

A Novel PLC Front-haul for 5G IoT Indoor Communication using Split C-RAN Architecture

by

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Author's Declaration

This thesis consists of material all of which I authored or co-authored: see Statement of Contributions included in the thesis. This is a true copy of the thesis, including any required final revisions, as accepted by my examiners.

I understand that my thesis may be made electronically available to the public.

Statement of Contributions'

Mai M. H. S. Ibrahim was the sole author for Chapters 1, 2, 5 and 6 which were written under the supervision of Dr. Pin-Han Ho and were not written for publication.

This thesis consists in part of three manuscripts written for publication. Exceptions to sole authorship of material are as follows:

Research presented in Chapter 3:

This research was conducted at the University of Waterloo by Mai Ibrahim under the supervision of Dr. Dr. Pin-Han Ho. Dr. Pin-Han Ho and Mai Ibrahim conceived of the presented idea. Mai Ibrahim developed the analytical framework. Dr. Pin-Han Ho supervised the project. All authors discussed the results and contributed to the final manuscript. Mai Ibrahim was the primary coder, contributing to coding development and analysis of all results. She performed the required calculations and developed simulations, analysed the results and wrote the manuscript. Dr. Pin-Han Ho contributed to the final version of the manuscript and supervised the project. Mai Ibrahim took the lead in writing the manuscript. Mai Ibrahim wrote the draft manuscript, which all co-authors contributed intellectual input on. All authors provided critical feedback and helped shape the research, analysis and manuscript.

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Research presented in Chapter 4:

Dr. Pin-Han Ho, Dr. Hesham G. Moussa and Mai Ibrahim originated the presented notion. Mai Ibrahim developed the theory and performed the computations. Dr. Pin-Han Ho and Dr. Hesham G. Moussa verified the analytical methods. Dr. Hesham G. Moussa encouraged Mai

Ibrahim to investigate the delay analysis over the radio interface and managed the findings of this work. All authors discussed the results and contributed to the final manuscript.

Mai Ibrahim developed the theoretical formalism, performed the analytic calculations and performed the numerical simulations. She designed and developed the codes, derived the models and analysed the results. Dr. Hesham G. Moussa assisted with calculations and results. Dr. Pin-Han Ho was in charge of overall direction and planning.

Mai Ibrahim worked out almost all of the technical details and performed the numerical calculations for the suggested performance analysis. Dr. Hesham G. Moussa and Mai Ibrahim contributed to the design and implementation of the research, to the analysis of the results and to the writing of the manuscript.

Mai Ibrahim took the lead in writing the manuscripts. Mai Ibrahim wrote the manuscripts with support from Dr. Pin-Han Ho and Dr. Hesham G. Moussa. Dr. Pin-Han Ho, and Dr. Limei Peng contributed to the final version of the manuscripts. Dr. Pin-Han Ho supervised the project. All authors provided critical feedback and helped shape the research, analysis and manuscript.

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Abstract

The demand for efficient telecommunications in the era of Fifth Generation (5G) and Internet of Things (IoT) necessitates innovative approaches to network architecture and communication technologies. Recently, split Centralized Radio Access Network (C-RAN) architecture, characterized by Central Unit (CU), geographically dispersed Distributed Unit (DU), and indoor Radio Units (RUs), has presented opportunities for optimizing communication links in indoor environments. Yet, the adaptation of this innovative architecture to enable massive indoor IoT applications is still deemed inefficient due to the associated cost of deployment. Accordingly, this research investigates Power-Line Communication (PLC) as a cost-efficient alternative solution for C-RAN front-haul. Specifically, the focus is on exploring the utilization of indoor low-voltage power lines in the context of 5G New Radio (NR) indoor IoT applications.

First, to ensure that standard protocols like Common Public Radio Interface (CPRI) and Enhanced Common Public Radio Interface (eCPRI) can run on PLC, we introduce two novel patented components to the architecture, namely the CPRI-PLC-Gateway (CPG) and Enhanced CPRI-PLC-Gateway (eCPG). These are a plug and play components that come in pairs. They are used to create a virtual PLC front-haul link ensuring transparent transportation of unmodified CPRI or eCPRI frames between DU and RUs, even under challenging PLC channel conditions. As such, they set the foundation for optimizing the PLC front-haul and help resolve various challenges, including PLC time-varying nature and susceptibility to additive white Gaussian noise (AWGN).

Furthermore, investigations are extended to study the impact of the proposed PLC based split C-RAN system in the context of the Radio access network (RAN). For that, an indoor multi-story service building that houses a large number of air-interfaced IoT devices is considered. To ensure that the reported results apply to real-life applications, we consider a PLC network that encompasses typical indoor low-voltage 3-phase power lines and follows TN-S earthing configuration. Accordingly, it is shown that through the incorporation of the CPG and eCPG components, the implementation of In-band full-duplex (IBFD) communication over the multiple Input - multiple output (MIMO) PLC channel, and the integration

of the hybrid circuit-based isolation, the system can support a considerable number of air-interfaced IoT devices at standardized rates. It is also shown that the self-interference over the power line segment is mitigated which ensures robust bidirectional communication in the system.

Moreover, a significant aspect of the thesis revolves around conducting a comprehensive performance analysis for the proposed PLC front-haul for IoT indoor communications. Mathematical models, rooted in queuing theory, Markov modelling, and stochastic geometry, are developed to assess the end-to-end delay performance of the indoor front-haul solution. Analytical expressions are derived for various performance metrics, including radio coverage probability, the number of served devices, and system delay. Wireless spatial models, path-loss models, and interference considerations are meticulously analysed in terms of multiple factors such as the number of wireless IoT devices, radio and PLC bandwidth, and transmission technology, in regard to the delay performance of the proposed system. These models are rigorously validated through extensive simulations, demonstrating compliance with stringent 5G, CPRI, and eCPRI bit error rate (BER) and delay requirements.

Last, as the thesis further aims to examine the optimization challenge of maximizing throughput in a split-RAN system that includes a PLC front-haul link within a multi-story building. The goal is to optimize the number of fulfilled IoT devices while simultaneously satisfying their quality of service (QoS) criteria. The optimization problem is defined as a mixed-integer non-convex problem, which includes several objectives: maximizing the number of satisfied devices, minimizing operating cost, minimizing device transmit power, and minimizing PLC access delay. The thesis further explores the application of an Evolutionary Multi Objective Optimization (EMO) algorithm, specifically Non-dominated Sorting Genetic Algorithm II (NSGA-II), to address the issue of conflicting objectives in communication systems. The method operates by systematically generating successive iterations of solutions using tournament selection, single-point binary and simulated binary crossovers, and polynomial mutation operators. The system outcomes present a Pareto front consisting of non-dominated solutions for the issue defined using multi-objective optimization (MOO) showing a trade-off between the system objectives.

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Dedication

Praise be to Allah

This thesis is dedicated to my family who always loved me, believed in me and supported me, my Mother, Mervat, my sister, Mayada, my beautiful son, Omar and my deceased grandparents Mohamed and Habiba

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Table of Contents

List of Figures	xiv
List of Tables	xvii
List of Acronyms	xviii
List of Symbols	xxiv
1 Introduction	1
1.1 Massive M2M Communication	2
1.2 5G Split Centralized Radio Access Network	4
1.2.1 C-RAN Front-haul	6
1.2.2 Current Front-haul Challenges	6
1.2.3 Front-haul Deployment Cost	9
1.3 Research Motivation and Objective	10
1.3.1 Objective and Contributions	11
1.3.2 Publications Produced	13
1.4 Thesis Outline	14
2 Background and Literature Review	15
2.1 Power-Line Communication Channel Model	15
2.1.1 PLC Channel Bandwidth	16

2.1.2	MIMO PLC Channel Model	17
2.2	PLC Noises Models	20
2.2.1	Background Noise Model	20
2.2.2	Impulsive Noise Characteristics	21
2.3	Impulsive Noise: Challenges and Mitigation	22
2.3.1	Impulsive Noise Reduction Techniques	22
2.3.2	Reviewing Performance Modeling in PLC Networks	23
2.4	Overview of CPRI and eCPRI	24
2.4.1	Common Public Radio Interface (CPRI)	24
2.4.2	Enhanced Common Public Radio Interface (eCPRI)	25
2.5	Narrow band Internet of Things (NB-IoT)	27
2.6	Chapter Summary	29
3	PLC based Front-haul System	30
3.1	PLC Front-haul System Design	30
3.2	Network Architecture	33
3.2.1	eCPRI System Functional Split	35
3.2.2	In-Band Full-Duplex (IBFD)	36
3.3	PLC Front-haul System Model	38
3.4	Proposed Front-haul using CPG	39
3.4.1	Singular Value Decomposition (SVD) Processing	40
3.4.2	Impulsive Noise Detection & Re-transmission (IND)	40
3.4.3	Virtual eCPRI Link by eCPG Pair	42
3.4.4	CPRI Data Rate Matching	44
3.5	Simulation Outcomes	47
3.6	Chapter Summary	53

4	Performance Analysis of PLC Front-haul for IoT Indoor Communications	54
4.1	Radio Access Network (RAN) Architecture	55
4.1.1	Wireless Spatial Model	56
4.1.2	Wireless Transmission Model	57
4.2	Performance modelling	59
4.2.1	Coverage Probability over Wireless Link	59
4.2.2	Delay over the Air Interface	63
4.2.3	Radio to Access Network Data Rate Conversion	66
4.3	Access Delay Calculation	67
4.3.1	Gilbert-Elliot PLC Channel Model	69
4.3.2	Incoming Queuing Delay	70
4.3.3	Transport Delay	77
4.3.4	Selective Repeat Re-sequencing Delay	78
4.3.5	Effect of Impulsive Noise on Access Delay	81
4.4	Case Studies and Numerical Outcomes	83
4.4.1	Generating PLC GE Model	83
4.4.2	Numerical Results	84
4.5	Chapter Summary	92
5	Multi-objective Optimization of IoT RAN using NSGA-II Algorithm	93
5.1	Radio Access Physical Network	93
5.1.1	Wireless Transmission Model	95
5.1.2	Transmission over PLC-based Front-haul	96
5.2	Problem Formulation	99
5.2.1	Optimization Objectives	99

5.2.2	Optimization Outcomes	100
5.2.3	Optimization Constraints	100
5.2.4	PLC Access Delay Calculations	104
5.3	Use of Evolutionary Multi-objective Optimization (EMO)	107
5.3.1	NSGA-II for Multi-Objective Optimization in Communication Systems	108
5.3.2	Binary Tournament Selection	113
5.3.3	Single-Point Binary and Simulated Binary Crossover	114
5.3.4	Multi-tuple Gene	118
5.4	System Simulations	119
5.5	Chapter Summary	125
6	Conclusion and Future Work	126
6.1	Conclusion	126
6.2	Future Work and Research Directions	128
	References	130
A	CPG Virtual link flow chart in UL	154
B	IoT cellular indoor network	156
B.1	Laplace transform of Inter-cell interference	156
B.2	Average active IoT devices in one RU Voronoi	158
B.3	Average achievable bit rate by a device in coverage	160
C	PGF of CPG incoming queue length	161

List of Figures

1.1	Projected user data traffic per month from 2020 to 2030 [1, 2]	1
1.2	Growth of Connected Devices from 1950 to 2050 [11]	3
1.3	The Split C-RAN architecture	5
1.4	Systematic sketch of LTE-over-copper in the downstream direction [30] . . .	8
1.5	Total cost breakdown for sparse and dense areas for (a) 1:4 PON dense area, (b) 1:8 PON dense area, (c) 1:16 PON dense area, (d) 1:4 PON sparse area, (e) 1:8 PON sparse area, and (f) 1:16 PON sparse area [2]	9
1.6	Histogram of front-haul optical fibre deployment cost reduction [35]	10
2.1	Extended SISO PLC Channel Model to 300 MHz - CTF and Channel Phase - Same circuit	17
2.2	A 3x3 PLC MIMO channel model	19
2.3	Background noise PSD - Worst and Best cases [44]	21
3.1	TN-S earthing system: PE & N are not connected after the power source . .	31
3.2	Differential signal over Power line	32
3.3	Proposed split C-RAN system architecture for 5G indoor	33
3.4	The virtual link created using the CPG pair over the power line	34
3.5	The functional diagram of the proposed system architecture	35
3.6	System Functional Split	36
3.7	Echo suppression in PLC where transmission and reception are coupled on the same conductor pair showing the hybrid circuit used [61]	37

3.8	Single Phase MIMO PLC channel	39
3.9	Functional block diagram for Ethernet based front-haul CPG	43
3.10	Distribution fitting for MIMO PLC channel - Log-normal 1-300 MHz BW . .	45
3.11	Link Level simulator block diagram	48
3.12	PSD Mask for extended bandwidth up to 300MHz	49
3.13	Simulations outcomes of Bit error rate (BER) for PLC	51
3.14	Simulations outcomes of Frame error rate (FER) per hyper-frame for PLC .	52
3.15	Net throughput outcomes of MIMO PLC (6x6) vs. Eb/N0	52
4.1	Radio access network architecture	55
4.2	TR structure	58
4.3	HFs transmission flow over the radio access network	64
4.4	Gilbert Elliot Channel Markov Model	69
4.5	A state transition diagram based on the number of consecutive receive failures of HF	75
4.6	An illustration of SR window	79
4.7	Coverage Probability vs. RU & IoT device densities for different SINR thresh- olds (τ_a)	87
4.8	Delay over air interface (T_{air}) vs. RU & IoT device densities for different SINR thresholds (τ_a)	88
4.9	Total radio access delay (T_{e2e}) over air and PLC links interface vs. RU & IoT device densities for different SINR thresholds (τ_a)	89
4.10	Data rate from on RU into the PLC link through the CPG incoming buffer (R_p^j) vs. RU & IoT device densities for different SINR thresholds (τ_a)	89
4.11	Total access delay over PLC link (T_{access}) for different RS coding rate	90
4.12	Total access delay over PLC link (T_{access}) vs HARQ re-transmission Limit (L_r)	91

5.1	Illustration of Radio access network architecture	94
5.2	Binary Tournament Selection	113
5.3	Binary Crossover	115
5.4	Simulated Binary Crossover	116
5.5	Multi-tuple Interdependent Gene	119
5.6	3D Plot of the Pareto-front for Problem P1 solution for different generations	121
5.7	2D Plots of Pareto- front for Problem P1 solution	122
5.8	Comparison of system outcomes between NSGA-II across 150 generations vs. multiple single runs of Max-RSS	124
A.1	Flow chart for the PLC based communication system - RU act as transmitter	154
A.2	Flow chart for the PLC based communication system - DU act as receiver . .	155

List of Tables

2.1	Statistical models of channel parameters [58]	18
2.2	The values of $\Delta\varphi$ to generate φ_n to each defined path [58]	19
2.3	Background noise parameters	20
2.4	Impulsive Noise statistical noise model [44]	22
2.5	KPIs values for NB-IoT [77]	28
3.1	Log-normal Parameters for MIMO PLC Channel	46
3.2	System simulation parameters	50
3.3	Confidence Interval for BER and FER for all scenarios at SNR = 20dB and confidence level 95%	51
3.4	Effective Throughput outcomes for $E_b/N_0 = 35dB$ MIMO PLC (6×6), Same Circuit (Gbps)	53
4.1	System simulation fixed parameters	85
4.2	System simulation variable parameters	86
5.1	System simulation parameters	120
5.2	NSGA-II simulation parameters	121

List of Acronyms

3GPP	3 rd generation partnership group.
5G	Fifth Generation.
ACK	acknowledgment.
ARQ	automatic repeat request.
AWGN	additive white Gaussian noise.
BBU	Base-band unit.
BER	bit error rate.
BF	Beam forming.
BS	Base station.
BW	bandwidth.
C-RAN	Centralized Radio Access Network.
C&M	Control and Management.
CA	Carrier Aggregation.
CAPEX	capital cost.
CBR	constant bit rate.
CDF	cumulative density function.
CoMP	Coordinated Multi-Point.
CoS	class of service.
CPG	CPRI-PLC-Gateway.

CPRI	Common Public Radio Interface.
CPU	Central Processing Unit.
CRC	cyclic redundancy check.
CTF	Channel Transfer Function.
CU	Central Unit.
D-DIC	digital interference cancellation.
DF	data frames.
DU	Distributed Unit.
e2e	end-to-end.
eCPG	Enhanced CPRI-PLC-Gateway.
eCPRI	Enhanced Common Public Radio Interface.
EMC	Electro-Magnetic Compatibility.
EMO	Evolutionary Multi Objective Optimization.
FCS	frame check sequence.
FDMA	frequency division multiple access.
FEC	Forward Error Correction.
FER	frame error rate.
FEXT	Far end crosstalk.
FFT	fast Fourier transform.
FH	front-haul.
FIR	finite impulse response.
FPC	fractional power control.
FSMC	Finite State Markov Chain.

GA	genetic algorithm.
GB	giga-bit.
GE	Gilbert Elliot.
HARQ	Hybrid automatic repeat request.
HetNet	Heterogeneous networks.
HF	CPRI hyper-frames or eCPRI data frames.
HPF	high-pass filter.
HPPP	homogeneous Poisson point process.
i.i.d	Independent and identical distribution.
IBFD	In-band full-duplex.
ICT	Information and communication technology.
IEEE	Institute of Electrical and Electronics Engineers.
IETF	Internet Engineering Task Force.
IND	Impulsive noise detection & re-transmission.
IoT	Internet of Things.
KPI	key performance indicator.
LDPC	low-density parity-check.
LLS	link level simulator.
LOS	Line of sight.
LPF	low-pass filter.
LTE	Long Term Evolution.
LV	low voltage.

M2M	machine to machine.
max-RSS	maximum received signal strength.
Mbps	Mega bits per second.
MIMO	multiple Input - multiple output.
mMTC	massive machine type communication.
mmWave	Millimetre Wave.
MNO	Mobile Network Operators.
MOO	multi-objective optimization.
MTC	machine type communication.
N	Neutral.
NACK	non-acknowledgment.
NB-IoT	Narrow band Internet of Things.
NOMA	Non-Orthogonal Multiple Access.
NR	New Radio.
NSGA-II	Non-dominated Sorting Genetic Algorithm II.
OFDM	orthogonal frequency division multiplexing.
OFDMA	orthogonal frequency division multiple access.
OPEX	operating cost.
P	Phase/Live.
P-NRZ	polar non-return to zero.
P2P	point-to-point.
PAPR	peak to average power ratio.
PDF	probability density function.

PDF	probability density function.
PDU	power distribution unit.
PE	Protective Earthing.
PGF	Probability Generating Function.
PLC	Power-Line Communication.
PMF	probability mass function.
PON	passive optical networks.
PPP	Poisson point process.
PSC	PLC FDMA sub-channel.
PSD	power spectral density.
PTP	Precision Time Protocol.
QoS	quality of service.
RAN	Radio access network.
RF	Radio frequency.
RRU	remote radio unit.
RS	Reed-Solomon.
RSC	Radio sub-channel.
RSS	received signal strength.
RU	Radio Unit.
RU_s	Radio Units.
RV	random variable.
SBX	Simulated Binary Crossover.
SdB	staging delay buffer.

SDMA	Space division multiple access.
SI	self-interference.
SINR	signal to interference and noise ratio.
SISO	single-input-single-output.
SLS	system level simulator.
SNR	signal to noise ratio.
SOI	signal-of-interest.
SR	Selective repeat.
SUCH	shared uplink channel.
SVD	singular value decomposition.
SyncE	Synchronous Ethernet.
Tbps	Tera bits per second.
TCO	Total Cost of Ownership.
TDM	time division multiplexing.
TR	transmissions round.
Ts	time slot.
Tx	transmit.
WDM-PON	Wavelength Division Multiplexed Passive Optical Network.

List of Symbols

$C_0^{n,t}$	The device-RU coverage probability for the device located at $d_0^{n,t}$ and served by the RU ₀
α_a	The path-loss exponent over the air interface
χ_e	The eCPRI data frame (HF) header redundancy factor and higher IP/Ethernet network layers
$\dagger$	Indicates the Hermitian operator of a matrix
Δf	The sampling frequency used (frequency step)
Δt	The duration of one CPRI hyper-frame or eCPRI data frame and is equivalent to one OFDM symbol period
ϵ_0	The bit error rate (BER) of the Gilbert Elliot (GE) Good State denoted as state 0
ϵ_1	The BER of the GE Bad State denoted as state 1
ϵ_a	The power control factor over the air interface
ϵ_{rs}	The BER expression after applying RS FEC
η_H	The effect of SR based HARQ over the data rate with L_r re-transmissions
η_I	The effect of IND over the data rate
η_{rs}	The overhead introduced due to RS(n_{rs}, k_{rs})
Γ	The background-to-impulsive noise ratio
γ	The instantaneous received signal to noise ratio (SNR)

γ_p^j	The mean arrival rate of HFs at the incoming buffer of each R-CPG
γ_d	The Poisson process parameter at which each IoT device generates DFs
γ_p	The total arrival rate to the SISO PLC link
γ_r	The total incoming data rate into RU _{<i>j</i>}
λ_a	The active IoT devices density per unit area
$\lambda_a^{n,t}$	The density of the inter-cell interfering active devices on the n^{th} sub-channel during the t^{th} TR
λ_b	The RU density per unit area
λ_d	The IoT devices density per unit area
$\lambda_i(f_n)$	The eigenvalues of $\mathbf{H}_{\text{PLC}}\mathbf{H}_{\text{PLC}}^\dagger$ at a given frequency f_n
Λ_{ap}	The conversion factor between the data rate over the air interface and the required data rate for the HF to carry such data over the PLC FH
$\mathbb{E}[\cdot]$	Denotes the expectation operation
$\mathbb{P}_{drop}$	The probability that an HF is dropped over the PLC link after $L_r + 1$ re-transmissions
$\tilde{\mathbf{s}}$	the CPRI/ eCPRI hyper-frames pre-processed signal from RU to different DU
$\tilde{\mathbf{y}}$	The post-processed through multiplication by $\mathbf{U}^\dagger$ at the receiver side
$\mathbf{C}$	The transition probability matrix of the PLC channel GE model
$\mathbf{H}_{1\mathbf{P}_1}$	The MIMO PLC channel over $i - th$ single-phase line
$\mathbf{H}_{\text{PLC}}$	The block PLC MIMO channel over the three-phase line
$\mathbf{n}_{bg}$	The PLC background noises
$\mathbf{n}_{imp}$	The PLC impulsive noises
$\mathbf{U}$	The left singular vector of $\mathbf{H}_{\text{PLC}}$

$\mathbf{V}$	The right singular vector of $\mathbf{H}_{\text{PLC}}$
$\mathbf{y}$	CPRI/eCPRI signal received at DU over the PLC FH (II)
$\mathcal{A}^t$	The probability that at least one DF is generated during the t^{th} transmissions round (TR)
$\mathcal{D}_a$	The number of bits in the DF sent over the radio link
$\mathcal{D}_p$	The number of bits within one HF
$\mathcal{F}_a^n$	The unconditional achievable bit rate per device derived using the Shannon's channel capacity formula over the air interface
$\mathcal{N}_a$	The number of the orthogonal sub-channels available over the air interface channel
$\mathcal{N}_{\text{PLC}}$	The number of FDMA sub-channels of one PLC port.
$\mathcal{P}_s$	The conditional scheduling probability given the number of active devices in the t^{th} TR
$\mathcal{R}_a^n$	The mean bit rate achievable by the i^{th} device
$\mathcal{R}_j$	The aggregated data rate received at the RU $_j$ from all scheduled devices within its coverage
$\mathcal{R}_p$	The total rate going into the PLC link for all the FDMA sub-channels
$\mathcal{R}_p^j$	The data rate of RU $_j$ relayed over to the incoming buffer at its CPG after conversion of the radio DFs into HF
$\mathcal{X}$	The scheduling status indicating if the device is scheduled
$\mathfrak{L}_a^n$	The lower bound by a fixed allocation of the bit rate per device over the air interface
$\mathcal{C}_a$	The general radio coverage probability for any scheduled device-RU pair
$\mathcal{L}_{I_a}\{x\}$	The Laplace transform of x with respect to inter-cell interference I_a

$\mu_a^{n,t}$	The device's service rate for the n^{th} sub-channel and during the t^{th} TR
μ_r	The number of times HF is received incorrectly at the receiver
μ_p	The mean of the Log-normal distribution representing the MIMO PLC channel
ν_p	The PLC link peak utilization
Ω_a	The total bandwidth of the air interface channel
ω_a	The bandwidth of one orthogonal sub-channels of the air interface channel
Ω_p	The bandwidth of the PLC channel over one port
ω_p	The bandwidth of one FDMA sub-channel of one PLC port.
ω_r	The slot number of the SR window
$\bar{\gamma}$	The average received SNR over PLC link
$\bar{\mathcal{T}}_r$	The mean throughput over the PLC link
$\bar{b}_b$	The mean number of HFs transmitted while in Bad state, i.e. the mean burst length
$\bar{b}_g$	The mean number of HFs transmitted while in Good state, i.e. the mean state holding time
$\bar{e}$	The mean FER over the PLC channel
$\bar{P}_e$	The average frame error rate (FER) over the PLC fading channel
$\bar{P}_{e_{rs}}$	The improved average FER over the log-normal received SNR after using RS FEC
$\bar{q}$	The total mean number of HFs at the RU side
$\Phi_a^{n,t}$	The locations of the inter-cell active device (i.e. in other clusters) interfering on the n^{th} sub-channel during the t^{th} TR
Φ_a	The set of the locations of the active IoT devices in the network
Φ_b	The set of the locations of the BSs' RUs

Φ_d	The set of the locations of the IoT devices connected to the network
π_{c_0}	The steady state probabilities that the channel is in the Good state
π_{c_1}	The steady state probabilities that the channel is in the Bad state
ρ_p	The mean traffic load of HFs into the j^{th} R-CPG
Σ	The diagonal matrix whose diagonal entries areThe singular values of $\mathbf{H}_{\text{PLC}}$
σ_a^2	The received additive white Gaussian noise power of the air interface
σ_p	The standard deviation of the Log-normal distribution representing the MIMO PLC channel
τ_a	The threshold of the received SINR defining the RU coverage
$\Theta_0^{n,t}$	The received SINR level at RU ₀ of the DF transmitted by device ₀ on the n^{th} sub-channel and during the t^{th}
$a_p(k)$	The number of new HF arrivals in time slot k
A_{imp}	The amplitude of the impulsive noise based on a Middleton's class A distribution
b_j	The j^{th} RU in the network's 2-D location
c_v	A constant defined for the Voronoi tessellation in $\mathbb{R}^2$
C_{MIMO}	The instantaneous capacity of the PLC MIMO channel
CS_k	The PLC channel GE current channel state
D	The propagation distance of the DF over the air interface
d_i	The 2-D location of the i^{th} device in the network
D_p	The number of bits in one CPRI hyper-frame or eCPRI data frame
e_0	The FER of the GE Good State denoted as state 0
e_1	The FER of the GE Bad State denoted as state 1

E_b	The power per transmit bit
$f_R(r_t)$	The PDF for the distance between a scheduled device and its serving RU
g_{ij}	The instantaneous power gain of the Rayleigh fading radio channel between the i^{th} device and the j^{th} RU
gb	The ratio of guard band
I_a	The inter-cell interference on the n^{th} sub-channel
k_{rs}	The Reed-Solomon channel coding message length
L_p	The maximum path length of the PLC link
L_r	The limit for the number of re-transmission for one hyper-frame in SR based HARQ
M_j^t	The number of devices associated with the j^{th} RU) at the t^{th} TR
$M_{a_j}^t$	The number of active devices in the j^{th} cluster at the t^{th} TR
$M_{c_j}^t$	The number of covered devices within the j^{th} RU cluster during the t^{th} TR
M_{max}	The maximum number of devices that a RU can schedule in a single TR
M_{tot}	The number of identical, battery-powered IoT devices connected onto the network
N	Neutral
N_0	The mean PLC noise power
N_f	The number of samples in the frequency range over the PLC link
N_p	The total number of differential ports over the three-phase power line
N_{p_i}	The number of ports in the i^{th} single-phase line
$N_{Rx}(f_n)$	The noise at the receiver of the PLC front-haul
n_{rs}	The Reed-Solomon channel coding block length

P	Phase/ Live
$P_e(\gamma)$	The average FER over AWGN channel
p_i	the nominal transmission power of any connected device over the air interface
P_s	The peak power of the sensed signal over the PLC link
P_{imp}	The probability that an impulse is detected using IND
P_{mean}	The stored average power sensed over the PLC link
$P_{T_x}(f_n)$	The transmitted power over the PLC link
$P_{w_{rs}}$	The probability of interleaved RS FEC decoder error
PE	Protective earth
$Q(.)$	The Gaussian Q-function
q_i	The power received at the RU from the device after fractional power control (FPC)
Q_F	The quantizer resolution in the frequency domain
$q_{i,j}(k+1)$	The number of HFs in the incoming queue at time slot $(k+1)$ with the channel state is j and the source state i at this time slot
$q_{i,j}[h]$	The steady state probability of the incoming queue length at the RU side
$r_j(k)$	The departure of an HF from the system while the channel is in a state
r_t	The distance between the device of interest and the origin during the t^{th} TR
R_{eff}	The required matched effective data rate along the PLC line corresponding to the required CPRI option (R_{opt})
r_j	The probability of departure given the current state channel is $CS_k = j$
T_a	The duration of TR which is equivalent of one 5G sub-frame (1 ms)
T_q	The incoming queuing delay at the R-CPG until the new HF is transmitted

T_t	The transmission delay over the PLC link
t_w	The impulse width
T_{PLC}	The time required to send the HF over the PLC FH
T_{air}	The total delay over the radio network
T_I	The delay introduced due to Impulsive noise detection & re-transmission (IND)
t_i	The inter-arrival time between the impulses
T_{on}	The mean length of the on period
T_{pH}	The HARQ functionality processing delay
T_{pg}	The propagation delay over the PLC link
T_{proc}	The constant processing time at the RU to convert the data frames (DF) into CPRI hyper-frames or eCPRI data frames (HF)
T_{rsq}	The re-sequencing delay at the D-CPG including the time required for re-transmissions
t_{rs}	The number of errors that the RS channel code can surely correct
T_{ts}	The transport delay through the PLC channel
Th	The PAPR threshold that indicates the occurrence of an impulse event
v	The speed of light in the PLC line is calculated as $v = 2/3c$
v_j	The probability of no departure (HF re-transmission) when the channel is in the state $CS_k = j$
W	The Selective repeat ARQ (SR) maximum window size
x_t	The distance between the device located at $d_i^{n,t}$ and using the n^{th} sub-channel to the origin
y	Signal received at the receiver side of the PLC link

y_t The distance between an arbitrary device located at $d_i^{n,t}$ and its serving RU located at b_i

$P1N$ The differential signal between 1-st (P1) phase and the neutral (N) conductors

$P1PE$ The differential signal between 1-st phase and the protective earth (PE) conductors.

Chapter 1

Introduction

With the rapid advancement of wireless devices, there is a surging demand for higher data transmission rates. By the year 2030, the average user mobile data per month is expected to increase from 5 giga-bit (GB) in 2020 to more than 250 GB per month in 2030 as shown in Figure 1.1 [1, 2]. The CISCO visual index report for 2018 - 2023 indicates in broadband data growth exceeding the double, from 45.9 Mega bits per second (Mbps) to 110.4 Mbps [3]. Furthermore, the continuous growth of mobile internet applications is expected to cause an increase in mobile data up to 1 Tera bits per second (Tbps) per device [2]. As a consequence, the new generation of communication systems must be equipped to handle this exponential surge in capacity [4].

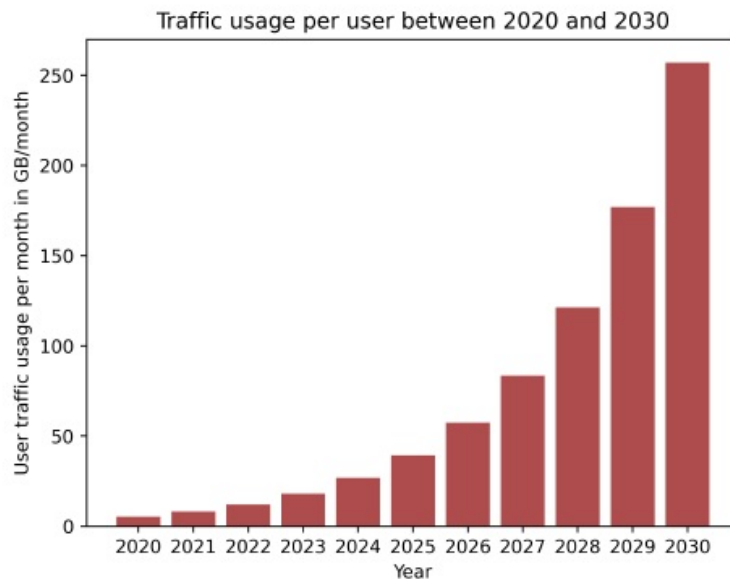


Figure 1.1: Projected user data traffic per month from 2020 to 2030 [1, 2]

In the context of Fifth Generation (5G) standards set by 3rd generation partnership group (3GPP) and Institute of Electrical and Electronics Engineers (IEEE), numerous cutting-edge technologies have been explored to handle the exponential increase in data rates. Solutions like Heterogeneous networks (HetNet), Carrier Aggregation (CA), Massive multiple Input - multiple output (MIMO), Beam forming (BF), Millimetre Wave (mmWave) and Coordinated Multi-Point (CoMP) are among the top contenders for 5G networks [5–7]. However, the implementation of these new technologies and the operational costs associated with multi-layered networks pose financial challenges for operators [8]. Additionally, the lack of cooperation mechanisms between independently operating Base stations (BSs) leads to severe inter-cell interference [9]. Moreover, the distributed layout of networks results in low hardware utilization since each BS is designed to handle peak rates, even if they are rarely achieved [5]. To address the high demands for data rates, the deployment of low-power small BSs is considered a highly effective solution, especially for indoor service provisioning where antennas need to be situated close to users to ensure Line of sight (LOS) transmissions.

1.1 Massive M2M Communication

The advent of 5G New Radio (NR) is set to usher in a new era of indoor services that will revolutionize daily life. These services will rely on vast networks of Internet of Things (IoT) devices, promising to impact various aspects of life significantly [5]. IoT is expected to facilitate intelligent applications such as environmental monitoring, smart cities, intelligent transport systems, e-health, industrial automation, and robotics, becoming an economic force that will transform the world in profound ways. By 2050, the number of connected devices is projected to exceed 170 billion, predominantly consisting of simple battery-powered appliances as shown in Figure 1.2 [10, 11]. This expansion has led to the rise of machine to machine (M2M) communication, also known as machine type communication (MTC), as a crucial enabling technology for many IoT applications.

M2M communication is an autonomous technology that allows machines to collaborate and communicate without human intervention, working together to achieve specific objectives [12]. The focus has shifted to Massive M2M applications, which present unique network-

ing challenges. These applications involve a large number of devices transmitting periodic or event-driven small data packets. Many of these devices are low-mobility, low-complexity, and battery-powered, equipped with limited radio capabilities. As a result, Massive M2M communication has become a prominent area of research and standardization in recent years [13].

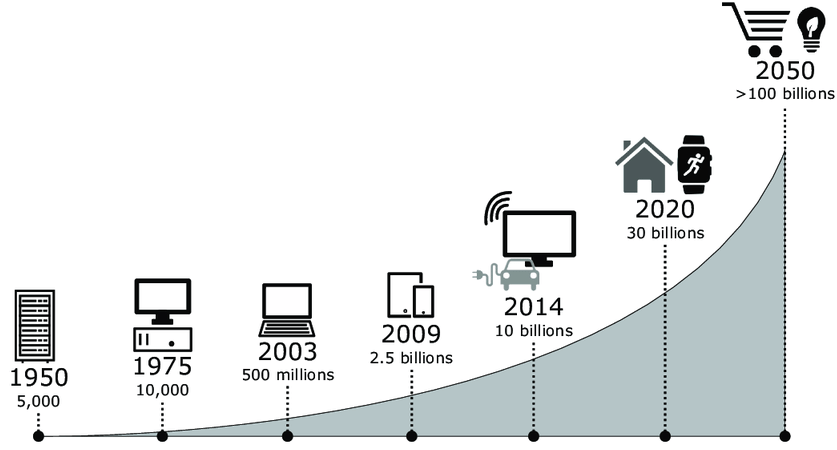


Figure 1.2: Growth of Connected Devices from 1950 to 2050 [11]

Various standardization bodies, including the 3GPP, the IEEE and the Internet Engineering Task Force (IETF), recognize M2M communication as a significant traffic source in future 5G networks [13, 14]. To address the expected surge in connected devices, these bodies have extensively redesigned future networks. Subsequently, substantial research efforts have focused on massive M2M technology, where numerous heterogeneous devices can communicate effectively, to meet diverse quality of service (QoS) requirements specified by different applications, all with minimal human intervention.

Hence, for 5G mobile networks, it is crucial to explore alternative layouts beyond the conventional distributed architecture. In 2009, China Mobile introduced the concept of Centralized Radio Access Network (C-RAN) [15]. C-RAN architecture is a defining feature of 5G NR. In the traditional distributed setup, the remote radio unit (RRU) and Baseband unit (BBU) of BSs are co-located at the same site, with the RRU positioned near the antenna outside, and the BBU located on-site. However, in C-RAN architecture, the radio transceivers are decoupled from the digital base-band processors. The RRUs are placed

closer to the antennas, and multiple BBUs from various BSs are consolidated in a centralized base-band pool to address the increased computational demands of the BBUs [16, 17].

C-RAN offers numerous advantages that make it an attractive choice for modern communication networks. One of its primary benefits is the possibility of implementing advanced technologies through resource virtualization, which enables more flexible and on-demand resource allocation compared to traditional networks. C-RAN also supports service deployment at the edge, improving resource utilization and reducing the pressure on backbone servers, leading to higher user satisfaction due to access to a larger pool of resources [15, 16]. C-RAN also facilitates advanced network optimization techniques that reduce latency, a crucial factor for real-time applications like video streaming and online gaming. Additionally, C-RAN simplifies network management through a centralized interface, reducing operational costs and enhancing overall network efficiency [15].

1.2 5G Split Centralized Radio Access Network

The split C-RAN architecture has revolutionized the traditional approach by separating the radio frequency and base-band processing functionality. This architecture consists of three logical nodes: the Central Unit (CU), the Distributed Unit (DU), and the Radio Unit (RU), each with its own 5G NR stack function. This innovative design enables the deployment of RUs within buildings to achieve highly effective CU at the central station connected through high-speed links known as front-haul (as depicted in Figure 1.3) [18]. Nevertheless, when implementing massive machine type communication (mMTC) applications in indoor environments with split C-RAN, a crucial balance must be struck between the cost of necessary infrastructural modifications and the achievable QoS levels. This trade-off becomes particularly significant when dealing with applications that demand stringent delay constraints [19].

