting that in case of conventional OTFS scheme without windowing ($W = 0$), $\mathbf{V} = \mathbf{I}_{MN}$, $\mathbf{W} = \mathbf{I}_{MN}$, and $\mathbf{B} \in \mathbb{C}^{(MN) \times (MN+G)}$, and thus, the effective channel under fractional Doppler in DD domain is given as

$$\mathbf{H}_{DD} = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{B} \mathbf{H}_T \mathbf{A} (\mathbf{F}_N^H \otimes \mathbf{I}_M). \quad (3.163)$$

Remark 5. Regarding (3.159), the statistical characteristics of the noise term in DD domain under Rx windowing are given as

$$\mathbf{C}_{n_w} = \sigma_o^2 \mathbf{P}_{DD}, \quad (3.164)$$

where σ_o^2 denotes the spectral density of additive white Gaussian noise in time domain. $\mathbf{P}_{DD} = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{P}_T (\mathbf{F}_N^H \otimes \mathbf{I}_M)$ and $\mathbf{P}_T = \mathbf{V} \mathbf{W} \mathbf{W}^H \mathbf{V}^H$ are diagonal Hermitian matrices, and the corresponding diagonal elements represent the power of the proposed Rx windowing in time domain and DD domain, respectively. Thus, the covariance matrix ($\mathbf{C}_{n_w}$) is not an eye matrix in DD domain, (i.e., $\mathbf{V} \mathbf{W} \mathbf{W}^H \mathbf{V}^H \neq \mathbf{I}_{MN}$), for the proposed RW-OTFS scheme. As a result, the received signal $\mathbf{y}$ with Rx windowing is exposed to colored noise, which increases the error in decision making for the data information while suppressing IDI.

In accordance with (3.162), the detection of the data symbols can be performed via either maximum likelihood (ML) technique or more practical Rx detectors such as minimum mean square error (MMSE), which are respectively given as follows

$$\hat{\mathbf{x}}_{ML} = \arg \min_{\mathbf{x} \in \mathbb{A}^{NM}} \|\mathbf{y} - \mathbf{H}_i \mathbf{x}\|^2, \quad (3.165a)$$

and

$$\hat{\mathbf{x}}_{\text{MMSE}} = \mathbf{H}_i^H (\mathbf{H}_i \mathbf{C}_{n_w}^{-1} \mathbf{H}_i^H + \mathbf{I}_{MN})^{-1} \mathbf{C}_{n_w}^{-1} \mathbf{y}, \quad (3.165b)$$

where $\hat{\mathbf{X}}_{\text{ML}} = \text{vec}^{-1}(\hat{\mathbf{x}}_{\text{ML}})$, $\hat{\mathbf{X}}_{\text{MMSE}} = \text{vec}^{-1}(\hat{\mathbf{x}}_{\text{MMSE}})$, and the effective channel with integer parameters as

$$\mathbf{H}_i = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{V} \mathbf{W} \mathbf{B} \mathbf{H}_{\text{T},i} \mathbf{A} (\mathbf{F}_N^H \otimes \mathbf{I}_M), \quad (3.166)$$

with $\mathbf{H}_{\text{T},i} = \sum_{i=0}^{L-1} h_i \mathbf{\Pi}^{l_i} \Delta^{k_i}$. Later, we demodulate the data symbols considering the modulation alphabet ($\mathbb{A}$) to recover the transmitted data bits.

In accordance with (3.165) and (3.166), it can be seen that both $\mathbf{H}_w$ and $\mathbf{H}_i$ have the same channel coefficients, delay spread and Doppler spread values. In addition, each channel tap of $\mathbf{H}_w$ suffers from fractional Doppler impairment (κ_i), which is relaxed with the aid of proposed Rx windowing technique RW-OTFS. Unfortunately, having CSI at the Rx side is a big challenge in practical scenarios, and it becomes more severe in case of OTFS transmission under fractional Doppler presence. Hence, the task of channel estimation at Rx becomes a crucial operation for signal detection. Among the existing channel estimation techniques for DD domain [128]–[131], this work utilizes an embedded pilot based approach [131]. Referring to **Figure 3.10(a)**, the reference pilot and data symbols in an OTFS frame are located as

$$x[k, l] = \begin{cases} x_p[k, l], & k = k_p, l = l_p, \\ 0, & k_p - G_k \leq k \leq k_p + G_k, \\ & l_p - G_l \leq l \leq l_p + G_l, \\ x_d[k, l], & \text{otherwise,} \end{cases} \quad (3.167)$$

where G_k and G_l represent the guard alongside the Doppler and delay axis, respectively, with $G_k \geq 2 \times \max\{k_i\}$ and $G_l \geq \max\{l_i\}$. Both the pilot and data symbols are placed in the same OTFS frame, and they are separated via guard to prevent interference between the pilot and data symbols in DD domain. Since $x_p[k, l]$ and its position (k_p, l_p) are known at the Rx side, the channel coefficients are estimated as discussed in [131]. Specifically, a threshold (Υ) method has been used to estimate the channel parameters. If the magnitude of DD data given in (3.162) is greater than the threshold, i.e., $|y[k, l]| \geq \Upsilon$, for $k_p - G_k/2 \leq k \leq k_p + G_k/2, l_p \leq l \leq l_p + G_l$, channel parameters are estimated as

$$\hat{h}_i[k - \hat{k}_i, l - \hat{l}_i] = y[k, l] / x_p[k, l], \quad (3.168)$$

where $\hat{h}_i$, $\hat{k}_i$, and $\hat{l}_i$ represents the estimates of h_i , k_i and l_i , respectively, with $i \in [1 \hat{L}]$. $\hat{L}$ denotes the number of detected channel taps. Otherwise, $\hat{h}_i[k - \hat{k}_i, l - \hat{l}_i] = 0$. In this regard, we extract the channel parameters by using the threshold $\Upsilon = 3\sigma_o$ as given in [131]. The pilot at position (k_p, l_p) can be spread alongside locations of $(k_p - \max\{k_i\}) \leq k \leq (k_p + \max\{k_i\})$ and $l_p \leq k \leq (l_p + \max\{l_i\})$ after passing the channel, as illustrated in **Figure 3.10(b)**. Since the sidelobes of $x[k, l]$ emerge due to κ_i , IDI occurs and limits the quality of channel estimation. As seen in **Figure 3.10(c)** and **Figure 3.10(d)**, RW-OTFS mitigates IDI around the desired pilot, resulting in better channel estimation. In this line, the estimated $\mathbf{H}_w$ is given as follows

$$\hat{\mathbf{H}}_w = (\mathbf{F}_N \otimes \mathbf{I}_M) \hat{\mathbf{H}} (\mathbf{F}_N^H \otimes \mathbf{I}_M), \quad (3.169)$$

with $\hat{\mathbf{H}} = \sum_{i=0}^{\hat{L}-1} \hat{h}_i \mathbf{\Pi}^{\hat{l}_i} \mathbf{\Delta}^{(\hat{k}_i)} \in \mathbb{C}^{(MN) \times (MN)}$. In this line, $\mathbf{x}$ is obtained by replacing $\mathbf{H}_i$ with $\hat{\mathbf{H}}_w$ in (3.165b).

3.7.2. Performance analysis of Rx windowing OTFS

This section presents the derived theoretical analysis for the bit error probability (BEP) in order to offer beneficial insights regarding the performance of the proposed RW-OTFS transmission scheme with practical pulse shaping.

To calculate the overall error performance of the proposed system, we rewrite the input-output relation in DD domain given in the Proposition 1 as follows

$$\mathbf{y} = \bar{\Phi}_{v,\tau}(\mathbf{x}) \mathbf{h} + \mathbf{n}_w, \quad (3.170)$$

where $\mathbf{h} = [h_0, h_1, \dots, h_{L-1}]^T$ and each element of $\mathbf{h}$ is an independent and identically distributed complex Gaussian random variable $h_i \sim \mathcal{CN}(\mu, 1/L)$. $\bar{\Phi}_{v,\tau}(\mathbf{x}) \in \mathbb{C}^{(MN) \times L}$ represents the concatenated matrix that includes the transmitted data $\mathbf{x}$, which is mathematically given as

$$\bar{\Phi}_{v,\tau}(\mathbf{x}) = [\bar{\Psi}_0 \mathbf{x} \mid \bar{\Psi}_1 \mathbf{x} \mid \dots \mid \bar{\Psi}_{L-1} \mathbf{x}], \quad (3.171)$$

where $\bar{\Psi}_i = (\mathbf{F}_N \otimes \mathbf{I}_M) \mathbf{V} \mathbf{W} \mathbf{B} \mathbf{\Pi}^{l_i} \mathbf{\Delta}^{(k_i + \kappa_i)} \mathbf{A} (\mathbf{F}_N^H \otimes \mathbf{I}_M)$. Moreover, τ and $\bar{v}$ denote the vector realization of delay values and Doppler values for the given wireless channel $\mathbf{h}$, respectively, with $\tau = [\tau_0, \tau_1, \dots, \tau_{L-1}]$, $\bar{v} = [\bar{v}_0, \bar{v}_1, \dots, \bar{v}_{L-1}]$, $\mathbf{v} = [v_0, v_1, \dots, v_{L-1}]$, and $\mathbf{v}_\kappa = [v_{\kappa,0}, v_{\kappa,1}, \dots, v_{\kappa,L-1}]$. Note that $\bar{v} = \mathbf{v} + \mathbf{v}_\kappa$, with $\tau_i = l_i/M\Delta f$, $v_i = k_i/NT$, and $v_{\kappa,i} = \kappa_i/NT$ considering (2.8). In addition, these indices follow a discrete uniform

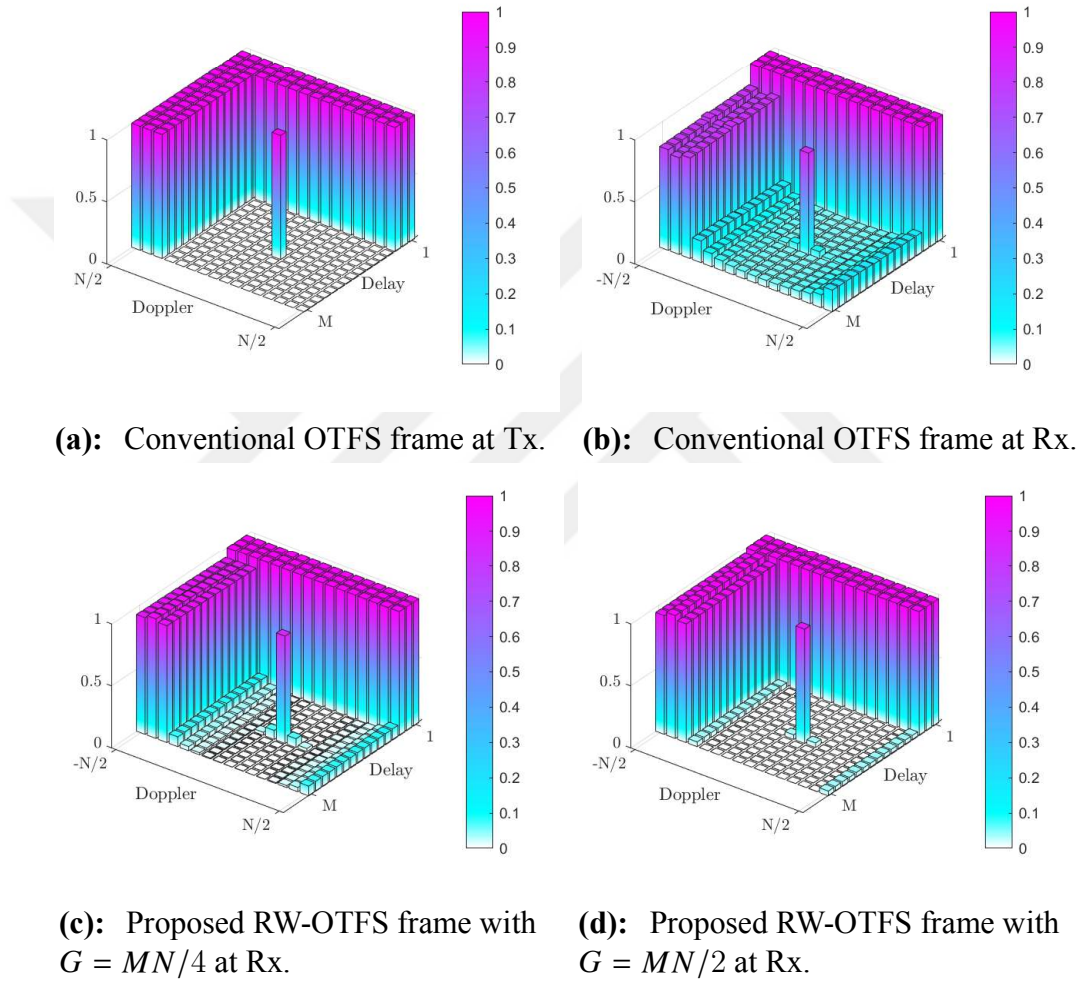


Figure 3.10: Illustration of X_{DD} and Y_{DD} with wireless channel parameters $L = 4$ and $\kappa = [0.15, 0.2, -0.15, -0.2]$.

distribution given as $\tau_i \in [0, \tau_{max}]$, $\nu_i \in [\nu_{min}, \nu_{max}]$, and $\kappa_i = [-\kappa_{max}, \kappa_{max}]$. Considering (3.171) and $\bar{\Psi}_i$, we can clearly see the impact of Rx windowing in the decision of data given in (3.170).

In line with (3.170) and (3.171), we consider the pairwise error probability (PEP) in order to derive the bit error rate (BER) performance of the RW-OTFS, assuming ML estimation at the Rx side. Particularly, we calculate the BER performance of the RW-OTFS under fractional Doppler impairment by taking into account the concept of the union bound as follows

$$\Pr\{e\} = \sum_{\mathbf{x}_i} \sum_{\mathbf{x}_j} \frac{1}{(MN2^{MN})} \Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j\} \varepsilon(\mathbf{x}_i, \mathbf{x}_j), \quad (3.172)$$

where $\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j\}$ represents the PEP for the decision when the symbol block $\mathbf{x}_i$ is transmitted and it is erroneously detected as $\mathbf{x}_j$. Moreover, $\varepsilon(\mathbf{x}_i, \mathbf{x}_j)$ denotes the corresponding bit-based pairwise error for the given symbol block pair. Moreover, the PEP in terms of averaging with respect to the joint PDF of $\mathbf{h}$ is given as

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j\} = \int \Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j | \mathbf{h}\} f_{\mathbf{h}}(\mathbf{h}) d\mathbf{h}, \quad (3.173)$$

where $f_{\mathbf{h}}(\mathbf{h})$ represents the joint probability density function (PDF) of $\mathbf{h}$, while $\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j | \mathbf{h}\}$ is the conditional PEP regarding $\mathbf{h}$. In accordance with (3.170) and the principle of the ML detector, the mathematical representation of the conditional PEP is given as

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j | \mathbf{h}\} = \Pr\{\|\mathbf{y} - \Phi_{\nu,\tau}(\mathbf{x}_i)\mathbf{h}\|^2 > \|\mathbf{y} - \Phi_{\nu,\tau}(\mathbf{x}_j)\mathbf{h}\|^2\}, \quad (3.174)$$

where $\Phi_{\nu,\tau}(\mathbf{x}_i) = [\Psi_0 \mathbf{x}_i | \Psi_1 \mathbf{x}_i | \dots | \Psi_{L-1} \mathbf{x}_i]$ and $\Psi_i = (F_N \otimes I_M) V W B \Pi^{l_i} \Delta^{k_i} A (F_N^H \otimes I_M)$, indicating that no fractional Doppler occurs, i.e., $\kappa = 0$.

After utilizing the norm expansion $\|\mathbf{y} - \Phi_{\nu,\tau}(\mathbf{x})\mathbf{h}\|^2 = \|\mathbf{y}\|^2 + \|\Phi_{\nu,\tau}(\mathbf{x})\mathbf{h}\|^2 - 2\Re\{\mathbf{y}^H \Phi_{\nu,\tau}(\mathbf{x})\mathbf{h}\}$, where $\Re\{\cdot\}$ denotes the real-part operator, (3.174) is rewritten as

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j | \mathbf{h}\} = \Pr\{\|\Phi_{\nu,\tau}(\mathbf{x}_i)\mathbf{h}\|^2 - \|\Phi_{\nu,\tau}(\mathbf{x}_j)\mathbf{h}\|^2 > 2\Re\{\mathbf{y}^H [\Phi_{\nu,\tau}(\mathbf{x}_i) - \Phi_{\nu,\tau}(\mathbf{x}_j)]\mathbf{h}\}\}. \quad (3.175)$$

Later, after performing some mathematical manipulations, the conditional PEP is calculated as follows

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j \mid \mathbf{h}\} = \Pr\{\mathbf{h}^H \mathbf{S} \mathbf{h} > 2\Re\{\mathbf{h}^H \mathbf{Z} \mathbf{h}\} + 2\Re\{\mathbf{n}_w^H [\Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j)] \mathbf{h}\}\}, \quad (3.176)$$

where $\mathbf{S} = \Phi_{v,\tau}(\mathbf{x}_i)^H \Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j)^H \Phi_{v,\tau}(\mathbf{x}_j)$ and $\mathbf{Z} = \bar{\Phi}_{v,\tau}(\mathbf{x}_i)^H [\Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j)]$. Consequently, we calculate the closed-form expression for the conditional PEP as given in the following proposition.