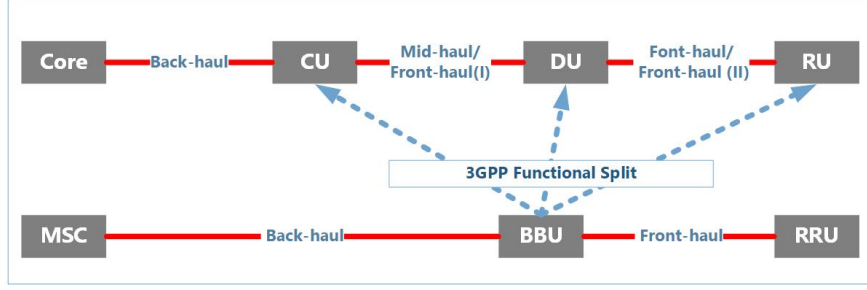


Figure 1.3: The Split C-RAN architecture

By further dividing the C-RAN into a split layout, we amplify these benefits through increased resource utilization, reduced energy consumption, and minimized interference via CoMP features [15, 16]. A split architecture, dividing the CU from the DUs coordinated performance features, load management, and real-time optimization. Moreover, this architecture enhances scalability, enabling the network to accommodate a vast number of RUs without additional hardware resources, thus simplifying expansion as the number of connected devices grows. Scalable and cost-effective network deployments rely on flexible hardware and software components that can seamlessly integrate across different vendors.

Centralizing processing resources in split C-RAN improves energy efficiency by reducing power consumption and energy costs. RUs are designed for low power consumption, while CUs and DU are optimized for energy efficiency [20]. C-RAN enables lower costs and greener communication, demonstrating that a centralized architecture can significantly reduce both operating cost (OPEX) and capital cost (CAPEX) [20]. For instance, China’s adoption of a C-RAN layout has achieved a 70% reduction in base station infrastructure OPEX [8].

This architecture allows the network to adapt to various use cases and QoS requirements, such as gaming, voice, and video, which have varying latency tolerances and transport dependencies [21]. It also enables adaptation to diverse deployment scenarios, including rural or urban environments with distinct access to transport technologies like fibre. Despite these advantages, the primary challenge in 5G C-RAN is developing a reliable front-haul link between the DU and the RUs. Additionally, this front-haul link must provide high capacity, low delay, and cost and energy efficiency to ensure the successful deployment of C-RAN in 5G networks [21].

1.2.1 C-RAN Front-haul

One of the primary challenges in split C-RAN is the front-haul, which serves as the communication link between each RU and the DU. This link is responsible for transmitting both the uplink (RU to DU) as well as the downlink (DU to RU) radio signals. Compared to the traditional distributed layout, the front-haul link in C-RAN is longer, introducing an additional, non-negligible latency component [22]. To meet the requirements of multiple RUs, the front-haul is expected to provide high bandwidth. Different transport means, such as optical fibre, wireless links, and copper cables, can be employed for Front-Hauls. Optical fibre is considered the ideal technology in C-RAN due to its large bandwidth and lower latency, but it is also the most expensive option.

Various communication media options have been considered for constructing front-hauls, including optical fibre, coaxial cable, and CAT 5/6/7 cables. However, each of these options faces limitations in the indoor scenario. For instance, the installation of optic fibres or coaxial cables can be inflexible, expensive, or even impossible due to the need for infrastructural alterations [23]. While CAT cables offer easier deployment and lower infrastructure costs compared to the previous options, they still lack the widespread availability found in power-lines. It is important to note that all the mentioned strategies may require external power sources for the indoor RUs.

1.2.2 Current Front-haul Challenges

When considering various transmission media options for the front-haul link, including wireless, copper-based wires, optical fibre, and power-lines, optical fibre is often regarded as the most desirable due to its high bandwidth and faster speed capabilities. However, using optical fibre comes with its own set of challenges.

While optical fibre is a suitable transmission media for back-haul and certain sections of the front-haul, the "last mile" connection to within buildings poses difficulties. The existing connection infrastructure in many areas is comprised of coaxial cables, phone lines, and copper cables. While it is preferable to have fibre as the primary media connecting homes

to the network, the last mile to homes may still need to be covered by coaxial cable or copper lines, making fibre-to-building (FTTB) a more practical solution [24]. Furthermore, in numerous buildings, the pathways or hidden ducts in the walls are either non-existent or already occupied, making it impossible to install new fibre cables especially in old homes [25].

In London, UK, building owners are strongly reluctant to install separate networks for each Mobile Network Operators (MNO). The absence of agreements or regulations for national roaming has resulted in a lack of mobile coverage in the London underground tunnels in the UK. Although a shared indoor infrastructure could significantly increase indoor capacity, there are currently no indications that the regulatory and commercial barriers (not technical) will be lifted during the 5G rollout [26]. In a developing countries, deploying underground fibre optic cables presents various technical hurdles [23]. The placement of underground fibre cables is typically done alongside highways and city roads. However, in areas where buildings are constructed without proper authorization, it becomes challenging to plan the fibre route and adhere to the initial plans. This makes the design and implementation of the fibre network more complex. Additionally, sharing the existing right of way by multiple MNO leads to frequent fibre cable cuts [23]. When one MNO requests to deploy its fibre optic cable by sharing right of way with another MNO, the existing MNO's fibre cable is at a higher risk of being cut during excavation MNO.

fibre optic networks offer numerous advantages but also come with certain drawbacks. One major concern is their vulnerability to physical damage due to their smaller and lighter nature. Accidental cuts during construction or rewiring can disrupt services for many users since fewer cables serve a larger population [27]. Additionally, fibres are sensitive to bending, making their installation around corners challenging. Moreover, fibre networks are susceptible to damage from radiation or exposure to chemicals. While fibre optic networks are cost-effective in the long run, their initial installation costs can be significant, requiring specialized test equipment and skilled installers [27]. Diagnosing and fixing issues in fibre optics networks can be expensive, adding to the short-term cost considerations. Another concern is the occurrence of "fibre fuse" at high power, which can rapidly destroy long lengths of

cable when excessive light meets an imperfection in the fibre. Additionally, fibre cables have limitations in propagating light in only one direction, necessitating the laying of two separate cables for bidirectional communication [27].

Limited bandwidth and short transmission distances are significant challenges when utilizing copper twisted-pair and coaxial cables for the front-haul link [28, 29]. These types of cables offer narrower bandwidth compared to other high-speed front-haul media, which can impede data transmission capacity, particularly in scenarios requiring high data rates, such as 5G networks. While CAT cables, especially higher categories like CAT6 and CAT6a, can support gigabit Ethernet speeds, they may not be suitable for extremely high data rates demanded by certain front-haul applications. Moreover, twisted-pair and coaxial cables are prone to signal degradation over longer distances, leading to increased error rates and reduced data transmission rates [28, 29].

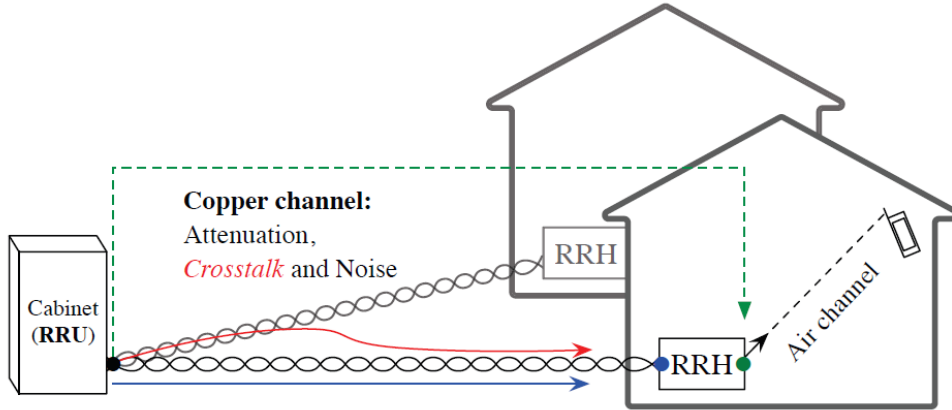


Figure 1.4: Systematic sketch of LTE-over-copper in the downstream direction [30]

In addressing these issues, a potential solution involves utilizing only the low-frequency portion (0 - 100 MHz) of each twisted-pair or cable for transmission [31, 32]. This low-frequency spectrum exhibits minimal Far end crosstalk (FEXT) and noise, enabling longer transmission distances of up to 200 meters as shown in Figure 1.4. However, implementing this solution is impractical when a high number of antennas need to be connected to the same RU. In 5G systems, a single RU may be connected to 64 antenna arrays, making it unfeasible to have 64 twisted-pair cables, as it is not scalable. Additionally, twisted-pair and coaxial cables are thicker and bulkier, making their installation more complex and challenging, especially in confined spaces such as indoor environments.

1.2.3 Front-haul Deployment Cost

The installation cost of fibre is the highest among all transmission media due to the need for specialized expertise in its installation, as well as the requirement for special test equipment and setup [20, 27]. Additionally, the use of optical fibre communication solely in front-hauling requires new expensive infrastructure. As such, re-using existing infrastructure, for example, currently deployed fibres and copper cables for the front haul system, is highly desirable. Even notable attempts, such as Google’s endeavour to reduce fibre installation costs in Louisville, USA, were unsuccessful [33]. Google fibre shut down its network in Louisville, Kentucky, and exited the city due to a series of fibre installation failures that resulted in exposed cables on the roads. The network went live in late 2017, shortly after construction began, using a shallow trenching strategy. However, the micro-trenching approach caused cables to be exposed, leading to inconvenience and safety concerns for residents. Google fibre attempted to fix the problem in August 2018, but was unsuccessful, ultimately deciding to leave Louisville [33].

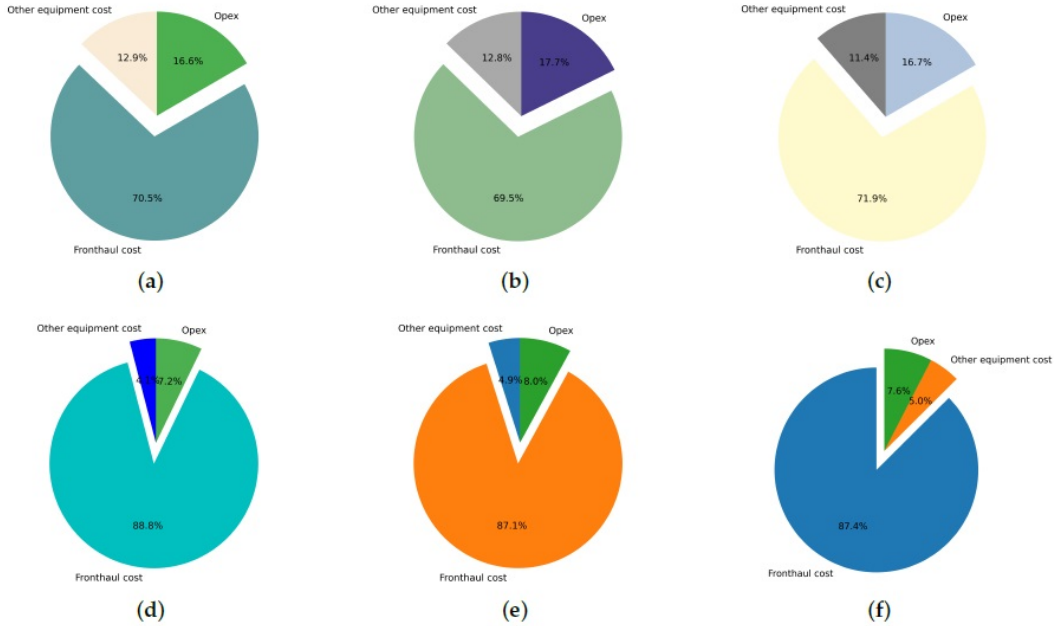


Figure 1.5: Total cost breakdown for sparse and dense areas for (a) 1:4 PON dense area, (b) 1:8 PON dense area, (c) 1:16 PON dense area, (d) 1:4 PON sparse area, (e) 1:8 PON sparse area, and (f) 1:16 PON sparse area [2]

Passive optical networks (PON) technology is considered the optimal solution for 5G and beyond network front-haul due to its ability to offer high-capacity and low-latency connections [2, 34]. Evaluating the cost associated with front-haul installation, the authors in [2] focus on three different PON splitting ratios (1:4, 1:8, and 1:16) for both dense and sparse areas. As shown in Figure 1.5, the most dominant cost is the expense of front-haul fibre deployment, which accounts for a significant percentage ranging from 69.5% to 88.8% which is equivalent to \$1.415M to \$4.56M (USD), respectively [2].

In [35], the authors show a possible reduction in front-haul deployment cost when using Wavelength Division Multiplexed Passive Optical Network (WDM-PON), wireless or a combination for front-haul link between the DU and the different RUs. As shown in Figure 1.6 a maximum cost reduction of 41% in comparison to point-to-point (P2P) fibre connection (highest cost) could be achieved, however this is still the most significant contributor to the Total Cost of Ownership (TCO).

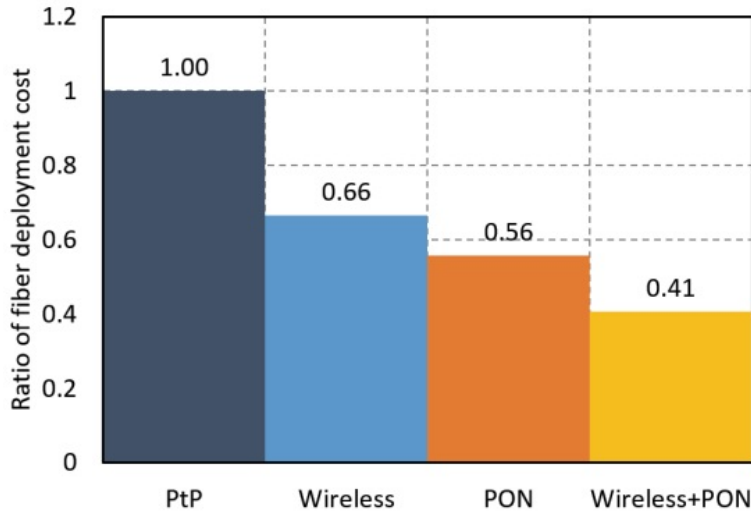


Figure 1.6: Histogram of front-haul optical fibre deployment cost reduction [35]

1.3 Research Motivation and Objective

Motivated by the limitations mentioned above and the desire to reduce the cost of deployment, this thesis aims to answer the following question: **Is low voltage indoor power lines the feasible and less expensive alternative front-haul media for provision-**

ing indoor coverage under 5G NR IoT? Alternatively, using Power-Line Communication (PLC) on in-building indoor power lines could be a promising counterpart, which has been proposed as a robust, cost-optimized solution in the fields of smart home, smart grid, broadband connections and cellular back-haul[36–38]. This is because it does not require infrastructural alteration thanks to the ubiquity of the indoor power lines and could support the traffic generated by the different IoT devices [39, 40]. Additionally, the power line medium is an open, unlicensed medium similar to the 2.4 GHz and 5 GHz radio frequency bands. Simply, power lines are installed everywhere, and this eliminates the need for any new infrastructure or licensing expenses for every customer as well as the need for any external power source. Moreover, when using PLC, there is full control over the communications as there is no licensing required and is ideal for point to point connections [41].

PLC has been taken as the supporting communication media for indoor broadband services. Although with numerous advantages against all the other counterparts thanks to its ubiquity in the indoor environment. The PLC channel is characterized by a log-normal frequency dependent attenuation affects the transmitted communication signal [42, 43]. Moreover, an indoor PLC link is subject to various types of noises. The power-line noises can be classified into background noises and impulsive noises based on their time-domain nature [44]. The spontaneous nature of the impulsive noise certainly causes a devastating corruption to the data transmitted over the power lines, and such corruptions can hardly be recovered by any coding mechanism.

1.3.1 Objective and Contributions

The main objective is to consider using low-voltage indoor power lines as part of C-RAN front-hauls to support the transportation of time division multiplexing (TDM) Common Public Radio Interface (CPRI) frames or Ethernet based Enhanced Common Public Radio Interface (eCPRI) frames for 5G NR indoor service provisioning. Particularly, we introduce two versions of a novel component, namely CPRI-PLC-Gateway (CPG) & Enhanced CPRI-PLC-Gateway (eCPG), for enabling the power lines to support CPRI/eCPRI streams over power-lines respectively. Specifically, the CPG/eCPG is deployed across the PLC based

front-haul line as a pair, to create a virtual tunnel to overcome the challenges of power-lines and meet the data rate, delay and BER requirements [45, 46]. It shall be shown that the proposed approach can shape the power lines into qualified CPRI/eCPRI links in supporting the 5G NR M2M indoor services.

Additionally, a general analytic framework is developed over the Radio access network (RAN) to derive accurate mathematical models on the radio coverage probability, the number of served devices, and the delay in the system. The derived expressions are thorough and can be used to quantify the inferences of each design parameter to the RAN end-to-end delay performance, such as the device density, transmission power, the bandwidth over the radio and PLC channels, as well as the number of devices the PLC front-haul can support under eCPRI. An in-depth analysis on the RAN end-to-end delay performance of the novel PLC based front-hauling system shall be illustrated [40]. The CPG/ eCPG pair provides a suite of functions that ensures high reliability and minimized delay jitter. Notably, a staging delay buffer (SdB) is used at the receiver CPG in order to achieve a constant bit rate (CBR) provisioning of CPRI HFs at the exit port of the PLC link to maintain its TDM nature [47].

Moreover, to further address the challenges of indoor QoS and accommodate the increasing number of connected devices, PLC has been proposed as an alternative front-haul medium due to its availability and built-in power sources in buildings. The research focuses on optimizing the association of devices with the appropriate RUs and connected PLC ports while efficiently utilizing radio and PLC front-haul resources. A multi-objective optimization problem is proposed using a centralized approach and the Non-dominated Sorting Genetic Algorithm II (NSGA-II). The main contributions include a methodology for efficient IoT C-RAN with PLC-based front-haul, a novel joint RAN problem formulation, and case studies demonstrating the benefits of the proposed algorithm in maximizing the number of supported devices, minimizing device energy consumption, and reducing access delay. The algorithm operates by systematically generating successive iterations of solutions through the utilization of tournament selection, single-point binary and simulated binary crossovers, and polynomial mutation operators. The system outputs yield a Pareto front consisting of non-dominated solutions for the issue defined using multi-objective optimization (MOO).

1.3.2 Publications Produced

The research included in this thesis has been used to produce the following publications and patent:

- **M. Hassan**, H. G. Moussa, P. -H. Ho and L. Peng, "On Power-line based Front-hauling for IoT Cellular Indoor Communications," in *IEEE Transactions on Communications*, 05 July 2024.
- **M. Hassan**, H. Moussa and P. -H. Ho, "Performance Analysis of Power Line Communication System for Massive Indoor IoT Applications," in *Journal of Networking and Network Applications*, vol. 3, issue 3, 18 Dec 2023.
- **M. Hassan**, S. H. R. Naqvi and Pin-Han Ho, "Method and System for Communicating Data Over Residential Power Lines" U.S. Patent 20220038321, February 3, 2022.
- H. Mei, **M. Hassan**, L. Peng and P. -H. Ho, "A Novel Front-hauling Architecture under Centralized Radio Access Network (C-RAN)," in *Journal of Communications and Networks*, vol. 24, no. 3, pp. 305-312, June 2022.
- **M. Hassan**, S. H. R. Naqvi and Pin-Han Ho, "On CPRI based Front-hauling over Residential Power Lines", *Proceedings of the International Conference on Networking and Network Applications (NaNA)*, December 2020.

1.4 Thesis Outline

Chapter 2 presents a thorough literature review of all considered topics in this research. It also includes a discussion, and a list of gaps extracted from previous work.

Chapter 3 proposes a system design where power-lines are used as the main communication media for the "last mile" front-haul link in 5G C-RAN.

Chapter 4 investigates in detail the effects of created virtual link on the system RAN end-to-end delay and presents a detailed performance analysis of the IoT cellular network using PLC based front-haul.

Chapter 5 proposes a MOO problem to maximize the system throughput in terms of the number of satisfied IoT devices while meeting their QoS requirements in terms of device power as well as PLC access delay.

Chapter 6 provides the conclusion of the thesis. It also articulates the platform for future research work.

Chapter 2

Background and Literature Review

Split C-RAN emerged as a promising architecture, where the CUs of the massively deployed BSs are centralized into a common CU pool. On the other hand, the RUs are geographically distributed over a wide area in order to provide radio transmission and reception and connected to the DU [48]. The centralisation of the base band processing functions allows for multiple gains in regard to programmability, scalability, cost reduction, and mutual cooperation among the CUs for interference mitigation. At the antennas site all the radio functions and minimal local processing are performed by the RUs. All base band functions such as modulation/demodulation and precoding are handled by the DUs or the CUs. As mentioned in the first chapter, the main key issue in split C-RAN architecture is the front-haul connection between the RUs and the DU. In the next section, we will review the necessary background and literature review.

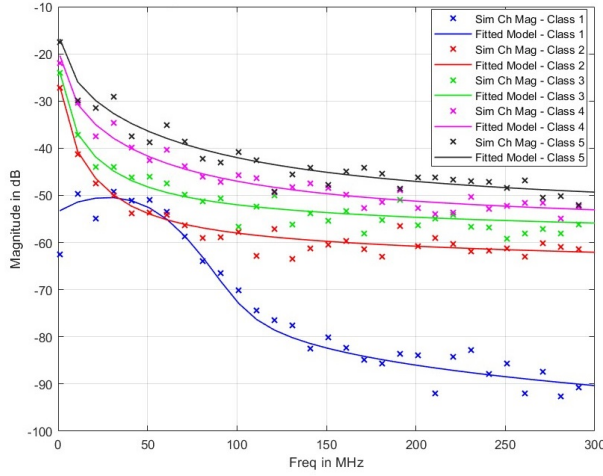
2.1 Power-Line Communication Channel Model

PLC has been deployed for broadband data transmissions for various applications such as smart grids, home entertainment and internet services [49–51]. The ubiquity of the indoor power-lines has made PLC as a promising candidate for any indoor wireless system. Furthermore, it allows the system to utilise the existing electrical power grid that can greatly ease the installation of indoor devices and significantly reduce installation cost [50]. Power lines are different from other wired media. The powerline channel has foundational features that are inherited from its media [51]. Therefore, it is essential to study the PLC channel model.

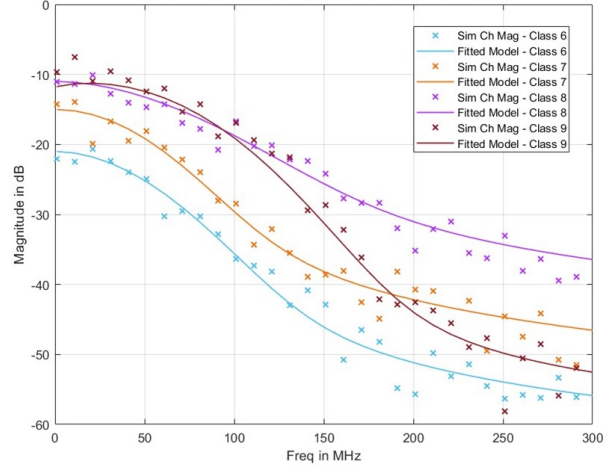
2.1.1 PLC Channel Bandwidth

The properties and interferences within indoor power line channels have been extensively examined in existing literature, adopting both bottom-up and top-down analytical approaches. Tlich *et al.* introduced a top-down statistical generator for the single-input-single-output (SISO) PLC channel [52]. Based on this generator, a fitting algorithm for nine distinct channel capacity classes was proposed in [53]. Although these findings could be relevant to our research objectives, they are limited to channel behaviour within the 100-MHz frequency range. However, for our specific study involving indoor front-hauling over power lines, we need to extend the PLC channel bandwidth to encompass the **1 MHz - 300 MHz** frequency range [54, 55]. This choice of extending the model to 300 MHz is grounded in the fact that currently available broadband PLC chipsets are capable of operating within this range. In particular, the broadband frequency range for PLC already spans from 1 MHz to 300 MHz [55, 56].

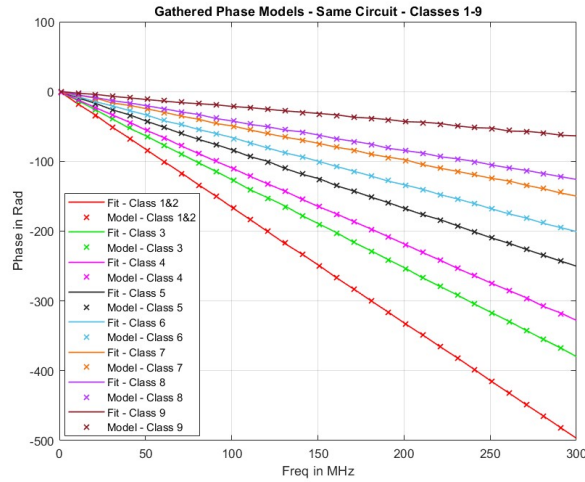
As comprehensive statistical channel models are scarce within this frequency range, our approach involves extending the statistical channel model from [53] and the fitting model from [52, 54] to encompass the entire 300 MHz bandwidth. This extension accommodates all nine rate classes identified in [53]. The transfer function and phase characteristics of the extended channel and fitting models exhibit consistency across the expanded frequency spectrum, as depicted in figures 2.1 and 2.1c. To validate the accuracy of our extended statistical model, we conducted a comparison with an extended-frequency version of the SISO PLC channel model sourced from the ICT OMEGA Project [57]. The results demonstrated alignment between our proposed model and the OMEGA PLC channel model within the extended 1-300 MHz frequency range. Consequently, we advocate employing the OMEGA PLC channel model within our system framework, encompassing the extended frequency range of 1-300 MHz.



(a) CTF - Classes 1 - 5



(b) CTF - Classes 6 - 9



(c) Channel Phase - Classes 1 - 9

Figure 2.1: Extended SISO PLC Channel Model to 300 MHz - CTF and Channel Phase - Same circuit

2.1.2 MIMO PLC Channel Model

Our study encompasses the adaptation of the ICT OMEGA SISO PLC channel model, which operates by transmitting the communication signal over the differential port formed by the Phase/Live (P) and Neutral (N) wires of the power-line. This model characterizes frequency-domain attenuation within the PLC channel and has undergone validation through statistical analysis of measurements collected from numerous test houses centred around PLC technology [58]. This adapted model has been extended to effectively encompass a frequency

range spanning from **1 MHz to 300 MHz** [58]. The multi-path SISO PLC Channel Transfer Function (CTF) can be mathematically expressed as follows:

$$H_{PN-PN}(f) = A \sum_{n=1}^{N_{pth}} g_n e^{-j \frac{2\pi d_n}{v_p} f} e^{-(a_0 + a_1 f^K) d_n} \quad (2.1)$$

where g_n is the n^{th} path gain, N_{pth} is the number of paths and d_n is the n^{th} path length. The speed of light in the PLC line is calculated as $v_p = 2/3c$ and $c = 3 \times 10^8$. The path lengths are modelled as uniformly distributed in the range $[0, L_p]$, where L_p is maximum path length and can reach up to 800m. The number of paths N_{pth} is modelled as a Poisson random variable (RV) with mean $\Lambda.L_p$ and with intensity $\Lambda = 0.2$. The distribution of the attenuation parameters A , a_0 , a_1 , and K are dependent whether the measurements were obtained over 2 power sockets situated on the same circuit or on two different circuits. The distributions of the different parameters are presented in table 2.1 [58].

Table 2.1: Statistical models of channel parameters [58]

	Same Circuit		Different Circuit	
	Co-Ch	Cross-Ch	Co-Ch	Cross-Ch
A	$Uni\ [0.005, 0.25]$	$Uni\ [0.005, 0.15]$	$Exp\ (\mu_A = 0.00238)$	
a ₀	$Exp\ (\mu_{a_0} = 0.00827) - 0.005$			
a ₁	$4e - 10$			
K	$\mathcal{N}\ (\mu_K = 1.01748, \sigma_K = 0.01955)$			

The potential for transmitting a MIMO signal exists through the utilization of distinct electrical wires or conductors within a single or multi-phase power line. In such arrangements, each power line typically comprises a minimum of three wires: P, N, and Protective Earthing (PE). When contemplating the PE wire for signal transmission, the signal can be effectively conveyed across three differential ports among P conductor, N conductor and PE conductor as P-N, P-PE, and N-PE enabling the establishment of a 3×3 MIMO channel as illustrated in Figure 2.2.

All other channel elements in the MIMO PLC channel can be generated from the SISO PLC channel using the below equation:

$$H(f) = A.\Delta A \sum_{n=1}^{N_{pth}} g_n e^{-j\varphi_n} e^{-j\frac{2\pi d_n}{v_p} f} e^{-(a_0+a_1 f^K)d_n} \quad (2.2)$$

where ΔA is the measure of the attenuation of the cross-channel in comparison to the PN-PN channel. In different circuit case, $\Delta A \simeq 1$ while for the same circuit case it is modelled as an exponential RV ($\exp(\mu = 0.3659) + 0.45$). φ_n are uniformly distributed in the range of $-\Delta\varphi/2$ and $\Delta\varphi/2$. The angle $\Delta\varphi$ represents the generated channel correlation with the SISO PN-PN channel. The correlation is larger with smaller values of $\Delta\varphi$ [58]. Table 2.2 shows the values of $\Delta\varphi$ for each channel path.

Table 2.2: The values of $\Delta\varphi$ to generate φ_n to each defined path [58]

	PN-PPE (h_{12})	PN-NPE (h_{13})	PPE-PN (h_{21})	PPE-PPE (h_{22})	PPE-NPE (h_{23})	NPE-PPE (h_{32})
Same circuit ($\Delta\varphi$)	π	π	π	π	π	$\pi/2$
Diff. circuit ($\Delta\varphi$)	π	π	π	2π	π	π

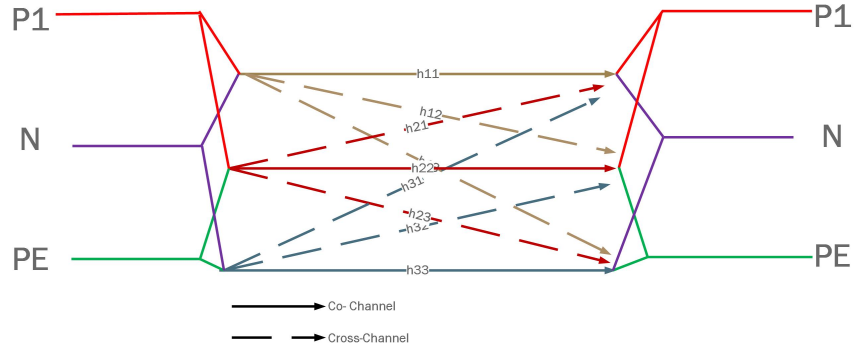


Figure 2.2: A 3x3 PLC MIMO channel model

2.2 PLC Noises Models

In PLC there are 5 types of additive noises. The source of noise at low voltage (LV) power line can be internal (inside the power network) or external (outside the power network). The noises are classified into 2 main classes: **Background noise** and **Impulsive noise**. Two distinct categories of noise are prevalent in Power Line Communication (PLC): background noise and impulsive noise. Background noise stems from the operation of household appliances and radio broadcasts, while impulsive noise arises from transient power surges triggered by shifts in the states of appliances, power supplies, and components with high cut-off voltage [44]. In scenarios involving data transmission through PLC, managing impulsive noise is paramount to ensuring the viability of the system. The modelling of the different PLC noises are discussed below.

2.2.1 Background Noise Model

The background noise in PLC is a kind of coloured Gaussian noise, which can be assumed to be cyclo-stationary with a level which strongly relies on the number and type of electrical devices connected to the network [44, 59]. Such noise can be characterized in terms of power spectral density (PSD) in unit of dBm/Hz as shown in Figure 2.3 and through a simple model with three parameters [60] as follows: $S_{bg}(f) = a_{bg} + b_{bg} * |f|^{c_{bg}}$ [dBm/Hz], where f is frequency in unit of MHz. The value of a_{bg} , b_{bg} and c_{bg} have been specified by [44] as shown in the Table 2.3 and have been adopted in [61].

Table 2.3: Background noise parameters

	a_{bg}	b_{bg}	c_{bg}
Best Case	-140	38.75	-0.72
Worst Case	-145	53.23	-0.337

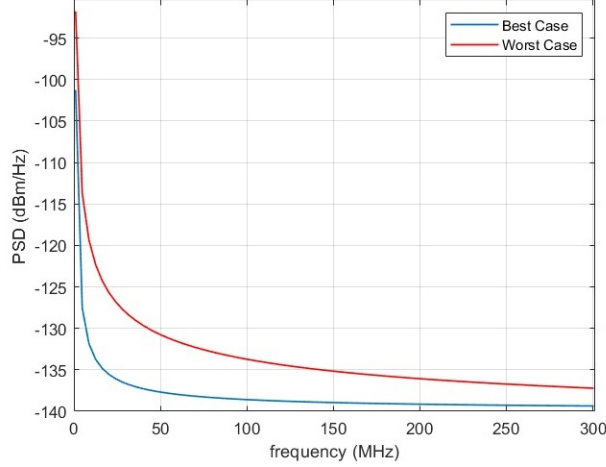


Figure 2.3: Background noise PSD - Worst and Best cases [44]

2.2.2 Impulsive Noise Characteristics

When an impulse noise event takes place, the noise PSD typically surges to a level at least 10-15 dB higher than the usual background noise for a time span of up to $100 \mu s$ [44, 62]. This impulsive noise effect can lead to a substantial power surge of up to 50 dB, resulting in significant data transmission losses [44, 62].

Impulse noise comprises three key elements: the amplitude of the impulse, the inter-arrival time between successive impulses, and the duration of each individual impulse [44]. The statistical modelling of impulse amplitude has been approached through various distributions. Existing models in the literature include Middleton's Class A, Double-sided Beta, and two-term Gaussian distributions. Research detailed in [44] reveals that Middleton's Class A model best suits the measurements, with an inter-arrival time following an exponential pattern with a mean of $\mu_i = 10.1 ms$ [44]. Although the impulse width can be regarded as a constant, such as $100 \mu s$, the two-term Gaussian model put forth in [44] offers a more accurate depiction of impulse width, with a mean of $4.7 \mu s$. Table 2.4 enumerates the specific statistical distributions and parameter values employed in this study.

Table 2.4: Impulsive Noise statistical noise model [44]

	PDF	Parameters
Amplitude	Middleton's class A	$A_{imp} = 5.8, \Gamma = 0.1$
Impulse Inter-arrival time	Exponential	$\mu_i = 10.1 \text{ ms}$
Impulse Width/ Duration	Two-terms	$P_1 = 0.0736, m_1 = 4.9, \sigma_1 = 0.2$
	Gaussian model	$P_2 = 0.0318, m_2 = 4.2, \sigma_2 = 0.25$

2.3 Impulsive Noise: Challenges and Mitigation

The impulsive noise [44, 62] is unique in the PLC environment, upon which the noise PSD may become 10-15 dB higher than that of usual background noise for a time duration up to 100 μs , causing a significant loss in data transmission [44]. Thus, although mitigating the effect of the impulsive noise has been widely reported in the literature, they are not effective as most of these techniques focus on the handling, detection and removal of the data error caused by the impulses rather than to avoid the exposure of data to a noise impulse in the first place.

2.3.1 Impulsive Noise Reduction Techniques

Prior research has extensively explored strategies for mitigating the impact of impulsive noise, which can be categorized into two main approaches: error handling and detection/removal techniques. Among error handling techniques, automatic repeat request (ARQ) stands out, involving the re-transmission of data segments affected by impulses. However, the use of ARQ can lead to an inherent reduction in spectral efficiency. To address this, Hybrid automatic repeat request (HARQ), with its inherent Forward Error Correction (FEC) capabilities, is often favoured as an alternative solution for handling impulsive noise-related errors [63].

Another effective approach for error handling is channel coding. In the context of the broadband PLC standard (IEEE P1901), diverse channel coding techniques are outlined, including convolutional encoding, Turbo code, and low-density parity-check (LDPC) code,

depending on whether the system is fast Fourier transform (FFT) or Wavelet based. Reed-Solomon (RS) coding is often used in conjunction with these techniques to counter impulsive noise [63]. Additionally, time-domain and time-frequency interleaving techniques are employed to address bursty impulsive noise occurrences [63].

Noise reduction techniques focusing on detection and removal leverage signal processing at the receiver end to counteract the impact of impulsive noise. One such approach is Blanking and Clipping, a time-domain method wherein the received signal affected by the noise impulse is either nullified or clipped. The effectiveness of these techniques depends on the chosen thresholds and the signal's peak to average power ratio (PAPR) [64]. An alternative strategy, known as Replacement, involves substituting the affected signal samples with the mean of the received signal. Clipping, replacement, or a combination of both can be employed, potentially alongside blanking [65]. However, a key limitation of these methods is the requirement for the receiver to possess knowledge about the characteristics of the impulsive noise.

Iterative threshold-based algorithms, time-frequency domain equalizers, and compressive sensing-based techniques also fall under the umbrella of detection and removal noise reduction methods [63, 65]. These approaches operate within a similar framework, aiming to rectify the impact of impulsive noise on the received signal.

2.3.2 Reviewing Performance Modeling in PLC Networks

Prior to investigating the newly proposed PLC-based system, it is imperative to delve into the existing literature on delay modelling in PLC networks. In studies such as [66, 67], the primary focus is on determining data delay through measurements conducted on a broadband PLC-based test-bed. These endeavours also involve ascertaining statistical delay models tailored for various applications. Furthermore, certain investigations centre on delay implications within broadband PLC-based smart grids, attributing these delays to elements like repeaters and transformers situated in both medium voltage and low voltage segments of the power grid [41, 68].

Various studies have explored delay dynamics in different networks, encompassing wire-

less and computer networks [69–72]. However, to the best of the authors’ knowledge, none have extensively addressed delay analysis within the context of a PLC-based network. Furthermore, a substantial proportion of these studies tend to narrow their focus to a specific delay type or make concessions on re-transmission rates and FEC techniques [69, 70]. In alignment with the objectives of this thesis, a comprehensive analysis of introduced delays in the context of the PLC virtual link-based front-haul architecture, as presented in [40, 47], is undertaken in chapter 4.