Proposition 2 (Conditional PEP). *The conditional PEP for the proposed RW-OTFS transmission when the transmitted symbol block $\mathbf{x}_i$ is erroneously detected as $\mathbf{x}_j$ under fractional Doppler is given with the aid of Q -function as [112]*

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j \mid \mathbf{h}\} = Q\left(\frac{\mathbf{h}^H \mathbf{T} \mathbf{h}}{\sqrt{2\sigma_o^2 \mathbf{h}^H \boldsymbol{\delta} \mathbf{h}}}\right), \quad (3.177)$$

where $\mathbf{T} = \mathbf{Z} + \mathbf{Z}^H - \mathbf{S}$, and

$$\boldsymbol{\delta} = (\Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j))^H \mathbf{P}_{DD} (\Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j)). \quad (3.178)$$

Proof. In accordance with (3.175), the distribution characteristics of the noise term are calculated as

$$2\Re\{\mathbf{n}_w [\Phi_{v,\tau}(\mathbf{x}_i) - \Phi_{v,\tau}(\mathbf{x}_j)] \mathbf{h}\} \sim \mathcal{CN}(0, 2\sigma_o^2 \mathbf{h}^H \boldsymbol{\delta} \mathbf{h}). \quad (3.179)$$

Moreover, after utilizing the norm expression, $2\Re\{\mathbf{h}^H \mathbf{Z} \mathbf{h}\}$ is simplified as

$$2\Re\{\mathbf{h}^H \mathbf{Z} \mathbf{h}\} = \|\mathbf{h}\|^2 + \|\mathbf{Z} \mathbf{h}\|^2 - \|\mathbf{h} - \mathbf{Z} \mathbf{h}\|^2, \quad (3.180a)$$

$$= \mathbf{h}^H \mathbf{h} + (\mathbf{Z} \mathbf{h})^H \mathbf{Z} \mathbf{h} - \mathbf{h}^H (\mathbf{I} - \mathbf{Z})^H (\mathbf{I} - \mathbf{Z}) \mathbf{h}, \quad (3.180b)$$

$$= \mathbf{h}^H (\mathbf{I} + \mathbf{Z}^H \mathbf{Z}) \mathbf{h} - \mathbf{h}^H (\mathbf{I} - \mathbf{Z})^H (\mathbf{I} - \mathbf{Z}) \mathbf{h}, \quad (3.180c)$$

$$= \mathbf{h}^H (\mathbf{Z} + \mathbf{Z}^H) \mathbf{h}, \quad (3.180d)$$

where after substituting (3.180) into (3.176), we complete the proof of Proposition 2. $\square$

Taking into account the fact that $\mathbf{T}$ and $\boldsymbol{\delta}$ are Hermitian matrices, we diagonalize these matrices with the aid of unitary transformation as follows

$$\mathbf{T} = \mathbf{U}^H \boldsymbol{\Gamma} \mathbf{U}, \quad (3.181a)$$

$$\delta = \mathbf{O}^H \mathbf{\Lambda} \mathbf{O}, \quad (3.181b)$$

where $\mathbf{U}$ and $\mathbf{O}$ are unitary matrices, i.e., $\mathbf{U}^H \mathbf{U} = \mathbf{I}$ and $\mathbf{O}^H \mathbf{O} = \mathbf{I}$, respectively, while $\bar{\mathbf{h}} = \mathbf{U}^H \mathbf{h}$ and $\tilde{\mathbf{h}} = \mathbf{O}^H \mathbf{h}$. Moreover, $\mathbf{\Gamma}$ and $\mathbf{\Lambda}$ are diagonal matrices, containing the singular values of $\mathbf{T}$ and δ . Since $\bar{\mathbf{h}}$ and $\tilde{\mathbf{h}}$ are related to $\mathbf{h}$ via the unitary transform, they have the same statistical characteristics, and thus, after substituting (3.181) into (3.177), we write (3.177) as follows

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j \mid \mathbf{h}\} = Q\left(\frac{\bar{\mathbf{h}}^H \mathbf{\Gamma} \bar{\mathbf{h}}}{\sqrt{2\sigma_o^2 \tilde{\mathbf{h}}^H \mathbf{\Lambda} \tilde{\mathbf{h}}}}\right), \quad (3.182)$$

where $\bar{\mathbf{h}}^H \mathbf{\Gamma} \bar{\mathbf{h}} = \sum_{i=0}^{L-1} \Gamma_i |\bar{h}_i|^2$ and $\tilde{\mathbf{h}}^H \mathbf{\Lambda} \tilde{\mathbf{h}} = \sum_{i=0}^{L-1} \Lambda_i |\tilde{h}_i|^2$, respectively. To further simplify (3.182) and express the conditional PEP as the ratio of independent and identically distributed (i.i.d) exponential random variables [132], we utilize the Cholesky decomposition as follows

$$\bar{\mathbf{h}}^H \mathbf{\Gamma} \bar{\mathbf{h}} = \bar{\mathbf{h}}^H \mathbf{C}^H \mathbf{C} \bar{\mathbf{h}} = \mathbf{K}^H \mathbf{K} = \sum_{i=0}^{L-1} \varphi_i, \quad (3.183a)$$

$$\tilde{\mathbf{h}}^H \mathbf{\Lambda} \tilde{\mathbf{h}} = \mathbf{K} \mathbf{Q} \mathbf{K} = \sum_{i=0}^{L-1} q_i \varphi_i, \quad (3.183b)$$

where $\mathbf{C}$ denotes the Cholesky factor of $\mathbf{\Gamma}$, i.e., $\mathbf{C} = \mathbf{\Gamma}^{\frac{1}{2}}$, while $\mathbf{C}^H \mathbf{C}$ is a diagonal matrix with $\mathbf{K} = \mathbf{C} \bar{\mathbf{h}}$ and $\bar{\mathbf{h}} = \mathbf{K}(\mathbf{C}^H)^{-1}$. Regarding (3.183b), $\mathbf{Q} = (\mathbf{C}^H)^{-1} \mathbf{\Lambda} (\mathbf{C})^{-1}$ denotes a diagonal matrix.

Proposition 3 (Conditional PEP). *We obtain the expression for the conditional PEP given in (3.182) as the sum of exponential random variables via Cauchy-Schwarz inequality, as follows*

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j \mid \mathbf{h}\} \leq Q\left(\sqrt{\frac{1}{2\sigma_o^2} \sum_{i=0}^{r-1} \frac{1}{q_i} \varphi_i}\right). \quad (3.184)$$

Proof. Considering (3.183), we rewrite (3.182) as

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j \mid \mathbf{h}\} = Q\left(\sqrt{\frac{1}{2\sigma_o^2} \frac{(\sum_{i=0}^{L-1} \varphi_i)^2}{\sum_{i=0}^{L-1} q_i \varphi_i}}\right), \quad (3.185)$$

and with the aid of Cauchy-Schwarz inequality, the inner part of $Q(\cdot)$ is upper bounded by

$$\left(\sum_{i=0}^{r-1} \varphi_i\right)^2 \leq \left(\sum_{i=1}^r \frac{\varphi_i}{q_i}\right) \left(\sum_{i=0}^{r-1} q_i \varphi_i\right). \quad (3.186)$$

Consequently, with the substitution of (3.186) into (3.185) yields (3.184) that proves Proposition 3. $\square$

After averaging over (3.184) with respect to φ_i by using the alternative Q -function, i.e., $Q(x) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} e^{\left(-\frac{x}{2N_0 \sin^2(\theta)}\right)} d\theta$ and moment generating function (MGF) [122], we give the upper bound of unconditioned PEP as follows

$$\Pr\{\mathbf{x}_i \rightarrow \mathbf{x}_j\} \leq \prod_{i=0}^{r-1} \left(1 + \frac{1}{4\sigma_o^2 q_i L}\right)^{-1}, \quad (3.187)$$

where r represents the rank of $\mathbf{Q}$.

Remark 6. Considering the fact that $\mathbf{Q} = (\mathbf{C}^H)^{-1} \mathbf{\Lambda} (\mathbf{C})^{-1}$ and $\mathbf{C}$ being the Cholesky factor of $\mathbf{\Gamma}$ that depends on $\mathbf{T}$, q_i values dependent on both on $\mathbf{T}$ and δ . The effect of the fractional Doppler in the proposed OTFS transmission is in $\mathbf{T}$, while the impact of windowing exists in δ due to $\mathbf{P}_{DD}$. Thus, in case of zero fractional Doppler $\mathbf{T}$ converts to δ , with q_i only relaying on δ .

Lastly, substituting (3.187) into (3.172) yields the upper bound BER of the proposed RW-OTFS.

CHAPTER 4

4. RESULTS AND DISCUSSION

In this section, we present mathematical and simulation findings, accompanied by thorough discussions, for the aforementioned studies.

4.1. SC Transmission: Adaptive Radio Resource Utilization

This specific work is dedicated to evaluate SC transmission parameters and to find a good balance between these parameters for achieving URLLC in particular.

4.1.1. Impact of α and ζ on η

Influence of ζ on transmission latency is investigated in the following. Oversampling rate for pulse shaping $OSR = 8$ and a number of data symbols $S = 1000$ are used to obtain simulation results.

The required α and ζ pairs for SC transmission to achieve $\gamma = 30$ dB desired SINR are shown in **Table 4.1**. γ is calculated by error vector magnitude (EVM) [133]. The desired SINR level can be achieved by different α and ζ pairs. Regarding (3.2) and (3.3), each α and ζ pair corresponds to different levels of system efficiency which is shown in **Table 4.2**.

If only α factor is considered to provide the desired SINR with minimum ζ , the calculated η values are shown by the first row in **Table 4.2**. Since $\alpha=1$ is required, η reduces to 49.89%. In other words, the system suffers from loss of 50% efficiency, while providing minimum latency by the lowest ζ . Likewise, a decline of η to 50% is observed for minimum $\alpha = 0$, which corresponds to maximum η_f . Regarding (3.1), $\alpha = 0.1$ and $\zeta = 42$ offer maximum $\eta = 87.23\%$ for URLLC applications with exact desired $\gamma = 30$ dB. Latency can be further

Table 4.1: Achieved SINR levels for various α and ζ pairs

ζ	α	0	0.1	0.2	0.4	0.6	0.8	1
2		8.33	9.22	10.30	13.18	17.39	23.68	29.95
6		13.77	16.67	21.59	29.44	29.66	29.66	30.00
10		16.17	21.79	29.83	30.00	30.00	30.00	30.00
18		18.82	29.32	29.88	29.97	29.99	30.00	30.00
26		20.37	29.80	29.95	30.00	30.00	30.00	30.00
34		21.50	29.87	30.00	30.00	30.00	30.00	30.00
42		22.30	30.00	30.00	30.00	30.00	30.00	30.00
50		22.93	29.97	30.00	30.00	30.00	30.00	30.00
10^3		29.56	30.00	30.00	30.00	30.00	30.00	30.00

Table 4.2: Achieved $\eta(\%)$ for various α and ζ pairs

ζ	α	0	0.1	0.2	0.4	0.6	0.8	1
2		99.79	90.72	83.16	71.28	62.37	55.44	49.89
6		99.39	90.36	82.83	70.99	62.12	55.22	49.70
10		99.00	90.00	82.50	70.71	61.87	55.00	49.50
18		98.22	89.29	81.85	70.16	61.39	54.57	49.11
26		97.45	88.59	81.21	69.61	60.91	54.14	48.73
34		96.70	87.91	80.58	69.07	66.44	53.72	48.35
42		95.96	87.23	79.96	68.54	56.97	53.31	47.98
50		95.23	86.57	79.36	68.02	59.52	52.90	47.61
10^3		50.00	45.45	41.66	35.71	31.25	27.78	25.00

reduced for $\alpha = 0.2$ and $\zeta = 34$ at the cost of 7% loss on system efficiency. Moreover, reducing γ from 30dB to 29.32dB by $\alpha = 0.1$ and $\zeta = 18$ offers both latency reduction and enhancement on system efficiency.

4.1.2. Impact of S on transmission latency

Since transmission latency in SC communication is decided by data size S as well as ζ value, **Table 4.3** illustrates the impact of S on system efficiency for the desired SINR = 30 dB. Lower S values, correspond to lower latency and require higher α to achieve the desired SINR. Therefore, it results in a loss of system efficiency as in (3.3). Despite higher system efficiency with higher S , the use of lower α and larger ζ values results in a high level of OOB and PAPR that destroys the system reliability. Therefore, the impact of α and ζ pair on OOB and PAPR should be considered to decide appropriate S , which provides low latency by maintaining reliability.

Table 4.3: Achieved system efficiency regarding different S values for SINR=30dB.

S	α	ζ	$\eta(\%)$
10^5	0.05	40	95.20
10^4	0.05	40	94.86
10^3	0.05	40	91.56
200	0.1	22	81.85
10^2	0.15	14	76.55
10	0.35	6	45.94

4.1.3. Impact of α and ζ on OOB

Applications and use cases with ultra-reliable requirements are highly sensitive to adjacent channel interference (ACI) due to OOB. For this reason, OOB is illustrated for various ζ and α in **Figure 4.1(a)** and **Figure 4.1(b)**, respectively.

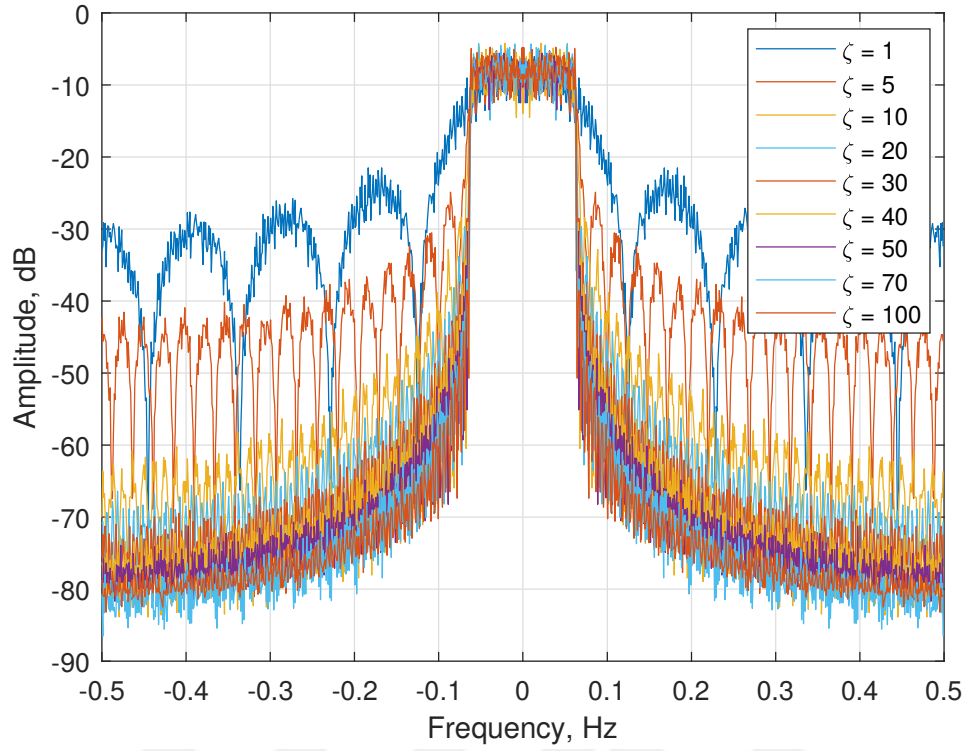
In **Figure 4.1(a)**, it is shown that the more ζ increases, the more OOB decreases. Furthermore, larger ζ causes more spread in time domain for RRC pulse, which is associated with better localization in frequency domain. As seen in the figure, the impact of ζ on OOB reduces for ζ larger than 70 while $\alpha = 0$. In case of higher α values for a given ζ , OOB further reduces, as visualized in **Figure 4.1(b)**. As a result, increase of both α and ζ reduces OOB, which is shown in **Table 4.4**.

Table 4.4: Calculated OOB for various α and ζ pairs

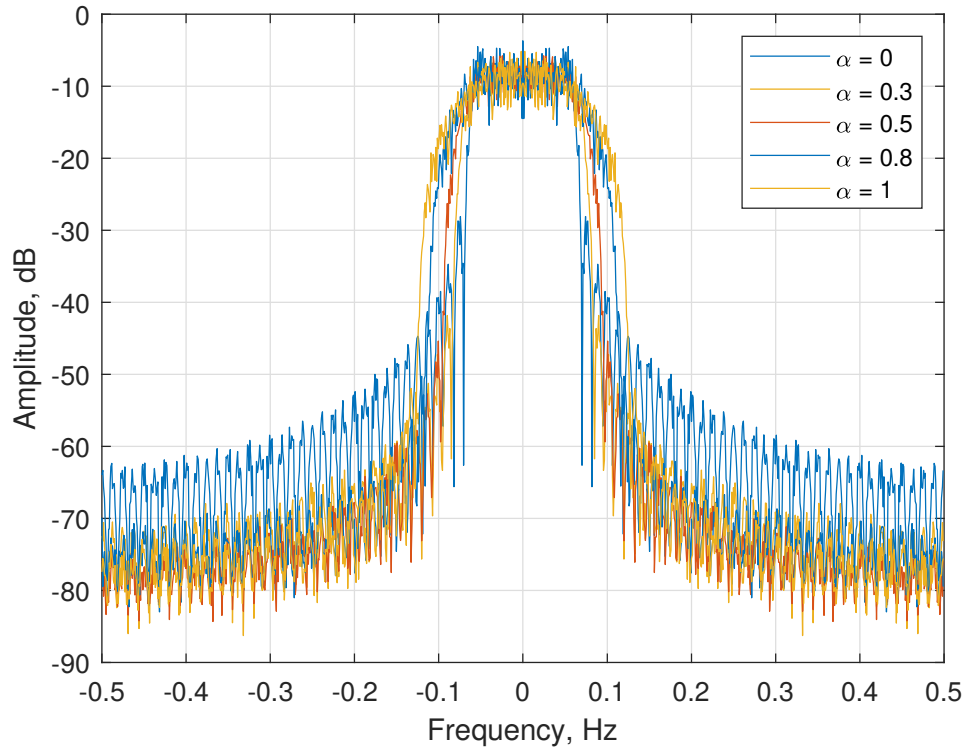
$\zeta \backslash \alpha$	0	0.1	0.2	0.4	0.6	0.8	1
2	-13.44	-14.89	-16.59	-19.27	-21.55	-24.61	-28.56
6	-18.16	-22.50	-27.40	-36.91	-37.51	-41.42	-42.44
10	-20.62	-27.87	-35.96	-41.83	-43.78	-45.77	-47.21
18	-23.74	-36.33	-42.88	-47.04	-47.97	-49.14	-49.89
26	-26.10	-41.60	-45.47	-47.21	-47.47	-47.80	-47.98
34	-27.96	-43.27	-47.02	-48.37	-49.14	-49.86	-50.32
42	-29.37	-44.79	-47.10	-48.65	-49.24	-49.83	-50.35
50	-30.88	-45.46	-47.62	-48.57	-49.19	-49.89	-50.34
10^3	-46.89	-49.14	-50.49	-51.55	-51.87	-52.45	-53.03

4.1.4. Impact of ζ and α on PAPR

PAPR levels due to different α and ζ pairs are summarized in **Table 4.5**. A larger ζ value not only increases transmission latency but also PAPR since more symbols are added



(a): OOBE for various ζ values for minimum $\alpha = 0$.



(b): OOBE for various α factors for $\zeta = 10$.

Figure 4.1: OOBE in conventional SC transmission considering various α and ζ pairs.

constructively or destructively at a given data sample. In general, an increase of α factor up to 0.4 decreases PAPR since the side lobes of the time domain pulse reduce. For $\alpha \geq 0.4$, the impact of pulse side lobes becomes insignificant, and the PAPR level saturates.

Table 4.5: Calculated PAPR in dB scale for various α and ζ pairs

$\zeta \backslash \alpha$	0	0.1	0.2	0.4	0.6	0.8	1
2	2.54	2.42	2.21	1.99	2.92	3.17	2.15
6	5.97	5.68	5.01	3.21	3.32	3.32	3.53
10	7.24	6.72	5.55	3.47	3.40	3.43	3.63
18	8.17	7.12	5.68	3.60	3.45	3.47	3.67
26	8.26	7.18	5.71	3.57	3.46	3.47	3.67
34	8.34	7.20	5.72	3.59	3.46	3.48	3.68
42	8.38	7.20	5.70	3.60	3.47	3.48	3.68
50	8.52	7.17	5.71	3.59	3.47	3.48	3.69
10^3	9.02	7.54	6.11	3.98	3.86	3.88	4.10

In order to achieve URLLC by optimized α and ζ pair for a desired OOB and PAPR, both Table 4.4 and Table 4.5 should be taken into consideration. For instance, ultra reliable system with maximum $\eta = 80.58\%$ for a desired OOB ≤ -40 dB and PAPR ≤ 6 is provided by $\alpha = 0.2$ and $\zeta = 34$ pair, according to Table 4.2. In case of applications with critical transmission latency, ζ value can be reduced with the penalty of η loss. Furthermore, regarding Table 4.3 data size $S \geq 200$ is not suitable to maintain desired latency and reliability.

The proposed SC transmission with adaptive α within a block provides more η with the increase of η_f for the desired OOB and PAPR, as illustrated in Figure 4.2.

4.2. Frequency Domain IM: Simulation Results And Discussions

In this section, a simulation study is adopted to investigate the performance of OFDM-IM, OFDM-GIM and OFDM-SNM under the effect of RF-impairments. System parameters used in all simulations are $N = 128$, $L = 10$, $CP = 16$, and subblock size of $n = 4$. Different RF-impairment parameters are used for CFO and IQ imbalance which are faced in practical scenarios. For all the simulations, BPSK modulation is used for utilized subcarriers. In order to provide a fair comparison between the aforementioned IM techniques two different numbers of active subcarriers are used for OFDM-GIM ($k = [1, 2]$) and ($k = [1, 2, 3]$). By considering the number of utilized subcarriers, OFDM-GIM

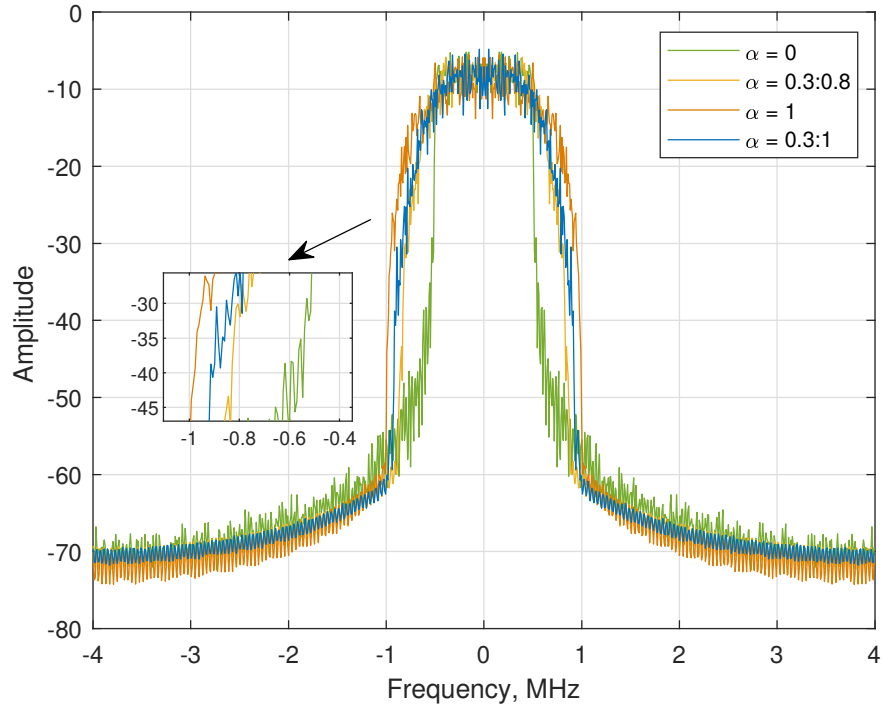


Figure 4.2: Maximizing η_f with adaptive α usage while satisfying requirements of URLLC.

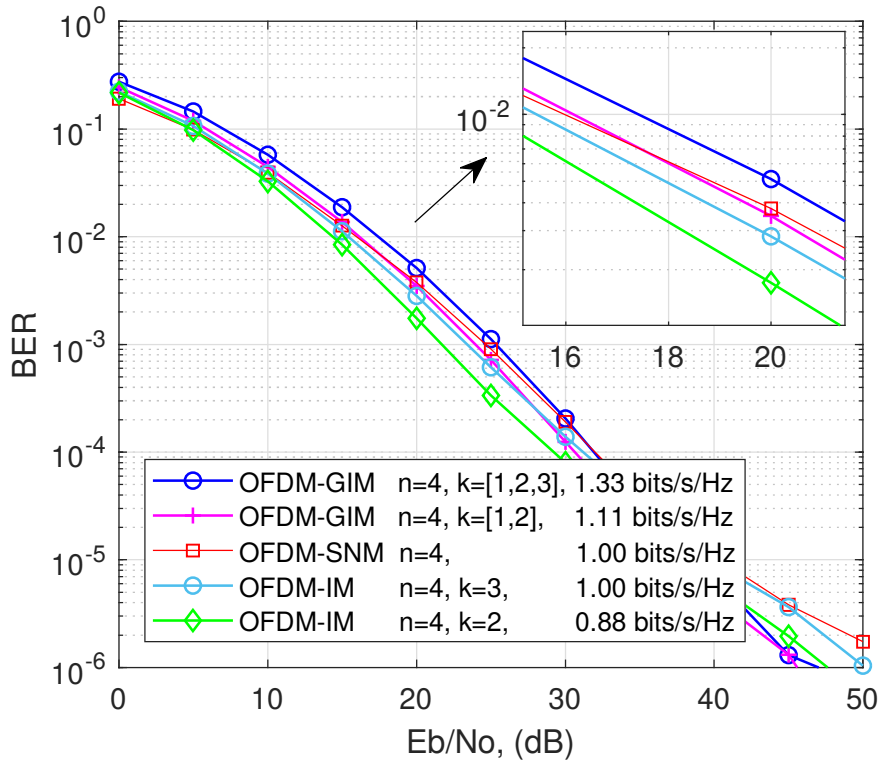


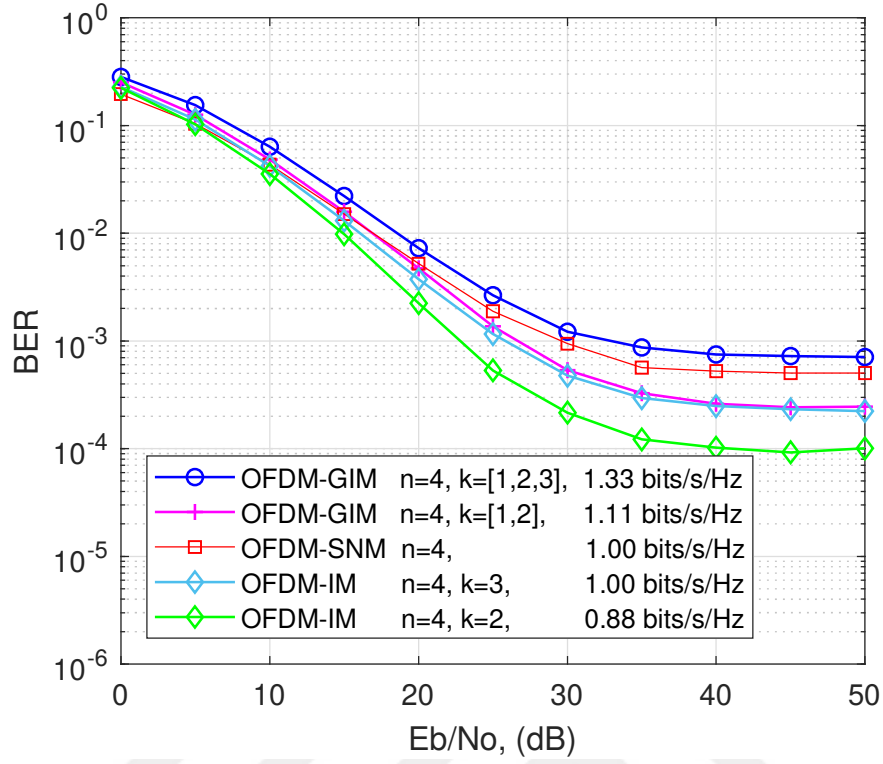
Figure 4.3: BER vs SNR for OFDM-IM, OFDM-GIM and OFDM-SNM.

($k = [1, 2]$) and ($k = [1, 2, 3]$) offer similar selectivity with OFDM-IM and OFDM-SNM, respectively.

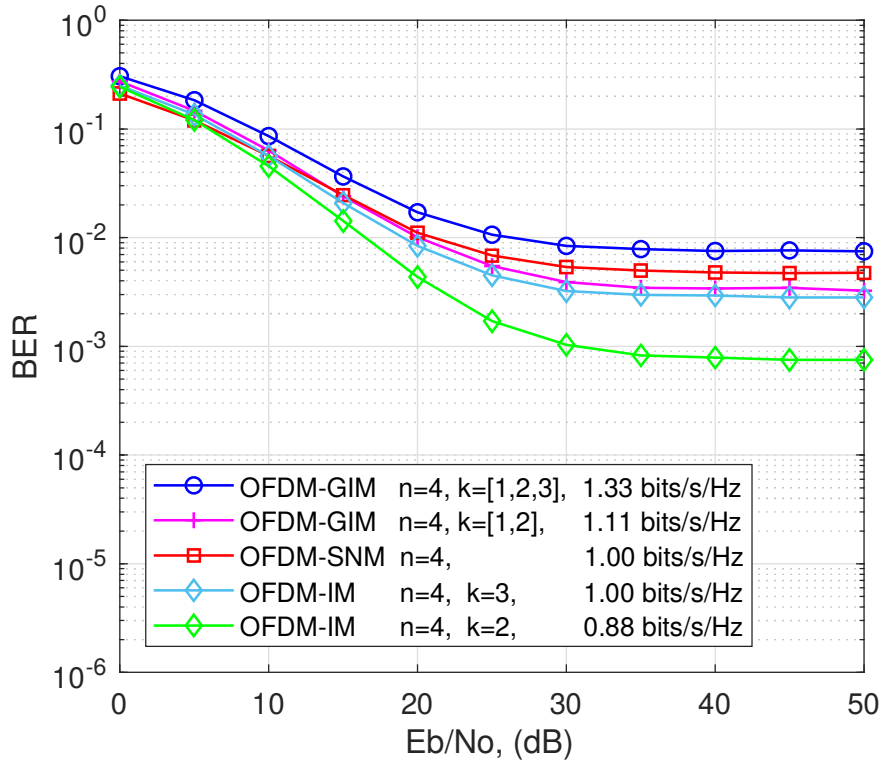
In **Figure 4.3**, simulation results of OFDM-IM, OFDM-GIM, and OFDM-SNM with ideal RF Rx are illustrated [32], [36], [37]. OFDM-IM scheme provides a slight better error performance. p_1 bits in OFDM-IM and OFDM-SNM are mapped to subcarrier indices and number of active subcarriers, respectively. Due to inherent successive subcarrier activation of OFDM-SNM, frequency selectivity can not be fully exploited as in OFDM-IM. The worst performance belongs to OFDM-GIM since both indices and number of active subcarrier are used to carry p_1 data bits. OFDM-GIM can provide the best spectral efficiency with a modest loss on error performance.

In **Figure 4.4**, the impact of CFO on error performance of the IM techniques is illustrated. Regarding (3.12), each subcarrier suffers not only from interference of all the remaining active subcarriers but also from their placement, too. OFDM-IM ($k = 2$) and OFDM-GIM ($k = [1, 2]$) have better performance compared with OFDM-SNM and OFDM-GIM ($k = [1, 2, 3]$) since the number of utilized subcarriers is less, as illustrated in **Figure 4.4(a)**. Even though, the average number of used subcarriers per subblock in OFDM-GIM ($k = [1, 2]$) is 1.5, it provides worse error performance than OFDM-IM ($k = 2$) and ($k = 3$) due to usage of two stage Rx which results in higher error propagation. In **Figure 4.4(b)**, the advantage of OFDM-IM against carrier frequency mismatch is more visible with the increase of CFO ($\omega = 0.1$). Moreover, performance of OFDM-GIM ($k = [1, 2]$) became similar with OFDM-SNM since the Rx performance gets worse.

In **Figure 4.5**, it is shown the error performance of frequency domain IM schemes under the impact of IQ imbalance. Considering (3.18), each subcarrier is only affected by its mirror subcarrier. Therefore, the distance between active subcarriers becomes less important than that of CFO scenario. OFDM-IM still provides the best error performance under IQ imbalance among the evaluated IM techniques, as in **Figure 4.5(a)**. Different from the CFO scenario, OFDM-SNM offers better BER performance even for OFDM-GIM ($k = [1, 2]$) due to less importance of successive subcarrier activation in the presence of IQ imbalance. As expected, in **Figure 4.5(b)** use of higher IQ imbalance parameters ($\epsilon = 0.1, \Delta\phi = 10^\circ$) further reduces BER performances of the IM techniques. For further insights, the aforesaid frequency domain IM techniques offer the same peak-to-average



(a) Normalized CFO ($\omega = 0.05$)



(b) Normalized CFO ($\omega = 0.1$)

Figure 4.4: BER versus SNR performance for IM schemes under CFO.

power ratio (PAPR) since total energy is equally distributed among the active subcarriers.

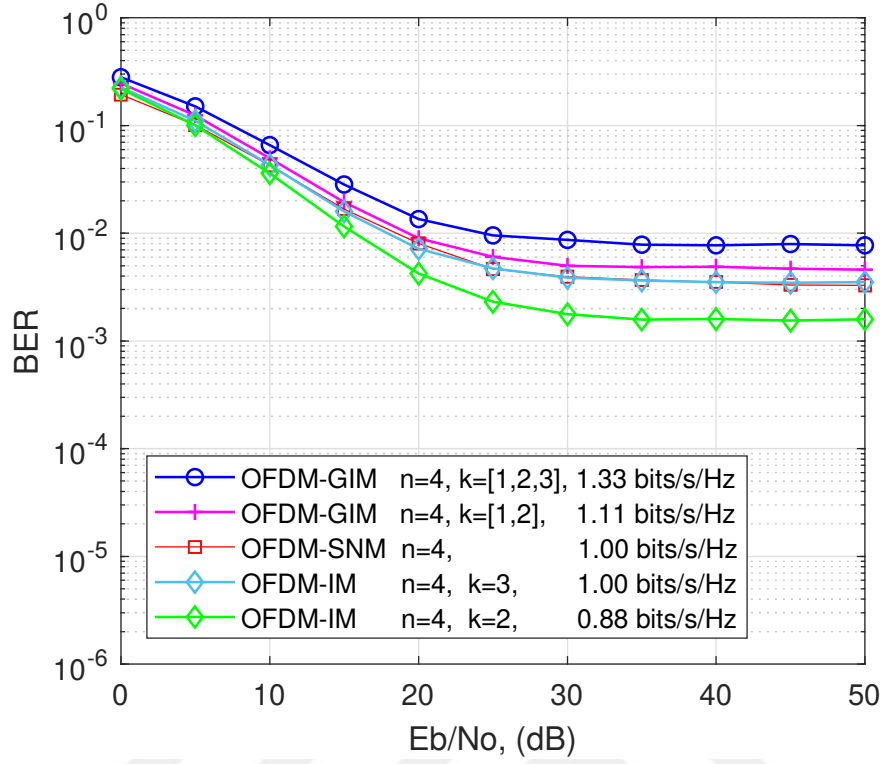
In all simulation results, it is observed that OFDM-IM with non-ideal Rx is superior to OFDM-GIM and OFDM-SNM owing to its one stage Rx and non-successive subcarrier allocation, respectively.