2.4 Overview of CPRI and eCPRI

Since this thesis is focused on the transmission of CPRI hyper-frames as well as eCPRI data-frames over the PLC-based front-haul, below is an overview of both standards:

2.4.1 Common Public Radio Interface (CPRI)

The CPRI is a widely adopted protocol standard used in wireless communication systems, particularly in cellular networks. It plays a pivotal role in enabling the efficient communication between different network components, primarily the BBU and the RRU, within a C-RAN architecture [73]. CPRI serves as a vital link that facilitates the exchange of digital data and control information between the BBU, which processes and generates the base-band signals, and the RRU, which is responsible for the Radio frequency (RF) processing and transmission. This protocol is designed to accommodate the high-speed and low-latency requirements of modern wireless networks, particularly those operating in 4G and 5G environments [73].

CPRI operates over dedicated optical fibres or high-speed electrical interfaces, depending on the deployment scenario and the requirements of the network. It employs a point-to-point communication architecture, where each BBU is connected to one or more RRUs via separate CPRI links. This architecture offers flexibility and scalability by allowing multiple RRUs to be connected to a single BBU, forming a coherent and synchronized network [73]. CPRI defines a set of specifications and protocols that govern the transmission of various types

of data between the BBU and RRU. These data types include digitized base-band signals, control signals, synchronization signals, and management information [73]. The CPRI protocol stack encompasses multiple layers, each responsible for specific functions such as data framing, synchronization, and error checking. The interface specifications provided by CPRI are categorized into different rates, such as CPRI Line Rates 1 to 10 each corresponding to a specific maximum data rate. These rates cater to various deployment scenarios, ranging from small cells to macro cells, and provide the required bandwidth to support the growing demands of data-intensive applications [73].

The adoption of CPRI offers several benefits to wireless networks. It enables the decoupling of the base-band processing from the RF processing, allowing for centralized or cloud-based base-band processing, which in turn enhances resource utilization and optimization. This decoupling also results in easier maintenance and upgrades, as well as simplified network management [73]. Moreover, CPRI's standardized interface ensures interoperability between equipment from different vendors, promoting a competitive ecosystem and reducing vendor lock-in. This standardization also aids in future-proofing the network, as CPRI allows for the seamless integration of new technologies and radio standards without significant changes to the infrastructure [73]. In summary, CPRI plays a crucial role in modern wireless networks, enabling efficient and high-speed communication between BBUs and RRUs. Its standardized interface and scalable architecture contribute to the flexibility, performance, and interoperability of cellular networks, making it a fundamental component in the evolution of wireless communication systems [73].

2.4.2 Enhanced Common Public Radio Interface (eCPRI)

The eCPRI is an evolution of the traditional CPRI protocol, specifically designed to address the requirements of next-generation 5G networks and beyond. As wireless networks continue to advance, the need for higher bandwidth, lower latency, and more efficient communication between network components has become increasingly critical. eCPRI aims to meet these evolving demands by enhancing the capabilities of the CPRI protocol [74].

eCPRI retains the fundamental principles of CPRI, such as enabling the communica-

tion between the BBU and the RRU in RANs. However, it introduces modifications and optimizations to the CPRI protocol to accommodate the specific requirements of 5G deployments, including M2M, massive MIMO, ultra-reliable low-latency communication (URLLC), and enhanced mobile broadband (eMBB) applications [74]. One of the primary objectives of eCPRI is to reduce the latency introduced by the communication interface between the BBU and the RRU. This reduction is essential for meeting the stringent latency requirements of 5G applications, such as vehicle-to-everything (V2X) communication, industrial automation, and augmented reality/virtual reality (AR/VR) services. By streamlining the data exchange and control signalling processes, eCPRI aims to achieve lower end-to-end latency compared to its predecessor [74].

eCPRI introduces several key enhancements over traditional CPRI. eCPRI introduces a flexible front-haul split, allowing for more efficient separation of base-band and radio functions [74]. This split enables different levels of processing to be performed at the RRU, minimizing the amount of data that needs to be transmitted over the front-haul link. eCPRI enables a finer-grained functional partitioning of the protocol layers, allowing for more optimized data transport and reduced overhead. eCPRI incorporates header compression techniques to minimize the overhead associated with control and management information, thereby increasing the effective data throughput [74]. eCPRI supports adaptive rate control mechanisms, allowing the data rate to be dynamically adjusted based on the traffic requirements, channel conditions, and available resources. eCPRI defines higher line rates to accommodate the increased bandwidth demands of 5G networks, ensuring that the protocol can handle the larger volumes of data generated by advanced applications [74].

eCPRI holds the potential to significantly enhance the efficiency, latency, and overall performance of 5G networks. Its focus on reducing latency and optimizing data transport aligns well with the requirements of applications that demand ultra-low latency and high throughput, such as real-time industrial automation, tactile internet, and mission-critical services [74]. As 5G deployments continue to expand and evolve, eCPRI's ability to seamlessly integrate with new technologies and standards positions it as a critical enabler for future wireless networks. Its enhancements in front-haul splitting, functional partitioning, and adaptive rate

control contribute to the evolution of radio access networks that can efficiently support the diverse and demanding communication needs of the digital era [74].

2.5 Narrow band Internet of Things (NB-IoT)

In this research, we focused on the use of the RAN in M2M communication. Therefore, it is essential to identify the parameters for the radio interface. Further in this research, Narrow band Internet of Things (NB-IoT) wireless communication technology is adapted. NB-IoT is designed to enable efficient and reliable connectivity for a wide range of IoT devices. It is part of the 3GPP cellular standards and operates within the licensed spectrum, making it suitable for applications that require secure and robust connectivity [75]. NB-IoT is specifically tailored to meet the connectivity needs of low-power, low-data-rate IoT devices that operate in remote or challenging environments. Unlike traditional cellular networks that focus on high data rates, NB-IoT prioritizes extended coverage, deep indoor penetration, and efficient power consumption. This makes it suitable for applications that require long battery life, such as smart metering, agricultural monitoring, asset tracking, and smart city deployments [75, 76].

NB-IoT offers several benefits for indoor applications. NB-IoT's signal has excellent indoor penetration capabilities, allowing it to penetrate walls, buildings, and other obstacles. This ensures that even in challenging indoor environments, where traditional cellular signals might struggle, NB-IoT devices can maintain reliable connectivity. NB-IoT is designed to operate with minimal power consumption, making it ideal for battery-operated devices. In indoor settings, where devices may not have easy access to power sources, NB-IoT's energy efficient operation extends the lifespan of batteries, reducing maintenance efforts [75, 77]. NB-IoT uses a narrow bandwidth, which allows it to operate in a more efficient manner, especially for devices that only need to transmit small amounts of data intermittently [75]. NB-IoT's narrow-band operation minimizes interference with other wireless devices operating in the same indoor environment. This ensures that IoT devices can coexist without disrupting other wireless services. In table 2.5 shows the network key performance indicators (KPIs) over the radio interface which enable it to support a high number of connected devices

within a cell. It also allows it to be scalable for large-scale IoT deployments [76]. NB-IoT's extended coverage, energy efficiency, and suitability for indoor environments make it a valuable technology for various applications ranging from asset tracking and smart buildings to healthcare and indoor agriculture [75, 76].

NB-IoT's ability to provide low-cost, low-power, and wide-area coverage for IoT devices positions it as a key enabler for the expansion of IoT applications. Its integration with existing cellular networks and support from mobile network operators contribute to its widespread adoption. As IoT deployments continue to grow, NB-IoT is poised to play a vital role in connecting a diverse range of devices and services, contributing to the development of a smarter and more connected world. NB-IoT is the best performing technology for IoT applications supporting operations in extreme coverage and a massive number of devices [75, 76].

Table 2.5: KPIs values for NB-IoT [77]

Performance Level	RSRP (dBm)	RSRQ (dB)	RSSI (dBm)	SINR (dB)
Excellent	≥ -80	≥ -5	≥ -85	≥ 20
Good	≥ -90 to < -80	< -5 to ≥ -10	< -85 to ≥ -95	≥ 10 to < 20
Fair	≥ -100 to < -90	< -10 to ≥ -15		≥ -5 to < 10
Poor	≥ -110 to < -100	< -15 to ≥ -20	< -95	≥ -20 to > -5
Very poor	≥ -130 to < -110	< -20		< -20

2.6 Chapter Summary

In this chapter, the literature review findings were presented, covering various topics relevant to the thesis. The emergence of split C-RAN architecture involves centralizing CUs and distributing RUs for radio transmission. PLC channel properties, noise, and impulsive noise challenges were studied. Adapting PLC for indoor front-hauling required extending the channel bandwidth to 300 MHz. The MIMO PLC channel model, based on the ICT OMEGA SISO PLC model, characterized attenuation and was adapted for a 1-300 MHz frequency range. Impulsive noise mitigation techniques like ARQ, HARQ, and channel coding were explored. Delay modelling in PLC networks was reviewed, with limited research addressing PLC-based networks. The CPRI was introduced as a protocol standard facilitating communication in cellular networks, while the eCPRI evolution tailored for 5G addressed latency and efficiency needs. NB-IoT was highlighted as a specialized wireless technology within 3GPP standards, focusing on efficient and reliable connectivity for low-power IoT devices with features such as extended coverage, narrow bandwidth, and scalability. These insights collectively contribute to a comprehensive understanding of the technological landscape relevant to the thesis.

Chapter 3

PLC based Front-haul System

To use power lines as the media of 5G NR front-hauls, many issues arise due to the harsh nature of power lines which were originally used for electricity transportation. The PLC channel is time-varying with Log-normal deep frequency dependent attenuation, which directly impacts the signal transmission [42, 43, 78]. A PLC channel is subject to many types of noises that cannot be described as additive white Gaussian noise (AWGN). This research considers using indoor power lines as part of C-RAN front-hauls to support the transportation of TDM CPRI HF's or eCPRI Ethernet based data frames for 5G NR indoor service provisioning.

3.1 PLC Front-haul System Design

While the developed system can be adaptable to various scenarios, its primary focus is on indoor service buildings, such as virtual reality office spaces, airports, shopping malls, or stadiums. These buildings are equipped with a 3-phase power line configuration (including 3 P conductors in addition to N and PE conductors which are distributed throughout the building to serve different areas of the building. The earthing system within the building adheres to the TN-S earthing configuration, where PE and N are distinct conductors that are interconnected only near the power source, as depicted in Figure 3.1. The communication dependent distributed RUs are linked to the power line network to facilitate communication.

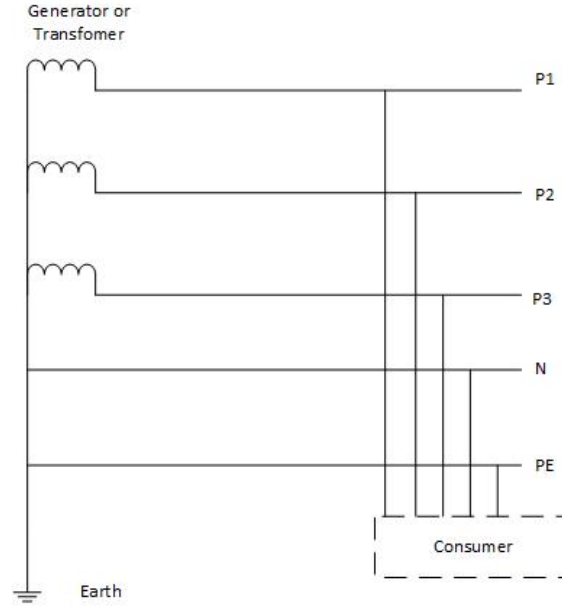


Figure 3.1: TN-S earthing system: PE & N are not connected after the power source

The power line grid functions as a mesh network, although our focus lies primarily in understanding its communication topology. The communication aspects pertain exclusively to low voltage power lines. Communication is facilitated through a gateway unit, which incorporates a low-pass filter (LPF) for signals below 1 MHz, effectively filtering out electric signals at 50 or 60 Hz, and a high-pass filter (HPF) to allow only communication signals above 1 MHz. Appliances connected to the network introduce impulsive noise due to their plugging and unplugging, disrupting the line load and causing mismatches [79]. Background noise is also factored into the model as coloured Gaussian noise instead of White Gaussian noise. The communications signal is sent between the transmitter and receiver in differential mode. Given that each power outlet possesses a single-phase power line, employing 2 differential ports between the P and N conductors and the P and PE conductors permits the application of Space division multiple access (SDMA), enabling the transmission of more than one signal. The system's channel bandwidth design spans through 1 to 300 MHz range, allowing up to three RU to consume nearly 300 MHz bandwidth each over the power-line, utilizing the 3 available ports per line, assuming no access technique applied. Within the building, the main power line maintains its low voltage nature, with an extended bandwidth of 1 to 300 MHz. Furthermore, this main power line consists of 3-phase wiring, including 3 live conductors (P1,

P2, P3) alongside N and PE conductors, totalling 5 conductors. The communications signals are sent in differential mode over the 6 differential ports P_1N , P_2N , P_3N , P_1PE , P_2PE , P_3PE as illustrated in Figure 3.2. The NPE differential conductor is not used in this system for simplicity.

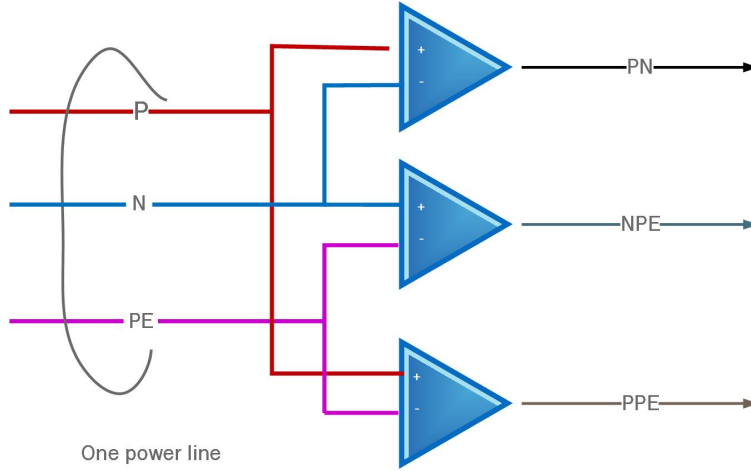


Figure 3.2: Differential signal over Power line

In our devised system, the power-line network within the service building serves as a last-mile transmission medium for the second segment of the front-haul link, adhering to the split C-RAN architecture discussed earlier. This strategic approach obviates the need for installing new communication links within the building and capitalizes on the pre-existing, widely distributed power-line infrastructure. Each floor's PLC channel assumes the form of a multi-path MIMO channel, taking advantage of the dual ports available on each line and leveraging the entire bandwidth (1 to 300 MHz) to convey CPRI/eCPRI frames. Although the PLC channel model detailed in [58] was originally designed for a channel bandwidth up to 100 MHz, our design extends its validity up to 300 MHz. This assumption is grounded in the observation that the newly generated channel path loss model still adheres to the fitting-based propagation loss model established in [53] as discussed in chapter 2.

3.2 Network Architecture

A split C-RAN scenario is adopted as demonstrated in Figure 3.3, where a conventional BS is divided into three entities: CU, DU, and RU. The CU is located at the central office and connected to the DU via the first segment of front-haul(I)/mid-haul [18], mostly on an optical fibre link carrying the CPRI or eCPRI signal. The DU are further connected to the RUs distributed in each indoor chamber via the second segment of the front-haul (II) carrying digital CPRI HFs or eCPRI data frames. In this dissertation we shall refer forward to CPRI HFs or eCPRI data frames as HFs and the second segment of the front-haul (II) is further referred to as simply front-haul.

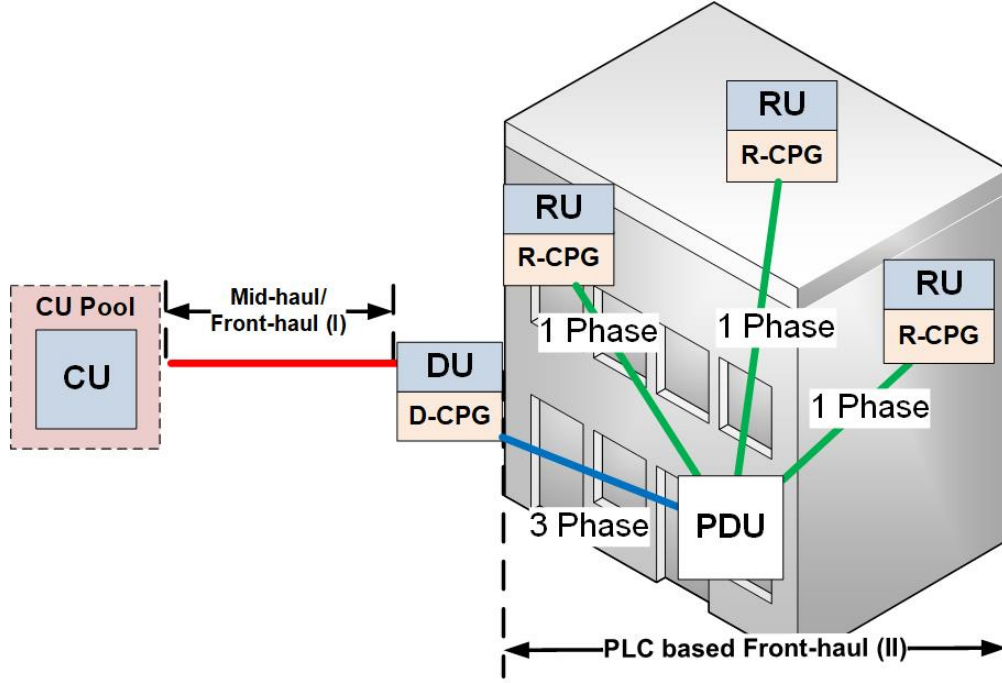


Figure 3.3: Proposed split C-RAN system architecture for 5G indoor

This study explores the utilization of low voltage indoor power lines within C-RAN front-haul configurations for transmitting TDM CPRI frames or Ethernet-based eCPRI frames in the context of indoor 5G NR services. To facilitate this, a pioneering component known as the CPG or eCPG is introduced which serves the purpose of enabling power lines to effectively support CPRI/ eCPRI streams, respectively, amidst the inherently noisy PLC

environment. Both CPG or eCPG will further be referred to as CPG. More specifically, the CPG is strategically placed at both the DU and RU, forming a complementary CPG pair. This configuration establishes a virtual tunnel that ensures the seamless transport of HFs with favourable data rates and minimal BER, as illustrated in Figure 3.4. This accomplishment is achieved through the integrated functionalities of the proposed CPG, primarily driven by two key modules: singular value decomposition (SVD) and IND, both of which are complemented by the incorporation of HARQ mechanisms.

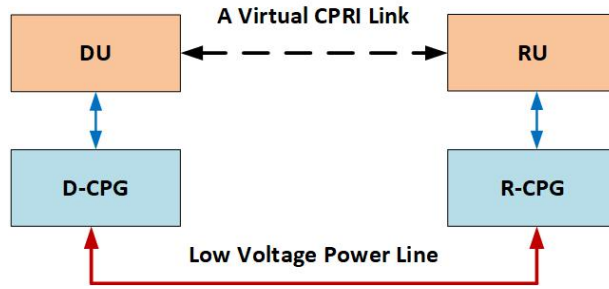


Figure 3.4: The virtual link created using the CPG pair over the power line

In the system model, each DU and RU is connected with a CPG, denoted as D-CPG and R-CPG respectively. The CPG is a plug-and-play component that can be deployed to interface between the front-haul system and the power lines without any modification to the DU and RU. It uses a HPF to obtain the PLC signals. Specifically, the D-CPG firstly connects to the indoor power distribution unit (PDU) via a 5-conductor three-phase line. Beyond the PDU, each power socket receives one of the single-phase lines containing 3 conductors: the P, N and PE conductors, by which a RU attached with the R-CPG is connected (see Appendix A for complete system flow charts).

Figure 3.5 illustrates the detailed functional diagram of the proposed PLC based front-hauling system. Here, the DU serves as a hub that multiplexes/de-multiplexes the CPRI/eCPRI streams between the CU and the RUs. It may also bear some split functions from the original BBU. The PDU, on the other hand, is a dummy component of the electric grid simply interfacing the three-phase line with the single-phase line. Each antenna of the RUs is connected to the PDU via one of the shared single-phase lines.

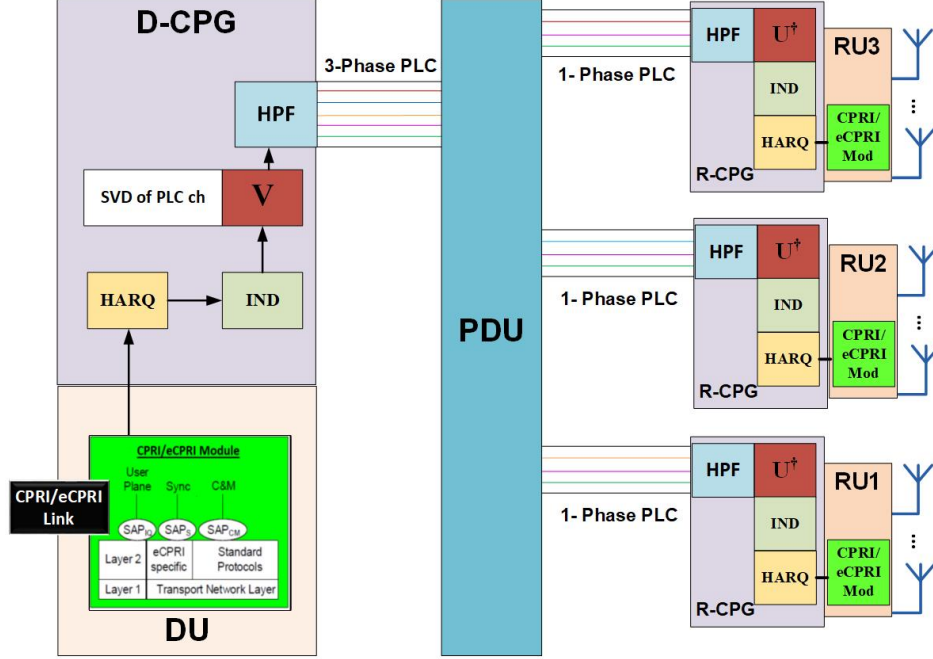


Figure 3.5: The functional diagram of the proposed system architecture

3.2.1 eCPRI System Functional Split

The system is centred around a service building scenario, wherein the building is outfitted with a 3-phase power-line infrastructure. A pivotal aspect of the system's architecture lies in the split of functionality among the various components within the BS. Illustrated in Figure 3.6, the foundational assumption is that the functional split between the RU and DU adheres to the eCPRI option 7-2, known as Splits II_D & I_U , which represent IQ-oriented splits for both downlink and uplink transmissions. Notably, this particular split entails demanding requirements in terms of high-capacity data transfer across the power-line section of the front-haul, while concurrently necessitating stringent latency thresholds. Conversely, the precise functional division between the CU and DU is contingent upon the technology employed for the front-haul segment, and this determination is typically made by the network's vendor/operator. However, the specifics of this functional split are not addressed within the scope of this research study.

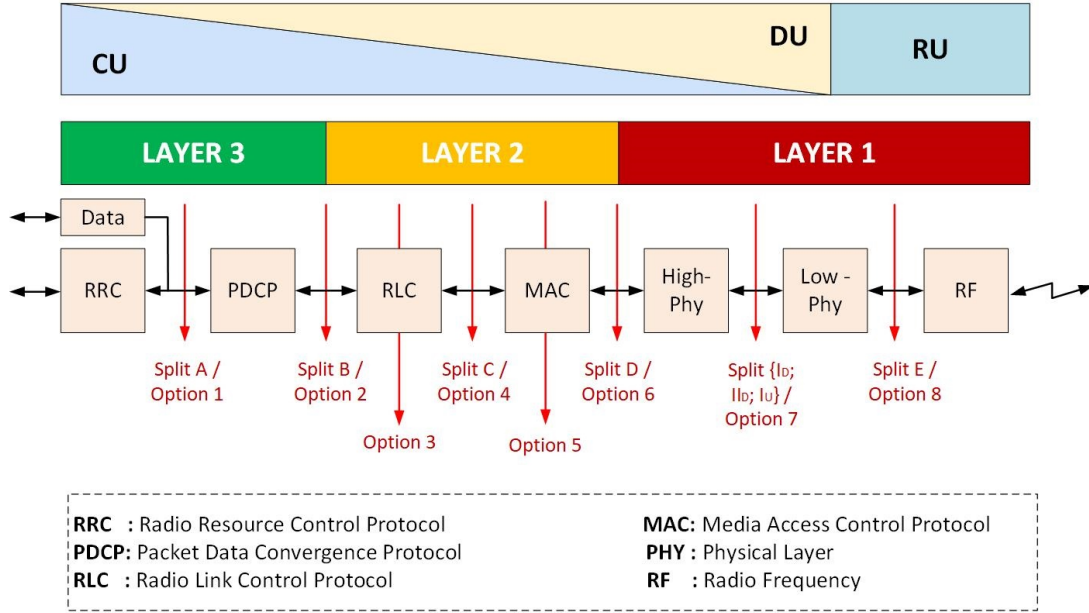


Figure 3.6: System Functional Split

3.2.2 In-Band Full-Duplex (IBFD)

A three-phase power-line configuration is employed as the final link for 5G NR front-haul communication. Within each phase, we consider a 2x2 MIMO PLC channel, focusing on the utilization of the 2 differential ports PN and PPE, omitting the NPE port. Facilitating full-duplex communication, our design incorporates In-band full-duplex (IBFD), enabling simultaneous transmission and reception of signals through the same ports. This IBFD approach holds the potential to double the data rate and facilitate bidirectional point-to-point communication over the PLC segment of the front-haul. To realize effective IBFD-based communication, it is imperative to mitigate self-interference (SI). The reduction of SI involves two main categories of techniques: echo suppression and interference cancellation. Echo suppression involves isolating the transmitted signal from the incoming signal-of-interest (SOI), effectively minimizing the SI within the composite received signal. On the other hand, interference cancellation encompasses the process of reconstructing the echo from the known transmitted signal, subsequently eliminating it from the received signal to achieve SI reduction [61, 80–82].

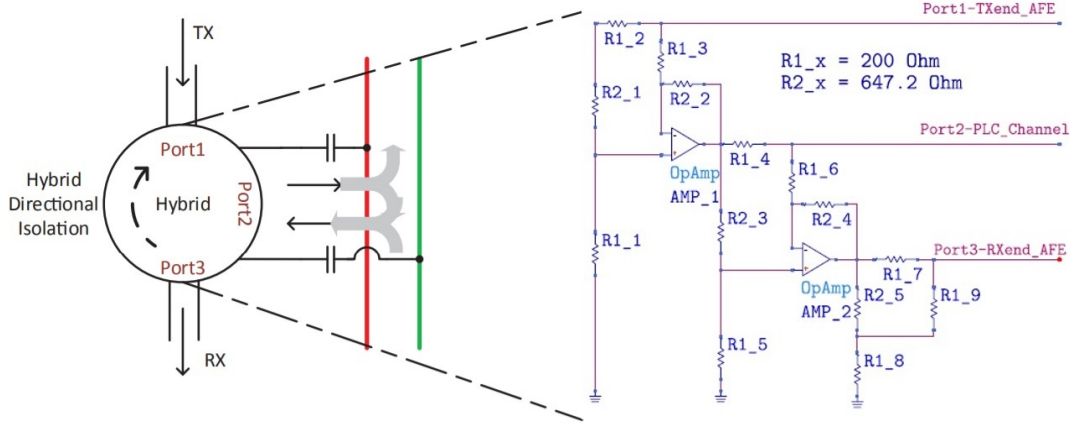


Figure 3.7: Echo suppression in PLC where transmission and reception are coupled on the same conductor pair showing the hybrid circuit used [61]

A hybrid-based isolation technique illustrated in Figure 3.7 is used, to enable comprehensive utilization of all power-line conductors for MIMO IBFD communication. The hybrid-based isolation aims to achieve a suitable level of SI reduction, a critical factor in enabling successful IBFD communication. If line impedance matching is established, the hybrid-based isolation is designed to attain the necessary 80 dB SI reduction within the transmit signal level [80]. However, achieving impedance matching within PLC is challenging due to the variability and unknown nature of the access impedance. Consequently, to ensure effective SI reduction, interference cancellation becomes essential. In the context of interference cancellation, the well-known transmit signal is utilized to replicate the echo that interferes with the incoming SOI in the digital domain, employing a technique referred to as digital interference cancellation (D-DIC). This involves tapping the transmitted signal and passing it through an finite impulse response (FIR) filter, which effectively mimics the SI channel, enabling the reproduction of the echo. Given the dynamic nature of the PLC channel, where conditions change rapidly, the adoption of D-DIC is advantageous. This technique taps the transmit (Tx) signal and digitally filters it, leading to enhanced performance efficiency [80].

3.3 PLC Front-haul System Model

The CPRI/ eCPRI signal received at DU over front-haul (II) is affected by the PLC channel and can be modelled as:

$$\mathbf{y} = \mathbf{H}_{\text{PLC}}\tilde{\mathbf{s}} + \mathbf{n}_{\text{bg}} + \mathbf{n}_{\text{imp}} \quad (3.1)$$

where $\mathbf{H}_{\text{PLC}} \in \mathbb{C}^{N_p \times N_p}$ is the block PLC MIMO channel over the three-phase line, N_p is the total number of differential ports over the three-phase power line and $\tilde{\mathbf{s}} \in \mathbb{C}^{N_p \times 1}$ is the CPRI/ eCPRI HF from different RUs to DU in uplink direction. The CPRI/ eCPRI HFs are subject to the presence of both PLC background noise $\mathbf{n}_{\text{bg}}$ and impulsive noises $\mathbf{n}_{\text{imp}}$, both of which are elements of the complex vector space $\mathbb{C}^{N_p \times 1}$. By considering a balanced three-phase line with symmetrical components, it's reasonable to disregard the mutual coupling among the three primary phase lines, as outlined in [83]. Additionally, the shunt admittance can be excluded from consideration, as their contribution remains below 5% for power lines shorter than 80 km [84]. Hence, the interference between distinct single-phase lines can be reasonably overlooked when compared to the interference occurring among different ports of the same single-phase line. Given this premise, the matrix $\mathbf{H}_{\text{PLC}}$ can be represented as a block diagonal matrix as follows:

$$\mathbf{H}_{\text{PLC}} = \begin{bmatrix} \mathbf{H}_{1\text{P}_1} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_{1\text{P}_2} & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{H}_{1\text{P}_3} \end{bmatrix} \quad (3.2)$$

where, $\mathbf{H}_{1\text{P}_i} \in \mathbb{C}^{N_{p_i} \times N_{p_i}}$ is the MIMO PLC channel over i -th single-phase line, and N_{p_i} is the number of ports in the same single-phase line, yielding the total number of ports over three-phase lines as $N_p = \sum_{i=1}^3 N_{p_i}$. A single-phase MIMO PLC channel with 2 ports used for data transmission can be expressed as:

$$\mathbf{H}_{1\text{P}_1} = \begin{bmatrix} P_1N - P_1N & P_1N - P_1PE \\ P_1PE - P_1N & P_1PE - P_1PE \end{bmatrix} \quad (3.3)$$

where P_1N is the differential signal between 1-st (P1) phase and the neutral (N) conductors, P_1PE is the differential signal between 1-st phase and the protective earth (PE) conductors. The diagonal elements represent the co-channel coefficients, and the off-diagonal elements represent the cross-channel among the different ports of the single-phase line as shown in Figure 3.8.

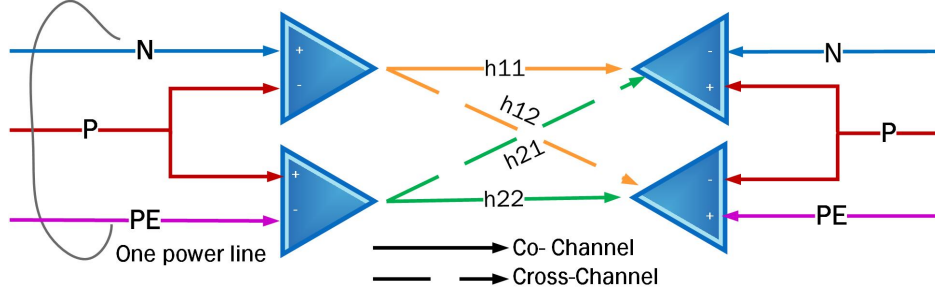


Figure 3.8: Single Phase MIMO PLC channel

In our study, we employ the Information and communication technology (ICT) OMEGA SISO PLC channel model introduced in the preceding chapter, extending its applicability to the frequency range of 1 - 300 MHz. To facilitate MIMO communication, we further extend this SISO PLC channel model to encompass the characteristics of a single-phase PLC setup, as detailed in [58]. Noise within indoor PLC systems takes the form of five distinct types, classified into background noise and impulsive noise categories [44]. The background noise is represented as coloured Gaussian noise, with worst-case scenario parameters considered as outlined in the previous section [44].

3.4 Proposed Front-haul using CPG

The introduced CPG constitutes a novel element designed to facilitate the utilization of the power line as the front-haul link while maintaining complete transparency to the CPRI protocol interactions between the DU and RU. The core objectives of the CPG include:

1. Reduction of BER through the implementation of SVD processing and IND, working collaboratively with HARQ.

2. Execution of rate matching to ensure that the effective transmission rate across the power lines is not lower than the rate of the CPRI transmission.

3.4.1 Singular Value Decomposition (SVD) Processing

With the multiple transmit/receive ports of the PLC system, the space dimension can be exploited using spatially multiplexing of several data streams onto the PLC MIMO channel, leading to increased capacity. Assuming perfect PLC channel estimation, SVD is applied on the MIMO PLC channel such that $\mathbf{H}_{\text{PLC}} = \mathbf{U}\mathbf{\Sigma}\mathbf{V}^\dagger$, where the $\dagger$ symbol indicates the Hermitian operator of a matrix. The columns $\mathbf{U}$ and $\mathbf{V}$ consist of left and right singular vectors of $\mathbf{H}_{\text{PLC}}$, respectively, and $\mathbf{\Sigma}$ is a diagonal matrix whose diagonal entries are the singular values of $\mathbf{H}_{\text{PLC}}$ [85]. The HF are pre-processed through multiplication by $\mathbf{V}$ at the transmitter side such that $\tilde{\mathbf{s}} = \mathbf{V}\mathbf{s}$ and post-processed through multiplication by $\mathbf{U}^\dagger$ at the receiver side such that $\tilde{\mathbf{y}} = \mathbf{U}^\dagger\mathbf{y}$. The received CPRI/ eCPRI signal $\tilde{\mathbf{y}}$ after the post-processing can be expressed as:

$$\begin{aligned}\tilde{\mathbf{y}} &= \mathbf{U}^\dagger\mathbf{y} = \mathbf{U}^\dagger\mathbf{U}\mathbf{\Sigma}\mathbf{V}^\dagger\tilde{\mathbf{s}} + \mathbf{U}^\dagger\mathbf{n}_{\text{bg}} + \mathbf{U}^\dagger\mathbf{n}_{\text{imp}} \\ &= \mathbf{U}^\dagger\mathbf{U}\mathbf{\Sigma}\mathbf{V}^\dagger\mathbf{V}\mathbf{s} + \mathbf{U}^\dagger\mathbf{n}_{\text{bg}} + \mathbf{U}^\dagger\mathbf{n}_{\text{imp}} = \mathbf{\Sigma}\mathbf{s} + \widetilde{\mathbf{n}_{\text{bg}}} + \widetilde{\mathbf{n}_{\text{imp}}}\end{aligned}\tag{3.4}$$

3.4.2 Impulsive Noise Detection & Re-transmission (IND)

Given the significant strength of impulsive noise, which surpasses background noise by several magnitudes within a time span of 4 to 100 microseconds, it can lead to the corruption of one or more HFs even when SVD processing is applied. This results in the introduction of bit errors in the data plane, causing notable delays and imposing additional processing overhead.

To counteract the detrimental impact of impulsive noise, IND is strategically employed to recover lost HFs by means of lightweight re-transmissions following the detection of impulsive noise events. Specifically, the proposed IND is orchestrated through the cooperation of transmitter and receiver CPGs, namely D-CPG and R-CPG. These components consistently

monitor the PAPR, computed as $10 \log_{10}(P_s/P_{mean})$, over each HF duration or an individual orthogonal frequency division multiplexing (OFDM) symbol period ($\Delta t = 66.67 \mu s$). Here, P_s signifies the peak power of the observed signal across the PLC link, and P_{mean} denotes the stored average power sensed over the same link. Whenever the PAPR surpasses a specific threshold ($Th = 10dB$, selected based on a background-to-impulsive noise ratio of $\Gamma = 0.1$), this signals the occurrence of an impulsive noise event. Consequently, as shown in Algorithm 3.1 below, the transmitter halts its transmission, while the receiver discards the presently received HF.

Upon sensing the impulsive noise subsiding below the established threshold Th , the transmitter triggers the re-transmission of the HF that was held back due to the impulsive noise interference. Simultaneously, the receiver advances to process the succeeding received HF, facilitating the resumption of smooth communication operations.

Algorithm 3.1: Pseudo-code for IND

Input: PLC link mean power (P_{mean}), Background to impulsive noise ratio (Γ) and HF duration (Δt)

1. Set $Th = 10 \log_{10}(1/\Gamma)$, $\Delta t = 66.67 \mu s$
2. Measure P_s
3. **While** $10 \log_{10}(P_s/P_{mean}) > Th$, Impulsive event detected **Then Loop**
4. Transmitter: Hold HF transmission
5. Receiver: Discard incoming HF
6. Measure P_s over time = Δt
7. **end While**
8. Transmitter: Transmit held HF
9. Receiver: Accept incoming HF

Output: None

With HARQ, the receiver acknowledges each received CPRI HF in the Vendor Specific Fields of the HF in the reverse direction [73]. The readily present frame check sequence (FCS) based on cyclic redundancy check (CRC) in the Control and Management (C&M) information of the CPRI HF is used for HF error checking [73].

3.4.3 Virtual eCPRI Link by eCPG Pair

To enable the utilization of PLC-based front-haul as the final link, a comprehensive set of transmission mechanisms is deployed across the PLC link. These mechanisms encompass IND and Selective repeat (SR)-HARQ strategies, designed to counteract the challenges presented by the often-hostile PLC channel environment [40, 47]. This suite of transmission strategies is implemented through a pair of devices situated at both ends of the PLC link, referred to as ECPG. Specifically, the RU-attached eCPG is termed R-eCPG, while the one connected to the DU is designated as D-eCPG.

Given that eCPRI operates on Ethernet-based principles, the IND functionality can be efficiently realized through the inherent Ethernet collision sensing mechanism. This coordination is managed by the HARQ-IND controller, as depicted in Figure 3.9. This enhancement streamlines the eCPG architecture, while still preserving the CPG functionality for eCPRI data frames that rely on UDP/IP. Analogous to the CPG introduced in [40], the D-eCPG and R-eCPG are easily deployable plug-and-play devices, establishing connections with the DU and RU, respectively.