4.3. Hybrid NOMA: Numerical And Simulation Results

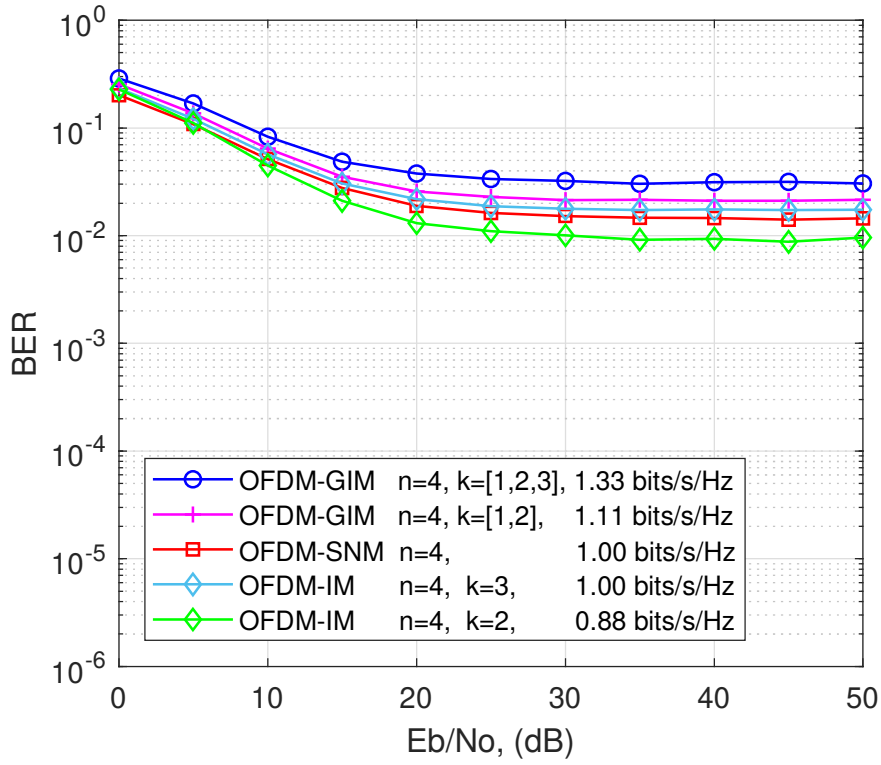
In this section, a simulation study through the Monte Carlo model is adopted to investigate the performance of the proposed IM-NOMA for different configurations and make a comparison with the classical OFDM-NOMA. System parameters used in all simulations are $N = 64$, $L = 10$, $L_{CP} = 16$, and BPSK modulation.

Figure 4.6 provides the achievable rate of OFDM-NOMA and IM-NOMA with two different configurations ($b = 4, a = 2$) and ($b = 8, a = 4$) under target of $R_{FUE} = 2$ bps/Hz as a function of transmitted power. Average channel gains of N_{UE} and F_{UE} are set to $\Omega_{f,1} = 20$ dB and $\Omega_{f,2} = 10$ dB, respectively. Full search algorithm is used to split transmit power (P) between UEs while meeting strict rate constraints and maximizing total achievable sum rate. In this study, the step size between two consecutive power factors is 0.01. Both IM-NOMA and OFDM-NOMA satisfy the rate need of F_{UE} . Moreover, IM-NOMA with ($b = 8, a = 4$) provides 2 dB gain over the OFDM-NOMA for a total system rate of $R = 6$ bps/Hz. IM-NOMA with ($b = 4, a = 2$) offers the same sum rate with OFDM-NOMA since equal number of data bits is transmitted over the same resources.

In **Figure 4.7**, the average BER performances of IM-NOMA and OFDM-NOMA are presented. The power allocation coefficients in **Figure 4.7(a)** correspond to $\beta_1 = 0.2$ and $\beta_2 = 0.8$ to meet R_{FUE} for the given $\Omega_{f,1}$ and $\Omega_{f,2}$. The theoretical calculation for IM-NOMA with ($b = 4, a = 2$) is obtained to validate computer-based simulations. It is clearly seen that OFDM-NOMA offers the worst performance due to full overlapping between UEs. Specifically, to achieve 10^{-4} BER, N_{UE} and F_{UE} require 30 dB and 40 dB SNR, respectively. IM-NOMA with ($b = 4, a = 2$) provides the same error performance for N_{UE} and F_{UE} at 22 dB and 25 dB SNR, while maintaining the same system sum rate. IM-NOMA with ($b = 8, a = 4$) increases the achievable rate at the cost of reducing reliability with respect to IM-NOMA with ($b = 4, a = 2$), but still much better than OFDM-NOMA. Lastly, the performance of the proposed Rx is evaluated when similar power is



(a) ($\epsilon = 0.05, \Delta\phi = 5^\circ$)



(b) ($\epsilon = 0.1, \Delta\phi = 10^\circ$)

Figure 4.5: BER versus SNR performance for IM schemes under IQ imbalance.

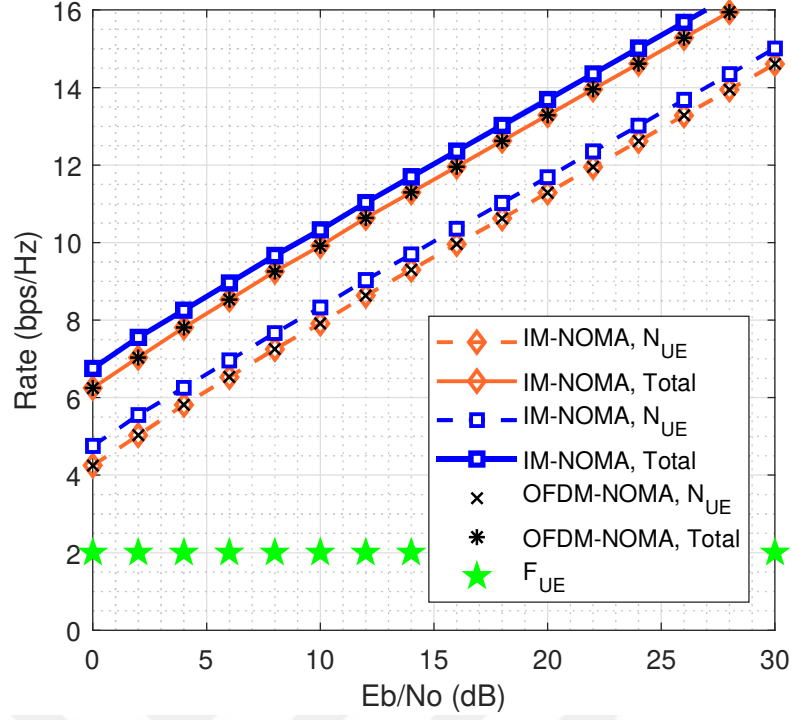
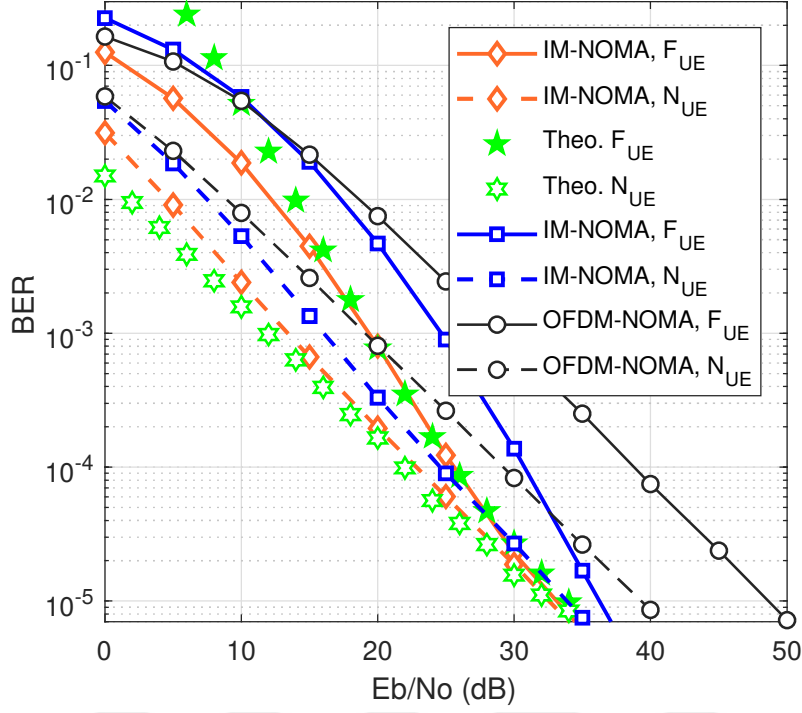


Figure 4.6: Achievable rate vs. SNR for proposed IM-NOMA and OFDM-NOMA schemes. Orange and blue lines correspond to F_{UE} with $(b=4, a=2)$ and $(b=8, a=4)$, respectively.

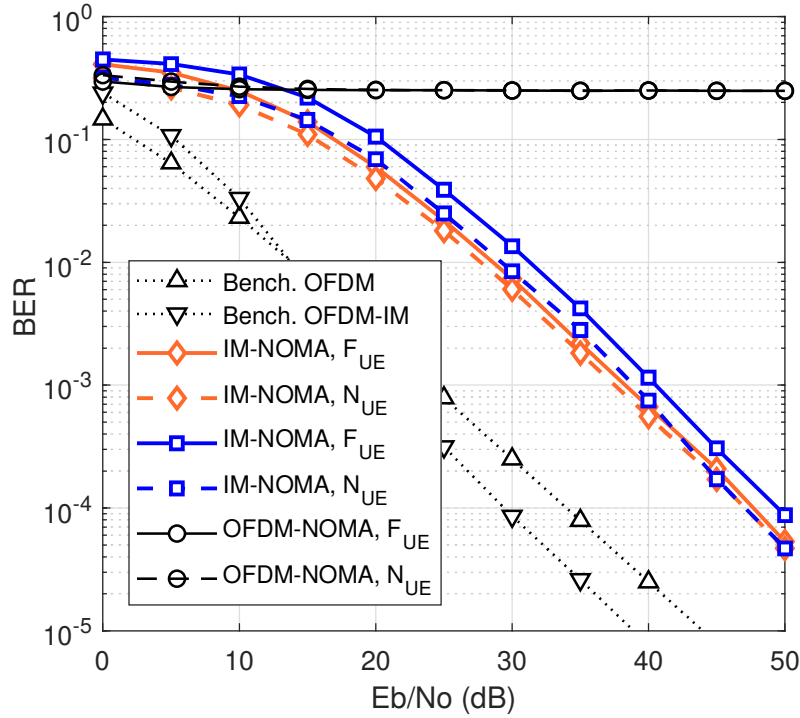
assigned to UEs for a given $(\Omega_{f,1}, \Omega_{f,2})$ and $R_{F_{UE}}$, as shown in **Figure 4.7(b)**. In OFDM-NOMA, neither F_{UE} nor N_{UE} is able to detect its own signal since MUD is not feasible in scenarios with zero SINR. On the other hand, the proposed Rx exploits not only power multiplexing but also partial spectrum utilization. Hence, it offers significant BER gain for reliable communication even when power multiplexing fails. Note that the use of interleaved IM-NOMA will further improve the achieved reliability [117]. Consequently, the IM-NOMA concept can provide a higher achievable sum rate and system reliability by different configurations of power allocation and subcarrier activation of OFDM-IM. Moreover, instead of ML, log-likelihood ratio (LLR) detector can be used to achieve the same complexity compared with the conventional OFDM-NOMA system with a slight compromise on error performance.

4.4. Multiuser OTFS: Numerical Results And Discussions

In this section, we assess the performance of the proposed multi-user OTFS scheme via Monte-Carlo-based simulations considering various practical scenarios. In this work,



(a): $\beta_1 = 0.2, \beta_2 = 0.8$.



(b): $\beta_1 = 0.5, \beta_2 = 0.5$.

Figure 4.7: BER vs. SNR for proposed IM-NOMA and OFDM-NOMA schemes. Orange and blue lines correspond to F_{UE} with $(b=4, a=2)$ and $(b=8, a=4)$, respectively.

we assume the delay index and Doppler index are uniformly generated from the set $\{0, \dots, M - 1\}$ and $\{0, \dots, N - 1\}$, respectively. To have a better understanding of the derived mathematical equations and to assess their validity, we first simulate the BER performance of the proposed OTFS-based multiple accessing scheme with binary phase shift keying (BPSK) modulation, considering ML Rx. It is worth mentioning that since the complexity of ML detector increases exponentially with the parameters of OTFS frame. Hence, we select small values of N and M to assess the multi-user diversity coming from the time/frequency dispersion of the wireless channel of each UE in OTFS-based systems.

Figure 4.8 shows the BER performance of three different multi-user OTFS schemes with $N = 2, M = 4$ for BPSK transmission. The solid-line with diamond marker represents the performance of multi-user OTFS scenario in case the RR are randomly assigned to a single UE, whilst the number of channel taps P_i for each UE is uniformly distributed in the range of $[P_{min}, \dots, P_{max}]$. In other words, this result shows the performance of OTFS in case no scheduling mechanism exists to exploit the multi-user diversity. Alternatively, an intelligent assignment of the RR with the aid of our proposed scheduling scheme enhances significantly the total error performance of OTFS systems as illustrated via the solid-line with circle marker. Moreover, the error performance of the system get worse while the UE with minimum channel taps performs transmission via the solid-line with square marker. Thus, **Figure 4.8** explicitly illustrates the wide range on the performance difference between the two possible user scheduling scenarios with extreme channel tap configurations. In line with this, the dashed-line with circle marker validates the theoretical results, which matches with the simulation results for the proposed scheduling scheme. We numerically average the conditional error performance subjected to all possible delay and Doppler realizations in order to obtain the multi-user diversity gain. The proposed scheme enhances the reliability of OTFS by at least 10dB compared with the random scheduling scenario at 10^{-4} BER. It is worth noting that the performance of OTFS transmission with low values of N, M and P_u does not guarantee the full diversity within all possible realizations of $\mathbf{\Gamma}$, resulting in gap between simulation results and theory due to the union bound technique (3.49).

Different from **Figure 4.8**, **Figure 4.9(a)** and **Figure 4.9(b)** illustrate the error performance of the proposed scheduling scheme with practical values of (N, M) and MMSE detector. Even in the case of small OTFS parameters ($M = 2, N = 4$), our proposed scheme offers

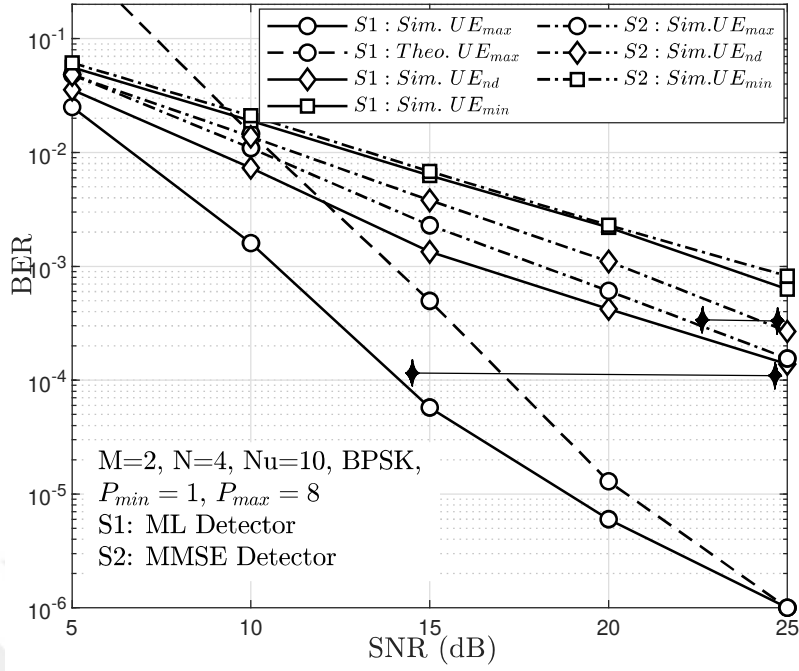
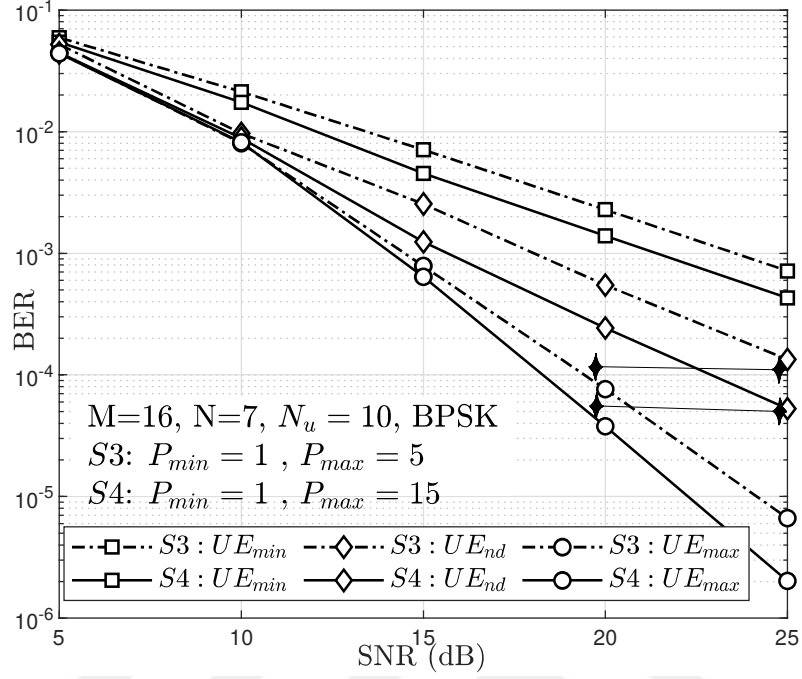


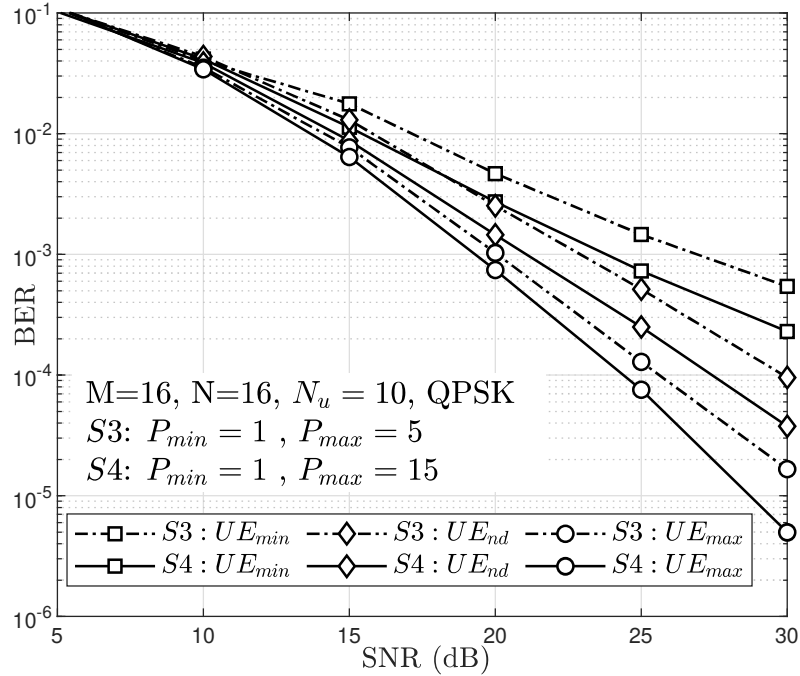
Figure 4.8: Theoretical vs. simulation results of the proposed scheme.

at least 2dB more gain compared to the random scenario. It is important to notice that, since the diversity of OTFS transmission with small parameters is significantly reduced, ML and MMSE detectors offer similar system error performance. **Figure 4.9(a)** shows the performance of an OTFS system with practical parameters ($M = 16, N = 7$) where the proposed scheme offers at least 5dB better BER performance compared with the random scenario for BPSK modulation. Moreover, in **Figure 4.9(b)**, similar effect occurs on the error performance for OTFS system with QPSK modulation.