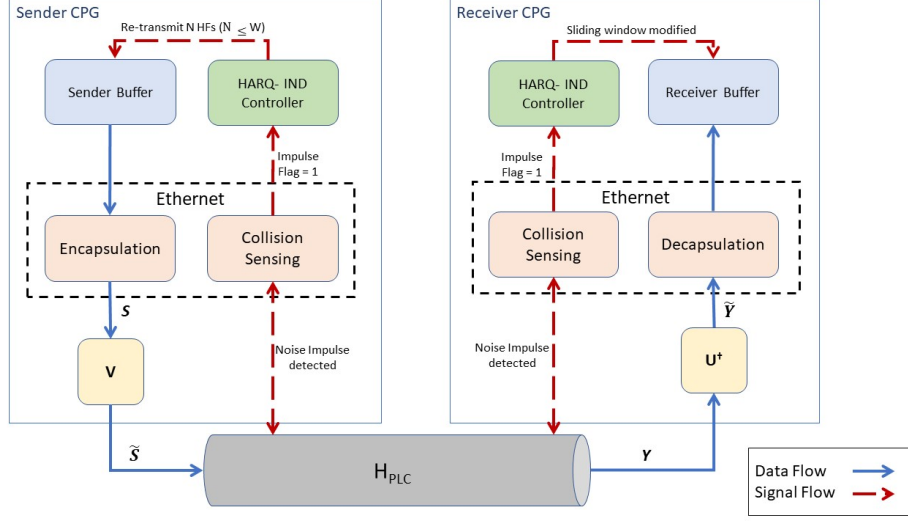


Figure 3.9: Functional block diagram for Ethernet based front-haul CPG

Within this framework, the eCPG pair located at each end of the PLC channel operates as a mediator, creating a virtual eCPRI link seamlessly connecting the DU and RU. This setup enables the transparent transportation of unmodified eCPRI frames between the two endpoints. The core objective of the eCPG is to achieve the desired eCPRI data rates even in the presence of the challenging PLC channel conditions. Crucially, the eCPG maintains complete transparency with regard to eCPRI protocol operations, as the un-encrypted data frames can be detected and processed by standard eCPRI modules.

An inherent advantage of both eCPG and CPG lies in their plug-and-play nature, ensuring complete transparency to eCPRI and CPRI protocol operations, respectively. This transparency stems from the fact that eCPRI frames remain un-encrypted, rendering them detectable by standard eCPRI modules. The proposed eCPG operates by extracting control words from eCPRI data frames, procuring essential information to guide CPG pairs in transporting the frames through the power line channel with requisite QoS.

Upon receiving eCPRI data frames from the RU (or from the DU in the case of downlink), the incoming buffer of the eCPG serves as a repository. The HARQ-IND controller engages in efficient buffer management, ensuring dependable data transfer between the eCPG pair via SR HARQ. In the reverse direction, the DU acknowledges each received data frame using vendor-specific reserved bits embedded in the eCPRI frame header. To detect impulse

events, the Ethernet collision sensing module interprets voltage levels on the PLC link, effectively emulating collisions. This process involves comparing the sensed voltages against a predetermined threshold, with values surpassing the threshold signifying an impulse event. Detection of such events triggers a notification to the controller [86].

For error checking, the readily available FCS based on CRC, embedded in the header of the eCPRI Ethernet frame, is employed [74, 74]. To counteract PLC cross-channel interference, a pre-coding function relies on modules V and $U^\dagger$ for transmit and receive signals, respectively, utilizing the SVD of the detected PLC channel. It is important to note that the term “CPG” is used interchangeably to denote both CPG and eCPG throughout this dissertation. The flowchart depicting the CPG/ eCPG virtual link in the uplink direction is presented in (*Appendix A*)

3.4.4 CPRI Data Rate Matching

To make the PLC link transparent to the DU and RUs, a critical requirement is that the PLC data rate should be larger than the corresponding CPRI option of the current CPRI stream. Nonetheless, the effective data rate along the PLC line is strongly affected by the re-transmissions of the lost HFs, which is in turn determined by the FER. In the rest of this section, we provide some analysis for finding the required system data rate to meet a given CPRI option. This will enable us to create a lookup table which matches the CPRI different options data rates with the effective data rate required over the PLC system.

We firstly estimate the average BER $\overline{P}_b$ of the PLC line by integrating the BER in AWGN for polar non-return to zero (P-NRZ), defined as $P_b = Q(\sqrt{2\gamma})$ over the log-normal distribution of the PLC channel $f_H(h)$ such that [87]:

$$\overline{P}_b = \int_0^\infty P_b f_H(h) dh = \int_0^\infty Q(\sqrt{2\gamma}) \frac{e^{-(\ln(h)-\mu_p)^2/(2\sigma_p^2)}}{h\sigma_p\sqrt{2\pi}} dh \quad (3.5)$$

where $Q(\cdot)$ is the Gaussian Q-function, $\gamma = |h|^2 E_b / N_0$ is the instantaneous received SNR, E_b is the power per transmit bit and N_0 is the mean PLC noise power. The log-normal parameters of the three-phase PLC channel with length $L_p = 800m$ and $L_p = 100m$, are obtained

through distribution fitting as shown in Figure 3.10 as well as Table 3.1. Subsequently, the average FER over the PLC fading channel $\overline{P_e}$, is computed by integrating the average FER in AWGN channel, $P_e(\gamma) = 1 - (1 - P_b)^{D_p}$, over the log-normal distribution of the average received SNR, $f(\gamma; \bar{\gamma})$ [88]:

$$\overline{P_e} = \int_0^\infty P_e(\gamma) f(\gamma; \bar{\gamma}) d\gamma = \int_0^\infty (1 - (1 - Q(\sqrt{2\gamma}))^{D_p}) \frac{e^{-(\ln(\gamma/\bar{\gamma}) - \mu_p)^2 / (2\sigma_p^2)}}{2\gamma\sigma_p\sqrt{2\pi}} d\gamma \quad (3.6)$$

where $\bar{\gamma} = \mathbb{E}[|h|^2]E_b/N_0$ is the average received SNR with $\mathbb{E}[|h|^2] = e^{(\mu_p + \frac{\sigma_p^2}{2})}$ and $\mathbb{E}[\cdot]$ denotes the expectation operation. The instantaneous received SNR is $\gamma = |h|^2 E_b/N_0$ and D_p is the number of bits in one HF.

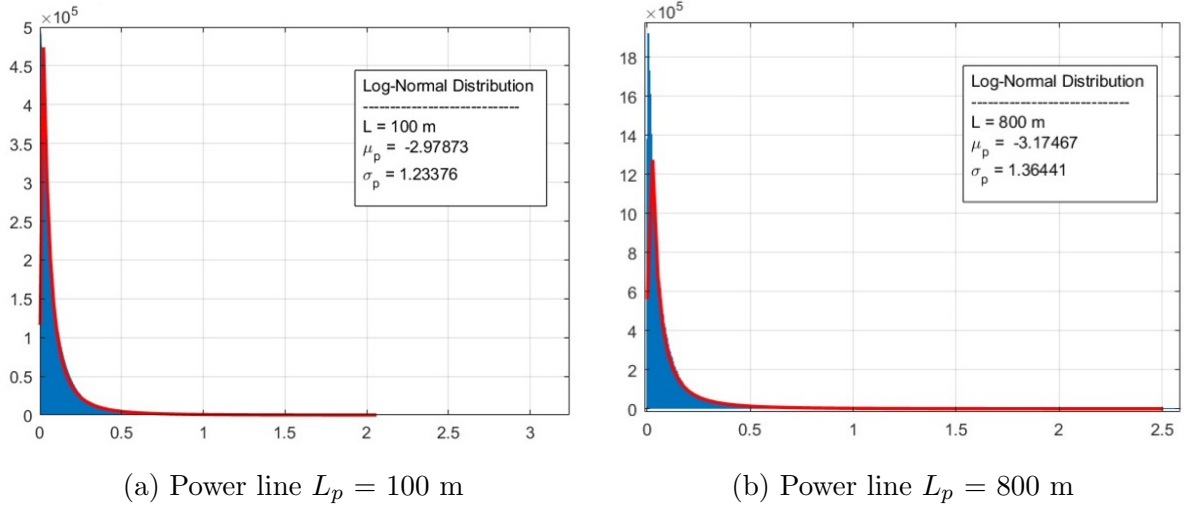


Figure 3.10: Distribution fitting for MIMO PLC channel - Log-normal 1-300 MHz BW

Table 3.1: Log-normal Parameters for MIMO PLC Channel

PLC Ch. Log-normal Parameters		
Power line Length	Mean (μ_p)	St. deviation (σ_p)
$L_p = 100 \text{ m}$	-2.97873	1.23376
$L_p = 800 \text{ m}$	-3.17467	1.36441

To find the effect of HARQ on the data rate we need to find the effect of the interleaved RS FEC as well as the effect of SR using the maximum window size ($W = 1/2$ CPRI frame size or the duration for one 5G frame (10ms)). When applying RS(n_{rs}, k_{rs}), the improved average FER over the log-normal received SNR can be approximated as [89]:

$$\overline{P_{ers}} = \int_0^\infty P_{w_{rs}} \frac{e^{-(\ln(\gamma/\bar{\gamma}) - \mu_p)^2 / (2\sigma_p^2)}}{2\gamma\sigma_p\sqrt{2\pi}} d\gamma \quad (3.7)$$

where $P_{w_{rs}}$ is the probability of interleaved RS FEC decoder error [90]:

$$P_{w_{rs}} = \sum_{i=t_{rs}+1}^{n_{rs}} \binom{n_{rs}}{i} P_b^i (1 - P_b)^{n_{rs}-i} \quad \& \quad t_{rs} = \frac{(n_{rs} - k_{rs})}{2} \quad (3.8)$$

Using the FER after RS FEC, the effect of SR based HARQ over the data rate with L_r re-transmissions is defined as[91]:

$$\eta_H = \frac{1}{1 + \mathbb{E}[L_r]} = (1 - \overline{P_{ers}})(1 - \overline{P_{ers}}) \quad (3.9)$$

This is irrespective of the number L_r as only one frame copy is transmitted at a time and each frame is acknowledged by one acknowledgment (ACK). The probability of loss of ACK is the same as FER because it is part of the frame header.

The effect of IND over the data rate is calculated as:

$$\eta_I = 1 - P_{imp} \quad (3.10)$$

where P_{imp} is the probability that an impulse is detected is defined as $P_{imp} = \mathbb{P}[A_{imp} > \sqrt{P_{mean}/\Gamma}] \times \mathbb{P}[t_i \leq \Delta t]$, where A_{imp} is the amplitude of the impulsive noise and t_i is the inter-arrival time between the impulses. Based on the distribution models mentioned in section II.C P_{imp} can be expressed as [44]:

$$P_{imp} = \left(1 - \int_0^{A_{imp}} \sum_{z=0}^{\infty} \frac{e^{-A} A^z}{z!} \frac{e^{-x^2/2\sigma_z^2}}{\sqrt{2\pi\sigma_z^2}} dx\right) \times (1 - e^{-\Delta t/\mu_i}) \quad \text{with } \sigma_z^2 = \sigma^2 \left(\frac{(z/A) + \Gamma}{1 + \Gamma}\right) \quad (3.11)$$

Consequently, the required matched effective data rate (R_{eff}) along the PLC line corresponding to the required CPRI option (R_{opt}) can be expressed as:

$$R_{eff} = R_{opt}/(\eta_{rs} \times \eta_H \times \eta_I) \quad (bps) \quad (3.12)$$

where $\eta_{rs} = k_{rs}/n_{rs}$ is the overhead introduced due to RS(n_{rs}, k_{rs}) FEC. Using the same methodology the effective system throughput is calculated as:

$$Thrpt = C_{MIMO} \times \eta_{rs} \times \eta_H \times \eta_I \quad (bps) \quad (3.13)$$

where C_{MIMO} is the instantaneous capacity of the PLC MIMO channel.

3.5 Simulation Outcomes

In order to simulate our system, first a link level simulator (LLS) of the physical layer signal over the PLC link was created using MATLAB. Then the outcomes of the LLS were used into a MATLAB based system level simulator (SLS) to simulate the full functioning system including HARQ and IND. Figure 3.11 illustrates the block diagram of the simulator created.

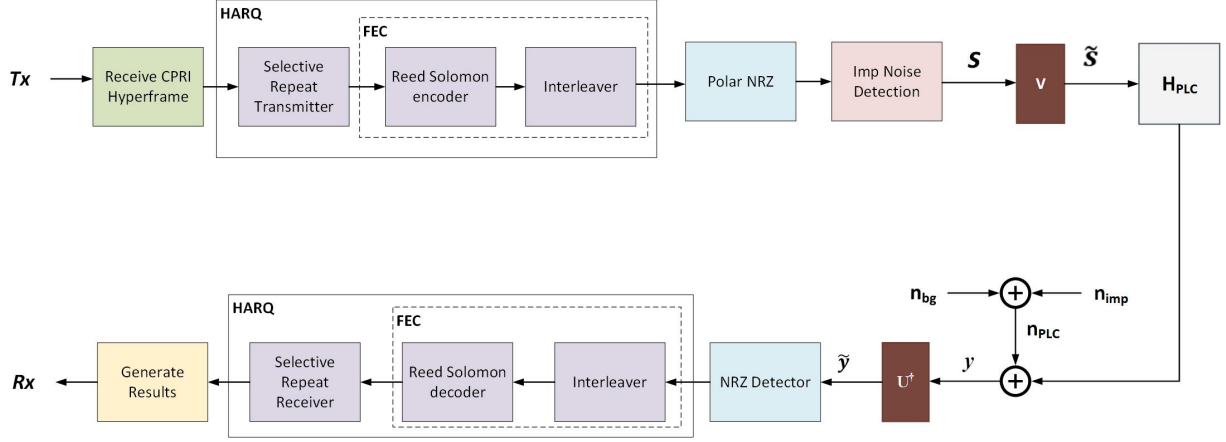


Figure 3.11: Link Level simulator block diagram

Four scenarios are simulated:

Sc1: HFs Only: Sending HFs without any modifications onto the PLC lines.

Sc2: HFs + HARQ: HFs are sent over the power lines while enabling the HARQ type-I technique.

Sc3: HFs + IND: HFs are sent onto the power lines while applying IND.

Sc4: HFs + HARQ+ IND: HFs are sent onto the PLC lines while enabling HARQ type-I and IND.

In all scenarios, polar non-return to zero (P-NRZ) is applied. In the developed system the regulations of the general Electro-Magnetic Compatibility (EMC) have been respected. In which the PSD is set to -50dbm/Hz up to 30 MHz and to -80dbm/Hz for higher frequencies as shown in Figure 3.12 [92].

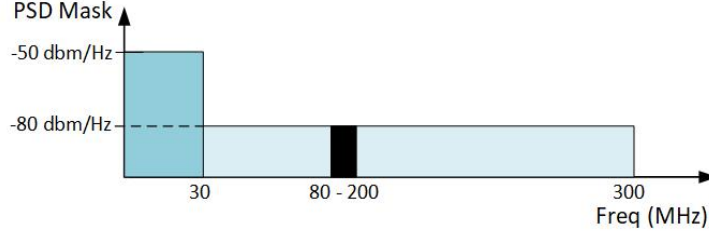


Figure 3.12: PSD Mask for extended bandwidth up to 300MHz

The main evaluating metric for our system is the BER and FER. In eq. (3.14), the instantaneous capacity (C_{MIMO}) of the PLC MIMO channel is calculated [79].

$$C_{MIMO} = \sum_{n=1}^{N_f} \Delta f \sum_{i=1}^{N_p} \log_2 \left(1 + \frac{P_{T_x}(f_n) \lambda_p(f_n)}{N_{R_x}(f_n) N_p} \right) \quad (bps) \quad (3.14)$$

where $\lambda_i(f_n)$ is the eigenvalues of $\mathbf{H}_{PLC} \mathbf{H}_{PLC}^\dagger$ at a given frequency f_n . $P_{T_x}(f_n)$ is the transmitted power over the PLC link and $N_{R_x}(f_n)$ is the noise at the receiver of the PLC front-haul. N_f is the number of samples in the frequency range, Δf is the sampling frequency used (frequency step) and N_p is the number of transmit/receive ports used [79].

The simulation parameters utilized are detailed in Table 3.2. It's noteworthy that adherence to the regulations of general EMC has been ensured, with the maximum PSD fixed at -50 dBm/Hz up to 30 MHz and lowered to -80 dBm/Hz for higher frequencies, in line with established guidelines [92]. The simulation encompasses a bandwidth that extends to encompass the very high-frequency range ($\leq 300MHz$). Moreover, to maintain a conservative stance, the maximum length of a PLC system recorded in the literature (800 m) has been adopted, alongside the length of power lines within a multi-story service building that rarely exceeds 100 m.

Table 3.2: System simulation parameters

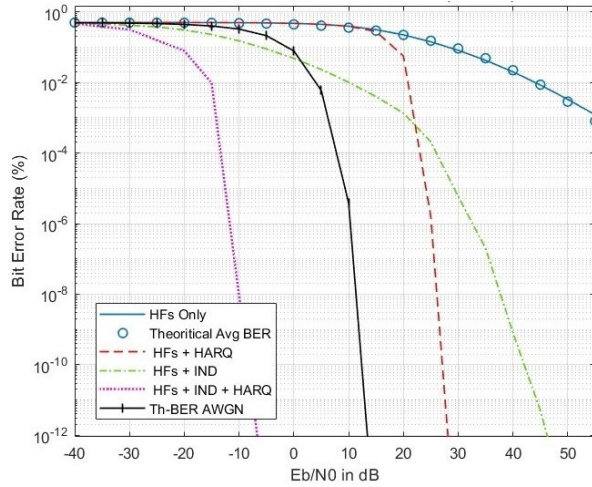
Parameter	Value
E_b/N_0 range	-20 - 50 dB
Power Line Length	100 m and 800 m
MIMO PLC Ports (3-Phase)	6 Tx/RX ports
PLC BW / Port	1 - 300 MHz
Chip Rate (f_c)	3.84 MHz
HARQ limit/frame (L_r)	4 re-transmission
RS-FEC rate (n_{rs}, k_{rs})	(127, 121)
Inter-leaver rate	32
Selective Repeat Window (w)	75 Hyper frame
IND Threshold (Th)	10 dB
IND Mean power (P_{mean})	-115 dB
CPRI Hyper frame size	256 basic frames
CPRI Hyper frame duration	66.67 μs

Figure 3.13 presents the BER of the system, while Figure 3.14 displays the HF FER for all four scenarios. When the CPRI/eCPRI signal is directly transmitted over the PLC-based front-haul without modifications, both BER and FER exhibit high values, even for substantial SNRs. Thus, implementing the system as is proves infeasible. The introduction of HARQ significantly enhances system performance, with a set limit of four re-transmissions per HF. For SNR values surpassing 20 dB, the residual bit error rate records zero with a simulated bit count of 10^{13} (*i.e.*, $BER < 10^{-12}$). In regard to eCPRI the one-way frame loss ratio requirement (R-26) *i.e.* $FER < 10^{-7}$ is achieved with the proposed system even for very low SNR values. Relative to the unmodified CPRI scenario, the integration of IND

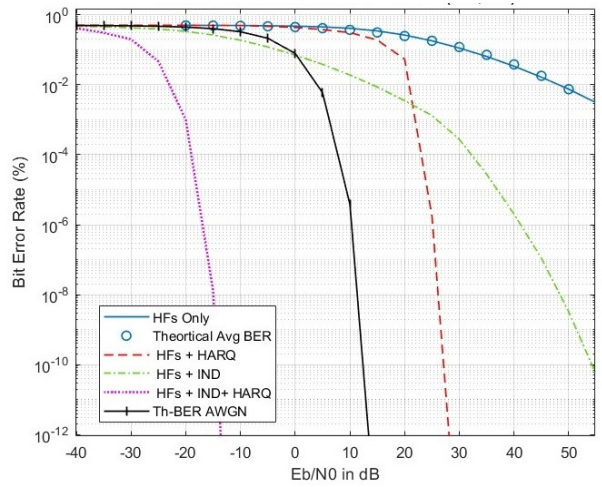
distinctly improves both BER and FER. With IND in play, and for SNR values exceeding 40 dB, the error rate for received bits dwindles to zero without any re-transmissions through higher layers. Remarkably, through the employment of both HARQ and IND, it's evident that residual BER and FER of zero can be achieved for $SNR > -10$ dB. These results are derived from simulations comprising 10^{13} bits and 10^8 HF's simulated per SNR value. For a confidence level of 95%, the BER confidence intervals for the simulated scenarios at SNR of 20dB are shown in Table 3.3 below. For scenario 4, both BER and FER recorded zero value for the simulated number of bits and frames.

Table 3.3: Confidence Interval for BER and FER for all scenarios at SNR = 20dB and confidence level 95%

Scenario	BER	FER
Sc. 1	$0.2248 \pm 1.53e^{-3}$	$0.9620 \pm 2.12e^{-3}$
Sc. 2	$0.0562 \pm 3.83e^{-4}$	$0.2406 \pm 5.3e^{-4}$
Sc. 3	$0.0014 \pm 2.79e^{-5}$	$0.1181 \pm 1.68e^{-5}$
Sc. 4	$< 10^{-12} \pm 0$	$< 10^{-7} \pm 0$

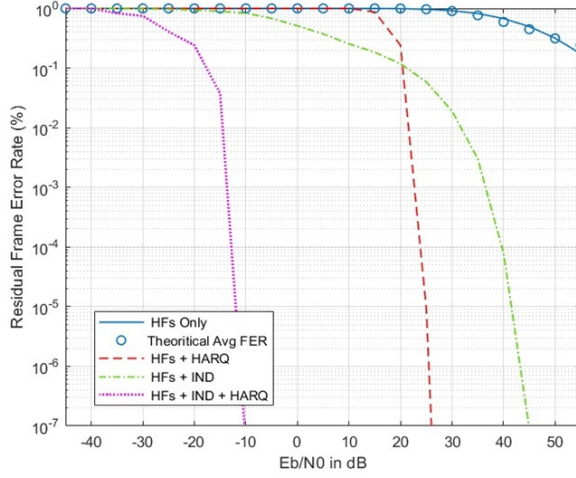


(a) Power line Length ($L_p = 100\text{m}$)

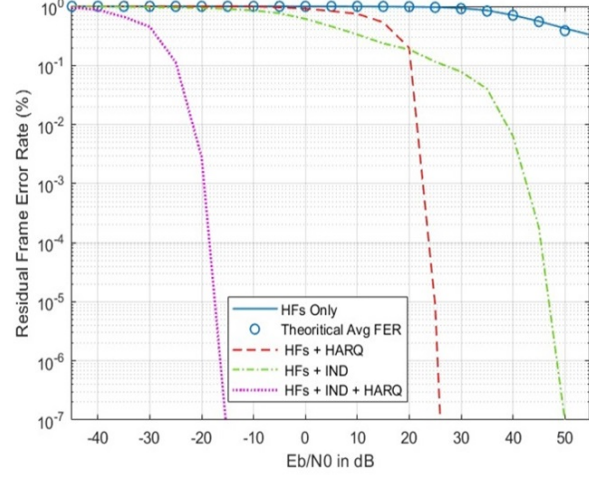


(b) Power line Length ($L_p = 800\text{m}$)

Figure 3.13: Simulations outcomes of Bit error rate (BER) for PLC

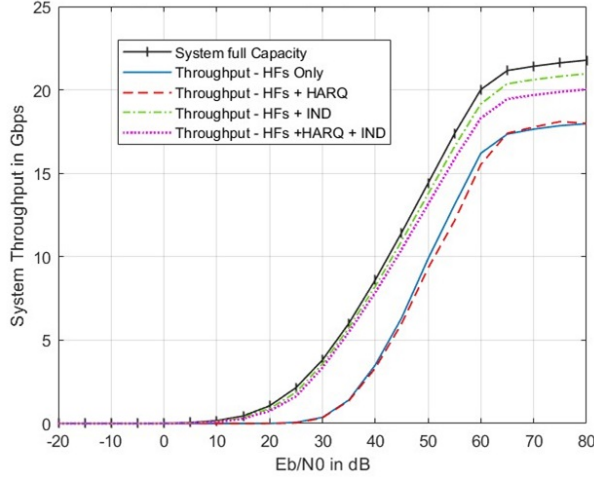


(a) Power line Length ($L_p = 100\text{m}$)

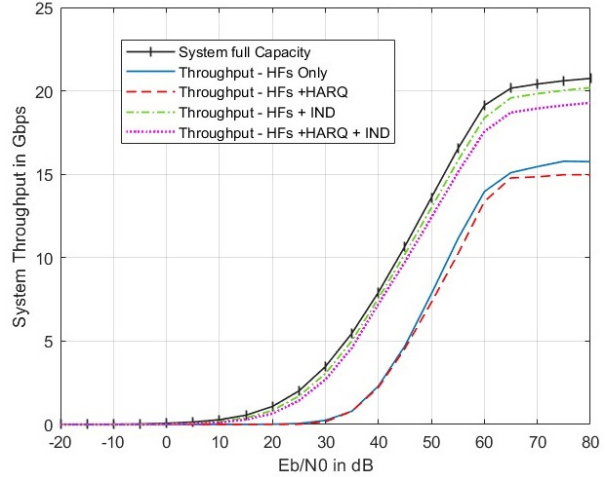


(b) Power line Length ($L_p = 800\text{m}$)

Figure 3.14: Simulations outcomes of Frame error rate (FER) per hyper-frame for PLC



(a) Power line Length ($L_p = 100\text{m}$)



(b) Power line Length ($L_p = 800\text{m}$)

Figure 3.15: Net throughput outcomes of MIMO PLC (6x6) vs. E_b/N_0

Each of the four scenarios was executed 100 times per port, and the resultant mean and maximum capacity values, along with the effective throughput at $E_b/N_0 = 35$ dB, have been tabulated in Table 3.4 and visualized in Figure 3.15. It's worth noting that even higher data rates can be attained by expanding the bandwidth into the very high-frequency range (> 300 MHz) [93]. The effective throughput, observed when both HARQ and IND are employed, constitutes a significant portion, ranging from approximately 85% to 90%,

of the system's full capacity. This is valid across different power-line lengths, varying from $L_p = 800$ m to 100 m.

Table 3.4: Effective Throughput outcomes for $E_b/N_0 = 35dB$
MIMO PLC (6×6), Same Circuit (Gbps)

Length (L_p)	Outcomes	Sc. 1	Sc. 2	Sc. 3	Sc. 4
800 m	Mean Sys. Cap.	5.471			
	Max Sys. Cap	12.765			
	Mean Thrpt	0.797	0.7933	5.016	4.6
	Max Thrpt	1.8582	0.8305	11.705	11.2354
100 m	Mean Sys. Cap.	6.003			
	Max Sys. Cap	13.2			
	Mean Thrpt	1.406	1.358	5.734	5.4610
	Max Thrpt	3.091	2.986	12.57	11.97

3.6 Chapter Summary

This chapter delves into the exploration of leveraging indoor LV PLC as the final segment of the front-haul for indoor users. The focus is on seamlessly transporting unmodified TDM CPRI HFs and Ethernet-based eCPRI frames between the outdoor DU and the indoor RU. The core challenge lies in mitigating the challenging power-line environment, and to address this, we introduce a groundbreaking network component called the CPG/eCPG. These components function as intermediaries, creating virtual CPRI/eCPRI links between the DU and RU. The central idea behind the CPG is to facilitate the realization of desired CPRI options even in the presence of the noisy PLC channel, all while maintaining complete transparency to the underlying CPRI protocol operations. Through comprehensive simulations, the capabilities and effectiveness of the CPG are thoroughly demonstrated, showcasing its potential to revolutionize power-line communication in the context of front-haul.

Chapter 4

Performance Analysis of PLC Front-haul for IoT Indoor Communications

In this chapter, the main aim is to extend on the earlier system design by constructing a thorough mathematical model based on the system presented in Chapter 4 [40]. The model defines RAN end-to-end (e2e) delay performance using queuing theory and stochastic geometry. The investigation includes wireless radio interface delay and the new PLC-based front-haul system. The number of wireless IoT devices, number of installed RUs, and PLC-based front-haul bandwidth and transmission technologies determine the e2e uplink delay. The model can be easily modified to include downlink data traffic e2e latency inside the same system framework.

In this chapter, an analytic framework on e2e delay performance of the proposed indoor front-haul solution, including radio coverage probability, number of served devices, and system delay is introduced. Furthermore, we shall derive closed-form expressions for the e2e delay performance according to a number of key system parameters, including the device density, transmission power, bandwidth on the radio and PLC channels, and the maximum number of devices supported by the proposed PLC-based front-haul solution. Moreover, the analytical models are validated through simulations, demonstrating that the proposed indoor front-haul solution meets the stringent 5G and eCPRI delay requirements.

4.1 Radio Access Network (RAN) Architecture

In an indoor multi-story service building environment, a massive number of wireless IoT devices are randomly distributed. The IoT devices are used to collect error- and delay-sensitive data, to be transmitted to a number of service owners. A split C-RAN system is considered in the network, in which the BS is divided into three units: CU, DU, and RU. As shown in Figure 4.1, the FH link is divided into 2 segments: a Mid-haul/front-haul (I) segment that connects the CU and DU using optical fibre, and a front-haul (II) segment that connects the DU with different indoor RUs. The technology used over front-haul (I) is out of the scope of this research.

To support the IoT devices, a number of RUs are randomly distributed across the building. Each RU is connected to a power socket that provides power as well as access to the PLC front-haul. The front-haul (II) segment carries CPRI or eCPRI messages frames to and from the indoor RUs. Throughout the rest of the chapter, CPRI or eCPRI messages are referred to simply as HFs unless otherwise indicated.

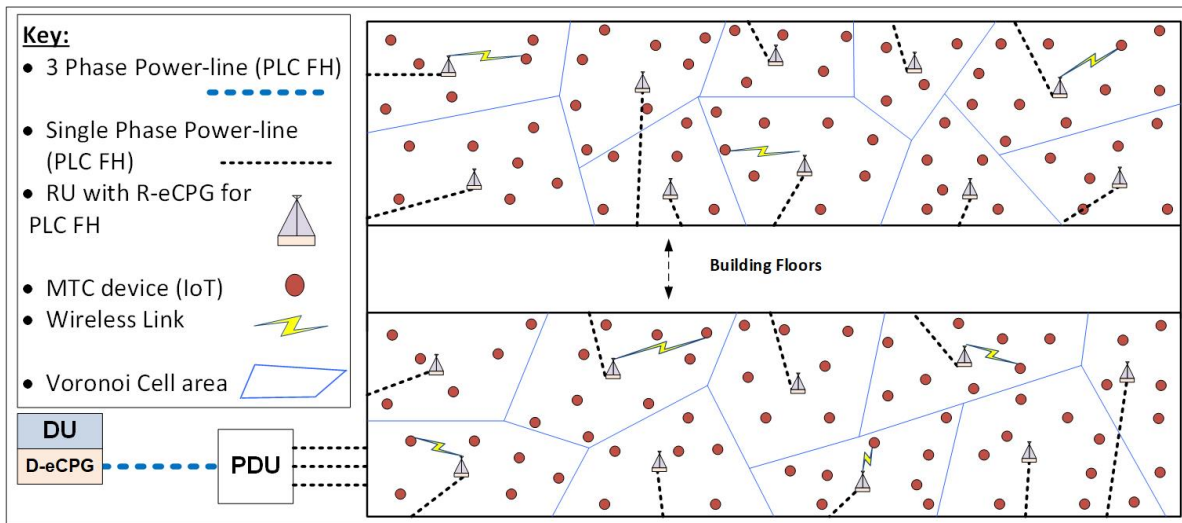


Figure 4.1: Radio access network architecture

The front-haul transmission technology considered over the PLC link is eCPRI option 7-2, namely Splits II_D & I_U , which are IQ-oriented splits (downlink and uplink) and are the most demanding splits in terms of front-haul throughput and latency. As for the air

interface, a 5G OFDM wireless radio channel is considered, where the service band is shared among all the IoT devices, hence, devices associated with different RUs and scheduled on the same sub-channel experience inter-cell interference.

4.1.1 Wireless Spatial Model

In the following analysis, we focus on evaluating the system performance for a single story of the building at a time, assuming no inter-story interference. Let's consider a layer of cellular RUs connected to a DU then to the CU of the BS, where the locations of the RUs within the building can be represented by a Homogeneous Poisson point process (HPPP) denoted as $\Phi_b = \{b_0, b_1, b_2, \dots\}$. The RU density per unit area is denoted as λ_b , and b_j represents the 2-D location of the j^{th} RU. The network serves a large number (M_{tot}) of identical IoT devices that are randomly distributed across the specific story of the building under consideration. The locations of these devices are modelled as an independent HPPP denoted as $\Phi_d = \{d_0, d_1, d_2, \dots\}$, with a device density per unit area of λ_d . Here, d_i represents the 2-D location of the i^{th} device in the network. The IoT devices can be either in an inactive or active state, depending on data availability for transmission.

Each active IoT device generates s according to a Poisson process with a parameter $\gamma_d \in (0, \infty)$. The size of each data-frame transmitted over the air interface is $\mathcal{D}_a$ bits. As our primary focus is on active devices, it is important to model their locations. Using the concept of "Independent Thinning" from stochastic geometry, we can model the locations of the active devices as a thinned independent HPPP derived from the original HPPP representing all devices in the network. We denote the set of active device locations as $\Phi_a \subseteq \Phi_d$, which follows an independent HPPP with a density of $\lambda_a = \mathcal{A}^t \times \lambda_d$, where $\mathcal{A}^t$ represents the probability that at least one data-frame is generated during the t^{th} TR.

Once a device is active, it attempts to associate with a RU based on the maximum received signal strength (max-RSS) association rule [94, 95]. Devices associate with the RU to form a cluster. Each RU is responsible for relaying the data-frames generated by the devices in its cluster to the DUs over the PLC-based front-haul, which are then transmitted to the core network. However, as each IoT device produces a small-sized data-frame, it

would be a waste of front-haul resources to adopt a per-data-frame relay framework. Thus, to improve the efficiency at the FH, in this work, the assumption is that each RU has the capability to aggregate the small-sized data-frames generated from its associated IoT devices and convert them into the appropriate eCPRI transmission format before sending them over the PLC front-haul.

4.1.2 Wireless Transmission Model

Orthogonal frequency division multiple access (OFDMA) is used over the air interface channel at the RUs, where a shared uplink channel (SUCH) of a total bandwidth of Ω_a and is divided into $\mathcal{N}_a$ equal bandwidth orthogonal sub-channels. Let time be divided into TR, which include the time for one 5G sub-frame $T_a = 1ms$, the constant processing time at the RU to convert the data-frame into HF set as $T_{proc} = 10\mu s$ and the time required to send the HF over the PLC front-haul T_{PLC} which is defined later [96]. Figure 1 depicts the transmission round structure. Furthermore, assume that at the beginning of each TR, all active IoT devices associated with the j^{th} RU will have one data-frame to transmit. For every TR, the RU converts the received data from all the devices within its coverage into CPRI or eCPRI HFs and launches the HFs to the DU through the PLC-based front-haul.

Allow M_j^t denote the number of devices in the j^{th} cluster (i.e., associated with the j^{th} RU) at the t^{th} TR, where $\sum_j M_j^t \leq M_{tot}$. It should be noted that M_j^t is a RV because it is affected by the association conditions, the locations of both devices and RUs, their densities, and the transmission power. Each RU schedules its associated devices across the sub-channels until one device is scheduled on each sub-channel. Let M_{max} define the maximum number of devices that a RU can schedule in a single TR and is limited such that $M_{max} = \min\{M_{a_j}^t, \mathcal{N}_a\}$. $M_{a_j}^t$ denotes the number of active devices in the j^{th} cluster at the t^{th} TR and $M_{a_j}^t \leq M_j^t$.

Due to the inter-cell interference among devices in different clusters and scheduled on the same sub-channel, not every scheduled device is able to successfully transmit its data-frame to its serving RU at the current TR. Devices that are able to transmit are said to be in coverage, whereas devices with failed transmissions are said to be in blockage. A data-frame

that fails to be transmitted in the current TR is dropped. As such, let $M_{c_j}^t$ define the number of scheduled devices associated with the j^{th} RU that are in coverage and can successfully transmit their data in the t^{th} TR, where $M_{c_j}^t \leq M_{max}$.

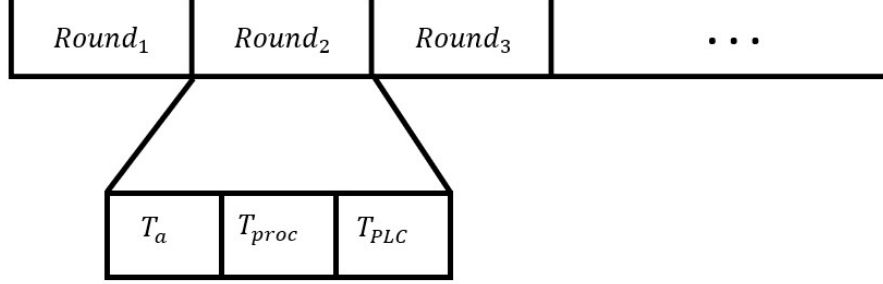


Figure 4.2: TR structure

A general power-law path-loss model describes how transmitted signals over the air interface are attenuated during propagation. The rate of signal power decay is given by $D^{-\alpha_a}$, where D is the propagation distance and α_a is the exponent of the path-loss. Consider a Rayleigh fading radio channel with an instantaneous power gain, g_{ij} , which follows an exponential distribution with unity mean (i.e., $g_{ij} \sim \exp\{1\}$). The radio channel also introduces additive white Gaussian noise with received noise power σ_a^2 . Radio channel gains are uniformly distributed, independent of distance, and independent of one another (Independent and identical distribution (i.i.d)). Due to the vast number of devices, the radio network is limited by interference. When broadcasting their payload over the scheduled uplink sub-channel, devices are deemed to be using FPC. As such, consider a single RU located at the origin and an associated IoT device that is located at d_i^t . Therefore, the power received at the RU from the device after FPC can be calculated as $q_i = p_i ||d_i^t||^{(\epsilon_a - 1)\alpha_a} g_{i0}$, where p_i is the nominal transmission power of any connected device over the air interface, $||d_i^t||$ is the Euclidean distance between the i^{th} device and its serving RU during the t^{th} TR, and $\epsilon_a \in \{0, 1\}$ is the power control factor.

4.2 Performance modelling

The section describes our analysis and performance modelling approaches for RAN network of the proposed PLC based front-haul system.

4.2.1 Coverage Probability over Wireless Link

Coverage probability describes the probability that a data-frame from an IoT device is successfully received by its serving RU. Successful reception of a data-frame is a function of the signal to interference and noise ratio (SINR), where, for the data-frame to be correctly decoded at the RU, it must be received with a SINR greater than a certain threshold (τ_a). The number of devices in each cell is random, as are the quantity and distribution of active devices within the network at the start of each TR. However, it is presumed that at the start of each TR, each RU will have an equal number of active devices situated inside its Voronoi cell. Although this assumption is controversial, it will help in simplifying the Laplace transforms. The effect on the accuracy of these assumptions is verified through simulations.

4.2.1.1 Scheduling Probability Calculations

In what follows, concentrating on a single RU (RU_0). The number of active devices associated with RU_0 during the t^{th} TR is denoted by $M_{a_0}^t$. It is a RV with probability mass function (PMF) calculated by approximating the probability density function (PDF), over a Voronoi cell of area $\mathcal{V}_a$, as a generalized gamma distribution and is given by [94, 97]:

$$\mathbb{P}(M_{a_0}^t = m) = \frac{(c_v \cdot \lambda_b)^{c_v} \lambda_a^m \Gamma(m + c)}{\Gamma(c_v) (\lambda_a + c_v \cdot \lambda_b)^{m+c_v} m!}, \quad 0 \leq m < \infty \quad (4.1)$$

where $c_v = (3d + 1)/2 = 7/2$ is a constant defined for the Voronoi tessellation in $\mathbb{R}^2$ with dimensionality $d = 2$ and $\Gamma(c_v) = \int_0^\infty t^{c_v-1} e^{-t} dt$. Thus, the average number of active devices would be calculated as [97, 98] (*see: Appendix B.2*):

$$\overline{M}_a^t = \mathbb{E}_m[M_a^t] \approx \left\lceil \frac{\lambda_a \Gamma(c_v + 1)}{\lambda_b \Gamma(c_v)} \right\rceil \quad (4.2)$$

As RUs are equipped with omni-directional antennas, transmissions from inter-cell interfering devices cause interference at RU_0 . When the RU density is high enough and each RU's Voronoi cell area is sufficiently small, the modelling is valid [94, 97].

Given the number of active devices in the t^{th} TR, $M_{a_0}^t = m$, let $\mathcal{P}_s$ be the conditional scheduling probability. The scheduling probability of an active IoT device is obtained by averaging across the PMF supplied in eq. (4.2) [99–101]:

$$\begin{aligned} \mathcal{P}_s &= \mathbb{E}_{M_{a_0}^t} \left[\frac{\mathcal{N}_a}{\max\{M_{a_0}^t, \mathcal{N}_a\}} \middle| M_{a_0}^t = m \right] \\ &= \sum_{m=0}^{\mathcal{N}_a} \frac{(c_v \cdot \lambda_b)^{c_v} \lambda_a^m \Gamma(m+c)}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{m+c_v} m!} + \sum_{m=\mathcal{N}_a+1}^{\infty} \frac{\mathcal{N}_a}{m} \frac{(c_v \cdot \lambda_b)^{c_v} \lambda_a^m \Gamma(m+c)}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{m+c_v} m!} \end{aligned} \quad (4.3)$$

4.2.1.2 SINR Calculations

The device-RU transmission success probability, i.e. for a device to have successful transmission, the received signal at its serving RU should have SINR above τ_a i.e. lies within the coverage probability range of its serving RU. Although the upcoming analysis is focused on the RU at the origin (with index $j = 0$), it may be applied to any device-RU pair in the network. Now, let $\Theta_0^{n,t}$ denote the received SINR level at RU_0 of the data-frame transmitted by device₀ on the n^{th} sub-channel and during the t^{th} TR and is given by:

$$\Theta_0^{n,t} = \frac{p_i \|d_0^{n,t}\|^{(\epsilon_a-1)\alpha_a} g_{00}}{I_a + \sigma_a^2}, \quad n \in \{1, 2, \dots, \mathcal{N}_a\} \quad (4.4)$$

where σ_a^2 is the received power of the radio channel additive Gaussian noise and I_a is the interference experienced by each device.

Devices scheduled on the same sub-channel as other devices serviced by different RUs within the network encounter inter-cell interference. As a result, the locations of the inter-cell active device (i.e., in different clusters) interfering on the n^{th} sub-channel during the t^{th} TR are represented by an independent thinned Poisson point process (PPP) ($\Phi_a^{n,t} = \{d_1^{n,t}, d_2^{n,t}, \dots\}$) with density $\lambda_a^{n,t} = \lambda_b$. The interference I_a in eq. (4.4) is provided by:

$$I_a = \sum_{d_i^{n,t} \in \Phi_a^{n,t}} P_a ||d_i^{n,t} - b_j||^{\epsilon_a \alpha_a} ||d_i^{n,t}||^{-\alpha_a} g_{i0} \quad (4.5)$$

where the radio channel gain between the i^{th} device and RU₀ is denoted by g_{i0} . The Euclidean distance between the i^{th} inter-cell interfering device and the origin is indicated by $||d_i^{n,t}||$, and that between the i^{th} inter-cell interfering device and its serving RU (the j^{th} RU) is shown by $||d_i^{n,t} - b_j||$. Keep in mind that $||d_0^{n,t}||$, $||d_i^{n,t}||$ and $||d_i^{n,t} - b_j||$ depend on where the device and the RU are located, therefore, they are all random distances. Given the exponential distribution of g_{00} and g_{i0} with unity mean, the planned device-RU coverage probability for the device at $d_0^{n,t}$, which is serviced by the RU₀, may be represented as follows:

$$\begin{aligned} C_0^{n,t} &= \mathcal{P}_s \cdot \mathbb{P}(\Theta_0^{n,t} > \tau_a) \\ &= \mathcal{P}_s \cdot \mathbb{P}\left(\frac{P_a ||d_0^{n,t}||^{(\epsilon_a - 1)\alpha_a} g_{00}}{I_a + \sigma_a^2} > \tau_a\right) \\ &= \mathcal{P}_s \cdot \mathbb{E}_{r_t, I_a} \left[\mathbb{P}\left(g_{00} > \frac{\tau_a}{P_a} (I_a + \sigma_a^2) r_t^{(1 - \epsilon_a)\alpha_a}\right) \right] \\ &= \mathcal{P}_s \cdot \exp\left\{-\frac{\tau_a}{P_a} r_t^{\alpha_a(1 - \epsilon_a)} \sigma_a^2\right\} \cdot \mathcal{L}_{I_a} \left\{\frac{\tau_a}{P_a} r_t^{\alpha_a(1 - \epsilon_a)}\right\} \end{aligned} \quad (4.6)$$

where $r_t = ||d_0^{n,t}||$ is used to denotes the distance between the device of interest and the origin during the t^{th} TR, $\mathcal{L}_{I_a}\{x\}$ is the Laplace transform of x with respect to inter-cell interference I_a and $\mathcal{P}_s$ is defined in eq. (4.3).

4.2.1.3 Inter-cell Interference Calculations

Let $s = \frac{\tau_a}{P_a} r_t^{\alpha_a(1 - \epsilon_a)}$, then the conditional coverage probability conditioned on the distance r_t to the nearest RU can be written as:

$$C_0^{n,t}(\sigma_a, \tau_a, \lambda_d^{n,t}, \alpha_a, \epsilon_a | r_t) = \mathcal{P}_s \cdot \exp\{-s\sigma_a^2\} \mathcal{L}_{I_a}\{s\} \quad (4.7)$$

where, $\mathcal{L}_{I_a}\{s\}$ is given by [94, 102] (*see: Appendix B.1*)

$$\begin{aligned}
\mathcal{L}_{I_a}\{s\} &= \mathbb{E}_{I_a}[\exp\{\{-sI_a\}\}] \\
&= \mathbb{E}_{g_{i0}, s, \|d_i^{n,t} - b_j\|, \|d_i^{n,t}\|} \left[\exp\{-sP_a \sum_{\|d_i^{n,t}\| \in \Phi_a^{n,t}} \|d_i^{n,t}\|^{-\alpha_a} \|d_i^{n,t} - b_j\|^{\epsilon_a \alpha_a} g_{i0}\} \right] \quad (4.8)
\end{aligned}$$

Let $y_t = \|d_i^{n,t} - b_j\|$ denote the distance between an arbitrary device located at $d_i^{n,t}$ and its serving RU located at b_j , and $x_t = \|d_i^{n,t}\|$ denotes the distance between the device located at $d_i^{n,t}$ and using the n^{th} sub-channel to the origin and during the t^{th} TR. All the Euclidean distances r_t , y_t , and x_t are random distances because they depend on where the device and RU are placed. Using Campbell's theorem on eq. (4.8) and given that the distances r_t , y_t and x_t all follow Rayleigh distributions with parameters λ_b , λ_a and $\lambda_d^{n,t}$ respectively, the Laplace transform of the interference can be expressed as [94, 102]:

$$\begin{aligned}
\mathcal{L}_{I_a}\{s\} &= \mathcal{L}_{I_a} \left\{ \frac{\tau_a}{p_i} r_t^{\alpha_a(1-\epsilon_a)} \right\} \\
&= \mathbb{E}_{g_{i0}, r_t, y_t, x_t} \left[\exp\{-\tau_a r_t^{\alpha_a(1-\epsilon_a)} \cdot \sum_{x_t \in \Phi_a^{n,t}} x_t^{-\alpha_a} y_t^{\epsilon_a \alpha_a} g_{i0}\} \right] \\
&\approx \exp \left(-2\pi \lambda_d^{n,t} \int_{x_t > 0} \left[1 - \int_{y_t > 0} \frac{2\pi \lambda_a y e^{(-\pi \lambda_a y_t^2)}}{1 + \tau_a r_t^{(1-\epsilon_a)\alpha_a} y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} dy_t \right] x_t dx_t \right) \quad (4.9)
\end{aligned}$$

Substituting eq. (4.9) into eq. (4.6), the conditional coverage probability conditioned on r_t , can be expressed as:

$$\begin{aligned}
C_0^{n,t}(\sigma_a, \tau_a, \lambda_d^{n,t}, \alpha_a, \epsilon_a | r_t) &= \mathcal{P}_s \cdot \exp\{-s\sigma_a^2\} \mathcal{L}_{I_a}\{s\} \\
&\approx \mathcal{P}_s \cdot \exp \left(-\frac{\tau_a}{p_i} r_t^{(1-\epsilon_a)\alpha_a} \sigma_a^2 - 2\pi \lambda_d^{n,t} \cdot \right. \\
&\quad \left. \int_{x_t > 0} \left[1 - \left(\int_{y_t > 0} \frac{2\pi \lambda_a y e^{(-\pi \lambda_a y_t^2)}}{1 + \tau_a r_t^{(1-\epsilon_a)\alpha_a} y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} dy_t \right) \right] x_t dx_t \right) \quad (4.10)
\end{aligned}$$

Since the preceding analysis is applicable to any device-RU pair, let the general radio coverage probability for any scheduled device-RU pair be denoted by $\mathcal{C}_a$. For simplicity, let R_t be the random distance between a device and its serving RU. The average unconditional

coverage probability as a function of σ_a , τ_a , $\lambda_d^{n,t}$, α_a and ϵ_a at any TR is defined as:

$$\begin{aligned}
\mathcal{C}_a &= \mathcal{P}_s \cdot \mathbb{E}_{R_t} [C_0^{n,t}(\sigma_a, \tau_a, \lambda_b, \alpha_a, \epsilon_a | R_t)] \\
&= \mathcal{P}_s \cdot \int_{r_t > 0} C_0^{n,t}(\sigma_a, \tau_a, \lambda_b, \alpha_a, \epsilon_a, r_t) 2\pi\lambda_b r_t e^{-\pi\lambda_b r_t^2} dr_t \\
&\approx 2\pi\lambda_b \cdot \mathcal{P}_s \cdot \int_{r_t > 0} \exp \left(-\pi\lambda_b r_t^2 - \frac{\tau_a}{p_i} r_t^{(1-\epsilon_a)\alpha_a} \sigma_a^2 - 2\pi\lambda_b \right. \\
&\quad \left. \int_{x_t > 0} \left[1 - \int_{y_t > 0} \frac{2\pi\lambda_a y_t e^{(-\pi\lambda_a y_t^2)}}{1 + \tau_a r_t^{(1-\epsilon_a)\alpha_a} y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} dy_t \right] x_t dx_t \right) \cdot r_t dr_t
\end{aligned} \tag{4.11}$$

where the PDF for the distance between a scheduled device and its serving RU is defined as $f_R(r_t) = 2\pi\lambda_a r_t e^{-\pi\lambda_a r_t^2}$.

4.2.2 Delay over the Air Interface

Within our system, the comprehensive delay of a data-frame is comprised of two primary components. The first component is the initial link delay, which represents the amount of time the data-frame spends in the queue at the source device before it reaches the serving RU via the air interface. The second component is the subsequent link delay, which encompasses the duration that the data-frame remains in transmission queues at the R-CPG (edge of the PLC-based front-haul network) and re-sequencing queues prior to being transmitted to the serving DU as shown in Figure 4.3.

Initially, attention is directed towards the radio link, where the mean radio transmission delay captures the time from data-frame generation at an IoT device to its reception at the serving RU. The rate of data-frame generation at a device is γ_d , while the departure rate from the serving RU to the PLC-based front-haul network, after conversion into the eCPRI format as a super HF, is μ_a . The arrival and departure processes at an IoT device are modelled using M/G/1 queues, assuming ample queue space. The service rate at a device depends on the scheduling probability (likelihood of being scheduled on a sub-channel) and achievable SINR. An RU's service rate is affected by PLC bandwidth constraints and technology used over the link.

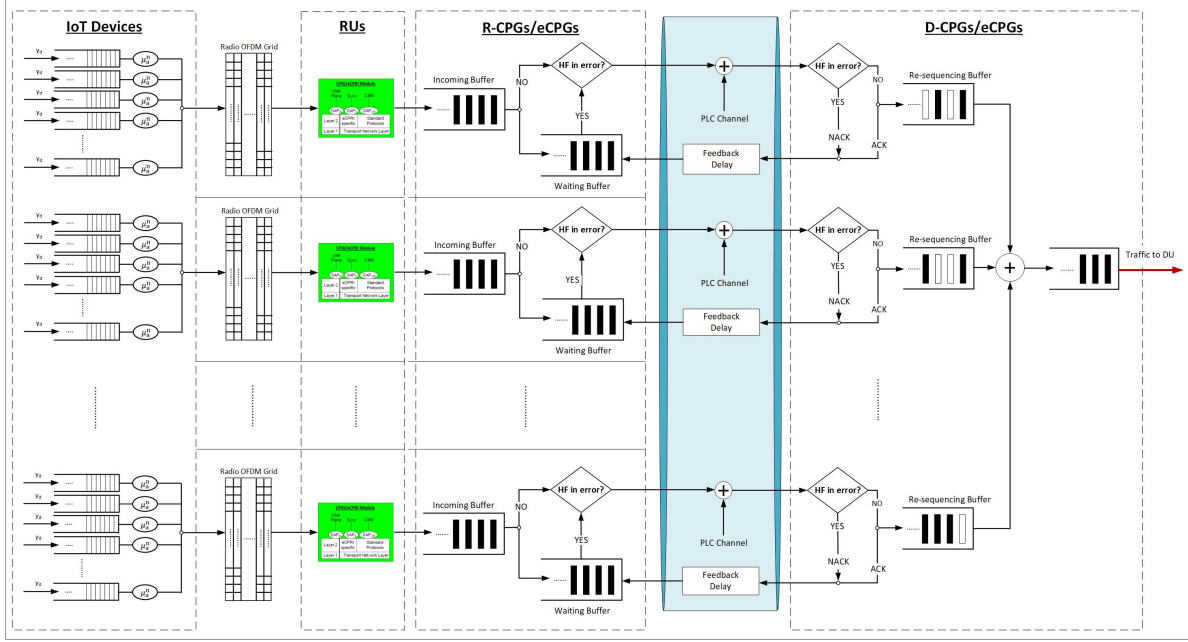


Figure 4.3: HF's transmission flow over the radio access network

4.2.2.1 Device-RU Queue Analysis

In order for any IoT device to effectively transmit its data-frame, it is necessary for the device to establish an association with a RU, be allocated a schedule on a sub-channel based on OFDM and be positioned within the coverage range of the RU. The average queuing delay at an IoT device, which is defined as the average number of 5G frames' time (T_a) required for the device to establish a successful connection with its serving RU and transmit a single data-frame, may be represented as $\overline{\Gamma}_a$. The coverage probability of the device-RU, as expressed in equation eq. (4.11), is used to model the geometric distribution of $\overline{\Gamma}_a$, as described in the work of Hesham et al. [99]. As such, the mean over the air delay is given as $\overline{\Gamma}_a = (T_a/\mathcal{C}_a)$ [99].

On the other hand, let the device's service rate for the n^{th} sub-channel and during the t^{th} TR be defined as [99]:

$$\mu_a^{n,t} = \left\lfloor \frac{T_a \mathcal{R}_a^n}{\mathcal{D}_a} \right\rfloor \quad n = 1, 2, \dots, \mathcal{N}_a \quad (4.12)$$

where, $\mathcal{D}_a$ represents the number of bits in the data frame sent via the radio connection.

Meanwhile, $\mathcal{R}_a^n$ is used to describe the average bit rate that may be achieved by the i^{th} device. This device is inside the coverage area, situated at $d_i^{n,t}$, and is scheduled to use the n^{th} sub-channel of the RU located at the origin. The floor function $\lfloor \cdot \rfloor$ ensures that only data-frame that are received at the RU successfully during the t^{th} TR are considered.

It should be noted that $\mathcal{R}_a^n$ is conditioned on the probability that the device is scheduled on a sub-channel and can successfully transmit its data (i.e it is within the RU coverage $\Theta_0^{n,t} > \tau_a$). Additionally, in eq. (4.12), the flooring function is to ensure that only data-frames that have been fully transmitted and were received by the RU within the current TR are taken into account. Using Shannon's formula, $\mathcal{R}_a^n$ is given by:

$$\mathcal{R}_a^n = \mathbb{E}[\omega_a \log_2(1 + \Theta_0^{n,t}) \mid \Theta_0^{n,t} > \tau_a \ \& \ \mathcal{X} = 1] \quad (4.13)$$

where $\mathcal{X} = [0, 1]$ defines the scheduling status such that $\mathcal{X} = 0$ indicates the device is not scheduled and $\mathcal{X} = 1$ indicates a scheduled device.

Since there is no closed-form probability density function (PDF) expression for the SINR; therefore, the maximum of two values is used to calculate $\mathcal{R}_a^n$. First, a lower bound by a fixed allocation of the bit rate given by $\mathfrak{L}_a^n$ is considered

$$\mathfrak{L}_a^n = \omega_a \log_2(1 + \tau_a) \quad (4.14)$$

The second approximation for $\mathcal{R}_a^n$ is the unconditional mean achievable bit rate. Its value can be calculated using the definition of the expectation $\mathbb{E}[X] = \int_{\zeta>0} \mathbb{P}(X > \zeta) \, d\zeta$, as the logarithm function is non-negative. Therefore, using eq. (4.11), the unconditional achievable bit rate $\mathcal{F}_a^n$ can be derived using the Shannon's channel capacity formula, defined as (*see: Appendix B.3*):

$$\begin{aligned} \mathcal{F}_a^n &= \mathbb{E} [\omega_a \log_2(1 + \Theta_0^{n,t})] \\ &= \int_{\zeta>0} \left(2\pi\lambda_b \cdot \int_{r_t>0} r_t e^{-\pi\lambda_b r_t^2} \cdot e^{-\frac{\beta_a}{p_i} r^{\alpha_a(1-\epsilon_a)} \sigma_a^2} \mathcal{L}_{I_a} \left\{ \beta_a r^{\alpha_a(1-\epsilon_a)} \right\} \, dr_t \right) \, d\zeta \end{aligned} \quad (4.15)$$

where, $\beta_a = 2^{(\zeta/\omega_a)} - 1$ and $\omega_a = (1 - gb) \cdot \Omega_a / \mathcal{N}_a$ and gb is the ratio of guard band.

Therefore, the mean achievable bit rate by a device is $\mathcal{R}_a^n = \max \{\mathcal{L}_a^n, \mathcal{F}_a^n\}$. Based on the above analysis and considering first in first out service rule, the total delay over the radio network, can be calculated as:

$$T_{air} = \frac{1}{\mu_a^{n,t} - \gamma_d \cdot \overline{\Gamma}_a} \quad (4.16)$$

4.2.3 Radio to Access Network Data Rate Conversion

The calculation of the aggregated data rate received at RU_j from all scheduled devices within its coverage is performed as follows:

$$\mathcal{R}_j = \sum_{n=1}^{M_{c_j}^t} \mathcal{R}_a^n \quad (4.17)$$

Consequently, the total incoming data rate into RU_j is set as:

$$\gamma_r = \sum_{n=1}^{M_{c_j}^t} \mu_a^{n,t} = \sum_{n=1}^{M_{c_j}^t} \left\lfloor \frac{T_a \mathcal{R}_a^n}{\mathcal{D}_a} \right\rfloor = \left\lfloor \frac{T_a \mathcal{R}_j}{\mathcal{D}_a} \right\rfloor \quad (4.18)$$

Following the consolidation of data-frame into CPRI hyper-frames or eCPRI data frames, the resulting HF is of size $\mathcal{D}_p$, which corresponds to that of an eCPRI message. Specifically, the size of the HF would range from 4 – 65539 bytes, as stated in the eCPRI specification [74]. Therefore, the data rate of RU_j sent to the incoming buffer at its CPG, after the conversion of radio data-frames into HF, may be described as $\mathcal{R}_p^j = \Lambda_{ap} \cdot \mathcal{R}_j$. The symbol Λ_{ap} is the conversion factor that relates the data rate sent via the air interface to the data rate necessary for the HF to transport the same data over the PLC front-haul system. The calculation of Λ_{ap} is derived from the eCPRI split I_U (or II_D in the downlink) and is expressed as $\Lambda_{ap} = 2 \times Q_F \times \chi_e \times \nu_p$ [46, 103]. The quantizer resolution used in the frequency domain is denoted as Q_F and has a specific value of 9. This particular quantizer resolution is chosen and implemented for Long Term Evolution (LTE) systems [46]. The redundancy factor of the HF header and the upper layers of the IP/Ethernet network are denoted as χ_e . The peak usage of the PLC link, denoted as ν_p , is recommended to be set at $\nu_p = 1$ under the

assumption of a worst-case scenario [46, 103]. The data rate produced at this stage, (split I_U), is dependent on the level of utilization of the radio resource blocks at the RUs. In other words, the front-haul data rate is directly proportional to the load on the cell.

Frequency division multiple access (FDMA) is used to split the PLC channel between the different RUs such that the bandwidth is divided into $\mathcal{N}_{PLC}$ sub-channels each of bandwidth $\omega_p = (1 - gb) \cdot \Omega_p / \mathcal{N}_{PLC}$ (i.e., $\mathcal{N}_{PLC}$ corresponds to the number of RUs that the link can support during t^{th} TR).

Hence, the mean arrival rate of HF's at the incoming buffer of each R-CPG denoted as (γ_p^j) can be defined as:

$$\gamma_p^j = \left\lfloor \frac{(T_a + T_{proc}) \cdot \mathcal{R}_p^j}{\overline{\mathcal{D}_p}} \right\rfloor \quad (4.19)$$

where $\overline{\mathcal{D}_p}$ is the mean HF size.

The total rate going into the PLC link for all the FDMA sub-channels is calculated as $\mathcal{R}_p = \sum_{j=1}^{\mathcal{N}_{PLC}} \mathcal{R}_p^j$. This is based on the assumption that all PLC sub-channels shall be allocated. The total arrival rate to the SISO PLC link is defined as:

$$\gamma_p = \sum_{j=1}^{\mathcal{N}_{PLC}} \gamma_p^j = \left\lfloor \frac{(T_a + T_{proc}) \cdot \mathcal{R}_p}{\overline{\mathcal{D}_p}} \right\rfloor \quad (4.20)$$

4.3 Access Delay Calculation

To find the access system delay, a SVD function module is placed at the CPG which transforms the originally 6x6 MIMO PLC channel of the three-phase line to 6 SISO parallel channels [40]. Thus, each virtual eCPRI link is modelled by a SISO channel via a single PLC port. The use of SVD simplifies the analysis to be based on a SISO link rather than a MIMO link.

The transmission flow diagram of the proposed radio access system is given in Figure 4.3; where the R-CPG is subject to an input of HF stream. At the edge of the D-CPG, the HF's are received at a generally random rate due to PLC channel errors and re-transmissions. Therefore, in the case of eCPRI, the access system from the RU edge to the D-CPG edge

can be presented as a G/G/1 queue. For the PLC line to be virtual to the RU-DU pair, the RU requires the incoming traffic to be of a generic nature due to the air interface nature.

As a part of the Selected Repeat (SR) process, once an HF is launched into the PLC channel, a copy of it is stored in the waiting buffer at the R-CPG. If a non-acknowledgment (NACK) is received, the HF is re-transmitted instantaneously before any other new HF. Once an ACK is received, the HF is removed from the waiting buffer and a new HF is transmitted. To deal with any possible out-of-sequence delivery of the HFs at the DU, when an HF with a higher identifier is received correctly it must wait in a re-sequencing buffer until all the HFs of the lower identifier are received correctly or dropped.

The access delay (T_{PLC}) of the considered system consists of 5 main parts: 1) incoming queuing delay at the R-CPG until the new HF is transmitted (T_q), 2) transport delay through the PLC channel (T_{ts}), 3) re-sequencing delay at the D-CPG including the time required for re-transmissions (T_{rsq}), 4) delay introduced due to IND (T_I) [40, 47], and 5) HARQ functionality processing delay T_{pH}

Thus the HF access delay can be defined as:

$$T_{PLC} = T_q + T_{ts} + T_{rsq} + T_I + T_{pH} \quad (4.21)$$

Finite State Markov Chain (FSMC) is applied to describe the HF transmission flow behaviour in each state for a single time slot (T_s). It is essential to enable the analysis of the delay throughout the transmission path of the HF. Each T_s is equivalent to the transmission time of a single eCPRI HF. With the use of eCPRI, both the RU and DU are synchronized with the use of synchronization protocols such as Synchronous Ethernet (SyncE) and Precision Time Protocol (PTP).

To facilitate the analytical modelling, the channel state transitions are considered to occur at the end of T_{ss} , where a T_s corresponds to a HF transmission time over the PLC link. It is assumed that the state transition time is very small compared to the total system delay. HFs arriving at the RU are firstly stored in the source buffer of an infinite capacity and served on a first-come-first-serve basis. An ideal SR approximation is also assumed such

that the dependence between the queuing process and the history of the HF transmission process does not exist. This approximation does not imply the disregard of the feedback delay, but rather a removal of its influence on the queuing process.

Further, the feedback channel for ACK/NACK is assumed to be error-free. In other words, at time t , the ACK/NACK that arrives at the RU is for the HF that was transmitted at time $t - t_f$, where t_f is the feedback delay over the PLC link. Such an ACK/NACK message has the same probabilistic nature as the ACK/NACK message associated with the HF to be transmitted at time t i.e., ideal SR HARQ approximation is assumed. Note that this approximation does not mean that the feedback delay is ignored, but that its impact on the queuing process is considered. Lastly, it is assumed that the error in the channel transition time and the feedback delay are both negligible compared to the sojourn time of a channel state.

4.3.1 Gilbert-Elliot PLC Channel Model

The PLC channel is modelled as the Gilbert-Elliot channel model which is a binary two-state finite-state Markov Chain model [104, 105] for the non-stationary time-varying channel as shown in Figure 4.4. The Good State denoted as state 0 is characterized by low FER e_0 , while the Bad or Burst state denoted as state 1 is characterized by high FER denoted (e_1) such that $e_0 \ll e_1$ with the transition probability matrix $\mathbf{C}$ defined as $\mathbf{C} = \begin{bmatrix} c_{00} & c_{01} \\ c_{10} & c_{11} \end{bmatrix}$.

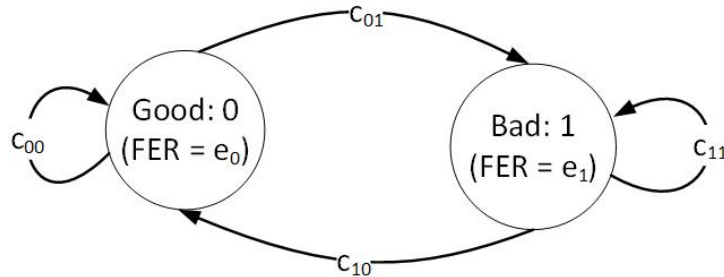


Figure 4.4: Gilbert Elliot Channel Markov Model

The mean number of HFs transmitted while in Bad state, i.e. the mean burst length ($\overline{b_b}$) and the mean number of HFs transmitted while in Good state, i.e. the mean state holding time ($\overline{b_g}$) are respectively defined as [106]:

$$\overline{b_b} = \frac{1}{1 - c_{11}} \quad (4.22)$$

$$\overline{b_g} = \frac{1}{1 - c_{00}} \quad (4.23)$$

The probability a HF is received in error given that the channel was in state CS_k , which is the instantaneous FER without implementing any error correction mechanism and, is defined as:

$$\mathbb{P}\{e_k|CS_k\} = \begin{cases} 1 - (1 - \epsilon_0)^{\mathcal{D}_p}, & \text{if } CS_k = 0 \\ 1 - (1 - \epsilon_1)^{\mathcal{D}_p}, & \text{if } CS_k = 1 \end{cases} \quad (4.24)$$

where ϵ_0 and ϵ_1 are the BER in Good state and Bad state, respectively. $\mathcal{D}_p$ is the number of bits within one HF. Given e_0 & e_1 then the mean FER can be calculated as $\bar{e} = \frac{c_{10} e_0 + c_{01} e_1}{c_{10} + c_{01}}$ [107].

4.3.2 Incoming Queuing Delay

The incoming queuing delay is the waiting time due to the re-transmission of other HFs in error at the RU. Looking into the access side of the network, the source of HFs at the RU side is the aggregated data-frames received from the air interface after conversion. Therefore, the source is modelled as an N-state Discrete Markov process, where $N_{src} = 2$ represents the **ON-OFF** behavior of the eCPRI data flows [69, 74]. Based on the **ON-OFF** Markovian source model, the transition probabilities can be defined as follows:

1. $a_{01} = \mathbb{P}\{\text{Prob. of arrival of at least one new HF in current } Ts \text{ with no arrival in previous } Ts\}$
2. $a_{10} = \mathbb{P}\{\text{Prob. of no arrival of new HF in current } Ts \text{ with one new arrival in previous } Ts\}$

3. $a_{00} = 1 - a_{01} = \mathbb{P}\{\text{Prob. of no new HF arrival in current } Ts \text{ with no arrival in previous } Ts\}$
4. $a_{11} = 1 - a_{10} = \mathbb{P}\{\text{Prob. of arrival of new HF in current } Ts \text{ with one new arrival in previous } Ts\}$

Accordingly, the transition probability matrix for the modelled source can be calculated using the following [108, 109]:

$$\rho_p^j = \gamma_p^j \cdot \frac{a_{01}}{a_{01} + a_{10}} \quad \text{and} \quad T_{on} = \frac{1}{a_{10}} \quad (4.25)$$

where T_{on} is the mean length of the **ON** period, and $0 < \rho_p^j < 1/\mathcal{N}_{plc}$ is the mean traffic load of HFs into the j^{th} R-CPG. Thus, the total mean traffic load into one SISO PLC link is given by $\rho_p = \rho_p^j \times \mathcal{N}_{plc} \dots (0 < \rho_p < 1)$.

At $T_s = k$, let the incoming queue length at the R-CPG, given that the PLC channel state is y , and the incoming HF source is in state u , be denoted as $q_{u,y}^k$. Accordingly, the length of the incoming queue at the next time slot (i.e., $q_{x,y}^{k+1}$), given there is no change in the PLC channel state and that the incoming HF source is in a new state (x) where $x, \& y \in \{0, 1\}$, can be defined as:

$$\therefore q_{x,y}^{k+1} = \begin{cases} q_{u,y}^k + a_x^k - r_y^k, & \text{if } q_j^k \neq 0 \\ q_{u,1-y}^k + a_x^k - r_{1-y}^k, & \text{if } q_{1-y}^k \neq 0 \\ a_x^k, & \text{if } q_j^k \text{ or } q_{1-y}^k = 0 \end{cases} \quad (4.26)$$

where a_x^k is the number of new HF arrivals to the R-CPG incoming queue buffer in time slot k , and $r_y^k \in \{0, 1\}$ is the number of HF departures from the R-CPG incoming queue buffer onto the PLC channel when in state $y \in \{0, 1\}$. It should be noted that, since the assumption is one HF is sent per time slot and that the arrival process is an **ON-OFF** process, then we can deduce that $a_x^k \in \{0, 1\}$.

Let CS_k denote the PLC channel state at the k^{th} time slot, where $CS_k \in \{y, 1 - y\}$. Also, let L_r denote the re-transmission limit of an HF before it gets dropped. Therefore, the conditional probability of departure given CS_k at time-slot k can be defined as

$$r_y = \mathbb{P}\{r_y^k = 1\} = \mathbb{P}\{\text{HF deliver} \mid CS_k = y\} \times \mathbb{P}\{\text{HF drop after } L_r \mid CS_k = y\}$$

$$r_{1-y} = \mathbb{P}\{r_{1-y}^k = 1\} = \mathbb{P}\{\text{HF deliver} \mid CS_k = 1 - y\} \times \mathbb{P}\{\text{HF drop after } L_r \mid CS_k = 1 - y\} \quad (4.27)$$

Furthermore, we can also derive the conditional probability of no departure i.e. HF re-transmission given CS_k as

$$v_y = \mathbb{P}\{r_y^k = 0\} = \sum \mathbb{P}\{\text{HF re-trans} < L_r \mid CS_k = y\} \quad (4.28)$$

$$v_{1-y} = \mathbb{P}\{r_{1-y}^k = 0\} = \sum \mathbb{P}\{\text{HF re-trans} < L_r \mid CS_k = 1 - y\}$$

The receiving of an HF in error or successfully is only dependent on the current channel state and is considered independent of the transition in the channel state. For a certain current channel state ($CS_k = y$), the received HF state transition diagram based on the number of consecutive receive failures of an HF at the DU side can be described as in Figure 4.5. These states are referred to as HF states/ HF steady states S_l , where l indicates the number of HF re-transmissions. Defining the probability of receiving a transmitted HF at the DU as α_p and referring to Figure 4.5, the probability of an HF departure from the system (steady state S_0) given the PLC channel state is expressed as:

$$r_y = \mathbb{P}\{S_0 \mid CS_k = y\} = \frac{1}{1 + \sum_{l=1}^{L_r} \alpha_p \cdot e_j^l} = \frac{1 - e_j}{1 - e_j + \alpha_p \cdot e_j - \alpha_p \cdot e_j^{L_r+1}}$$

and for current channel state ($CS_k = 1 - y$)

$$r_{1-y} = \mathbb{P}\{S_0 \mid CS_k = 1 - y\} = \frac{1}{1 + \sum_{l=1}^{L_r} \alpha_p \cdot e_{1-y}^l} = \frac{1 - e_{1-y}}{1 - e_{1-y} + \alpha_p \cdot e_{1-y} - \alpha_p \cdot e_{1-y}^{L_r+1}} \quad (4.29)$$

Accordingly, the probability of no HF departure from the system (steady states $S_1, S_2 \dots S_{L_r}$) given the current channel state being $CS_k = y$ can be expressed as:

$$v_y = \sum_{l=1}^{L_r} \mathbb{P}\{S_l | CS_k = y\} = \frac{\alpha_p \cdot (1 - e_j) \cdot \sum_{l=1}^{L_r} e_j^l}{1 - e_j + \alpha_p \cdot e_j - \alpha_p \cdot e_j^{L_r+1}} = \frac{\alpha_p \cdot e_j \cdot (1 - e_j^{L_r})}{1 - e_j + \alpha_p \cdot e_j - \alpha_p \cdot e_j^{L_r+1}}$$

and for current channel state ($CS_k = 1 - y$)

$$v_{1-y} = \sum_{l=1}^{L_r} \mathbb{P}\{S_l | CS_k = 1 - y\} = \frac{\alpha_p \cdot (1 - e_{1-y}) \cdot \sum_{l=1}^{L_r} e_{1-y}^l}{1 - e_{1-y} + \alpha_p \cdot e_{1-y} - \alpha_p \cdot e_{1-y}^{L_r+1}} \quad (4.30)$$

$$= \frac{\alpha_p \cdot e_{1-y} \cdot (1 - e_{1-y}^{L_r})}{1 - e_{1-y} + \alpha_p \cdot e_{1-y} - \alpha_p \cdot e_{1-y}^{L_r+1}}$$

With the assumptions that all launched HFs are received at the DU, i.e. no HF lost over the channel and that after enough transit time, there shall be always a new HF in the incoming queue buffer to be sent over the channel, it is reasonable to set $\alpha_p = 1$. Thus, the term $q_{x,y}^{k+1}$ in (4.26), given channel state is y and source state x , can be expressed as follows [69]:

$$q_{x,y}^{k+1} = \begin{cases} q_{u,y}^k + x & \rightarrow \text{Case 1, with probability } (c_{y,y} \cdot a_{u,x} \cdot v_y) \\ q_{u,y}^k + x - 1 & \rightarrow \text{Case 2, with probability } (c_{y,y} \cdot a_{u,x} \cdot r_y) \\ q_{u,1-y}^k + x & \rightarrow \text{Case 3, with probability } (c_{1-y,y} \cdot a_{u,x} \cdot v_{1-y}) \\ q_{u,1-y}^k + x - 1 & \rightarrow \text{Case 4, with probability } (c_{1-y,y} \cdot a_{u,x} \cdot r_{1-y}) \\ q_{0,y}^0 & \rightarrow \text{Case 5, with probability } (c_{y,y} \cdot a_{0,x}) \\ q_{0,1-y}^0 & \rightarrow \text{Case 6, with probability } (c_{1-y,y} \cdot a_{0,x}) \end{cases} \quad (4.31)$$

where, $x \in \{0, 1\}, y \in \{0, 1\}$

In all cases, the source state changes (i.e. HF arrivals) from u HF's at time slot k to x HF's at time slot $k + 1$, where u & $x \in \{0, 1\}$. Below, we define the possible six cases as captured in (4.31).

Case 1 There are $x \in \{0, 1\}$ new arrivals & no departures and channel stays in state y

Case 2 There are $x \in \{0, 1\}$ new arrivals & 1 new departure and channel stays in state y

Case 3 There are $x \in \{0, 1\}$ new arrivals & no departures and channel moves from state $1 - y$ to y

Case 4 There are $x \in \{0, 1\}$ new arrivals & 1 new departure and channel moves from state $1 - y$ to y

Case 5 $q_j^k = 0$ happens at system start i.e. $t = 0$ and channel state remains y .