Figure 4.10 assesses the performance of the proposed OTFS scheduling scheme under fractional Doppler ($|\kappa| \in [0, \dots, 0.1]$) and imperfect CSI, considering channel errors as $\hat{h}_u^i = h_u^i + \Delta h_u^i$, where Δh_u^i is modeled as an $CN(0, \epsilon 10^{-\frac{\gamma_{dB}}{10}})$ and $\hat{h}_u^i = h_u^i$ if $\epsilon = 0$. The error performance of the proposed technique is robust against CSI error, leading to a shift for all the results in accordance with the level of error (Δh_i). It is worth noting that the effect of channel impairments increases with the diversity of the OTFS system. For this reason, we assess the performance of the system under imperfect CSI in case of extreme $\epsilon = 0.1$ value and practical $\epsilon = 0.05$ value, shown via dotted line and dot-dashed line, respectively. The minimum error channel impact occurs in case of minimum channel spread, since the UE diversity is minimized. Even though the proposed scheme suffers the most from channel



(a): MMSE detector for BPSK transmission.



(b): MMSE detector for QPSK transmission.

Figure 4.9: BER performances for the proposed multi-user OTFS scheme considering various practical scenarios.

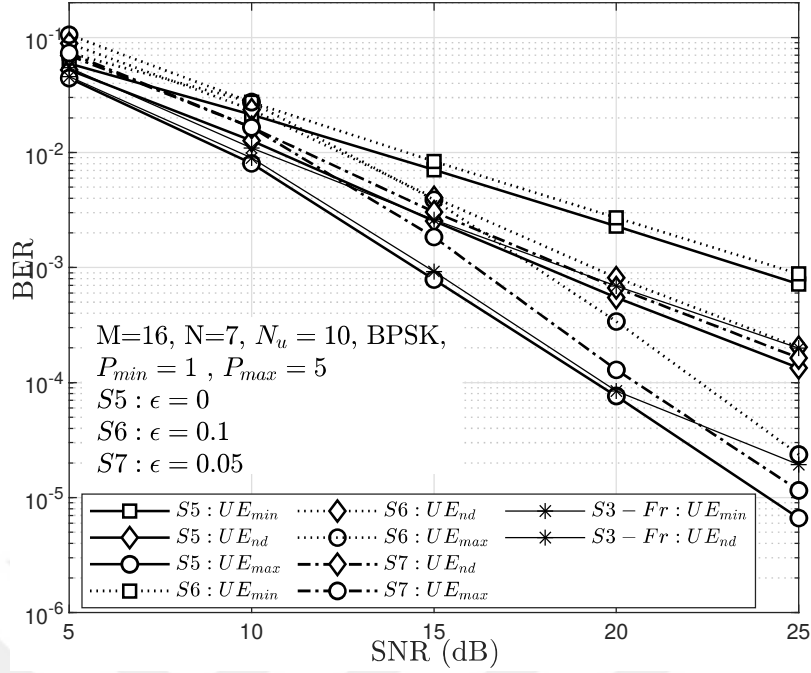


Figure 4.10: The proposed scheme under imperfect CSI and fractional Doppler.

errors, it provides the best BER performance. Moreover, the superiority of this scheme becomes more and more prominent considering practical CSI error values (i.e., $\epsilon = 0.05$). Likewise, the proposed scheduling mechanism outperforms the random RR assignment in the presence of the fractional Doppler, as illustrated via the starred lines. Besides, the Rx complexity of the OTFS-based transmission depends on the dimensions of $\mathbf{H}_{u,eff}$ matrix for the MMSE reception. Since the size of $\mathbf{H}_{u,eff} \in \mathbb{C}^{MN \times MN}$ is irrelevant of the number of channel taps P_u of the desired UE, the proposed multi-user OTFS scheme does not bring extra complexity. Specifically, the number of complex multiplications (CM) needed by the conventional MMSE Rx (3.39b) is $\mathcal{O}(M^3N^3)$. In addition, CSI determines the user fairness for the real systems. However, in this work, we consider that the channel characteristics are similar for each UE since the impact of channel fading reduces with the increase of the number of channel taps in OTFS systems.

4.5. IQ Imbalance In The Delay-Doppler Domain: Numerical Results And Discussions

In this section, we investigate and discuss the effect of IQ imbalanced Tx in OTFS transmission for various practical configurations and scenarios considering EVA channel model

[134]. We perform Monte-Carlo-based simulations to evaluate the error performance of IQ imbalanced Tx in OTFS transmission under ideal and realistic wireless channel conditions. Since the design of OTFS transmission is composed of several parameters, it is worthy to revise the way how these parameters are related in realistic OTFS transmission. For instance, in case we have the maximum supportable delay spread and Doppler spread as $\tau_{max} = M - 1/M\Delta f$ and $\nu_{max} = N - 1/NT$, respectively, we need to choose the system parameters i.e., bandwidth $B = M\Delta f$ and transmission time $T_f = NT$ considering the constraints of $\tau_{max} < T$ and $\nu_{max} < \Delta f$ of the wireless multipath channel. As an example, let's assume a fast time-varying wireless channel model with $\tau_{max} = 1 \mu s$ and $\nu_{max} = 100$ Hz, and the communication system characterized by bandwidth of $B = 10$ MHz within latency $T_f = 1$ ms, hence we need to use a subcarrier spacing of $\Delta f = 40$ kHz with $M = 250$ and $N = 40$ to support the communication of the system. Additionally, to explicitly show the effects of the IQ imbalanced Tx in OTFS transmission, we also take the values of the delay taps ν_i and Doppler taps τ_i as multiples of the delay step size $1/M\Delta f$ or Doppler step size $1/NT$. In all simulations, we assume M-ary symbols with constant envelope for various configurations of N , M , and channel taps L under ML detector and minimum mean square error (MMSE) detector as follows.

In **Figure 4.11**, we show the BER performance of both ideal and IQ imbalanced OTFS Tx with the ML detection in order to illustrate the effect of IQ imbalance on the diversity gain of OTFS transmission. Since the complexity of the ML Rx severely increases with the increase of N and M values, we perform OTFS ($M = 2, N = 4$) transmission to evaluate the mathematical analysis. To observe the effect of IQ imbalance for small N and M values, we choose IQ parameters as $\Delta\phi = 15^\circ$ and $\epsilon = 0.2$ for BPSK transmission. As shown in **Figure 4.11**, IQ imbalanced Tx limits the diversity gain of the OTFS transmission only with respect to Doppler, leading to overall limited system diversity but does not cause an error floor as in conventional one-dimensional communication systems, i.e., OFDM [135]. Importantly, the simulation results match with the theoretical results as in (3.112), which clearly shows that BER is not only a function of rank (r) but also a function of the SNR. Although the rank gets smaller under IQ imbalance effect, the system performance does not saturate with the increase of SNR. Additionally, we show the performance of IQ imbalanced Tx for OTFS transmission with imperfect CSI ($e = 0.25$),

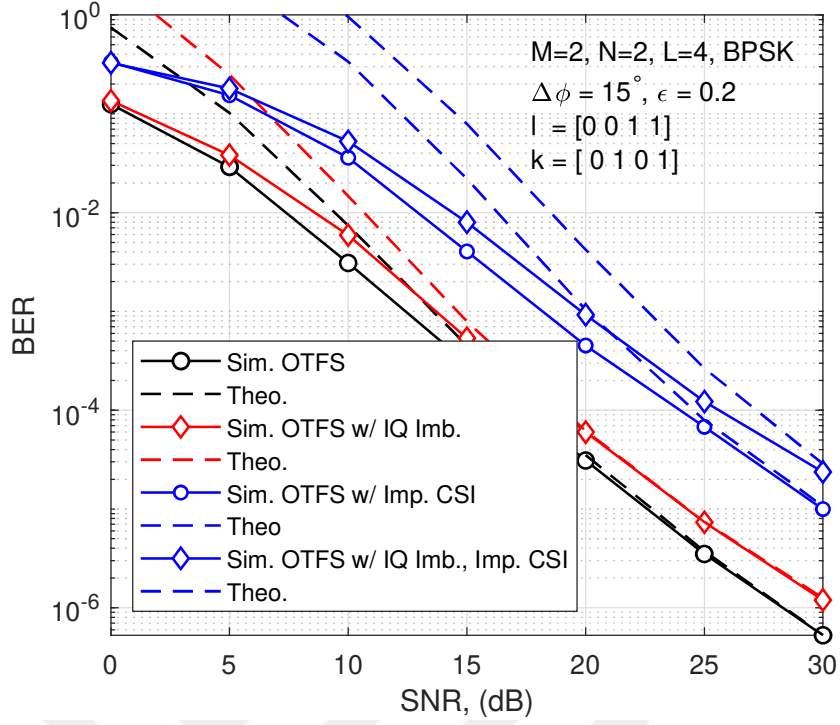


Figure 4.11: BER performance of OTFS with ML Rx under IQ imbalanced Tx.

which causes a linear degradation in the error performance compared with ideal CSI scenario. We observe the same physical effect of IQ imbalance on the OTFS transmission for high order modulation, i.e., QPSK modulation, as shown in **Figure 4.12**.

In **Figure 4.11** and **Figure 4.12**, we use small values of N and M , which are not practical for standards of wireless communication technologies, in order to perform a feasible ML detection. Therefore, we evaluate the performance of IQ imbalanced Tx for OTFS transmission with practical values of M and N in **Figures 4.13**, **4.14**, and **4.15**. In particular, we consider two different wireless channel scenarios, i.e., scenario 1 (S1) and scenario 2 (S2), for realistic OTFS transmission [50], [71], [102]. S1 and S2 are characterized by different τ values, as follows. In Scenario 1 (S1), we adopt Extended Vehicular A (EVA) model that is the accepted channel model in LTE. The model defines delay profile for an environment with mobility, where the number of channel taps is $L = 9$. In Scenario 2 (S2), to provide comprehensive evaluation of OTFS performance under IQ imbalance impact, we do not limit τ values to EVA model, but we generate τ values considering the relation between OTFS transmission parameters and wireless channel characteristics, and thus l_i can take any value in delay axis between 0 and $M - 1$ that is explained above. In S2, we assume $L=9$ as in S1. In both scenarios, we generate Doppler values (v_i) considering maximum

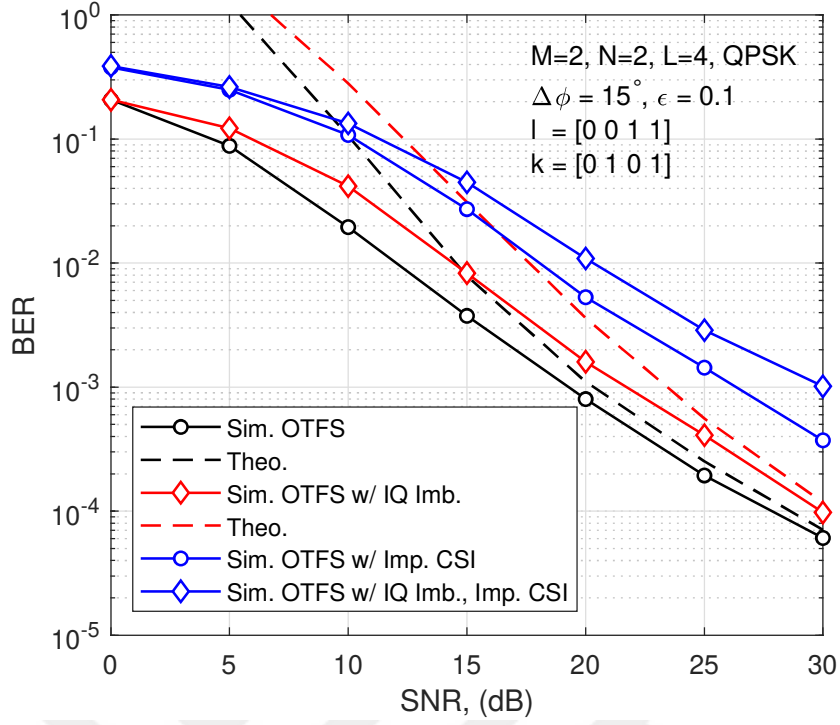


Figure 4.12: BER performance of OTFS with ML Rx IQ imbalanced Tx.

Table 4.6: The system parameters for realistic OTFS transmission

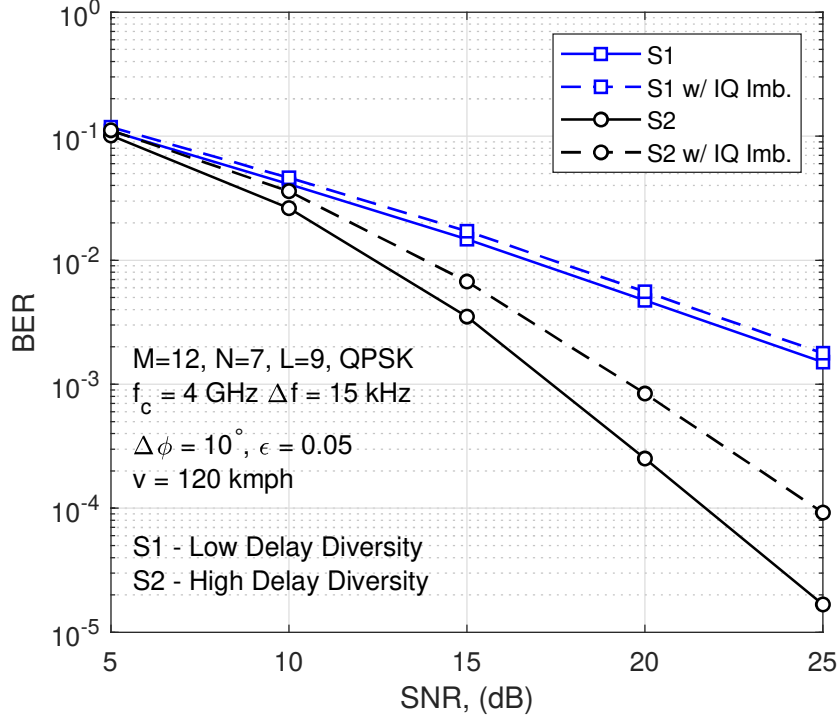
System Parameters	
Carrier Freq.	4 GHz
SCS (Δf)	15 kHz
User Speed	120 kmph, 500 kmph
(M,N)	(12,7), (128,7)
L	9
Rx	MMSE

Doppler shift (v_{max}), given that $v_i = v_{max} \cos(\Delta\phi_i)$, where $\Delta\phi_i$ refers to angle of arrival that is uniformly distributed $\Delta\phi_i \in [-\pi, \pi]$ regarding the Jake's model. Moreover, v_{max} depends on the user speed (v) and operating frequency (f) by $v_{max} = fv/c$, where c is the propagation speed of the signal.