Case 6 $q_{1-y}^k = 0$ happens at system start i.e. $t = 0$ and channel state changes to $1 - y$.

To ensure that the probability of the number of HF's in the incoming queue length is independent of time, we shall consider the steady-state form. As such, let $q_{x,y}^h$ denote the steady-state of the incoming queue length at the RU side where $q_{x,y}^h \triangleq \lim_{k \rightarrow \infty} \mathbb{P}[q_{x,y}^k = h]$. Thus, by taking the limit, $q_{x,y}^h$ can be derived to yield:

$$\begin{aligned}
q_{x,y}^h = & \sum_{u=0}^{\min(N_{src}, h+1)} \left(c_{y,y} \cdot a_{u,x} \cdot r_y \cdot q_{u,y}[h-u+1] + c_{\bar{y},y} \cdot a_{u,x} \cdot r_{\bar{y}} \cdot q_{u,\bar{y}}[h-u+1] \right) \\
& + \sum_{u=0}^{\min(N_{src}, h)} \left(c_{y,y} \cdot a_{u,x} \cdot v_y \cdot q_{u,y}[h-u] + c_{\bar{y},y} \cdot a_{u,x} \cdot v_{\bar{y}} \cdot q_{u,\bar{y}}[h-u] \right) \\
& + \sum_{u=0}^{\min(N_{src}, 1)} \left(c_{y,y} \cdot a_{u,x} \cdot r_y \cdot q_{u,y}[1-u] + c_{\bar{y},y} \cdot a_{u,x} \cdot r_{\bar{y}} \cdot q_{u,\bar{y}}[1-u] \right) \\
& + c_{y,y} \cdot a_{0,x} \cdot q_{0,y}^0 + c_{\bar{y},y} \cdot a_{0,x} \cdot q_{0,\bar{y}}^0
\end{aligned} \tag{4.32}$$

where, $\bar{y}$ denotes $1 - y$ and $x \in \{0, 1\}$.

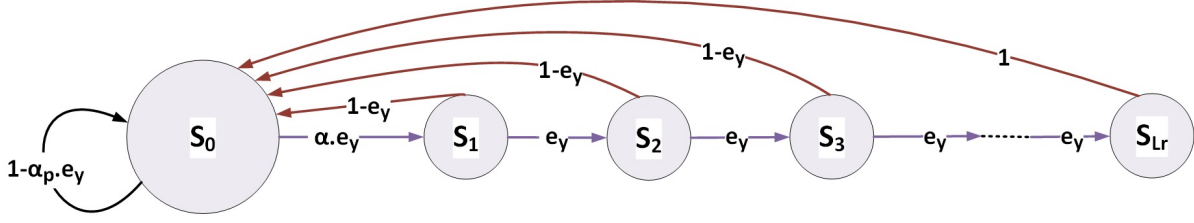


Figure 4.5: A state transition diagram based on the number of consecutive receive failures of HF

Let $Q_{x,y}(z)$ denote the Probability Generating Function (PGF) for the queue length $q_{x,y}^h$ when the source state is x and the channel state is y . Therefore, the PGF of the queue length for all channel and source states $\mathbf{Q}(z)$ is set as:

$$\mathbf{Q}(z) = \sum_{m=0}^{\infty} [\mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z)]^{m+1} \cdot [\mathbf{I} \otimes [\mathbf{E}(z)^{-1} - \mathbf{I}]] \mathbf{Q}_0$$

where, $\mathbf{Q}(z) \triangleq [Q_{0,0}(z), Q_{0,1}(z), \dots, Q_{N_{src},1}(z)]^\dagger$,

$$\mathbf{Q}_m(z) \triangleq [Q_{m,0}(z), Q_{m,1}(z)]^\dagger, \quad m = 0, 1, \dots, N_{src},$$

$$\mathbf{E}(z) \triangleq \text{diag}[\xi_0(z), \xi_1(z)], \quad \xi_y(z) \triangleq v_y + r_y z^{-1},$$

$$\mathbf{Q}_0 \triangleq [q_{0,0}^0, q_{0,1}^0, 0, 0, \dots, 0]^\dagger, \tag{4.33}$$

$$\text{diag}[z^x] \triangleq \text{diag}[1, z, z^2, \dots, z^{N_{src}}],$$

$\mathbf{A}$ is the transition probability matrix for HF arrival at RU,

$\mathbf{C}$ is the transition probability matrix for PLC channel state, and

$\dagger$ is the matrix transpose

Let $\mathbf{Q}(z) \triangleq \sum_{x,y} Q_{x,y}(z)$ be the PGF of the queue length. The initial values $q_{0,0}^0$ & $q_{0,1}^0$ may be found by locating all poles of $\mathbf{Q}(z)$ in a closed unit disc since $\mathbf{Q}(z)$ is analytic in a stable system. By spectral decomposition of the matrix $\mathbf{X} = \mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z)$, the poles can be found such that:

$$\mathbf{X} = \mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z) = \mathbf{G}(z) \mathbf{\Gamma}(z) \mathbf{G}^{-1}(z) \tag{4.34}$$

where, $\mathbf{\Gamma}(z) = \text{diag}[\lambda_0(z), \lambda_1(z), \dots, \lambda_{2N_{src}+1}(z)]$ is the eigen values of the matrix $\mathbf{X}$. $\mathbf{g}_m(z)$

and $\mathbf{l}_m(z)$ denote the corresponding left column and right row eigen vectors of eq. (4.34), respectively, where $\mathbf{G}(z)$ and $\mathbf{G}^{-1}(z)$ are defined as:

$$\begin{aligned} \mathbf{G}(z) &\triangleq [\mathbf{g}_0(z), \mathbf{g}_1(z), \dots, \mathbf{g}_{2N_{src}+1}(z)] \quad \text{so, } \mathbf{G}^{-1}(z) \triangleq [\mathbf{l}_0(z), \mathbf{l}_1(z), \dots, \mathbf{l}_{2N_{src}+1}(z)]^\dagger \\ \therefore \mathbf{X} &= \mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z) = \sum_{x=0}^{2N_{src}+1} \lambda_x(z) \mathbf{g}_x(z) \mathbf{l}_x(z) \end{aligned} \quad (4.35)$$

Substituting back in eq. (4.33) and applying geometric series, we get:

$$\mathbf{Q}(z) = \sum_{x=0}^{2N_{src}+1} \frac{\lambda_x(z)}{1 - \lambda_x(z)} \mathbf{g}_x(z) \mathbf{l}_x(z) [\mathbf{I} \otimes \mathbf{E}(z)^{-1} - \mathbf{I}] \mathbf{Q}_0 \quad (4.36)$$

Substituting $\mathbf{Q}_0$ into the previous equation, we obtain:

$$\mathbf{Q}(z) = \sum_{x=0}^{2N_{src}+1} \frac{\lambda_x(z)}{1 - \lambda_x(z)} \sum_{m=0}^{2N_{src}+1} g_{m,x}(z) \cdot \left(l_{x,0}(z) \frac{1 - \xi_0(z)}{\xi_0(z)} q_{0,0}^0 + l_{x,1}(z) \frac{1 - \xi_1(z)}{\xi_1(z)} q_{0,1}^0 \right) \quad (4.37)$$

since $\mathbf{g}_m = [g_{m,0}, g_{m,1} \dots g_{m,2N_{src}+1}]^\dagger$ and $\mathbf{l}_m = [l_{0,m}, l_{1,m} \dots l_{2N_{src}+1,m}]^\dagger$.

For each root $|z_x| < 1$ and as $\mathbf{Q}(z)$ is analytic for each root, we can set up the following boundary equation:

$$\left(l_{x,0}(z_x) \frac{1 - \xi_0(z_x)}{\xi_0(z_x)} q_{0,0}^0 + l_{x,1}(z_x) \frac{1 - \xi_1(z_x)}{\xi_1(z_x)} q_{0,1}^0 \right) = 0 \quad (4.38)$$

Secondly, the definition $\lim_{z \rightarrow 1} \mathbf{Q}(z) = 1$ can be used to solve both equations to obtain the values for $q_{0,0}^0$ and $q_{0,1}^0$ and a complete form for $\mathbf{Q}(z)$. Therefore, the mean queue length $\bar{q} = \mathbb{E}[q_{x,y}^h]$ in the incoming queue buffer overall source and channel states is set as $\bar{q} = \mathbf{Q}'(1)$, where X' indicated the first derivative of X .

Therefore, to obtain the incoming queuing delay, first we need to find the average number of HFs at the RU side $\mathbb{E}[q]$. With non-zero feedback delay, the average number of HFs on the RU side including both the incoming queuing and waiting buffers is calculated as $\mathbb{E}[q] = \bar{q} + t_f \overline{\mathcal{T}}_r \rho_p$, where $\overline{\mathcal{T}}_r$ denotes the mean throughput over the PLC link identified in eq. (4.40) later. Then using Little's Law, the mean queuing time for the ideal SR HARQ is

obtained as:

$$T_q = \frac{\mathbb{E}[q]}{\rho_p} \quad (4.39)$$

4.3.3 Transport Delay

The transport delay (T_{ts}) is a combination of two delays: the transmission delay and the propagation delay. The transmission delay denoted as (T_t) is determined by the average throughput ($\overline{\mathcal{T}_r}$) and can be mathematically represented as $T_t = t_f \cdot \overline{\mathcal{T}_r} - t_f/2$ [69]. Here, t_f represents the total feedback delay involved in transmitting and propagating the HF and the ACK/NACK message. The propagation delay (T_{pg}) is a fixed delay that is determined by the length (L_p) and propagation speed (V_p) of the power cable. Mathematically, the propagation delay may be expressed as $T_{pg} = \lceil (L_p/V_p)/Ts \rceil$, where the propagation speed for power lines is determined to be $173 \text{ m}/\mu\text{s}$ [110].

The mean throughput ($\mathcal{T}_r$) is determined based on the analysis of the channel model as follows:

$$\begin{aligned} \mathcal{T}_r &= \mathbf{1} + \mathcal{U} [\mathbf{I} - \mathcal{S}]^{-1} \mathcal{V} \\ \text{s.t. } \mathcal{U} &= \begin{bmatrix} \mathbf{1} & \mathbf{1} \end{bmatrix} \quad \mathcal{S} = \begin{bmatrix} c_{00}e_0 & c_{10}e_0 \\ c_{01}e_1 & c_{11}e_1 \end{bmatrix} \quad \mathcal{V} = \begin{bmatrix} \pi_{c_0}e_0 \\ \pi_{c_1}e_1 \end{bmatrix} \end{aligned} \quad (4.40)$$

where $\mathbf{I}$ represents the identity matrix. The symbols π_{c_0} and π_{c_1} represent the steady state probability of the channel being in the Good and Bad states, respectively. These probabilities are set as [107]:

$$\pi_{c_0} = \frac{c_{10}}{c_{10} + c_{01}} \quad \text{and} \quad \pi_{c_1} = \frac{c_{01}}{c_{10} + c_{01}} \quad (4.41)$$

Therefore, the mean transport delay is set as $T_{ts} = T_t + T_{pg}$ [107].

4.3.4 Selective Repeat Re-sequencing Delay

In the literature, there are different methodologies to find the re-sequencing delay for SR HARQ. Assuming heavy traffic, the authors in [69] approximated the mean re-sequencing delay and buffer occupancy with no limit on the number of re-transmissions per HF. In this section, the aim is to find the probability distribution of the re-sequencing delay including the time for re-transmissions while considering the IoT air traffic and with a limited number of re-transmissions per HF.

To evaluate the re-sequencing delay, the following assumptions are considered in the analysis [70, 111]:

1. The re-sequencing buffer at the DU side is assumed to be an infinite buffer.
2. HF errors occur at random on a transmission channel based on its state with the error rate e_y of the channel state ($CS_k = y$).
3. ACK or NACK is sent immediately when an HF is received correctly or in error respectively; i.e. ignoring processing delay at the DU side as used in the literature [69].
4. There is no error for the delivery of ACKs/NACKs.
5. Re-transmission has high priority compared to the transmission of a new HF and is performed immediately after a NACK is returned to the sender.
6. The maximum number of re-transmissions is limited to L_r . If an HF is re-transmitted for L_r times and still received in error, it is dropped.
7. If there is no re-transmission in a time slot, a new HF is transmitted from the ones existing at the incoming queuing buffer.
8. The SR window size is set to be W slots, each numbered from 0 to $W-1$ as shown in Figure 4.6.
9. t_f is defined as the number of time slots in the total trip time, i.e. the feedback delay.

The accuracy of the approximate results is confirmed by comparing them versus more realistic simulation outcomes.



Figure 4.6: An illustration of SR window

In calculating the re-sequencing delay, when an HF is sent on a certain slot $\# \omega_r$ within the SR window, the HF is re-transmitted in the same slot $\# \omega_r$. This continues until it is delivered successfully or it gets dropped after L_r re-transmissions. Consequently, the ideal SR HARQ can be modelled as W independent slots, each being simulated as a stop-and-wait case [70]. Accordingly, the probability that an HF is dropped is given by $\mathbb{P}_{drop} = e_y \cdot \mathbb{P}_s(S_{L_r}|y)$, where $\mathbb{P}_s(S_l|y)$ is the steady-state probability of the HF state $S_l|y$, and y is the channel state that the HF has experienced. $\mathbb{P}_s(S_l|y)$ can be expressed as per the Markov Chain in Figure 4.5 as:

$$\mathbb{P}_s(S_0|y) = \frac{(1 - e_y)}{1 - e_y + \alpha_p e_y - \alpha_p e_y^{L_r+1}} \quad y \in \{0, 1\} \quad (4.42)$$

$$\mathbb{P}_s(S_l|y) = \frac{\alpha_p \cdot (1 - e_y) \cdot e_y^l}{1 - e_y + \alpha_p e_y - \alpha_p e_y^{L_r+1}} \quad (4.43)$$

$$l \in \{1, 2, 3, \dots, L_r\} \text{ \& } y \in \{0, 1\}$$

$$\mathbb{P}_{drop} = \frac{\alpha_p \cdot (1 - e_y) \cdot e_y^{L_r+1}}{1 - e_y + \alpha_p e_y - \alpha_p e_y^{L_r+1}} \quad y \in \{0, 1\} \quad (4.44)$$

To find the re-sequencing delay including re-transmission time (T_{rsq}), first, define its PDF in which $\mathbb{P}_{rsq}(\mu_r, \omega_r|y) = \mathbb{P}\{T_{rsq} = \mu_r t_f + \omega_r | CS_k = y\}$ slots including the re-transmission delay of each HF received in error. In slot $\#0$, the HF is received successfully after being received incorrectly μ_r times. As mentioned earlier, t_f is defined as of time slots in the total trip time, i.e. the feedback delay. At time t_0 , we focus on slot $\# \omega_r$ which is the slot within the SR window with the maximum number of failed re-transmissions of a particular HF. Two cases are considered in finding $\mathbb{P}_{rsq}(\mu_r, \omega_r|y)$

A. Case $\omega_r \neq 0$ In this case, the HF in slot $\#\omega_r$ has the maximum number of received failures within the SR window. This means that the HF in slot $\#\omega_r$ is received correctly after all other HFs; i.e. it has been received correctly after $\mu_r + 1$ re-transmissions. The HF in slot $\#0$ is received successfully after being received in error for less than or equal to μ_r times. The HFs in slots 1 to $\omega_r - 1$ as well as the HFs in slots $\omega_r + 1$ to $W - 1$ have been received in error less times than the HF in slot $\#0$.

Therefore, $\mathbb{P}_{rsq}(\mu_r, \omega_r | y)$ can be expressed as [70, 111]:

$$\mathbb{P}_{rsq}(\mu_r, \omega_r | y) = \left(\sum_{l=0}^{\mu_r} \mathbb{P}_t(\mu_r | y) \right) \cdot \mathbb{P}_R(\mu_r + 1, \omega_r | y) \quad (4.45)$$

where $\mathbb{P}_t(\mu_r | y)$ is the probability that the HF in slot $\#0$ is received correctly after μ_r re-transmissions with channel state being y and it is defined as:

$$\mathbb{P}_t(\mu_r | y) = \frac{(1 - e_y) \cdot e_y^{\mu_r}}{1 - e_y^{L_r+1}} \quad \text{so,} \quad \sum_{l=0}^{\mu_r} \mathbb{P}_t(\mu_r | y) = \frac{1 - e_y^{\mu_r+1}}{1 - e_y^{L_r+1}} \quad (4.46)$$

and $\mathbb{P}_R(\mu_r + 1, \omega_r | y)$ is the probability that the HF on slot $\#\omega_r$ has been received incorrectly $\mu_r + 1$ times before being received successfully. Considering the independence among the different time slots within the SR window, it is defined as:

$$\begin{aligned} & \mathbb{P}_R(\mu_r + 1, \omega_r | y) \\ &= \mathbb{P}_s(S_{\mu_r} | y) \cdot \prod_{w=1}^{\omega_r-1} \sum_{l=0}^{\mu_r+1} \mathbb{P}_s(S_l | y) \cdot \prod_{w=\omega_r+1}^{W-1} \sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) \\ &= \mathbb{P}_s(S_{\mu_r} | y) \cdot \left[\sum_{l=0}^{\mu_r+1} \mathbb{P}_s(S_l | y) \right]^{\omega_r-1} \cdot \left[\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) \right]^{W-1-\omega_r} \end{aligned} \quad (4.47)$$

where, $\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) = \frac{1 - e_y + \alpha_p \cdot e_y \cdot (1 - e_y^{\mu_r})}{1 - e_y + \alpha_p \cdot e_y - \alpha_p e_y^{L_r+1}}$ and,

$$\sum_{l=0}^{\mu_r+1} \mathbb{P}_s(S_l | y) = \frac{1 - e_y + \alpha_p \cdot e_y \cdot (1 - e_y^{\mu_r+1})}{1 - e_y + \alpha_p \cdot e_y - \alpha_p e_y^{L_r+1}}$$

Accordingly, $\mathbb{P}_{rsq}(\mu_r, \omega_r|y)$ can be expressed as:

$$\mathbb{P}_{rsq}(\mu_r, \omega_r|y) = \left(\frac{1 - e_y^{\mu_r+1}}{1 - e_y^{L_r+1}} \right) \cdot \mathbb{P}_s(S_{\mu_r}|y) \cdot \left[\sum_{l=0}^{\mu_r+1} \mathbb{P}_s(S_l|y) \right]^{\omega_r-1} \cdot \left[\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l|y) \right]^{W-1-\omega_r} \quad (4.48)$$

B. Case $\omega_r = 0$ In this case, at time t_0 the HF in slot #0 has been received after μ_r re-transmissions while all other slots within the SR window have been received in error for less than μ_r times or have been dropped. In such situation $\mathbb{P}_{rsq}(\mu_r, 0|y) = \mathbb{P}\{T_{rsq} = \mu_r t_f|y\}$ can be expressed as [70]:

$$\mathbb{P}_{rsq}(\mu_r, 0|y) = \mathbb{P}_t(\mu_r|y) \cdot \left[\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l|y) \right]^{W-1} \quad (4.49)$$

Therefore, the mean re-sequencing delay can be calculated as:

$$\overline{T_{rsq}} = \int_0^W \int_0^\infty t \cdot (\mathbb{P}_{rsq}|\omega_r) dt d\omega_r = \int_0^{W-1} \int_0^{L_r} \frac{\mu_r}{t_f} \cdot (\mathbb{P}_{rsq}|\omega_r) d\mu_r d\omega_r \quad (4.50)$$

4.3.5 Effect of Impulsive Noise on Access Delay

As per [40], an IND module is equipped in the CPG to detect the PAPR along the PLC channel in real time. By jointly considering the carrier energy level, the IND calculates the PAPR in every HF duration and concludes whether an impulsive noise event has occurred. Once PAPR is found larger than a threshold, an instant notification is made to the HARQ module as a reference for the protocol operation.

The study follows the cross-layer design of IND by assuming that the Ethernet module encapsulates each HF in an Ethernet HF and can provide instant channel status to the HARQ module regarding whether an impulsive noise event is in place. When the PAPR is detected greater than a certain threshold, e.g., 10dB, it indicates the occurrence of an impulsive noise event, where the SR-HARQ protocol shall react as follows

- 1) It holds stops the on-going HF transmission, while the DU drops the HF that is being received.

- 2) As soon as the channel is found clear again, the SR-HARQ simply consider all the currently unacknowledged HFs in the wait buffer at the RU side as lost and flushes all these HFs to the DU.

To consider the delay introduced into the system due to the use of IND, the module is modelled as a single state MC model with a probability that an impulse is detected as P_{imp} which is defined as $P_{imp} = \mathbb{P}[A_{imp} > \sqrt{Th \times P_{mean}}] \times \mathbb{P}[t_i \leq \Delta t]$, where A_{imp} is the amplitude of the impulsive noise event with Middleton's Class A distribution. t_i is the inter-arrival time between the impulses with an exponential nature with mean $\mu_i = 10.1$ ms [44]. Therefore, P_{imp} can be expressed as:

$$P_{imp} = \left(1 - \int_0^{A_{imp}} \sum_{z=0}^{\infty} \frac{\exp\{(-A)\} \cdot A^z}{z!} \frac{\exp\left\{\left(\frac{-p^2}{2\sigma_z^2}\right)\right\}}{\sqrt{2\pi\sigma_z^2}} dp \right) \times \left(1 - \exp\left\{\left(\frac{-\Delta t}{\mu_i}\right)\right\} \right) \quad (4.51)$$

with $\sigma_z^2 = \sigma^2 \left(\frac{(z/A) + \Gamma_i}{1 + \Gamma_i} \right)$

It is essential to find the leakage probability of an impulse noise over more than one time slot where the impulse width is modelled by the two terms Gaussian model in [44] with mean $4.7\mu s$ in the following equation:

$$\mathbb{P}_{t_w} = P_1 \mathcal{N}(\mu_1, \sigma_1^2) + P_2 \mathcal{N}(\mu_2, \sigma_2^2) \quad [\mu s] \quad (4.52)$$

where $P_1 = 0.736$, $\mu_1 = 4.9$, $\sigma_1 = 0.2$ and $P_2 = 0.318$, $\mu_2 = 4.2$, $\sigma_2 = 0.25$. The HF time slot duration is set to be $66.67\mu s$ which is the OFDM symbol period in 5G technology [74]. By calculating $\mathbb{P}\{t_w > 66.67\mu s\}$, the leakage probability is found to be negligible since the cumulative density function (CDF) of t_w reach 99.9999% at $t_w < 5.5\mu s$. Consequently, whenever an impulse is detected, the delay introduced is approximated up to one T_s plus the transmission of all the HFs being transported during the impulse being flushed through the system. Therefore, the introduced delay due to IND T_I can be expressed as:

$$T_I = P_{imp} \times (W + 1) \quad [Ts] \quad (4.53)$$

where W is the number of time slots within the selective repeat window (assuming all the frames in the window were not acknowledged). One slot is added to the delay due to the wait time for the impulse width which as indicated above is less than one time slot.

4.4 Case Studies and Numerical Outcomes

In order to proceed with the system simulations, first it is essential to find valid Gilbert-Elliot parameters for the PLC channel, not just arbitrary values as found in the literature. We take the ICT OMEGA extended MIMO channel model [58] for the PLC channel in the simulation, with 100m in length and the frequency bandwidth in use as 1-300 MHz for the GE modelling.

4.4.1 Generating PLC GE Model

To obtain the Gilbert-Elliot model for the PLC channel, first, the three-phase ICT OMEGA extended MIMO PLC channel was fitted using MATLAB as a Log-normal distribution with mean $\mu_p = -2.97873$ and standard deviation $\sigma_p = 1.23376$. A MATLAB system-level simulator was created to calculate the mean holding time and the mean burst length. Consequently, the GE transition matrix values were calculated using equations eq. (4.22) and eq. (4.23). The channel model is implemented by launching 10^6 HF per frequency sample. The cut-off FER is chosen such that when the instantaneous FER is below the cut-off value then the channel is considered in the good state and vice versa. The system is run with no FEC implemented and the instantaneous FER is calculated using equation eq. (4.24). The GE transition matrix and state FER values were calculated to be as follows:

$$\mathbf{C} = \begin{bmatrix} 0.9466 & 0.0534 \\ 0.9090 & 0.0910 \end{bmatrix} \quad \mathbf{Fe} = \begin{bmatrix} e_0 \\ e_1 \end{bmatrix} = \begin{bmatrix} 2.8280 \times 10^{-05} \\ 0.7908 \end{bmatrix} \quad (4.54)$$

The system was run again however to find the channel GE model parameters in the presence of RS FEC. The BER at the input of the decoder for non-return to zero in case of coded signal is $\epsilon_{rsin} = Q(\sqrt{\eta_{rs} \times 2\gamma})$, where $Q(\cdot)$ is the Gaussian Q-function, $\gamma = |h|^2 E_b / N_0$

is the instantaneous received SNR, E_b is the power per transmit bit and N_0 is the mean PLC noise power [87]. η_{rs} is the RS coding rate and defined as $\eta_{rs} = k_{rs}/n_{rs}$; where n_{rs} is the block length after encoding and k_{rs} is the message length to be encoded. The maximum allowable transmit power is adopted as per the PSD requirements of the regulations of general EMC [92]. To illustrate the effect of RS FEC the BER expression in eq. (4.24) needs to be replaced by the one in [87]:

$$\epsilon_{rs} = \epsilon_{rs_{out}} = \sum_{m=t_{rs}+1}^{n_{rs}} \frac{m}{n_{rs}} \binom{n_{rs}}{m} \epsilon_{rs_{in}}^m (1 - \epsilon_{rs_{in}})^{n_{rs}-m} \quad (4.55)$$

where $t_{rs} = (n_{rs} - k_{rs})/2$ is the number of errors that the code can surely correct. Three different RS rates are considered: RS_1 (127, 121), RS_2 (127, 95), and RS_3 (127, 63), corresponding to the code rates ≈ 0.95 , 0.75 , and 0.5 respectively. For each of the RS rates, the GE transition matrix and state FER values were generated as follows:

$$\begin{aligned} \mathbf{C}_{rs_1} &= \begin{bmatrix} 0.9760 & 0.0240 \\ 0.9471 & 0.0529 \end{bmatrix} & \mathbf{Fe}_{rs_1} &= \begin{bmatrix} 9.6498 \times 10^{-06} \\ 0.8409 \end{bmatrix} \\ \mathbf{C}_{rs_2} &= \begin{bmatrix} 0.9891 & 0.0109 \\ 0.9677 & 0.0323 \end{bmatrix} & \mathbf{Fe}_{rs_2} &= \begin{bmatrix} 3.5994 \times 10^{-06} \\ 0.8635 \end{bmatrix} \\ \mathbf{C}_{rs_3} &= \begin{bmatrix} 0.9923 & 0.0077 \\ 0.9738 & 0.0262 \end{bmatrix} & \mathbf{Fe}_{rs_3} &= \begin{bmatrix} 2.9646 \times 10^{-06} \\ 0.8431 \end{bmatrix} \end{aligned} \quad (4.56)$$

4.4.2 Numerical Results

In this section, we aim to confirm the accuracy of the system mathematical model as well as study the impact of RU and IoT device densities on the e2e delay. To achieve this objective, a MATLAB full system-level simulator of the RAN system was set up as per the parameters in Tables 4.2 and 4.1. The parameters in Table 4.1 are fixed throughout all the simulations and set as per [40, 112]. The minimum arrival on-period (T_{on}) is set as per value in [109]. In the simulations, the system was run for 10^6 time rounds as shown in Figure 4.2. For the radio link, NB-IoT 5G technology parameters are adapted.

Table 4.1: System simulation fixed parameters

Parameter	Value
Radio bandwidth (Ω_a)	200 KHz
Radio AWGN variance (σ_a^2)	20 dBm
Radio sub-carrier spacing (Δf_a)	15 KHz
Radio Modulation scheme	QPSK
Radio interface Scheduler	Greedy scheduler
RU transmit power [113, 114]	33-36 dbm
IoT device transmit power (P_a)	23 dBm
IoT radio data frame size ($\mathcal{D}_a$) [112]	28 bits
Radio power control factor (ϵ_a)	0.8
Radio path-loss exponent (α_a)	4
Data packet arrival rate (γ_d)	1 HF/8.4s (33.3 b/s)
eCPRI Processing delay (T_{proc}) [96]	10 μs
HARQ Processing delay (T_{pH}) [96]	4 μs
Radio and PLC guard band ratio (gb)	5%
Power Line Length (L_p)	100 m
MIMO PLC Ports (N_p)	6 Tx/RX ports
PLC bandwidth (BW) / Port (Ω_p)	1 - 300 MHz
IND Threshold (Th)	10 dB
IND Mean power (P_{mean})	-115 dB
Mean eCPRI frame size ($\overline{\mathcal{D}_p}$)	32770 bytes
PLC GE cut-off FER (e_{cut})	10^{-2}
Mean eCPRI on-period (T_{on})	5 Ts
Feedback delay (t_f)	3 Ts

Table 4.2: System simulation variable parameters

Parameter	Range	Default
RUs density / Km^2 (λ_b) [115]	5 - 50 / Km^2	20 / Km^2
IoT devices density / Km^2 (λ_d) [116]	1K - 100K / Km^2	40K / Km^2
RS-FEC (n_{rs}, k_{rs})	(127,121), (127,95), (127, 63)	(127, 121)
Mean PLC traffic load (ρ_p)	0.2 - 0.7	0.5
HARQ retrans. Limit (L_r)	2 - 10	4
SR Window (W)	2 - 16 HF's	5 HF's

The radio link settings are based on NB-IoT 5G characteristics, which provide efficient and reliable connectivity for IoT devices. This 3GPP cellular technology is ideal for indoor use, as it can penetrate walls, buildings, and obstacles [75, 76]. It is power-efficient, suitable for battery-powered devices and operates efficiently with a narrow bandwidth. The values for τ_a are set based on the SINR key performance indicators (KPIs) in 5G standard for excellent, good, fair, and poor coverage [77].

The outcomes of the system simulations show that as the IoT device density increases, the amount of data ready to send in one slot increases. As more data is being sent in the same time round, the air delay decreases until the radio resources become the limiting factor. The decrease in the air interface delay in turn decreases the overall e2e delay. Furthermore, increasing the number of RUs, serving the same number of IoT devices, increases the number of served IoT devices, but interference becomes the hindering factor as the SINR threshold becomes more severe.

We evaluate the impact of employing PLC front-hauling on the system's ability to support massive IoT applications. Through simulations with varying numbers of devices and RUs per floor, our proposed system demonstrates the achievement of stringent one-way delay requirements with RS FEC. The detailed results presented in Figures 4.7 through 4.11 validate the accuracy of our mathematical models and their suitability for studying various aspects of the system.

Figure 4.7a shows the coverage probability over the air interface which, as expected, increases as the number of RUs increases. The decaying increase in the probability of coverage reflects the increase in interference despite the increased radio resources. At $\tau_a = -20$ dB, and using NB-IoT parameters, the maximum of devices are scheduled at a time based on the size of the serving radio bandwidth. However, as the value for τ_a increases, the interference becomes the main limiting factor for the number of devices being served. The simulations show that assumptions used to develop the model has minimal effect on the accuracy as proven by the close match between analysis and simulation findings.

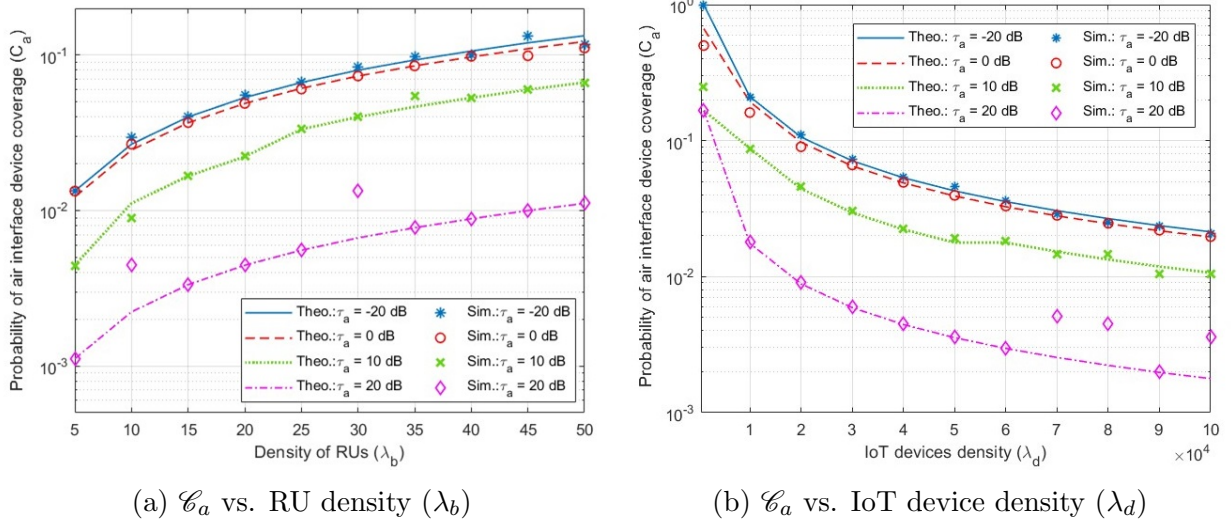


Figure 4.7: Coverage Probability vs. RU & IoT device densities for different SINR thresholds (τ_a)

Figure 4.7b shows the probability of coverage on the air interface that, as expected, decreases as the number of devices increases. The decrease in the probability of coverage reflects the increase in interference and the radio bandwidth limitation as more devices compete for the shared resources at the RUs. At $\tau_a = -20$ dB, and using NB-IoT parameters, a maximum of 12 devices can be scheduled at a time due to the size of the serving radio bandwidth. However, as the value for τ_a increases, the interference becomes the main limiting factor for the number of devices being served.

Since the delay over the air interface is determined by the coverage probability and the inter-cellular interference, Figure 4.8a illustrates that T_{air} decreases with the increase of radio resources. The reasoning behind this is more data is being received from the devices each time-round. Nevertheless, at higher SINR thresholds the effect of increasing the density of RUs and the radio resources is not significant on the delay due to the increased interference.

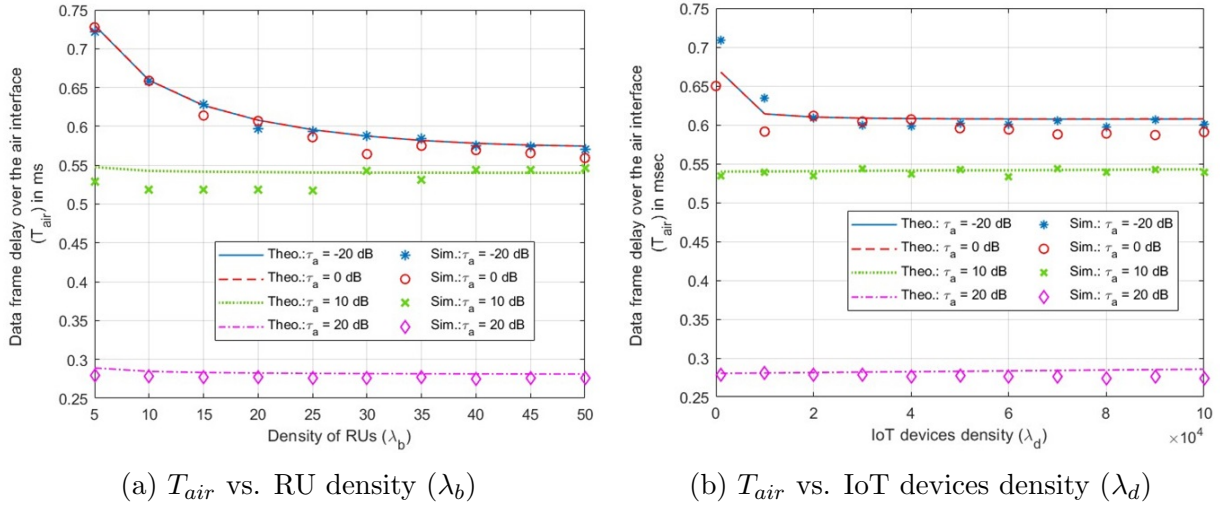
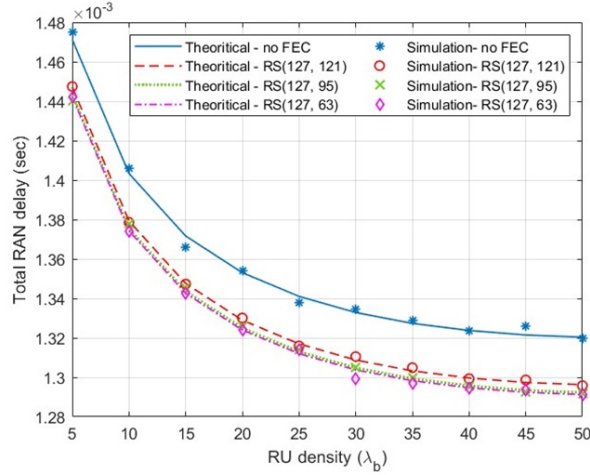


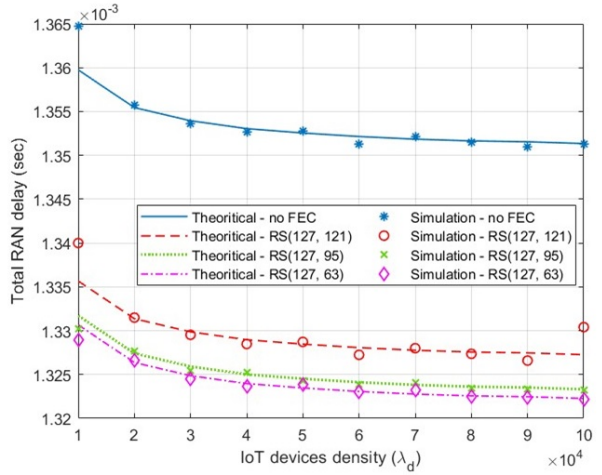
Figure 4.8: Delay over air interface (T_{air}) vs. RU & IoT device densities for different SINR thresholds (τ_a)

Furthermore, in Figure 4.8b, the increase of the number of IoT devices within the network does not have a great impact on the air interface delay (T_{air}). The explanation behind it is that the system is limited by the radio resources i.e. the number of RUs in use. As the SINR thresholds decrease the delay over the radio link increases due to interference and saturation of radio resources. However, the system consistently meets the delay requirements ($T_{air} < T_a$) in both Figures 4.8a and 4.8b.

Looking into the e2e radio access delay (T_{e2e}), as shown in Figures 4.9a and 4.9b, the system meets the delay requirements for 5G and eCPRI [117]. Despite, the expectation that the e2e delay would increase, the delay over the radio channel is the dominating factor and therefore, the e2e delay behave accordingly. Moreover, the ceil function when calculating the access delay components to one T_s overshadows the increase in the access delay due to the increase in the RU density.



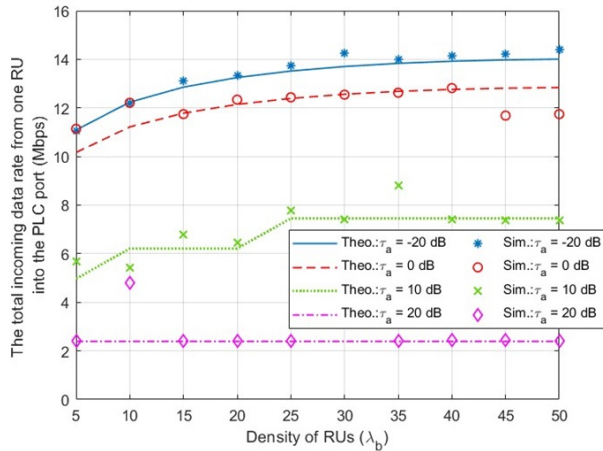
(a) T_{e2e} vs. RU density (λ_b)



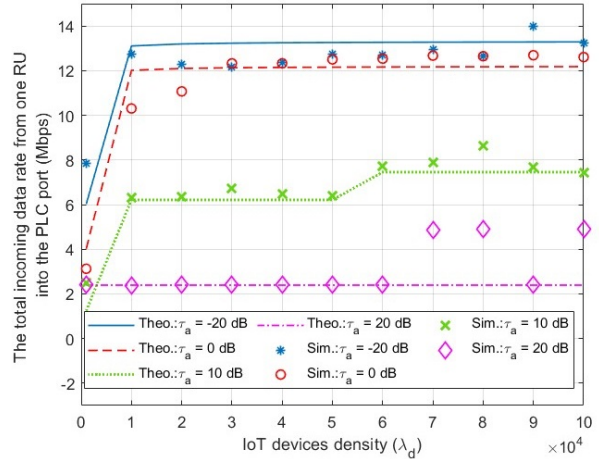
(b) T_{e2e} vs. IoT devices density (λ_d)

Figure 4.9: Total radio access delay (T_{e2e}) over air and PLC links interface vs. RU & IoT device densities for different SINR thresholds (τ_a)

Figures 4.10a and 4.10b, show that the total incoming data rate into on PLC SISO link from one RU. The PLC-based front-haul is capable to handle the total traffic incoming after conversion into HF even with increasing the RU density or the IoT device density. As SINR threshold increases (τ_a), the incoming data rate decreases as the number of covered devices per RU decreases due to interference limitations.



(a) R_p^j vs. RU density (λ_b)



(b) R_p^j vs. IoT device density (λ_d)

Figure 4.10: Data rate from on RU into the PLC link through the CPG incoming buffer (R_p^j) vs. RU & IoT device densities for different SINR thresholds (τ_a)

Focusing on the access part of the network, in Figure 4.11a, the use of RS FEC ensures

that the delay meets the eCPRI requirements. The PLC-based front-haul link could achieve the eCPRI low class of service (CoS) maximum one-way delay constraint of $100ms$ with no RS coding. However, the delay requirement of $1ms$ may be satisfied for medium CoS only with the use of RS coding. It is worth mentioning that increasing the coding rate does not have a major improvement in the delay. The discrepancies between the analytical and system simulation results are due to the imposed ceil function, yet, it is shown that these discrepancies are not severe as they do not exceed one T_s .

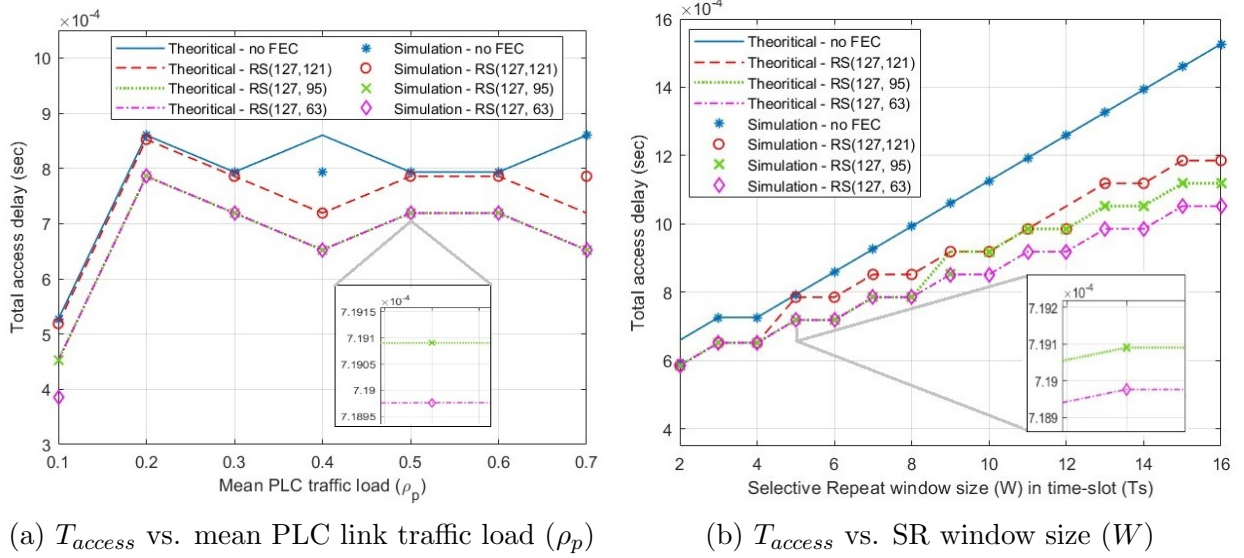


Figure 4.11: Total access delay over PLC link (T_{access}) for different RS coding rate

It is essential to mention that access delay requirements for eCPRI High CoS were found to be achievable using a different 5G numerology. That is, for eCPRI High500 ($500\mu s$), High200 ($200\mu s$), and High100 ($100\mu s$), the CoSs can be met when the T_s is set to the 5G OFDM symbol duration using symbols/ slots ($\mu = 1, 2$, & 3), respectively. The PLC-based front-haul, however, does not meet the requirements for eCPRI High25 requirements of ($25\mu s$) with such a high total processing delay of ($14\mu s$).

With CPRI Maximum round trip delay requirement (R-26) [73], considering the maximum I and Q sample width of 20 bits, one-way transmission is subject to the maximum delay of $9.71ms$ for option 10 with a fixed clock, which is equivalent to $\sim 146 T_s$ for option 10. It is shown in our simulation result that such requirements can be met even without the use of RS code and with any W and L_r values.

When the SR window is increased, the re-sequencing delay increases linearly thus affecting the total access delay. In Figure 4.11b, the access delay does not meet the eCPRI medium CoS (1ms) requirement when no RS coding once the window size increases above 10 slots. With the use of RS coding, the system could meet the requirement with a window size of up to 10 slots. Still, the increase in RS coding rate does not improve the system performance significantly. Therefore, it is recommended to keep the window size up to 10 slots while using RS coding.

It is noteworthy that increasing the re-transmission rate (L_r) increases the access delay linearly as shown in Figure 4.12. Moreover, as the re-transmission rate (L_r) increases the PDF for the re-sequencing delay changes. The effect of (L_r) over the total access delay is within the range of one T_s for each RS coding rate. If more than one HF is sent with every transmission, the effect of changing L_r would be of higher significance.

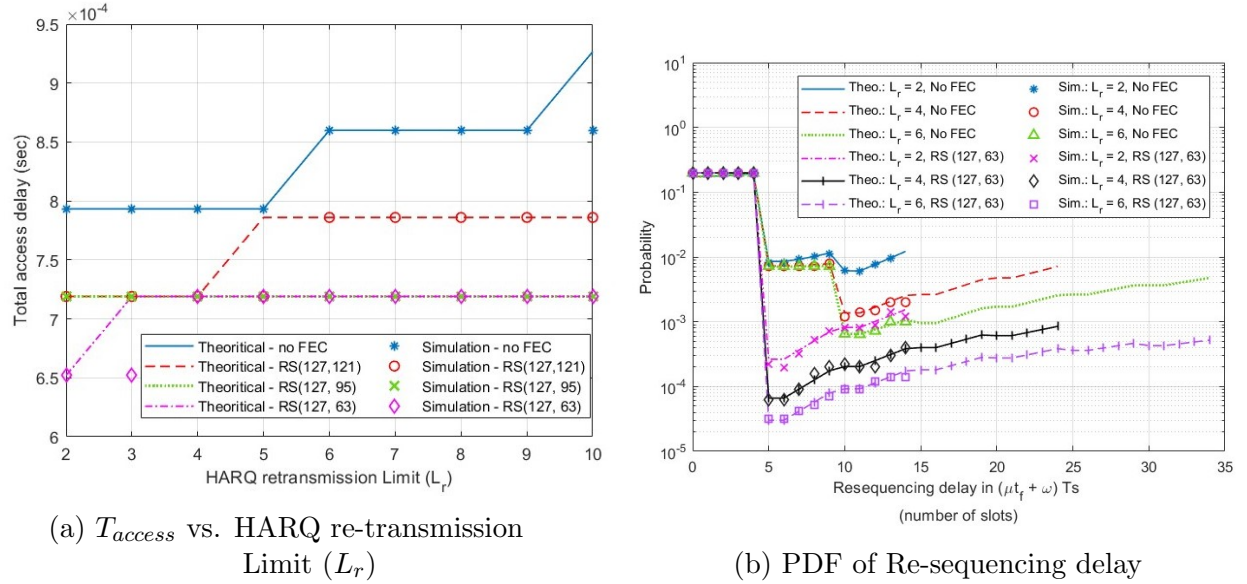


Figure 4.12: Total access delay over PLC link (T_{access}) vs HARQ re-transmission Limit (L_r)

The utilization of the Ideal SR approximation in the computation of queuing, transport, and re-sequencing delays has little impact, as shown by the simulation results presented above. The feedback delay has no impact on the calculation of the queue length; however, it has a direct effect on the re-transmission rate and consequently, the queue length in the waiting buffers as well as the transmission and re-sequencing delays. Since the feedback delay is shorter than the average duration of the PLC channel's bad period, a faulty packet will

be re-transmitted during the same bad period. Hence, it is probable that the re-transmitted packet may experience same faults in the PLC channel. With the increase in the feedback delay, the influence of erroneous correlations diminishes, leading to a decline in the average throughput ($\overline{\mathcal{T}_r}$) but the product $t_f \cdot \overline{\mathcal{T}_r}$ pretty much stays constant.

4.5 Chapter Summary

This chapter explores the use of low-voltage PLC links for facilitating eCPRI-based front-hauling within 5G IoT indoor mobile coverage environments, utilizing a split C-RAN architecture. The primary objective of this research is to examine the impact of various factors, such as the number of wireless IoT devices, bandwidth, and transmission technology, on the delay performance of the proposed system. To achieve this goal, we develop detailed mathematical models that draw insights from queuing theory, stochastic geometry, and Markov models. The accuracy of these models is substantiated through extensive system-level simulations, and the analytical results comprehensively address both the radio and access delay performances. Additionally, we validate the system's capability to efficiently support IoT indoor cellular applications and rigorously evaluate the feasibility of the proposed PLC-based front-hauling system, particularly in light of the stringent delay requirements set by the CPRI/eCPRI standards.

Chapter 5

Multi-objective Optimization of IoT RAN using NSGA-II Algorithm

5.1 Radio Access Physical Network

Consider a vast number of wireless low mobility, battery powered IoT devices, randomly dispersed across the inside of a multi-story service facility as in [101, 118]. The IoT devices are used to collect data that is error and delay-sensitive to be communicated to a variety of service owners. In the network, a split C-RAN system is investigated, wherein the BS is divided into three units: a CU, a DU, and RUs [18]. The front-haul link consists of two segments: a Mid-haul/front-haul (I) segment that connects the CU and DU, and a PLC based front-haul (II) segment that connects the DU with various indoor RUs as show in Figure 5.1. The front-haul (I) technology is beyond the scope of this research. RUs are randomly distributed around the building to serve the IoT devices. Each RU is hooked to a power outlet that provides both power and access to the PLC based front-haul. The front-haul (II) (referred to as front-haul) transports eCPRI/CPRI communications in regular Ethernet frames/ TDM hyper-frames between the indoor RUs and the DU. Unless otherwise specified, the remainder of this chapter refers to eCPRI/CPRI messages as HF [101].

In the following analysis, assuming there is no interference among the different floors, the focus is on maximizing the number of satisfied IoT devices whose data is carried over the PLC front-haul while decreasing devices' energy consumption and meeting the delay requirements. Consider an uplink network consisting of a layer of cellular RUs with locations modelled as

HPPP: $\Phi_b = \{b_0, b_1, b_2, \dots\}$ with RU density λ_b , where b_j indicates the 2D-location of the j^{th} RU in the network where $1 \leq j \leq B_{max}$ and B_{max} is the maximum number of RUs the MIMO 3-phase PLC front-haul supports. The network connects a large number (M_{tot}) of identical, battery-powered IoT devices that are randomly placed throughout the building. The devices' 2D-locations are modelled as an independent HPPP denoted as $\Phi_d = \{d_0, d_1, d_2, \dots\}$ with device density λ_d , where d_i indicates the location of the i^{th} device. Devices associate to a RU to form a cluster [94, 95]. Initially, each device is inactive until its data is ready for transmission at which point, the device becomes active. Data frames are generated based on a Poisson process with parameter $\gamma_d \in (0, \infty)$ and are sent over the air interface [101]. Active devices' 2D-locations (Φ_a) are modelled as a thinned independent HPPP of the devices' location independent HPPP, such that $\Phi_a \in \Phi_d$ with a density of $\lambda_a = \mathcal{A} \times \lambda_d$, where $\mathcal{A}$ is the probability that a device has a data-frame to send and becomes active, with total active devices (M_a).

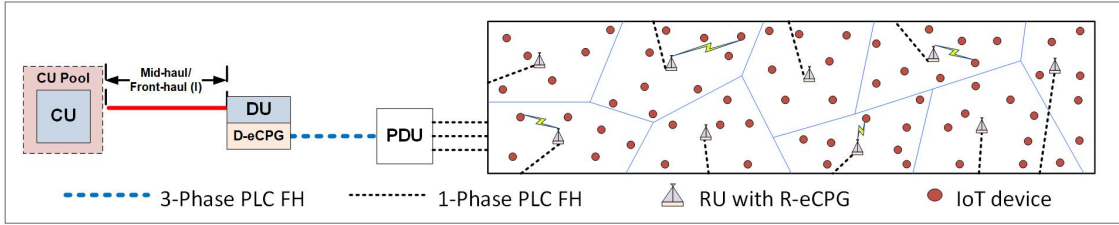


Figure 5.1: Illustration of Radio access network architecture

Each RU transforms the received uplink data from all associated active IoT devices within its coverage into HF and transmits them to the DU over the PLC-based front-haul. Let the set of devices associated with the j^{th} RU located at b_j be denoted as $\Psi_{b_j} = \{d_1, d_2, \dots, d_m\}$, where d_i is the location of the i^{th} device associated with the j^{th} RU. A binary indicator variable $x_{i,j}$ is employed which indicates the state of the i^{th} device's association with the j^{th} RU. If the device is associated with that specific RU, the variable is set to 1 and to 0 otherwise. The number of devices associated to the j^{th} RU is denoted as M_j , where $\sum_j M_j \leq M_a$ and $M_j = \sum_i x_{i,j}$. Lastly, let $\Psi_b = \{\Psi_{b_1}, \Psi_{b_2}, \Psi_{b_3}, \dots, \Psi_{b_B}\}$ denote the super location sets of the assigned active devices to RUs.

5.1.1 Wireless Transmission Model

At each RU, the uplink radio spectrum has bandwidth Ω_a which is divided into $n = \{1, 2, 3, \dots, \mathcal{N}_a\}$ orthogonal sub-channels of equal bandwidth (ω_a) which will be referred to as Radio sub-channels (RSCs). Each active device associated with a RU is scheduled upon a sub-channel until the maximum number of devices denoted as M_j^{max} is scheduled where $M_j^{max} = \min\{M_j, \mathcal{N}_a\}$. Another binary indicator $s_{i,j}^n$ is set to 1 to indicate if the i^{th} device is allocated to the n^{th} RSC of the j^{th} RU and is set to 0 to indicate otherwise. A device can only associate with a single RSC and the association of a device to a certain RU is dependent on the ability that this RU can provide the specified QoS level for the device.

The attenuation of transmitted signal over the air interface is governed by a general power-law path-loss model. The signal power degrades at rate $D^{-\alpha_a}$, where D is the propagation distance and α_a is the path-loss exponent. The radio channel is considered a Rayleigh fading with an exponentially distributed instantaneous power gain ($g_{i,j} \sim \exp\{1\}$). The radio channel capacity is an interference limited channel where the power gains are i.i.d as well as independent of the device-RU distance. The radio channel adds AWGN with received noise of power σ_a^2 . Since devices are assumed to be battery powered, therefore the transmit power $p_{i,j}^n$ is limited such that $0 < p_{i,j}^n \leq P_a^{max}$, where, P_a^{max} is the device maximum transmission power. The transmission power of each device is set such that the number of devices' goodput data over the PLC front-haul is maximized. Consequently, considering i^{th} device located at d_i connected to the j^{th} RU located at b_j , the received power at the RU from the device is calculated as:

$$q_{i,j}^n = p_{i,j}^n ||d_i - b_j||^{-\alpha_a} g_{i,j} \quad (5.1)$$

Using Shannon's calculation for radio channel capacity, the attainable data rate for the i^{th} device on the j^{th} RU on the n^{th} RSC is calculated as:

$$R_{i,j}^n = \omega_a \times \log_2 \left(1 + \frac{p_{i,j}^n ||d_i - b_j||^{-\alpha_a} g_{i,j}}{\sum_{b_h \in \Phi_b/b_j} \sum_{d_m \in \Psi_{b_h}} \frac{s_{m,h}^n p_{m,h}^n ||d_m - b_h||^{-\alpha_a} g_{m,h}}{p_{m,h}^n}} \right) \quad (5.2)$$

where Y/y_i represents all items in set Y excluding item y_i .

5.1.2 Transmission over PLC-based Front-haul

Each RU aggregates all data-frames received from all covered devices within its cluster and relays them over the PLC front-haul after converting them into HF [73, 74]. The PLC front-haul connecting the RUs to the DU is a 3-phase MIMO link, which has $N_p = 6$ ports. In the below analysis, it is assumed that all RUs have access to all 3-phases ports of the PLC link for simplification. Moreover, it is assumed that the maximum transmit power over each PLC port is used while ensuring that the regulations of general EMC have been respected; such that the total transmit power over each PLC port is below the maximum PLC PSD P_p^{mask} [92]. As in [40, 101] a SVD function module is placed at the CPG/eCPG which transforms the originally 6x6 MIMO PLC channel of the three-phase line to a diagonal 6 SISO parallel channels. Therefore, the MIMO PLC channel can be expressed such that $\Phi_p = \{\xi_1, \xi_2, \xi_3, \xi_4, \xi_5, \xi_6\}$ which denotes the ports for the 3-phase PLC link. Such that the ports are referred to as follows:

$$\begin{aligned}\xi_1 &= L_1N - L_1N, \quad \xi_2 = L_1PE - L_1PE, \\ \xi_3 &= L_2N - L_2N, \quad \xi_4 = L_2PE - L_2PE, \\ \xi_5 &= L_3N - L_3N, \quad \xi_6 = L_3PE - L_3PE\end{aligned}\tag{5.3}$$

where, L_1N is port between live connector of the 1st phase and the neutral connector, and L_1PE is port between live connector of the 1st phase and the protective earthing connector.

The bandwidth over each port is set to Ω_p divided using FDMA into $k = \{1, 2, 3, \dots, \mathcal{N}_{PLC}\}$ PLC FDMA sub-channels (PSCs) each of bandwidth $\omega_p = \Omega_p/\mathcal{N}_{PLC}$. ω_p is a multiple of the chip rate for CPRI and eCPRI, where $\Delta f = 3.84 \text{ MHz}$. Let the set of RUs associated with the p^{th} port, referred to as ξ_p , be denoted as $\Psi_{p_p} = \{b_1, b_2, \dots, b_j\}$, where b_j is the location of the j^{th} RU. The number of RUs associated to the p^{th} port is denoted as B_p , where $\sum_{p=1}^{N_p} B_p = B^{max}$. Each RU associated with a PLC port is scheduled upon at least one PSC until the maximum number of RUs denoted as B_p^{max} is scheduled where, $B_p^{max} = \min\{B_p, \mathcal{N}_{PLC}\}$. A binary indicator variable $y_{j,p}$ is employed which indicates the state of the j^{th} RU's association with the p^{th} port. If the RU is associated

to that specific PLC port, the variable is set to 1 and is set to 0 otherwise. Lastly, let $\Psi_P = \{\Psi_{p_1}, \Psi_{p_2}, \Psi_{p_3}, \dots, \Psi_{p_6}\}$ denote the super location sets of the assigned RU to PLC 3-phase ports. A fully loaded network of RUs is considered, in which all RUs have to be scheduled on at least one PSC at a PLC port [101]. A binary indicator variable $v_{j,p}^k$ indicates if the j^{th} RU is scheduled on the k^{th} PSC of the p^{th} PLC port.

For the j^{th} RU, the aggregated incoming data rate from all its covered devices after conversion into eCPRI is calculated as [101]:

$$R_{j,p}^k = \Lambda_{ap} \cdot \sum_{n=1}^{\mathcal{N}_a} R_{i,j}^n \quad (5.4)$$

where Λ_{ap} indicates the rate of conversion between the data rate over the air interface and the required data rate for the HF over the PLC link calculated as in [101]. Based on the eCPRI split I_U (or II_D in downlink), Λ_{ap} is calculated as $\Lambda_{ap} = 2 \cdot Q_F \times \chi_e \times \nu_p$ [46, 103]; where, $Q_F = 9$ is the quantizer resolution in the frequency domain for LTE [46]. χ_e is the header of the eCPRI-frame redundancy factor including higher IP/Ethernet network layers. The PLC link peak utilization ν_p is set to $\nu_p = 1$ assuming a worst-case scenario [46, 103].

As SVD of the PLC channel is used, the capacity per PSC eq. (5.5a), per port eq. (5.5b) and over the PLC eq. (5.5c) can be calculated respectively as follow [40, 79]:

$$C_{j,p}^{max} = \omega_p \cdot \log_2 \left(1 + \frac{p_p(f_k) \lambda_p(f_k)}{N_{R_x}^p(f_k) \cdot N_p} \right) \quad (bps) \quad (5.5a)$$

$$C_p^{max} = \sum_{k=1}^{\mathcal{N}_{PLC}} \omega_p \cdot \log_2 \left(1 + \frac{p_p(f_k) \lambda_p(f_k)}{N_{R_x}^p(f_k) \cdot N_p} \right) \quad (bps) \quad (5.5b)$$

$$C_{MIMO} = \sum_{p=1}^{N_p} \sum_{k=1}^{\mathcal{N}_{PLC}} \omega_p \cdot \log_2 \left(1 + \frac{p_p(f_k) \lambda_p(f_k)}{N_{R_x}^p(f_k) \cdot N_p} \right) \quad (bps) \quad (5.5c)$$

where $\lambda_p(f_k)$ is the eigen values of p^{th} port of the $\mathbf{H}_{PLC} \mathbf{H}_{PLC}^\dagger$ at frequency f_k . $\mathbf{H}_{PLC}$ is the 3-phase PLC MIMO channel and f_k is the carrier frequency of the k^{th} PSC. The transmitted power for each PLC port is $p_p(f_k)$, which is set to the maximum allowable value as per the PSD criteria of general EMC rules [92]. $N_{R_x}^p(f_k)$ is the noise at the receiver port, where $N_{R_x}^p = n_{imp}^p + n_{bg}^p$. The impulsive and background PLC noises at the receiver port are n_{imp}^p

and n_{bg}^p , respectively. $\mathcal{N}_{PLC}$ is the number of PSCs of one port in the frequency range, ω_p is the bandwidth of one PSC and N_p is the total number of used PSCs [79]. $\dagger$ represents the Hermitian operator of the PLC matrix.

Through data rate matching presented in [40], the higher limit of goodput that the PLC port can support in the presence of the functionalities of CPG/eCPG can be calculated as:

$$R_p^{max} = \eta_p \times C_p^{max} \quad (5.6)$$

The factor to match the required data rate over the PLC port due to SR-based HARQ, IND and RS FEC is denoted as $\eta_p = \eta_H^p \times \eta_I^p \times \eta_{rs}^p$. The increase in required data rate over the PLC link due to the SR HARQ re-transmission is defined as η_H^p . The effect of SR-based HARQ over the data rate with l_r^p re-transmissions per port is defined as [40]:

$$\eta_H^p = \frac{1}{1 + \mathbb{E}[l_r^p]} = (1 - \overline{P_{ers}}) (1 - \overline{P_{ers}}) \quad (5.7)$$

where $\overline{P_{ers}}$ is the improved mean FER over the log-normal PLC channel, with mean $\mu_p = -2.97873$ and standard deviation $\sigma_p = 1.23376$, due to the use of RS (n_{rs}^p, k_{rs}^p) [40]. This is irrespective of the number l_r^p as only one data-frame copy is transmitted at a time and each data-frame is acknowledged by one ACK.

The effect of the overhead introduced due to the use of RS (n_{rs}^p, k_{rs}^p) coding over the PLC port is $\eta_{rs}^p = k_{rs}^p/n_{rs}^p$. The effect of IND is defined as $\eta_I^p = 1 - P_{imp}^p$, where, P_{imp}^p is the probability that an impulse over the PLC port is detected and is defined as $P_{imp}^p = \mathbb{P}[A_{imp} > \sqrt{P_{mean}} / \Gamma_i] \times \mathbb{P}[t_i \leq \Delta t]$, where A_{imp} is the amplitude of the impulsive noise and t_i is the inter-arrival time between the impulses with an exponential nature with mean $\mu_i = 10.1ms$ [44]. Based on the distribution models mentioned in [40, 44], P_{imp}^p can be expressed as:

$$P_{imp}^p = \left(1 - \int_0^{A_{imp}} \sum_{z=0}^{\infty} \frac{\exp\{(-A)\} \cdot A^z}{z!} \frac{\exp\left\{\left(\frac{-a^2}{2\sigma_z^2}\right)\right\}}{\sqrt{2\pi\sigma_z^2}} da \right) \times \left(1 - \exp\left\{\left(\frac{-\Delta t}{\mu_i}\right)\right\} \right), \quad \text{with} \quad \sigma_z^2 = \sigma^2 \left(\frac{(z/A) + \Gamma_i}{1 + \Gamma_i} \right) \quad (5.8)$$

where the parameter A is the impulsive index and the parameter Γ_i is the ratio of the background-to-impulsive noise and are set such that $A = \Gamma_i = 0.01$ [44]. Δt is the rate at which the IND functionality listens for an impulse event [40].

5.2 Problem Formulation

In a split C-RAN system, the front-haul link is an essential high-speed, low-latency link. Given the limited bandwidth of the front-haul link, optimizing its usage means satisfying the largest number of IoT devices. In the following, we develop a MOO problem in which the goodput over the PLC front-haul, in terms of the number of satisfied IoT devices, is maximized, while meeting all the QoS requirements. In order to maximize the number of satisfied IoT devices, the following QoS requirements should be met:

- 1) The IoT device has to be active, associated to a RU, and scheduled onto a RSC.
- 2) The device should be within the coverage range of a RU and its data rate is above the minimum requirement.
- 3) The transmit power for each device should be below the maximum threshold to avoid battery life drainage.
- 4) The delay over the PLC access link is within the requirements of eCPRI standard regulations.
- 5) The data of the device is received at the DU successfully over the PLC front-haul link.
i.e. all the data from all RUs is received successfully at the DU.

5.2.1 Optimization Objectives

Consequently, this optimization problem, referred to as **P1**, is a joint device-RU, RU-PLC port association, RAN power control and allocation problem. The objectives of this MOO problem are:

- 1) Maximize the goodput in-terms of the number of satisfied served IoT devices.

- 2) Minimize the transmit power of all served IoT devices to prolong battery life.
- 3) Minimize the delay over the PLC front-haul by minimizing the mean delay of all PLC ports.

5.2.2 Optimization Outcomes

Furthermore, the MOO problem outcome is a set of matrices as follow:

- 1) IoT device to RU association matrix $\mathbf{X} = [x_{i,j}] \forall b_j \in \Phi_b, \& d_i \in \Phi_d$.
- 2) Device to RSC scheduling matrix in the associated RU $\mathbf{S} = [s_{i,j}^n] \forall b_j \in \Phi_b, d_i \in \Phi_a, \& 1 < n \leq \mathcal{N}_a$.
- 3) A matrix for the transmission power of all served devices $\mathbf{P}_a = [p_{i,j}^n] \forall b_j \in \Phi_b, d_i \in \Phi_a, \& 1 < n \leq \mathcal{N}_a$.
- 4) RU to PLC port association matrix $\mathbf{Y} = [y_{j,p}] \forall b_j \in \Phi_b, \& \xi_p \in \Phi_p$.
- 5) RU to PSC scheduling matrix in the associated port $\mathbf{V} = [v_{j,p}^k] \forall b_j \in \Phi_b, \xi_p \in \Phi_p \& 1 < k \leq \mathcal{N}_{PLC}$.
- 6) A vector for SR HARQ re-transmission rate per PLC port $\mathbf{L}_r = [l_r^p] \forall 0 < p \leq N_p$
- 7) A vector for the SR HARQ window size per PLC port $W = [w^p] \forall 0 < p \leq N_p$
- 8) A vector for the RS code rate per PLC port $\mathcal{H}_{rs} = [\eta_{rs}^p] \forall 0 < p \leq N_p$

5.2.3 Optimization Constraints

The optimization problem **P1** is governed by multiple constraints that are as follows:

C1: This constraint indicates that the i^{th} device located at d_i is in the coverage of the j^{th} RU.

C2: This constraint ensures that a device is associated only to one RU at a time.

- C3*: This constraint clarifies the connection between resource allocation and user association dynamics; where a device is only permitted to be scheduled on a RSC of a RU if it is associated with that specific RU.
- C4*: This constraint ensures that a device is scheduled on one RSC within its RU at a time.
- C5*: It ensures that a device is assigned service if it can acquire an average data rate over the radio channel that exceeds its minimum average rate requirement $R_a^{min} = \omega_a \times \log 2(1 + \tau_a)$; where τ_a is the received SINR threshold used to determine the RU coverage range.
- C6*: The constraint guarantees that no device is permitted to transmit at a higher power level than maximum device power P_a^{max} .
- C7*: This constraint only allows the data from the j^{th} RU located at b_j to be carried to DU over p^{th} PLC port of the front-haul.
- C8*: This constraint ensures that a RU is associated only with one PLC port at a time.
- C9*: This constraint confirms that a RU is scheduled a PSC of a port if it is associated with that specific PLC port.
- C10*: This constraint ensures that a RU has to be scheduled on at least one PSC within the PLC port at a time.
- C11*: The maximum allowable transmit power is used for each PLC port while respecting the regulations of the general EMC. In which the PSD is set to $-50 \text{ dbm}/\text{Hz}$ up to 30 MHz and to $-80 \text{ dbm}/\text{Hz}$ for higher frequencies as mentioned in previous chapters [92].
- C12*: It guarantees that the eCPRI-data from all RUs associated to a PLC port are assigned service and their total data rate over the PLC channel does not exceed the maximum rate R_p^{max} available on the PLC port where $R_p^{max} = \eta_p \times C_p^{max}$.
- C13*: The SR HARQ window w^p for each PLC port is set such that it is at least the size of 2 HFs up to 12 HFs; as per our findings in [101].

- C14:* This constraint ensures that HARQ re-transmission is enabled by setting the limit $1 \leq l_r^p \leq 10$. Although theoretically the limit of re-transmission can be infinite, if a HF is required to be re-transmitted more than 10 times, the system can depend on higher layers re-transmission mechanisms.
- C15:* Setting the RS coding rate η_{rs}^p for the PLC channel of each port is within the rate of 0.75 to 1, where the block length n_{rs}^p is set to 127 bits. The channel coding rate is set within such range since in the previous chapter, the outcomes show that further decrease in the code rate (i.e. increase in code word size) would not improve the system performance.
- C16:* This constraint is set to ensure that the eCPRI delay requirement ($T_p^{max} = 1ms$) over the access front-haul link is met (T_{PLC}^p) which is calculated as clarified in section 5.2.4 below [117].

$$\begin{aligned}
\mathbf{P1: } \mathcal{U} = & \max_{\mathbf{X}, \mathbf{Y}, \mathbf{V}, \mathbf{S}, \mathbf{P}_a} \sum_{b_j \in \Phi_b} \sum_{d_i \in \Psi_j} x_{i,j} + \min_{\mathbf{X}, \mathbf{Y}, \mathbf{V}, \mathbf{S}, \mathbf{P}_a} \sum_{b_j \in \Phi_b} \sum_{d_i \in \Psi_j} x_{i,j} \cdot \sum_{n=1}^{\mathcal{N}_a} s_{i,j}^n p_{i,j}^n + \min_{\mathbf{L}_r, \mathbf{W}, \mathcal{H}_{rs}} \frac{1}{N_p} \sum_{p=1}^{N_p} T_{PLC}^p \\
\text{s.t. } x_{i,j} = & \begin{cases} \in \{0, 1\}, & \text{if } d_i \in \Psi_j, \forall b_j \in \Phi_b \\ 0, & \text{otherwise} \end{cases} \Rightarrow C1 \\
\sum_{b_j \in \Phi_b} x_{i,j} \leq & 1 \quad \forall d_i \in \Phi_d \Rightarrow C2 \\
s_{i,j}^n = & [0, x_{i,j}] \quad \forall d_i \in \Phi_d \ \& \ b_j \in \Phi_b \Rightarrow C3 \\
\sum_{d_i \in \Psi_j} s_{i,j}^n \leq & 1 \quad \forall b_j \in \Phi_b \ \& \ 0 < n \leq \mathcal{N}_a \Rightarrow C4 \\
x_{i,j} \cdot \sum_{n=1}^{\mathcal{N}_a} s_{i,j}^n R_{i,j}^n \geq & R_a^{min} \quad \forall b_j \in \Phi_b \ \& \ 0 < n \leq \mathcal{N}_a \Rightarrow C5 \\
0 \leq x_{i,j} \cdot \sum_{n=1}^{\mathcal{N}_a} s_{i,j}^n p_{i,j}^n \leq & P_a^{max} \quad \forall d_i \in \Phi_d \ \& \ b_j \in \Phi_b \Rightarrow C6 \\
y_{j,p} = & \begin{cases} \in \{0, 1\}, & \text{if } b_j \in \Psi_{p_p}, \forall \xi_p \in \Phi_p \\ 0, & \text{otherwise} \end{cases} \Rightarrow C7 \\
\sum_{\xi_p \in \Phi_P} y_{j,p} = & 1 \quad \forall b_j \in \Phi_b \Rightarrow C8 \\
v_{j,p}^k = & [0, y_{j,p}] \quad \forall b_j \in \Phi_b \ \& \ \xi_p \in \Phi_p \Rightarrow C9 \\
\sum_{b_j \in \Psi_{p_p}} v_{j,p}^k \geq & 1 \quad \forall \xi_p \in \Phi_p, \ k \Rightarrow C10 \\
\sum_{k=1}^{\mathcal{N}_{PLC}} P_p(f_k) = & P_p^{mask}(f_k) \quad \forall \xi_p \in \Phi_P, \ 0 < p \leq N_p, \ \& \ 1 \leq f_k \leq 300 \text{ MHz} \Rightarrow C11 \\
y_{i,p} \cdot \sum_{k=1}^{\mathcal{N}_{PLC}} v_{j,p}^k R_{j,p}^k \leq & R_p^{max} \quad \forall 0 < p \leq N_p \ \& \ \xi_p \in \Phi_p \Rightarrow C12 \\
2 \leq w^p \leq & 12 \quad \forall 0 < p \leq N_p \Rightarrow C13 \\
0 < l_r^p < & 10 \quad \forall 0 < p \leq N_p \Rightarrow C14 \\
0.75 < \eta_{rs}^p < 1 \quad \text{where, } \eta_{rs}^p = & k_{rs}^p / n_{rs}^p \ \& \ n_{rs}^p = 127 \text{ bits, } \forall 0 < p \leq N_p \Rightarrow C15 \\
T_{PLC}^p \leq & T_p^{max} \Rightarrow C16
\end{aligned}
\tag{5.9}$$

As a result of the presence of indicative binary and discrete variables and continuous variables that represent the transmission power of the devices as well as PLC port coding rate and Access delay, Problem **P1** (5.9) is a complex multi-objective mixed-integer non-convex optimization problem. Additionally, it has nonlinear constraints that are presented in the logarithmic form of the devices' data rate. The fact that the optimization targets are a part of the constraints of each other also results in the creation of a duality, which adds an additional layer of complexity to the issue. As a result, it is impossible to solve **P1** using straight-forward convex optimization techniques.

5.2.4 PLC Access Delay Calculations

Building upon the findings of the preceding chapter, the PLC access delay (T_{PLC}^p) for each port is calculated as:

$$T_{PLC}^p = T_q^p + T_{ts}^p + T_{rsq}^p + T_I^p \quad (5.10)$$

where the incoming queuing delay at the R-CPG till transmission is (T_q^p), the transport delay through the PLC channel is (T_{ts}^p), the re-sequencing delay at the D-CPG re-transmissions is (T_{rsq}^p), and the delay introduced due to IND is (T_I^p) [101, 118]. In this optimization problem the constant processing delay shall be ignored.