Table I shows the used system parameters for OTFS transmission, where $M = 12$ and $N = 7$ corresponds to the resource unity in classical Long-Term Evolution (LTE) wireless broadband standard. Table II provides both the delay profile for LTE-EVA channel model and the calculation of Doppler values considering the user speed. In S1, since τ values are decided considering the EVA channel model, the wireless channel is not fully spread over the delay domain. Differently, in S2, l_i can take any value in delay axis between 0 and

Table 4.7: Delay and Doppler parameters in realistic OTFS transmission

LTE-EVA Channel τ_i μ s for $i = 1, \dots, L$								
0	0.03	0.15	0.31	0.37	0.71	1.09	1.73	2.51
$v_i = v_{max} \cos(\Delta\phi_i)$ Hz for $i = 1, \dots, L$								
$\Delta\phi_i \in [-\pi, \pi]$			$v_{max_{120}} = 0.44$ kHz			$v_{max_{500}} = 1.85$ kHz		

**Figure 4.13:** BER performance of OTFS with MMSE Rx under IQ imbalance considering vehicle speed 120 kmph.

$M - 1$, and thus wireless channel is well-spread in delay domain and low correlation exists in the received OTFS transmission. A moving vehicle with speed 120 kmph is considered. In **Figure 4.13**, we see that IQ imbalanced Tx slightly reduces the performance of OTFS transmission in S1, while it significantly degrades the system performance in S2 by around 3 dB at 10^{-4} BER. This is because IQ imbalance not only causes power degradation (α) and MDI interference on the desired signal as we prove in Theorem 1, but also introduces correlation in the OTFS transmission that limits the wireless channel diversity. In **Figure 4.14**, we consider the user speed 500 kmph that results in higher v_{max} . In this way, we observe the IQ imbalance impact not only on delay diversity but also on the Doppler diversity, which it can be easily seen in S1 while comparing **Figure 4.13** and **Figure 4.14**. For instance, the performance of IQ imbalanced Tx decreases by at least 2 dB in S1 with speed 500 kmph, while less than 1 dB performance degradation occurs in S1 with speed

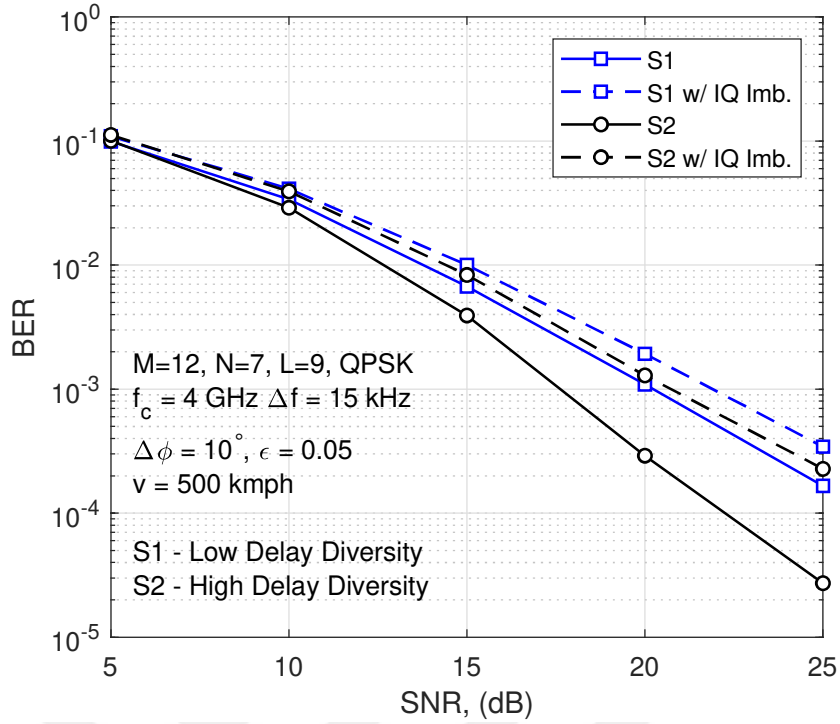


Figure 4.14: BER performance of OTFS with MMSE Rx under IQ imbalance considering vehicle speed 500 kmph.

120 kmph. **Figure 4.15** shows the results for high values of $N = 7$ and $M = 128$ in OTFS systems under practical IQ imbalance values $\Delta\phi = 2^\circ$ and $\epsilon = 0.15$ dB to provide a comprehensive evaluation of Tx IQ imbalance effect in delay-Doppler domain. As in the above discussions, although OTFS transmission with the user speed 500 kmph offers better BER performance than OTFS transmission with the user speed 120 kmph, it is more affected under the impact of IQ imbalance. In other words, as the system diversity of OTFS increases via either high (N,M) values or high velocity, the impact of IQ imbalance becomes significant. For instance, IQ imbalanced Tx limits the performance of OTFS user with speed 500 kmph two times more than OTFS user with speed 120 kmph. Therefore, the impact of IQ imbalance becomes more critical in OTFS with high channel dispersion, which is inevitable to avoid in beyond 5G vehicular wireless communications.

4.6. IM-aided IQ Imbalance Compensator: Numerical Results And Discussions

In this section, we assess and discuss the performance of our proposed scheme against the impact of IQ imbalance in OTFS transmission at Tx and Rx sides. We perform Monte-Carlo based simulations for various practical configurations and scenarios considering

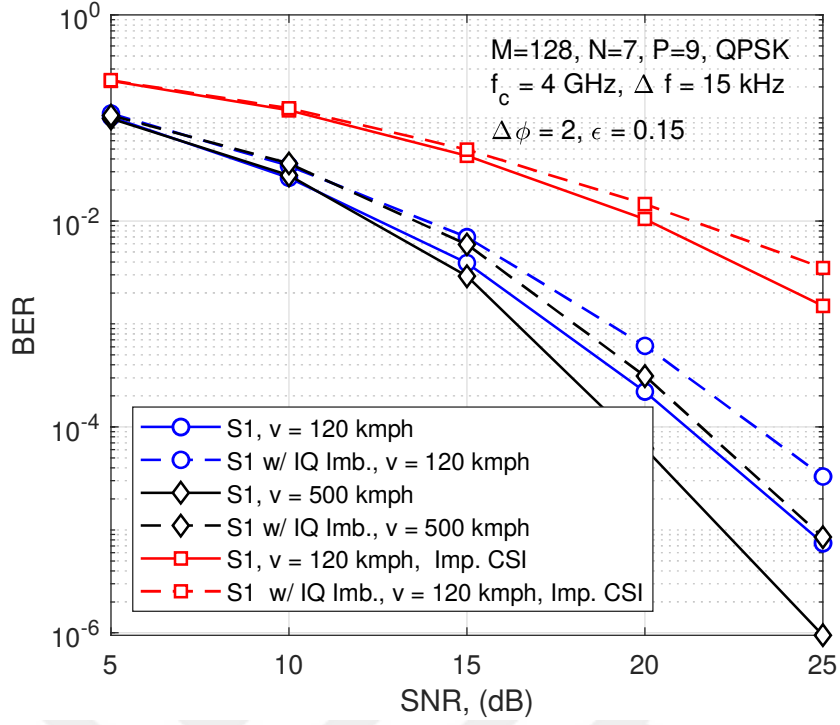


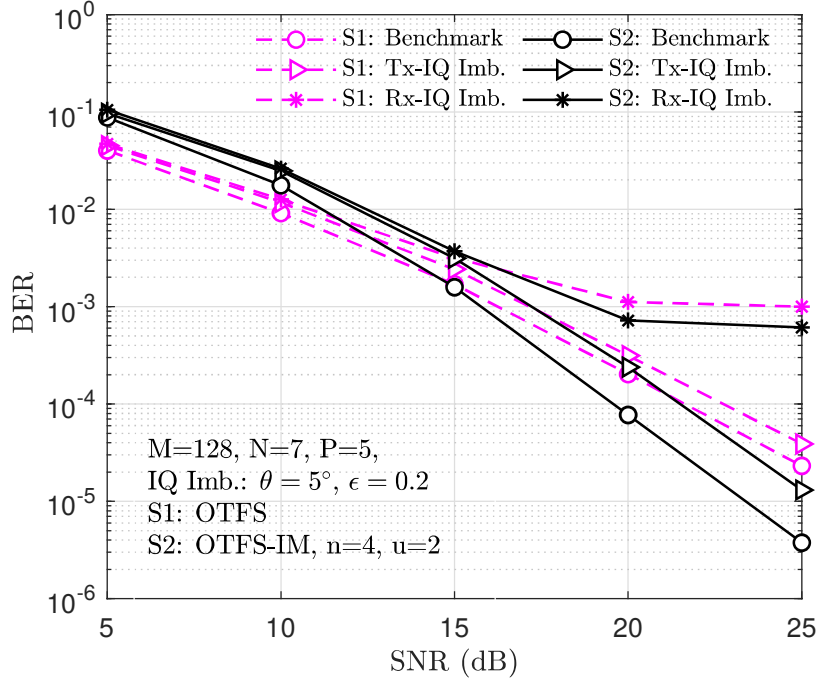
Figure 4.15: BER performance of OTFS with MMSE Rx under IQ imbalanced Tx considering high values of $M = 128$, $N = 7$ and practical IQ imbalance parameters $\Delta\phi = 2^\circ$, $\epsilon = 0.15$.

LTE-EVA channel model and OTFS with realistic pulse shape [102]. OTFS transmission is composed of several system parameters, hence it is important to revisit the way how these parameters are related in practical wireless communications. If the system bandwidth $B = M\Delta f$ and transmission time $T_f = NT$, the maximum supportable delay and Doppler spread equal $\tau_{max} = M - 1/M\Delta f$ and $\nu_{max} = N - 1/NT$, respectively, considering the constraints of $\tau_{max} < T$ and $\nu_{max} < \Delta f$ of the time variant wireless channel. We evaluate the performance of OTFS system under realistic pulse shaping where the system parameters are provided in **Table 4.8**. The user speed 500 km/h corresponds to a maximum Doppler shift of $\nu_{max} = 1.85$ kHz and Doppler tap values are generated as $\nu_i = \nu_{max} \cos(\phi_i)$, where $\phi_i \in \{-\pi, \pi\}$ is uniformly distributed. Moreover, we generate the delay taps considering LTE-EVA channel model. In all simulations, we assume Q -ary symbols with constant envelope including binary phase-shift keying (BPSK) and quadrature phase-shift keying (QPSK) for various configurations of IM parameters n and u . In all the simulations, we utilize the practical values of IQ imbalance parameters $\theta = 5^\circ$ and $\epsilon = 0.2$ for both Tx and Rx sides.

Table 4.8: System parameters of OTFS for realistic scenarios

System Parameters					
f_c	Δf	User Speed	(M,N)	L	Rx
4 GHz	15 kHz	500 km/h	(128,7)	5	MMSE

In **Figure 4.16**, we show the performance of conventional OTFS and OTFS-IM under practical Rx IQ imbalance for BPSK modulation. Although OTFS-IM is more affected due to the IQ imbalance, it offers better error performance compared with OTFS at high SNR values [92]. Moreover, the sparse structure of OTFS-IM allows us to compensate the impact of IQ imbalance in OTFS, as we show in the following results. For instance, **Figure 4.17** shows the performance of OTFS-IM under Tx IQ imbalance considering two different modulation schemes. To provide the same spectral efficiency with classical OTFS we choose OTFS-IM ($n = 4, u = 2$) and ($n = 8, u = 6$) for BPSK and QPSK transmission, respectively. “Benchmark” refers to the performance of OTFS-IM without IQ imbalance, “Tx-IQ Imb.” shows the system performance under Tx IQ imbalance, and “Compensator” shows the system performance with the proposed compensator. As seen in the figure by green lines, the proposed technique easily mitigates the effect of Tx IQ imperfection under both BPSK and QPSK transmission.

**Figure 4.16:** BER performance of OTFS and OTFS-IM with MMSE under Tx and Rx IQ imbalance considering BPSK.

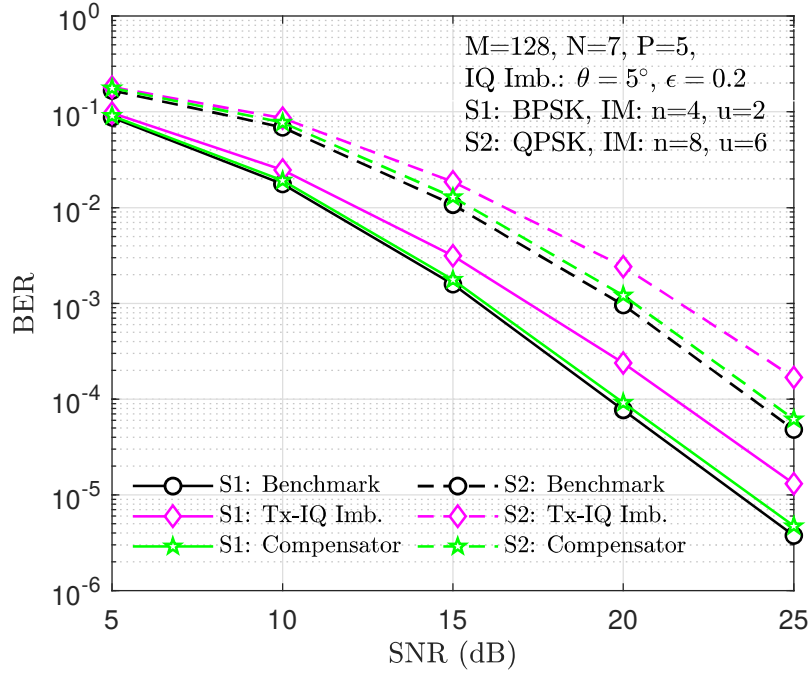


Figure 4.17: BER performance of OTFS-IM with MMSE under Tx IQ imbalance considering BPSK and QPSK modulation.

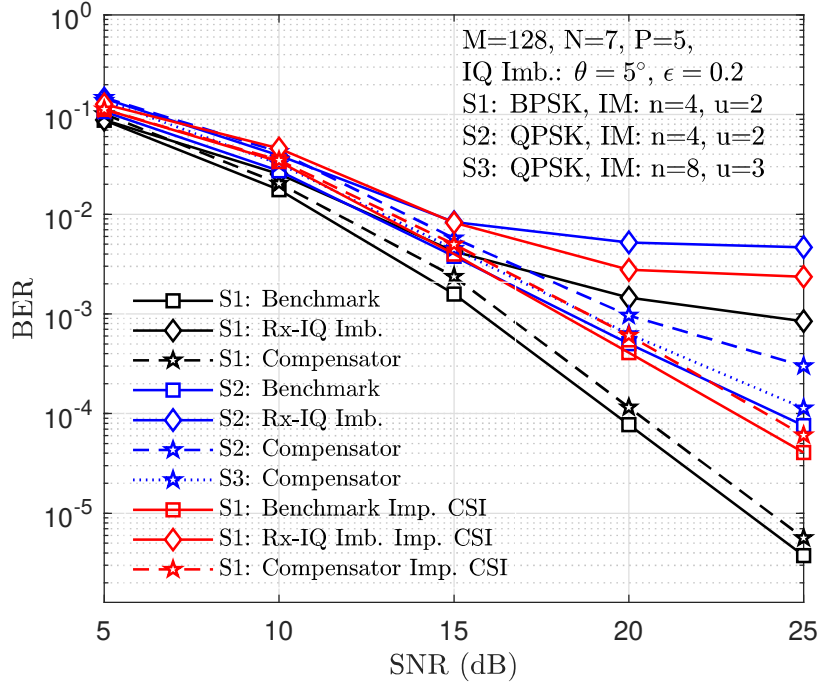


Figure 4.18: BER performance of OTFS-IM with MMSE under Rx IQ imbalance considering various practical scenarios.

In **Figure 4.18**, we discuss the performance of the proposed compensator under Rx IQ imbalance considering various activation ratios. It is worthy mentioning that Rx IQ imbalance severely affects the error performance due to additional CSI error, which is caused by mirror channel, given in (3.136). Therefore, Rx IQ imbalance saturates the error performance. The proposed compensator with BPSK modulation and $(n = 4, u = 2)$ is able to brake the saturation of error performance and offer a significant communication reliability without loss of spectral efficiency. In case of QPSK with the same activation ratio $(n = 4, u = 2)$, the performance of compensator gets worse due to the effect of imaginary part. Moreover, the data symbols become more sensitive to constellation rotation and MDI that decreases the reliability of correct estimation of active carriers. Hence, we increase the sparsity of received signal by utilizing $(n = 8, u = 3)$ to enhance estimation accuracy at the cost of reducing spectral efficiency. In addition, we consider channel estimation errors as $\hat{h}_i = h_i + \Delta h_i$, where Δh_i is modeled as an $CN(0, \iota 10^{\frac{-SNR_{dB}}{10}})$ and $\hat{h}_i = h_i$ if $\iota = 0$ in order to assess the performance of proposed scheme under imperfect CSI. Performance of the proposed scheme is immune against imperfect CSI, causing a shift for all the results regarding to the level of error (Δh_i). Briefly, the proposed IQ imbalance technique is able to compensate the effect of IQ mismatch and simultaneously provide service for future use-cases and applications. On the other hand, the complexity of classical IQ imbalance compensators (i.e., blind, pilot and DNN-based techniques) exponentially increases with the modulation order.

Beyond fifth-generation (5G) wireless communication systems are intended to offer solutions for the wireless communication society, taking into account a wide range of services and requirements in the near future, where high mobility wireless channels and small handsets will become common in our daily life. Consequently, the design of smart waveforms, enabling robustness against the wireless channel and hardware impairments will play a crucial role in the future of wireless communication society.

4.7. Rx Windowing OTFS: Numerical And Simulation Results

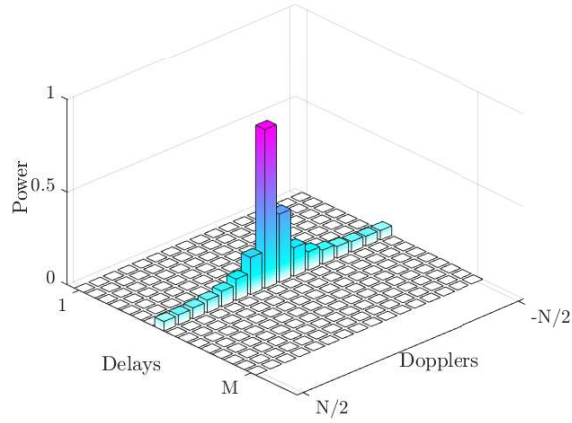
This section investigates and discusses the performance of the proposed RW-OTFS scheme for practical scenarios and use-cases. To illustrate the impact of the RC windowing in DD domain, we consider the RW-OTFS carrier under various CP length and

fractional Doppler configurations, i.e., G and κ . In addition, we utilize Monte-Carlo based simulations in order to assess the error performance of the proposed RW-OTFS transmission technique under wireless channels with fractional Doppler impairment. We consider M -ary symbols with constant envelope for all the simulation scenarios. Lastly, we evaluate the error performance of the RW-OTFS for the Extended Vehicular A (EVA) channel model, which is the most common channel model accepted for high-speed vehicles in the LTE standard [136].

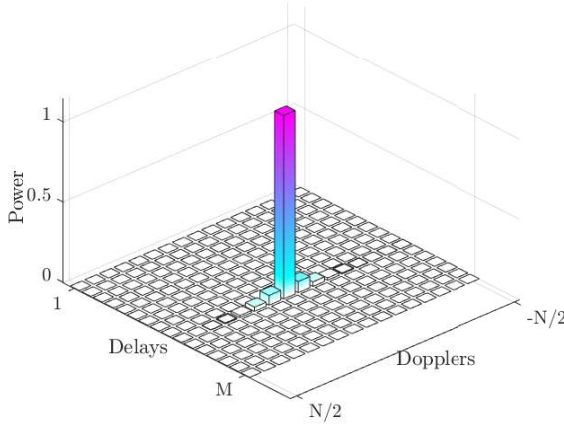
Figure 4.19 illustrates the importance of the proposed Rx windowing on mitigating IDI in OTFS transmission under fractional Doppler. To be consistent with **Figure 2.3**, we use the same transmission parameters $M = 16$ and $N = 16$ for the OTFS carrier in **Figure 4.19**. **Figure 4.19(a)** represents the scenario that corresponds to conventional OTFS transmission under fractional Doppler. In other words, no time guard is available for the windowing. As can be easily seen, the IDI is spread over the Doppler axis due to imperfect sampling stemming from fractional Doppler.

On the other hand, in the case of the proposed scheme considering $G = MN/4$ in **Figure 4.19(b)**, the IDI mostly reduces at the edges of the DD grid since the side-lobes of OTFS carrier fades faster with the aid of the proposed Rx windowing. Therefore, in **Figure 4.19(c)**, G is extended by half of the OTFS frame duration, i.e., $G = NM/2$. As the windowing guard increases, RC windowing scheme smooths the edges of OTFS frame, while performing a much slower transition. In this line, the proposed scheme achieves a stronger suppression of the side lobes of OTFS carrier compared with **Figure 4.19(b)**, and thus localizes the IDI only around the communication carrier. Moreover, it is important to mention that the localization of IDI can significantly increase the channel estimation performance under fractional Doppler. Consequently, the IDI is localized around the OTFS carrier of the interest, and thus, the received OTFS signal under fractional Doppler impairment appears even as $\kappa = 0$ in DD domain. Since the Doppler characteristics of OTFS signal exist globally, i.e., over the whole OTFS frame, the extension of the OTFS signal is used to manipulate the OTFS Dirichlet pulse shape.

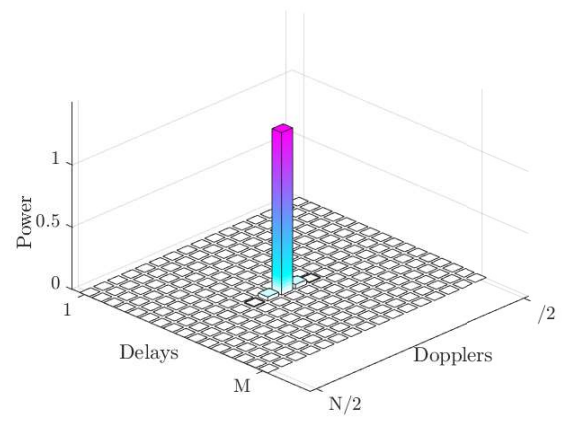
Figure 4.20 shows the BER performance of the proposed RW-OTFS scheme with ML detection in order to visualize the limiting impact that fractional Doppler causes on the diversity gain of OTFS transmission. Considering the fact that the Rx complexity of OTFS



(a): Conventional OTFS transmission



(b): The proposed RW-OTFS with $G = MN/4$



(c): The proposed RW-OTFS with $G = MN/2$

Figure 4.19: The OTFS carrier and the proposed RW-OTFS scheme under κ .

transmission with ML detector severely increases with the increase of M and N values, we choose small values of $M = 4$ and $N = 2$ in order to assess the theoretical analysis (3.184) for BPSK transmission. In particular, the dashed-lines denote the theoretical results, and thus, they validate the simulation results for the proposed RW-OTFS technique. The parameters of RW-OTFS used in **Figure 4.20** are small for practical wireless communications standards. Therefore, in **Figure 4.21**, we show the BER performance of RW-OTFS with $M = 12$ and $N = 7$, which represent the resource element in 4G wireless standard, while considering different guard lengths and MMSE detector at the Rx side.

In **Figure 4.21**, we investigate system performance under $\kappa_i \in [-0.5, 0.5]$ considering

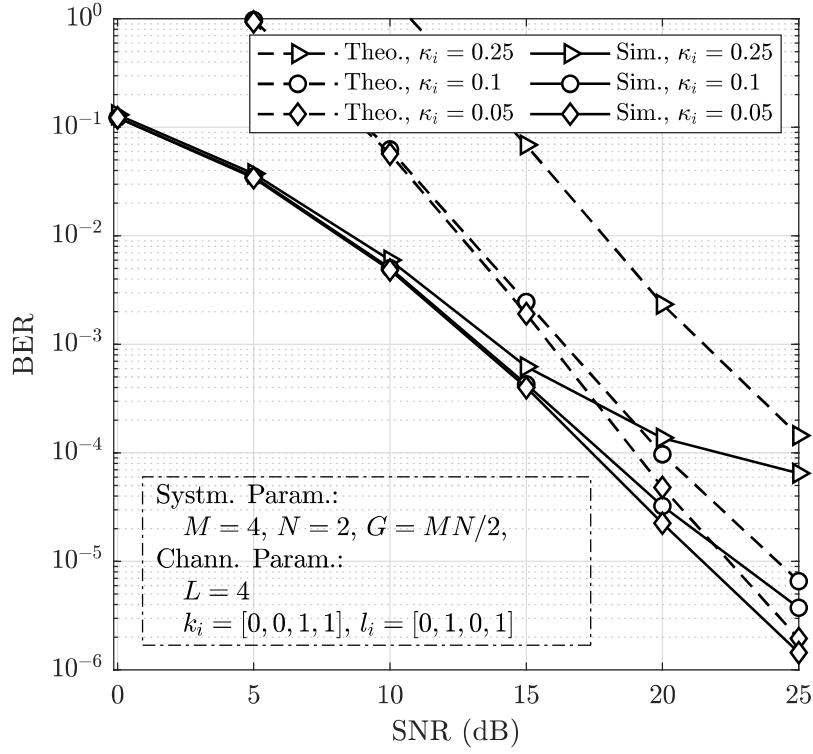


Figure 4.20: BER performance of the proposed RW-OTFS with ML detector under various fractional Doppler settings.

normalization factor, i.e., N , for all scenarios in order to explicitly show the importance of the proposed RW-OTFS under various G parameters. The solid-line with circle marker represents the benchmark for the performance of the conventional OTFS transmission with integer delay τ_i and integer Doppler ν_i wireless channel taps, while their corresponding values are uniformly distributed in the range of $[\tau_{min}, \tau_{max}]$ and $[\nu_{min}, \nu_{max}]$, respectively. On the other hand, the solid-line with square marker shows the error performance of the classical OTFS transmission under fractional Doppler. As seen in the figure, BER performance of the classical OTFS under the same fractional Doppler impairment does not exceed 10^{-3} . In the case of RW-OTFS with $G = MN/2$ where spectral efficiency equals $SE = \frac{MN}{G+MN} = 66.6\%$, the saturation of the BER is broken and at least a 10 dB gain is achieved at 25 dB SNR that further enhances with the increase of SNR as discussed in the above paragraph. Moreover, with the increase of the time duration of the RW-OTFS frame via $G = MN$, further suppression of the side-lobes occurs at the cost of 16.6% SE loss compared with the scenario $G = MN/2$. Thus, the error performance of the proposed RW-OTFS design approaches the scenario of classical OTFS transmission with integer Doppler values, i.e., free of IDI. Moreover, a difference by at least 3 dB exists between

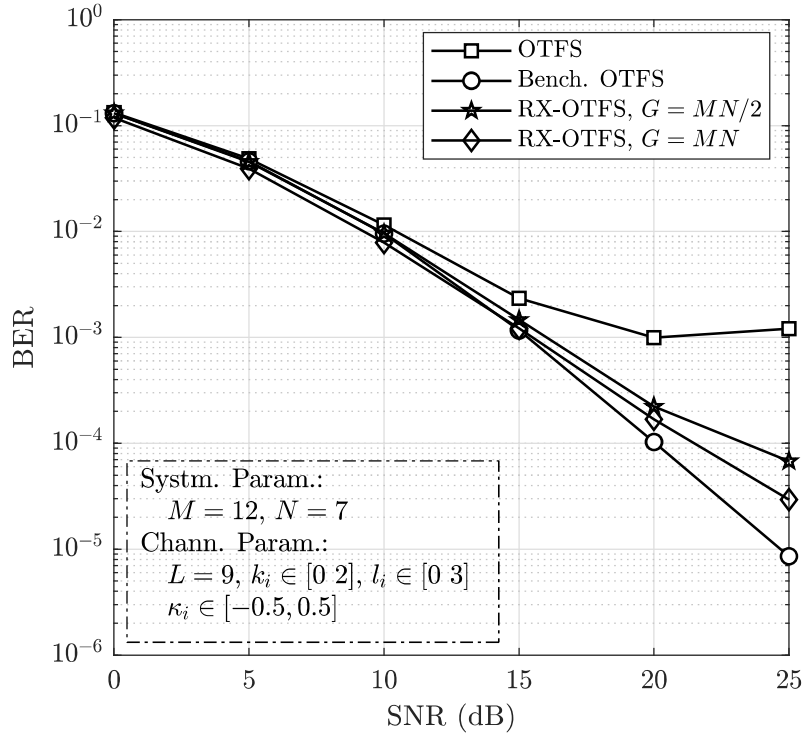


Figure 4.21: BER performance of the proposed RW-OTFS with MMSE detector under different guard lengths considering the maximum fractional Doppler.

$G = MN/2$ and $G = MN$ for the proposed RW-OTFS, since IDI decreases with the increase of RW-OTFS frame duration and vice-versa, as illustrated in **Figure 4.19**. In line with the characteristics of the Dirichlet shaped OTFS carrier, the impact of the IDI in RW-OTFS reduces as the CP length increases. In other words, localization of the RW-OTFS carrier gets better as the RW-OTFS frame time increases, leading to side-lobes with lower power compared to conventional OTFS scheme.

Since the feasibility of the ML decoder is limited due to its complexity, **Figure 4.22** shows the error performance of RW-OTFS with MMSE detector considering high M, N values and $\kappa_i \in [-0.5, 0.5]$. Importantly, the proposed RW-OTFS scheme is able to break the saturation even with $G = MN/4$ and error enhancement of the conventional MMSE detector. The error performance of RW-OTFS scheme enhances with the increase of the SNR since the IDI is controlled with the aid of Rx windowing. BER performance of the classical OTFS scheme heavily depends on the amount of IDI present. Moreover, the proposed RW-OTFS can further enhance the link reliability by means of a higher guard band, i.e., $G = MN/2$. In addition, **Figure 4.22** shows the simulation results for RW-OTFS with $M = 64, N = 7$, offering a slight performance gain.

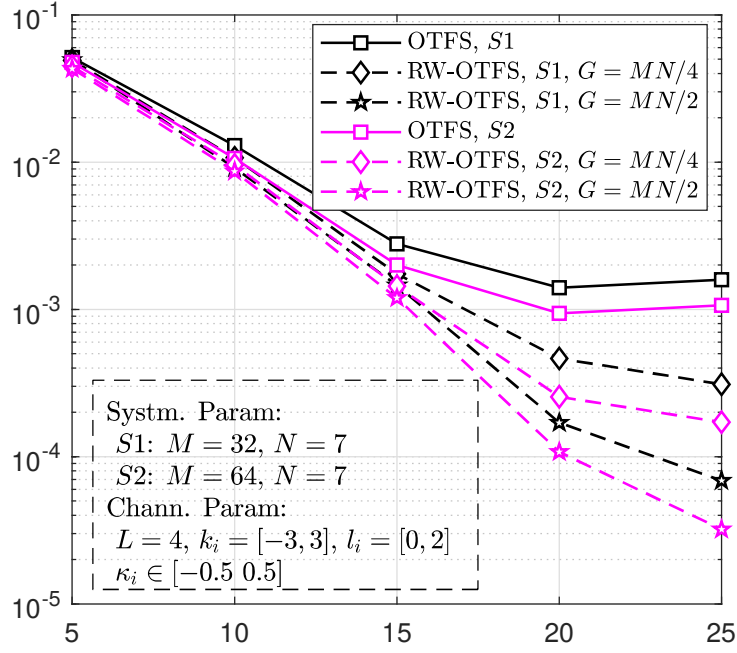


Figure 4.22: BER performance of the proposed RW-OTFS with MMSE detector under high M and N values.

Moreover, **Figure 4.23** shows the BER performance of the proposed RW-OTFS scheme with $G = MN/4$ considering various fractional Doppler ranges. To provide clear understanding of the robustness of the RW-OTFS under fractional Doppler, **Figure 4.23** considers the system parameters given in **Figure 4.22**. As we can see in **Figure 4.23**, although the proposed RW-OTFS scheme offers lower robustness for fractional Doppler ranges greater than 0.4, it still breaks the BER saturation caused by fractional Doppler even under CP size $G = MN/4$.

Figure 4.24 assesses the performance of the proposed RW-OTFS scheme under LTE EVA channel model, which defines the delay profile of the wireless environment with mobility, as given in **Table 4.9**. Hence, we generate Doppler parameters (ν_i) in accordance with the maximum Doppler shift (ν_{max}) while $\nu_i = \nu_{max} \cos(\Delta\phi_i)$. $\Delta\phi_i$ denotes the angle of arrival that is uniformly distributed within the range $\Delta\phi_i \in [-\pi, \pi]$ considering the Jake's model. In addition, the user speed (v) and operating frequency (f) determine the ν_{max} as $\nu_{max} = fv/c$, where c is the speed of light. In line with this, **Figure 4.24** presents the performance of the proposed scheme under different user mobility for LTE EVA channel model, as shown in **Table 4.10**. Moreover, we consider two different scenarios with fractional Doppler as S1: $\kappa_i \in [-0.4, 0.4]$ and S2: $\kappa_i \in [-0.2, 0.2]$. The solid-line

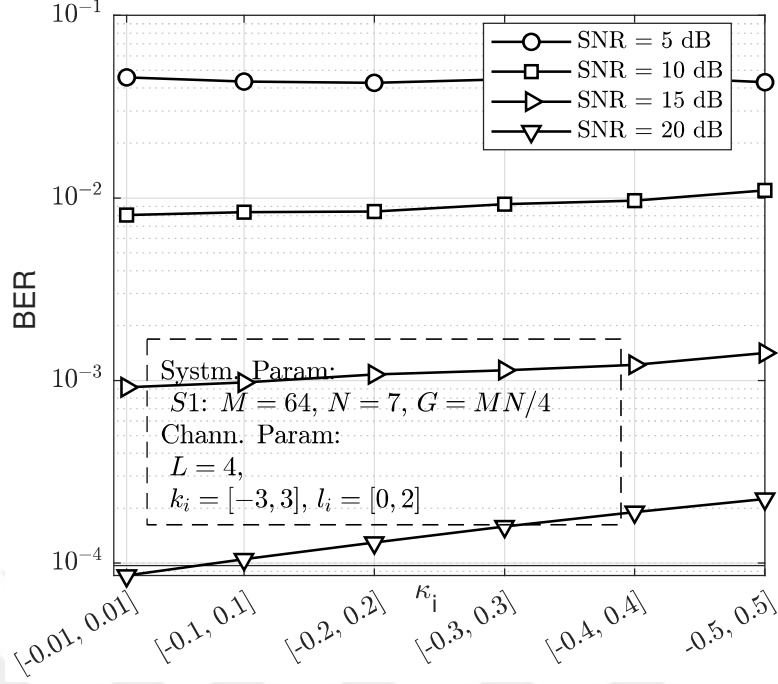


Figure 4.23: BER vs. fractional Doppler for the proposed RW-OTFS scheme considering different SNR values.

with circle marker and solid-line with square marker show the benchmark performance of conventional OTFS transmission for LTE EVA channel model with $v = 500$ kmph and $v = 120$ kmph, respectively. Since the diversity of the LTE EVA channel enhances with the increase of user speed, the error performance in case of $v = 500$ kmph is better than $v = 120$ kmph. Similarly, OTFS with $v = 500$ kmph offers better link reliability than OTFS with $v = 120$ kmph in both S1 and S2 under fractional Doppler impairment. The proposed RW-OTFS with $G = MN/2$ mitigates the effect of fractional Doppler for S2 with $v = 120$ kmph, as illustrated via dashed-line with the diamond marker. Differently, in case of S2 with $v = 500$ kmph, $G = MN/2$ is not enough to full-recover the link diversity and mitigate IDI, and thus, longer G is needed. In other words, the localization of IDI becomes difficult with the increase of channel diversity in DD domain.

To provide further useful insights about the performance of RW-OTFS scheme for practical implementations, Figure 4.25 illustrates the BER performance of the conventional OTFS transmission and the proposed RW-OTFS with Turbo coding under fractional Doppler. In order to verify the accuracy of the proposed RW-OTFS under practical scenarios, we consider both Turbo coding with hard decision and Turbo coding with soft decision with a coding rate of 1/2 under LTE EVA channel model. The error performance

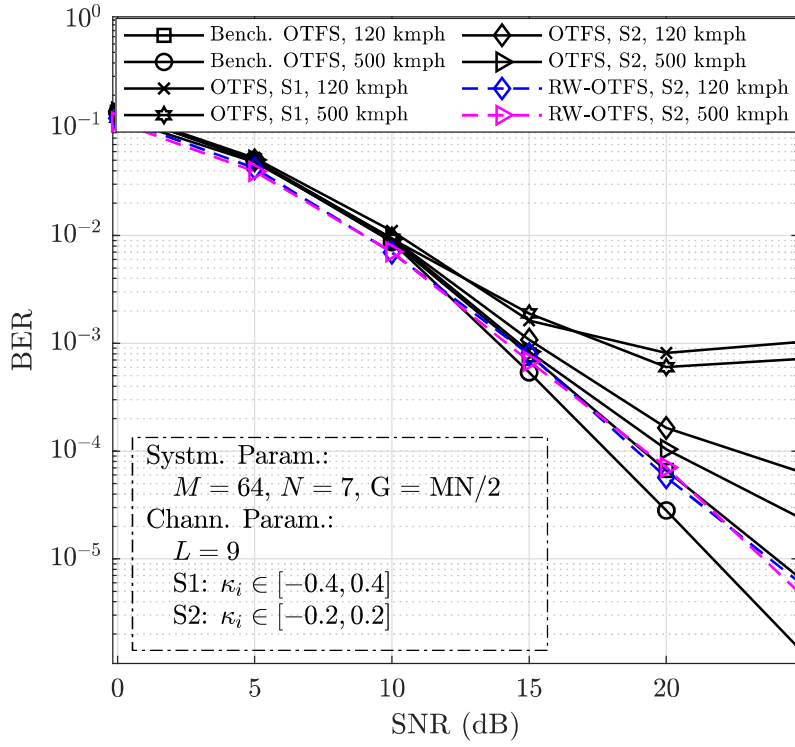


Figure 4.24: BER performance of the proposed RW-OTFS with MMSE detector under LTE EVA wireless channel model.

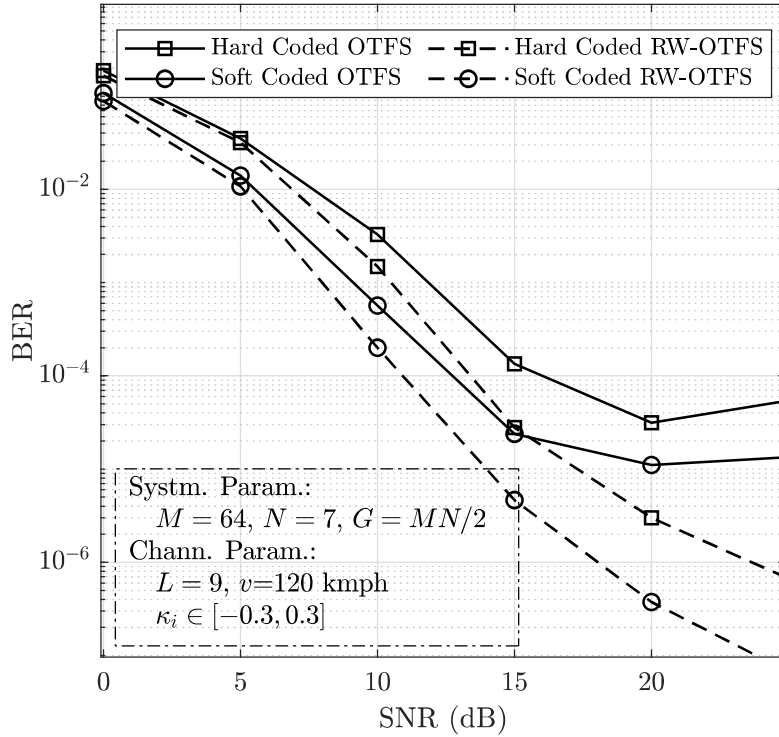


Figure 4.25: BER performance of the proposed RW-OTFS considering Turbo coding under LTE EVA wireless channel model.

Table 4.9: Delay and Doppler parameters in realistic OTFS transmission

LTE EVA Channel τ_i μ s with $i = 0, \dots, L-1$								
0	0.03	0.15	0.31	0.37	0.71	1.09	1.73	2.51
$v_i = v_{max} \cos(\Delta\phi_i)$ Hz with $i = 0, \dots, L-1$								
$\Delta\phi_i \in [-\pi, \pi]$			$v_{max_{120}} = 0.44$ kHz			$v_{max_{500}} = 1.85$ kHz		

Table 4.10: The system parameters for realistic OTFS transmission

System Parameters	
Carrier Freq.	4 GHz
SCS (Δf)	15 kHz
User Speed	120 kmph, 500 kmph
(M,N)	(64,7)
L	9
Detector	MMSE

of conventional OTFS saturates at 15 dB SNR, while the proposed scheme clearly avoids the error floor, and thus, it reaches to 10^{-6} BER. It is worth noting that satisfying the demands of services such as latency critical wireless communications requires to obtain a low BER within a short time period. As in the previous figures, the proposed RW-OTFS alleviates the effect of IDI and outperforms the conventional OTFS transmission under fractional Doppler, which is inevitable in beyond 5G communication systems.

In addition to the aforementioned discussions, **Figure 4.26** assesses the error performance of RW-OTFS with channel estimation. It is worth mentioning that in this work embedded pilot-based approach given in [131] is utilized to perform channel estimation. In both scenarios, the error performance of the conventional OTFS frame, which is represented via solid lines with circle markers, it is above 10^{-4} BER, since the level of IDI around the pilot location is significant, as illustrated in **Figure 3.10**. Different from classical OTFS, RW-OTFS with short CP configurations, i.e., $G = MN/8$ and $G = MN/4$, breaks the BER saturation. Moreover, $G = MN/4$ offer a reliable link communication, and the superiority of RW-OTFS becomes more visible with the increase of radio resources from $(M = 16, N = 16)$ to $(M = 32, N = 16)$. In addition to conventional OTFS, i.e., common CP-OTFS, local CP-OTFS has been discussed in the literature [137], where its performance under IDI is shown via solid line with star markers. Although the performance of CP-OTFS with $CP = M/2$ (i.e., total CP length for CP-OTFS frame is $M/2 \times N$) is close to RW-OTFS for small CP configuration, $G = MN/8$, its performance saturates under the fractional

Doppler regardless of the CP size. Considering $G = MN/4$ and $MN/2$, the proposed RW-OTFS provides considerable gain compared with the CP-OTFS under the fractional Doppler.

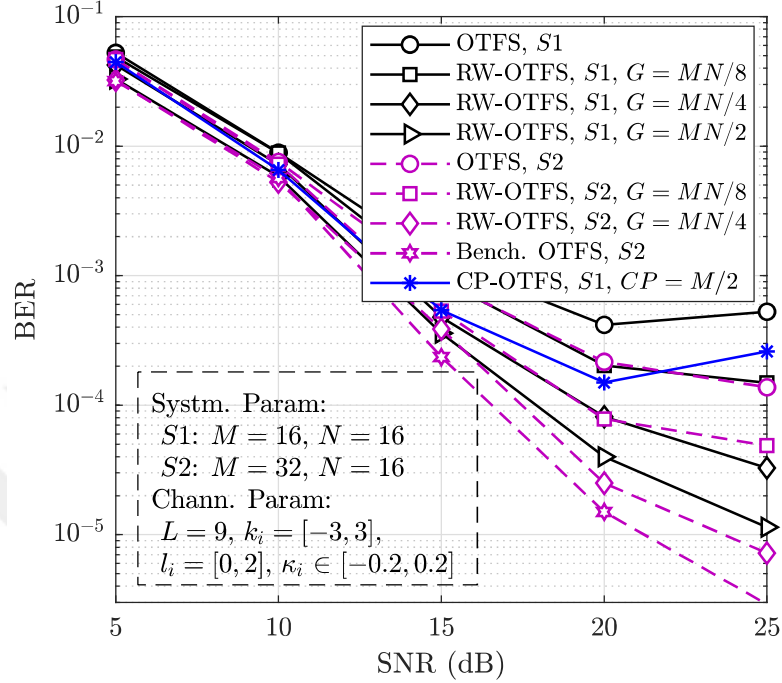


Figure 4.26: BER performance of RW-OTFS with 20dB SNR pilot.

In **Figure 4.27**, to show the obtained gain by the proposed RW-OTFS, normalized mean square error (NMSE) is calculated as $NMSE = \frac{\|H_w - H_i\|_F}{\|H_i\|_F}$, where $H_w = H_{DD}$ in case of conventional OTFS. It is observed that the proposed scheme outperforms the conventional scheme even under $G = MN/8$ configuration considering channel estimation at the Rx. RW-OTFS scheme increases the channel sparsity via Rx windowing and thus attempts to avoid the estimation of fractional Doppler shift since it introduces considerable system complexity [129], [130]. Specifically, the aim of RW-OTFS is to reduce the amount of Doppler-spread without introducing additional system complexity and relaxing the need for additional fractional Doppler channel estimation techniques. Hence, channel-spread mitigation and channel-spread estimation processes are two consecutive processes that can be exploited together to optimize spectral efficiency, BER performance, and system complexity of the OTFS based transmission. Therefore, the analysis, discussion, and evaluation of all these scenarios under a single framework are critical to investigate and highlighted as potential future work since it is difficult to include within one study.

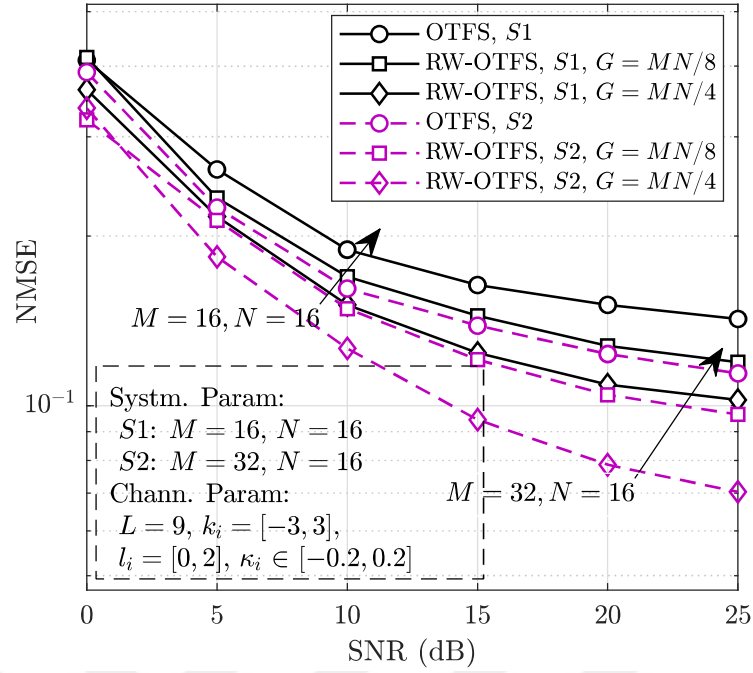


Figure 4.27: NMSE of RW-OTFS with 20dB SNR pilot.

In all figures, it is observed a significant degradation of the BER performance when fractional Doppler occurs in OTFS based transmission. The proposed RW-OTFS technique considerably mitigates the effect of IDI in DD domain for wireless communication systems. Moreover, as detailed in Section III, the proposed technique can be implemented via matrix multiplication in time domain. Thus, the proposed scheme controls the effect of fractional Doppler without inducing a heavy computational complexity and without requiring prior information.

CHAPTER 5

5. CONCLUSION AND FUTURE WORK

This thesis delves into three key research directions for wireless communications systems: 1) frame design and pulse shaping for interference management in time, frequency and delay-Doppler domains, 2) efficient radio resource assignment and utilization for single user and multi-user transmissions in delay-Doppler domain, and 3) analysis and mitigation of interference caused by RF impairments.

5.1. Frame Design In SC Transmission

This technique investigates the suitability of SC systems for URLLC by optimizing system parameters, which include data size, roll-off factor, and pulse length. Besides latency and reliability, optimization of these parameters is performed by considering time-domain efficiency, frequency-domain efficiency, OOB, and PAPR. Additionally, the proposed adaptive α utilization enables further system efficiency improvement. It is shown that the flexible structure of SC transmission provides opportunities to serve various use cases with URLLC. All the relationships investigated in this study are obtained through computer-based simulation results. Mathematical analysis of the relationships and optimization of the parameters under different channel conditions will be considered in our future study.

5.2. Frequency Domain IM

The concept of IM has been extensively studied in recent years. Several multicarrier transmission schemes with IM have been provided to serve the applications and use cases including eMBB, mMTC, and URLLC. Main issues related with multicarrier systems such

as CFO and IQ imbalance have not been studied considering IM. In this work, the performance of various IM techniques is evaluated under RF-impairments. Although OFDM-GIM offers the highest spectral efficiency, it is the most sensitive against non-ideal RF Rx effects. OFDM-SNM offers worse BER performance than OFDM-IM under CFO and IQ imbalance due to successive subcarrier activation. Hence, the impact of RF impairments should be considered in order to select an appropriate IM scheme for the desired application. As a future work, other existing IM schemes need to be evaluated in the presence of RF impairments. Moreover, the development of new methods for estimation and compensation of CFO and IQ imbalance is required.

5.3. Proposed Hybrid NOMA Scheme For Multiuser Detection

This thesis proposes a hybrid power domain NOMA scheme utilizing OFDM and OFDM-IM for downlink transmission with two UEs. It is shown that IM-NOMA is superior to OFDM-NOMA in terms of BER performance operating on frequency-selective channels and can further enhance the total system achievable rate. Moreover, the proposed Rx design considering the physical feature of IM-NOMA can make MUD feasible even for UEs with similar power factors. Future research will consider the presence of more than two UEs, use of various M-ary constellations and different subcarrier activation ratios will be considered.

5.4. Exploiting Multiuser Diversity In Delay-Doppler Domain

In this thesis, we have investigated the performance of OTFS transmission from the perspective of exploiting the multi-user diversity effect in downlink transmission. Specifically, we have proposed a novel scheduling algorithm with the assignment of the radio resources to the UE associated with the largest number of channel taps in order to improve the exploitation of the multi-user channel diversity. Moreover, we have derived the closed-form BER upper-bound expression for the proposed scheduling scheme. Accordingly, both the simulation and the theoretical results show that the proposed technique offers a significant improvement in error performance compared with random RR assignment scheme. It is worth mentioning that the generalization of this work by taking into account fading characteristics, various UE velocities, and inter-user interference in uplink

are left as a future work.

5.5. Impact Of IQ Imbalance In Delay-Doppler Domain

The physical effect of the IQ imbalance, which is a performance-limiting problem in the design of DCT, is investigated in the delay-Doppler domain for OTFS transmission. Specifically, the effect of IQ imbalanced Tx is mathematically modeled, where we show that IQ imbalance not only degrades the power of the desired signal but also leads to MDI occurring only along the Doppler axis. Moreover, the closed-form BER expression is derived considering the IQ imbalanced Tx model with and without the presence of channel estimation errors. Accordingly, both the simulation and the theoretical results show that an IQ imbalanced Tx causes a significant loss of error performance, especially in the case of the OTFS transmission with high diversity gain. It is worth noting that the generalization of this study taking into account the MIMO configuration, and the estimation and compensation of the IQ imbalance effect, are imperative to investigate, which remains as our future studies. Moreover, further studies that elaborate the effect of IQ imbalance on OTFS transmission considering both varying envelope M-ary symbols and different fading scenarios are an open research.

5.6. IQ Imbalance Compensator For OTFS Scheme

IM-aided OTFS-based wireless communications systems has been proposed as a practical scheme to mitigate the IQ imbalance effect in delay-Doppler domains. In particular, the sparsity of OTFS-IM signal in the delay-Doppler domain has been explored in order to localize and estimate the parameters of IQ imbalance. Hence, the primary information of OTFS-IM signal sparsity removes the need to use blind or pilot-based IQ imbalance compensation schemes, which suffer from high system complexity and high spectral efficiency loss. Differently, the proposed IQ imbalance compensator not only mitigates the IQ imbalance effect but also allows to communicate with low computational complexity and without compromising spectral efficiency for low-order modulation (i.e., BPSK and QPSK). The performance of the proposed scheme under different wireless channel characteristics (i.e., fading, delay-Doppler spreading, and user speed) and high-order modulation associated with the theoretical results have been left for further future studies.

5.7. Rx Windowing OTFS For Beyond 5G Wireless Systems

In this thesis, a global RC windowing at the Rx has been introduced in order to control the side-lobe energy of conventional OTFS carriers towards the Doppler axis in the delay-Doppler domain. In particular, a practical Rx windowing scheme named RW-OTFS has been proposed in order to suppress the effect of IDI in OTFS-based wireless communication technologies due to fractional Doppler impairment. The proposed RW-OTFS scheme prevents the spreading of IDI over the entire Doppler axis compared with conventional OTFS scheme under fractional Doppler. As such, RW-OTFS localizes IDI and contributes to OTFS-based technologies in ensuring the sparsity of received signal in delay-Doppler domain. In other words, RW-OTFS makes it feasible to perform Rx detection on integer delay-Doppler grids, thus ignoring the fractional Doppler grids in the DD domain. Moreover, different from the pilot-based channel estimation schemes and iterative interference suppression schemes, the proposed RW technique not only suppresses the effect of IDI but also does not introduce additional computational complexity, with a moderate compromise from the spectral efficiency. It is worth mentioning that the generalization of this work by considering proper pilot design, MIMO configuration, different fading scenarios, other pulse shaping Tx/Rx techniques, and joint utilization with various channel estimation schemes are imperative to investigate, which are left as open future research.

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