5.2.4.1 Incoming CPG Queue Delay

The incoming CPG queue delay is calculated as $T_q^p = \frac{\mathbb{E}[q]}{\rho_p}$; where the average number of HFs on the RU side including both the incoming queuing and waiting buffers is $\mathbb{E}[q] = \bar{q} + t_f \bar{\mathcal{T}}_r \rho_p$. $\bar{\mathcal{T}}_r$ is the mean throughput over the PLC link defined as:

$$\begin{aligned} \mathcal{T}_r &= 1 + \mathcal{U} [\mathbf{I} - \mathcal{S}]^{-1} \mathcal{V} \\ \text{s.t. } \mathcal{U} &= \begin{bmatrix} \mathbf{1} & \mathbf{1} \end{bmatrix} \quad \mathcal{S} = \begin{bmatrix} c_{00}e_0 & c_{10}e_0 \\ c_{01}e_1 & c_{11}e_1 \end{bmatrix} \quad \mathcal{V} = \begin{bmatrix} \pi_{c_0}e_0 \\ \pi_{c_1}e_1 \end{bmatrix} \end{aligned} \quad (5.11)$$

where $\mathbf{I}$ represents the identity matrix. The symbols π_{c_0} and π_{c_1} represent the steady state probability of the channel being in the Good and Bad states, respectively [101, 118]. These probabilities are set as [107]:

$$\pi_{c_0} = \frac{c_{10}}{c_{10} + c_{01}} \quad \text{and} \quad \pi_{c_1} = \frac{c_{01}}{c_{10} + c_{01}} \quad (5.12)$$

The mean incoming CPG queue length defined as $\bar{q}$ is calculated as $\bar{q} = \mathbf{Q}'(1)$, where X' indicated the first derivative of X and $\mathbf{Q}(z)$ represent the PGF of the queue length for all channel and source states. The PGF $\mathbf{Q}(z)$ is calculated as:

$$\begin{aligned} \mathbf{Q}(z) &= \sum_{m=0}^{\infty} [\mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z)]^{m+1} \cdot [\mathbf{I} \otimes [\mathbf{E}(z)^{-1} - \mathbf{I}]] \mathbf{Q}_0 \\ \text{where, } \mathbf{Q}(z) &\triangleq [Q_{0,0}(z), Q_{0,1}(z), \dots, Q_{N_{src},1}(z)]^\dagger, \\ \mathbf{Q}_m(z) &\triangleq [Q_{m,0}(z), Q_{m,1}(z)]^\dagger, \quad m = 0, 1, \dots, N_{src}, \\ \mathbf{E}(z) &\triangleq \text{diag}[\xi_0(z), \xi_1(z)], \quad \xi_y(z) \triangleq v_y + r_y z^{-1}, \\ \mathbf{Q}_0 &\triangleq [q_{0,0}^0, q_{0,1}^0, 0, 0, \dots, 0]^\dagger, \\ \text{diag}[z^x] &\triangleq \text{diag}[1, z, z^2, \dots, z^{N_{src}}], \\ \mathbf{A} &\text{ is the transition probability matrix for HF arrival at RU,} \\ \mathbf{C} &\text{ is the transition probability matrix for PLC channel state, and} \\ \dagger &\text{ is the matrix transpose} \end{aligned} \quad (5.13)$$

The initial values $q_{0,0}^0$ & $q_{0,1}^0$ are found by locating all poles of $\mathbf{Q}(z)$ in a closed unit disc by spectral decomposition of the matrix $\mathbf{X} = \mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z)$ [101, 118]. The poles can be found using the left column and right row eigen vectors of $\mathbf{X}$, respectively; where $\mathbf{G}(z)$ and $\mathbf{G}^{-1}(z)$ are defined as:

$$\begin{aligned} \mathbf{G}(z) &\triangleq [\mathbf{g}_0(z), \mathbf{g}_1(z), \dots, \mathbf{g}_{2N_{src}+1}(z)] \quad \text{so, } \mathbf{G}^{-1}(z) \triangleq [\mathbf{l}_0(z), \mathbf{l}_1(z), \dots, \mathbf{l}_{2N_{src}+1}(z)]^\dagger \\ \therefore \mathbf{X} &= \mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z) = \sum_{x=0}^{2N_{src}+1} \lambda_x(z) \mathbf{g}_x(z) \mathbf{l}_x(z) \end{aligned} \quad (5.14)$$

Substituting back in eq. (5.13) and applying geometric series [101, 118], we get:

$$\mathbf{Q}(z) = \sum_{x=0}^{2N_{src}+1} \frac{\lambda_x(z)}{1 - \lambda_x(z)} \mathbf{g}_x(z) \mathbf{l}_x(z) [\mathbf{I} \otimes \mathbf{E}(z)^{-1} - \mathbf{I}] \mathbf{Q}_0 \quad (5.15)$$

Substituting $\mathbf{Q}_0$ into the previous equation, we obtain:

$$\mathbf{Q}(z) = \sum_{x=0}^{2N_{src}+1} \frac{\lambda_x(z)}{1 - \lambda_x(z)} \sum_{m=0}^{2N_{src}+1} g_{m,x}(z) \cdot \left(l_{x,0}(z) \frac{1 - \xi_0(z)}{\xi_0(z)} q_{0,0}^0 + l_{x,1}(z) \frac{1 - \xi_1(z)}{\xi_1(z)} q_{0,1}^0 \right) \quad (5.16)$$

since $\mathbf{g}_m = [g_{m,0}, g_{m,1} \dots g_{m,2N_{src}+1}]^\dagger$ and $\mathbf{l}_m = [l_{0,m}, l_{1,m} \dots l_{2N_{src}+1,m}]^\dagger$. For each root $|z_x| < 1$ and as $\mathbf{Q}(z)$ is analytic for each root, the following boundary equation is set $\left(l_{x,0}(z_x) \frac{1 - \xi_0(z_x)}{\xi_0(z_x)} q_{0,0}^0 + l_{x,1}(z_x) \frac{1 - \xi_1(z_x)}{\xi_1(z_x)} q_{0,1}^0 \right) = 0$ [101, 118].

Secondly, the definition $\lim_{z \rightarrow 1} \mathbf{Q}(z) = 1$ can be used to solve both equations to obtain the values for $q_{0,0}^0$ and $q_{0,1}^0$ and a complete form for $\mathbf{Q}(z)$ [101, 118].

5.2.4.2 Transport Delay

The transport delay (T_{ts}^p) is a combination of two delays: the transmission delay and the propagation delay. The transmission delay denoted as (T_t^p) is determined by the average throughput ($\overline{\mathcal{T}_r}$) defined in eq. (5.11) and can be mathematically represented as $T_t^p = t_f \cdot \overline{\mathcal{T}_r} - t_f/2$ [101, 118]. Here, t_f represents the total feedback delay involved in transmitting and propagating the HF and the ACK/NACK message.

The propagation delay (T_{pg}^p) is a fixed delay that is determined by the length (L_p) and propagation speed (V_p) of the power cable. Mathematically, the propagation delay may be expressed as $T_{pg}^p = \lceil (L_p/V_p) \rceil$ (Ts), where the propagation speed for power lines is determined to be 173 m/ μ s [110]. Therefore, the mean transport delay is set as $T_{ts}^p = T_t^p + T_{pg}^p$ [101, 118].

5.2.4.3 Re-sequencing Delay

The mean re-sequencing delay is calculated as:

$$\overline{T_{rsq}^p} = \int_0^W \int_0^\infty t \cdot (\mathbb{P}_{rsq}|\omega_r) dt d\omega_r = \int_0^{W-1} \int_0^{L_r} \frac{\mu_r}{t_f} \cdot (\mathbb{P}_{rsq}|\omega_r) d\mu_r d\omega_r \quad (5.17)$$

where $\mathbb{P}_{rsq}|\omega_r$ is the PDF of the re-sequencing delay and is defined as [101, 118]:

$$\mathbb{P}_{rsq}(\mu_r t_f + \omega_r | y) = \begin{cases} \left(\frac{1 - e_y^{\mu_r + 1}}{1 - e_y^{L_r + 1}} \right) \cdot \mathbb{P}_s(S_{\mu_r} | y) \cdot \left[\sum_{l=0}^{\mu_r + 1} \mathbb{P}_s(S_l | y) \right]^{\omega_r - 1} \cdot \left[\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) \right]^{W - 1 - \omega_r} & \omega_r \neq 0 \\ \left(\frac{(1 - e_y) \cdot e_y^{\mu_r}}{1 - e_y^{L_r + 1}} \right) \cdot \left[\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) \right]^{W - 1} & \omega_r = 0 \end{cases} \quad (5.18)$$

where, $\mathbb{P}_s(S_{\mu_r} | y) = \frac{\alpha_p \cdot (1 - e_y) \cdot e_y^{\mu_r}}{1 - e_y + \alpha_p e_y - \alpha_p e_y^{L_r + 1}}$, $\sum_{l=0}^{\mu_r} \mathbb{P}_s(S_l | y) = \frac{1 - e_y + \alpha_p \cdot e_y \cdot (1 - e_y^{\mu_r})}{1 - e_y + \alpha_p \cdot e_y - \alpha_p e_y^{L_r + 1}}$ and,

$$\sum_{l=0}^{\mu_r + 1} \mathbb{P}_s(S_l | y) = \frac{1 - e_y + \alpha_p \cdot e_y \cdot (1 - e_y^{\mu_r + 1})}{1 - e_y + \alpha_p \cdot e_y - \alpha_p e_y^{L_r + 1}}$$

5.2.4.4 IND Introduced Delay

The delay introduced into the system due to the use of IND, the module is modelled as a single state MC model with a probability that an impulse is detected as P_{imp}^p defined in eq.(5.8) above [101, 118]. Whenever an impulse is detected, the delay introduced is approximated up to one Ts plus the transmission of all the HFs being transported during the impulse being flushed through the system. Therefore, the introduced delay due to IND T_I^p can be expressed as [101, 118]:

$$T_I^p = P_{imp}^p \times (w^p + 1) [Ts] \quad (5.19)$$

where w^p is the number of time slots within the selective repeat window of port p . One slot is added to the delay due to the wait time for the impulse width which as indicated above is less than one time slot [101, 118].

5.3 Use of Evolutionary Multi-objective Optimization (EMO)

In multi-objective optimization problems such as the one presented in this research, multiple conflicting objectives are simultaneously sought for optimization namely the number of satis-

fied IoT devices, their energy consumption and the access system delay. Evolutionary Multi Objective Optimization (EMO) is an optimization field that uses evolutionary algorithms to solve multi-objective problems [119]. EMO uses population-based techniques inspired by natural evolution, ensuring population diversity and avoiding converging to a single optimal solution.

EMO is preferred over other convex optimization methods like weighted sum and ϵ -constraint. EMO can handle non-convex and complex objective landscapes effectively. Moreover, unlike the weighted sum, there is no need to choose any weights in advance, which can be challenging in practice [120]. EMO aims to find the entire Pareto front, representing all trade-offs between conflicting objectives, while other methods may focus on a single solution. The Pareto front, also referred to as the Pareto frontier or Pareto curve, is the collection of solutions in a multi-objective optimization problem that are not dominated by any other solution. This implies that there is no other option that can enhance a certain objective function without simultaneously deteriorating at least one other objective. Furthermore, EMO avoids scalarization and directly works with multiple objectives, and it offers flexibility in problem formulation, accommodating various types of objectives and constraints. However, the choice of optimization method depends on the problem's characteristics and complexity [119, 120].

5.3.1 NSGA-II for Multi-Objective Optimization in Communication Systems

NSGA-II is a powerful evolutionary algorithm for widely used to solve multi-objective optimization problems, adept at tackling issues with conflicting objectives [121, 122]. NSGA-II efficiently classifies solutions into Pareto fronts based on their dominance relationships, maintaining diversity and using crowding distance to provide a well-distributed set of solutions. It converges quickly, adapts to complex landscapes, robust for real-world problems and its complexity is at most $O(MN^2)$ (where M is the number of objectives and N is the population size) [122]. The algorithm's easy implementation, scalability, and effectiveness make it a powerful choice for optimizing communication systems, where multiple conflicting ob-

jectives need to be considered simultaneously [122]. NSGA-II offers several advantages that make it particularly well-suited for multi-objective optimization problems in communication systems:

- 1) **Pareto-Optimal Solutions:** NSGA-II excels at identifying a diverse set of Pareto-optimal solutions in a single run. It achieves this by employing non-dominated sorting and crowding distance mechanisms. Non-dominated sorting categorizes solutions based on their dominance relationships, promoting solution diversity on the Pareto front. Crowding distance prioritizes solutions with a larger separation in the objective space, ensuring a well-balanced collection of non-dominated solutions that capture various trade-offs between objectives [122].
- 2) **Convergence and Diversity:** NSGA-II strikes a balance between exploration and exploitation. The algorithm leverages a fast non-dominated sorting approach and an elitist strategy to efficiently converge towards the Pareto front. Additionally, the crowding distance operator fosters diversity by favoring solutions positioned further apart in the objective space [122, 123].
- 3) **Absence of Weighting Factors:** Unlike some other multi-objective optimization techniques, NSGA-II does not require the pre-definition of weighting factors to transform the problem into a single-objective one. This eliminates the need for user-defined parameters, simplifying the implementation and avoiding the burden of assigning relative importance to different objectives [122].
- 4) **Parallelization:** NSGA-II is readily parallelizable, enabling it to leverage modern multi-core Central Processing Units (CPUs) and distributed computing systems. This capability translates to faster convergence and the ability to handle more complex optimization problems with larger solution spaces [123, 124].
- 5) **Global and Local Search Balance:** The integration of non-dominated sorting and crowding distance fosters a favorable balance between global and local search capabilities. NSGA-II effectively explores the entire solution space while maintaining a diverse set of solutions near the Pareto front [122].

- 6) **Constrained Multi-Objective Optimization:** NSGA-II can address constrained multi-objective optimization problems through the incorporation of constraint handling techniques. This feature enhances its adaptability to real-world optimization scenarios involving both objective and constraint functions [122].
- 7) **Ease of Implementation:** Compared to other intricate multi-objective optimization techniques, NSGA-II is relatively straightforward to implement, making it user-friendly. This accessibility benefits researchers and practitioners who may not be experts in the field [121, 122].
- 8) **Well-Established and Robust:** NSGA-II is a well-established and highly regarded method with a strong theoretical foundation. Its effectiveness in handling challenging optimization tasks has been validated through extensive testing on numerous benchmark problems and real-world applications [123, 124].

These attributes make NSGA-II a compelling choice for multi-objective optimization in communication systems, where engineers often face the challenge of optimizing multiple conflicting objectives simultaneously.

Reproduction operators play a pivotal role in guiding the search process of evolutionary algorithms (EAs). Specifically, the selection, crossover and mutation operators are closely linked to the diversity of solutions generated. Thus, the choice of these operators significantly impacts the effectiveness of the resulting solutions and a deeper look into those operators is essential.

The pseudo-code for NSGA-II (Algorithm 5.1) is shown below:

Algorithm 5.1: Pseudo-code for NSGA-II

Input: Population size: N_{pop} , Mutation Probability Pr_{mut} , Stopping criterion G_{max}

1. $P_0 \leftarrow$ Generate initialize population of N_{pop} individuals;
2. $Q_0 \leftarrow \emptyset$;
3. **Primary Loop:** For $g = 0 : G_{max}$
4. $R_g \leftarrow P_g \cup Q_g$; $P_{g+1} \leftarrow \emptyset$;
5. $\mathcal{F} \leftarrow$ **Fast non-dominated sorting (Algorithm 5.2)** (R_g) ;
6. $i = 1$; **while** $|P_{g+1}| + |\mathcal{F}_i| \leq N_{pop}$ **do**
7. $\mathcal{C}_i \leftarrow$ **Compute Crowding Distance (Algorithm 5.3)** of $\mathcal{F}_i$;
8. $P_{g+1} \leftarrow P_{g+1} \cup \mathcal{F}_i$;
9. Increment $i = i + 1$
10. **end while**
11. **if** $|P_{g+1}| < N_{pop}$ **then**,
12. $\mathcal{F}_i \leftarrow$ **Fast non-dominated sorting (Algorithm 5.2)** ($\mathcal{F}_i, \mathcal{C}_i$, “descending”);
13. $P_{g+1} \leftarrow P_{g+1} \cup \mathcal{F}_i [1 : (N_{pop} - |P_{g+1}|)] \rightarrow$ fill P_{g+1} with the $N_{pop} - |P_{g+1}|$ less crowded individuals of $\mathcal{F}_i$;
14. **end if**
15. $Q_{g+1} \leftarrow$ Apply **Selection** then **Crossover** (P_{g+1}, N_{pop});
16. $Q_{g+1} \leftarrow Q_{g+1}$ **Mutation** (Q_{g+1}, Pr_{mut});
17. Increment $g = g + 1$
18. **end for**

Output: $\leftarrow$ **Return** non-dominated solutions ($P_{G_{max}}$)

Algorithm 5.2: Pseudo-code for Fast Non-dominated sorting

Input: Joint population (i.e. main population + Offspring): $R_g = P_g \cup Q_g$

```
1. For each  $p \in P_g$ 
2.   For each  $q \in P_g$ 
3.     if  $(p \prec q)$                                 if  $p$  dominates  $q$  then
4.        $\mathcal{S}_p = \mathcal{S}_p \cup \{q\}$                 include  $q$  in  $\mathcal{S}_p$ 
5.     else if  $(q \succ p)$  then                    if  $p$  is dominated by  $q$  then
6.        $n_p = n_p + 1$                             increment  $n_p$ 
7.     end if
8.     if  $n_p = 0$  then                            if no solution dominates  $p$  then
9.        $\mathcal{F}_1 = \mathcal{F}_1 \cup \{p\}$                  $p$  is a member of the first front
10.    end if
11.  end for
12. end for
13.  $i = 1$ ; while  $\mathcal{F}_i \neq \emptyset$ 
14.   $\mathcal{H} = \emptyset$ 
15.    For each  $p \in \mathcal{F}_i$                         for each member  $p$  in  $\mathcal{F}_i$ 
16.      For each  $q \in \mathcal{S}_p$                     modify each member from the set  $\mathcal{S}_p$ 
17.         $n_q = n_q - 1$                         decrement  $n_q$  by one
18.        if  $n_q = 0$  then  $\mathcal{H} = \mathcal{H} \cup \{q\}$     if  $n_q$  is zero,  $q$  is a member of a list  $\mathcal{H}$ 
19.      end for
20.    end for
21.     $i = i + 1$ 
22. end while
23.  $\mathcal{F}_i = \mathcal{H}$                                 current front is formed with members  $\mathcal{H}$ 
```

Output: $\leftarrow$ **Return** non-dominated sorted Joint population ($\mathcal{F}$)

Algorithm 5.3: Pseudo-code Compute Crowding Distance of $\mathcal{F}$

Input: non-dominated sorted Joint population ($\mathcal{F}$)

```
1.  $N = \mathcal{F}$ 
2. for  $i = 1 \dots N$  do
3.    $\mathcal{C}[i] = 0$ 
4. end for
5. for  $m = 1 \dots M$  do
6.    $\text{SORT}(\mathcal{F}, m)$ 
7.    $\mathcal{C}[1] = \mathcal{C}[N] = \infty$ 
8.   for  $i = 2 \dots N - 1$  do
9.      $\mathcal{C}[i] = \mathcal{C}[i] + \frac{\mathcal{F}[i+1].m - \mathcal{F}[i-1].m}{f_m^{\max} - f_m^{\min}}$ 
10.  end for
11. end for
```

Output: $\leftarrow$ **Return** Crowding distance for all individuals in the population ($\mathcal{C}$)

5.3.2 Binary Tournament Selection

Selection is a method employed to choose the suitable individuals for participating in the process of reproduction. The selection system is crucial as it determines the individuals who have a direct impact on the creation of the following generation [125, 126]. It exerts a substantial impact on directing the search into a promising domain and achieving successful solutions rapidly. Conversely, it is necessary to preserve the variety of the population in order to prevent early convergence and achieve the global optimum.

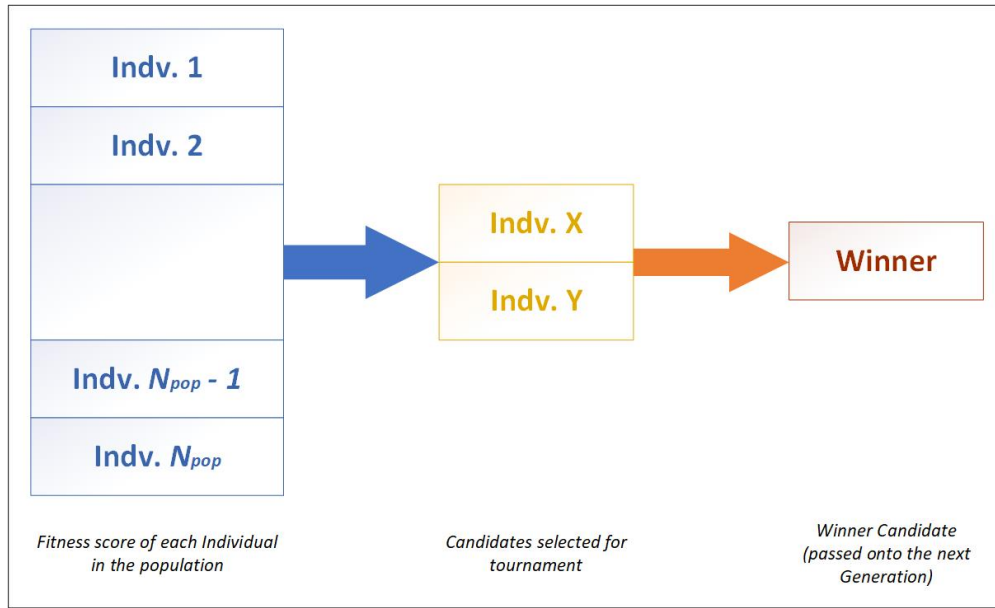


Figure 5.2: Binary Tournament Selection

The two most prominent techniques of selection are tournament selection and roulette wheel selection [125, 126]. Tournament selection is preferred over roulette wheel selection in NSGA-II because it has reduced complexity and more selective pressure, enabling it to better maintain population variety. Due to its independence from scale information, the tournament selection method is highly suitable for ranking-based fitness assignment, such as the non-dominated sorting technique employed in this research [125]. This approach entails organizing a competition among individuals and choosing the victor, with a focus on elitism to ensure that only the most exceptional individuals progress to the next generation [126]. When comparing two solutions with different non-domination ratings, we always choose

the option with the higher rank. Alternatively, if both points are part of the same front, we prioritize the point that is situated in a zone with a greater concentration of points (represented in the crowding distance) as shown in Figure 5.2 [125, 126].

5.3.3 Single-Point Binary and Simulated Binary Crossover

The reproductive phase of NSGA-II often relies on the application of traditional mutation and crossover operators, specifically the binary crossover and polynomial mutation [127, 128].

5.3.3.1 Single-Point Binary Crossover

Binary crossover is a highly effective strategy in the NSGA-II for several reasons. It preserves genetic information by combining parent solutions, resulting in offspring inheriting characteristics from both parents [122, 128]. This maintains a diverse population, preventing premature convergence to sub-optimal solutions. It also efficiently balances exploration and exploitation, a crucial balance in multi-objective optimization problems [122]. The binary crossover mechanism is easy to implement, integrating with other genetic operators like mutation and selection. It is also compatible with binary-coded problems, ensuring offspring solutions remain valid within the binary search space [128]. Furthermore, binary crossover can be adapted to work with different types of problems by adjusting parameters like crossover probability and selection strategies. Overall, binary crossover is a preferred choice for solving complex multi-objective optimization problems [122, 128].

In NSGA-II, for problems with discrete or integer variables, binary crossover with a mask can be employed to generate offspring. This technique involves two parent solutions and a binary mask [122]. The mask, with a length equal to the number of variables, determines which parent contributes a value at each position in the offspring. As illustrated in Figure 5.3, each bit in the mask is randomly generated as 0 or 1. If a bit is 1, the corresponding variable in the offspring comes from the first parent [129]. Conversely, a 0 in the mask indicates the offspring inherits the value from the second parent. This approach allows for the exchange of genetic information between parents while maintaining the discreteness or integer nature of the variables [128, 129].

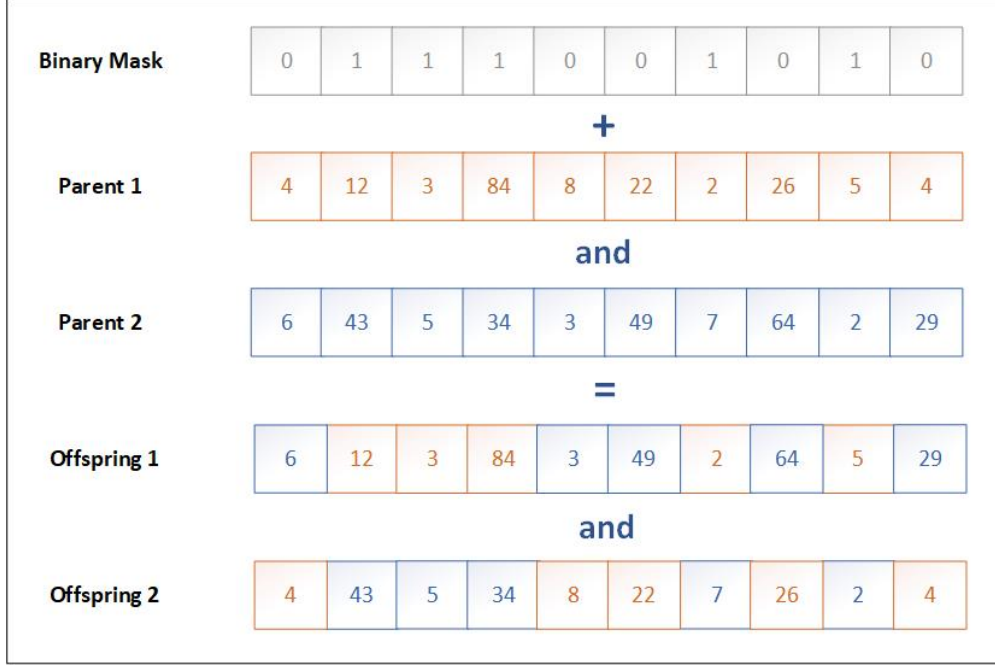


Figure 5.3: Binary Crossover

5.3.3.2 Simulated Binary Crossover

The success of binary genetic algorithms (GAs) in tackling discrete search spaces motivated the development of effective real-coded GAs. However, real-coded GAs faced challenges in overcoming limitations associated with variable coding. To address this, a real-coded crossover operator, Simulated Binary Crossover (SBX) [130, 131]. This operator draws inspiration from the single-point crossover used in binary GAs, enabling it to function independently of the underlying variable coding scheme [122, 130]. The search capability of a crossover operator refers to the probability distribution it employs to generate offspring from parent solutions. Initially, the search efficacy of the single-point crossover was established. Subsequently, SBX was designed to possess a comparable search capability, effectively mimicking the behavior of its binary counterpart [127, 130].

Within the NSGA-II framework, SBX has emerged as the preferred crossover operator for continuous multi-objective optimization problems. This preference stems from several key advantages offered by SBX. SBX mimicking the single-point crossover makes it well-suited for real-coded algorithms like NSGA-II [132]. This mechanism promotes genetic diversity by

ensuring offspring inherit characteristics from both parents, facilitating efficient exploration of the search space [128, 132]. Furthermore, SBX is classified as Parent-Centric, meaning that two children values (C_1 and C_2) are created around the parent values (P_1 and P_2), as shown in Figure 5.4, controlled by β which determines the magnitude of change introduced in the offspring during the crossover process. SBX has the ability to generate offspring within the boundaries of the parent variables, which helps preserve non-dominated solutions and contributes to a diverse Pareto front [128].

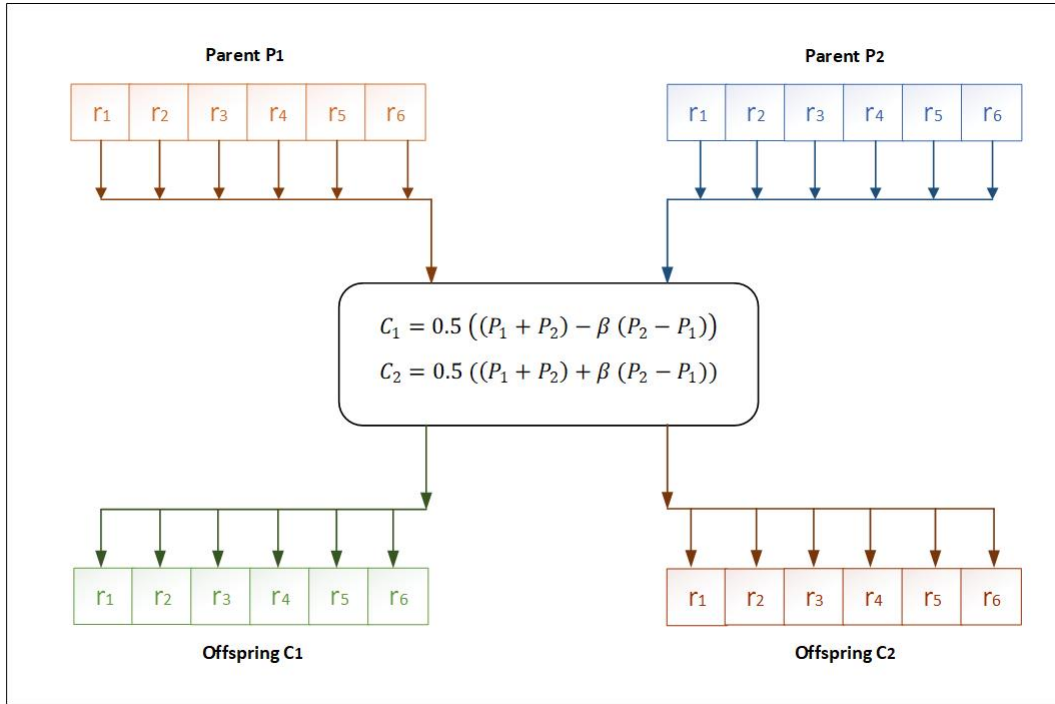


Figure 5.4: Simulated Binary Crossover

5.3.3.3 Polynomial Mutation

NSGA-II, a popular multi-objective evolutionary algorithm, requires a mutation operator that effectively introduces diversity into the population while maintaining the feasibility of solutions (i.e. keeping variables within defined bounds) [122]. Polynomial mutation is a well-suited choice for this purpose due to its ability to control the magnitude of changes introduced to individual solutions [123]. As shown in Algorithm 5.4 below, this mutation operator works by randomly selecting a variable and applying a random value drawn from a

probability distribution that favors small changes [124]. The specific distribution used is often a power law distribution, with the degree of the exponent controlling the likelihood of larger mutations. A higher exponent leads to a higher probability of smaller changes, ensuring a balance between exploration (finding new areas of the search space) and exploitation (refining promising solutions) as illustrated in the pseudo-code below [122].

Compared to other mutation operators like random walk or bit-flip, polynomial mutation offers better control over the magnitude of changes introduced. This is particularly beneficial for problems with real-valued variables and defined boundaries for these variables [122, 123]. Furthermore, the power law distribution used in polynomial mutation naturally favors smaller changes, promoting exploitation and convergence towards promising regions of the search space. However, the possibility of larger mutations allows for continued exploration and avoids premature convergence [123]. Polynomial mutation is a relatively simple concept to implement, requiring minimal computational resources compared to more complex mutation operators [122, 123]. These advantages make polynomial mutation a popular choice for NSGA-II, particularly when dealing with problems involving continuous variables and the need to maintain solution feasibility within defined bounds.

Algorithm 5.4: Pseudo-code for Polynomial Mutation

Input: Vector of individuals to be mutated (*individual*)

1. **for** $i = 1 \dots \text{size}(\text{individual})$ **do**
2. $p = \mathcal{U}[0, 1]$
3. **if** $p < \mathbb{P}_{mut}$ **then**
4. $\delta = p^{(1/(\eta_m+1))}$
5. $r = \mathcal{U}[0, 1]$
6. **if** $r < 0.5$ **then**
7. $\delta = -\delta$
8. **end if**
9. $\text{mutant}(i) = \text{individual}(i) + \text{round}(\delta \times (UB - LB + 1))$
10. $\text{mutant}(i) = \max(LB, \min(UB, \text{mutant}(i)))$
11. **end if**
12. **end for**

Output: $\leftarrow$ **Return** Vector of mutated individuals (*mutant*)

where $\mathbb{P}_{mut}$ is the mutation probability, η_m is the parameter that controls the degree of mutation, and LB and UB are the lower and upper bounds of the individual's values.

5.3.4 Multi-tuple Gene

In the context of our problem, for each network realization, one individual represents one set of parameters for the complete RAN entities, which include IoT device parameters, RU parameters and PLC port parameters. Moreover, it also carries the information interconnection entities such as the Device-RU Received signal strength (RSS) matrix, the Device-RU and the RU-PLC association and allocation matrices. This translates into a complex multi-layered interdependent gene which we shall refer to as "Multi-tuple" gene. A multi-tuple gene in the context of genetic algorithms can be defined as a gene that represents values or attributes of multiple entities simultaneously [133, 134]. Unlike a single-value gene, which holds a single piece of information, a multi-tuple gene comprises of a number of tuples (or vector / matrix) each contains multiple values, allowing it to encapsulate more complex data structures. This representation is particularly useful in optimization problems where each individual solution needs to encode parameters or features of multiple entities simultaneously [133, 134].

As shown in Figure 5.5, One individual gene consist of 3 tuples or sets of parameters, the IoT device, the RU and the PLC port parameters. Looking deeper into each tuple starting with IoT device tuple: The main manageable input matrix is the devices' transmit powers (P_a). SBX crossover and polynomial mutation are used on the devices' transmit powers matrix (P_a) to create new values every generation. All other IoT device parameters are obtained from the device transmit power. To clarify, the RSS matrix for device-RU pairs (RSS) is obtained using the Max-RSS association rule represented in eq. (5.1), then, the matrix of received data rates from devices at RU (R_a) is calculated using as sum of individual device's data rate found in eq. (5.2).

Looking into the RU tuple of the gene, the number of IoT devices with a certain cluster is obtained as a result of the total device's data rate (R_a). In regard to the PLC port tuple, the manageable parameters are the re-transmission limit (L_r), SR window size (W) and the channel coding rate ($\eta_r s$). All three of those parameters are initially obtained randomly within their respective bounds, then obtained through crossover as well as mutation operators. Further, the HF data rate per port is calculated as well as the percentage of

overhead per port (η_p) due to the use of CPG functionalities using eq. (5.4) and eq. (5.6). The RU-PLC port association as well as PSC allocation matrices are found through greedy scheduling based on the HF data rate per RU as well as the available capacity per PLC port. Finally, using the outcomes of the previous chapter, and using all the parameters generated regarding the PLC port, the PLC access delay (T_{PLC}) is calculated and maintained within 5G and eCPRI requirements.

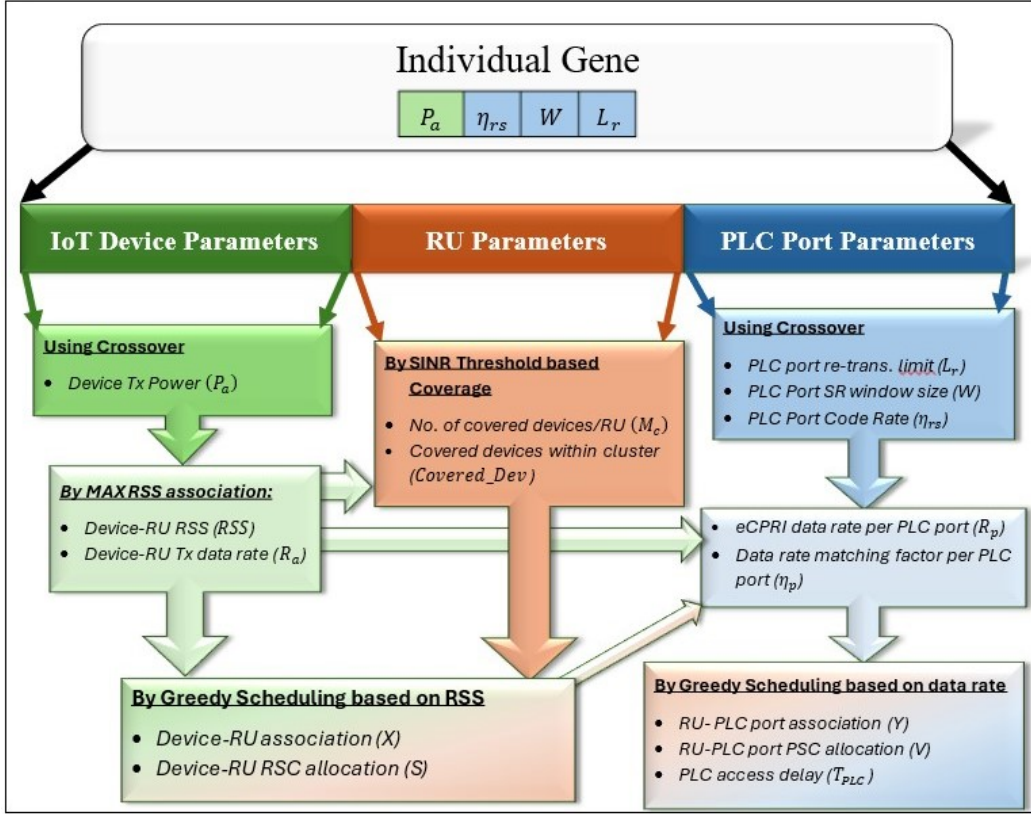


Figure 5.5: Multi-tuple Interdependent Gene

5.4 System Simulations

To address the joint MOO problem **P1** formulated above, a MATLAB-based NSGA-II system simulator has been developed. The simulator's RAN parameters were configured according to the specifications listed in Table 5.1. These parameters were set based on the used parameters as well as the outcomes of the previous chapter. Similarly, the NSGA-II parameters were set based on the values presented in Table 5.2. These parameters were selected because they are

commonly used in the literature; hence, they were chosen somewhat arbitrarily. However, it is important to note that these parameters can be adjusted to further refine and enhance the system's performance. The flexibility in parameter selection allows for tailored optimization that can adapt to varying conditions and requirements, thereby potentially improving the overall efficiency and effectiveness of the NSGA-II algorithm in solving the MOO problem.

Table 5.1: System simulation parameters

Parameter	Value
RUs density / Km^2 (λ_b)	20 / Km^2
IoT devices density / Km^2 (λ_d)	40,000 / Km^2
SINR Threshold / device (τ_a)	-5 dbm
Radio bandwidth (Ω_a)	200 KHz
Radio sub-carrier spacing (ω_a)	15 KHz
Number of RSCs / RU ($\mathcal{N}_a$)	12
Max. IoT device transmit power (P_a^{max})	500 mW
Radio path-loss exponent (α_a)	4
Radio and PLC guard band ratio (g_b)	5%
MIMO PLC Ports (N_p)	6 Tx/RX ports
PLC BW / Port (Ω_p)	1 - 300 MHz
PLC sub-carrier spacing (ω_p)	3.84 MHz
Number of PSCs / PLC port ($\mathcal{N}_{PLC}$)	74
PLC RS block length / port (n_{rs}^p)	127 bits
Mean PLC traffic load (ρ_p)	0.5
RS-FEC code rate (η_{rs}^p)	0.75 - 1
HARQ retrans. Limit (L_r)	2 - 10
SR Window (W)	2-12 HF _s

Table 5.2: NSGA-II simulation parameters

Parameter	Value
Population size/generation	50
Maximum number of generations	150
Probability of crossover	80%
Probability of mutation	40%

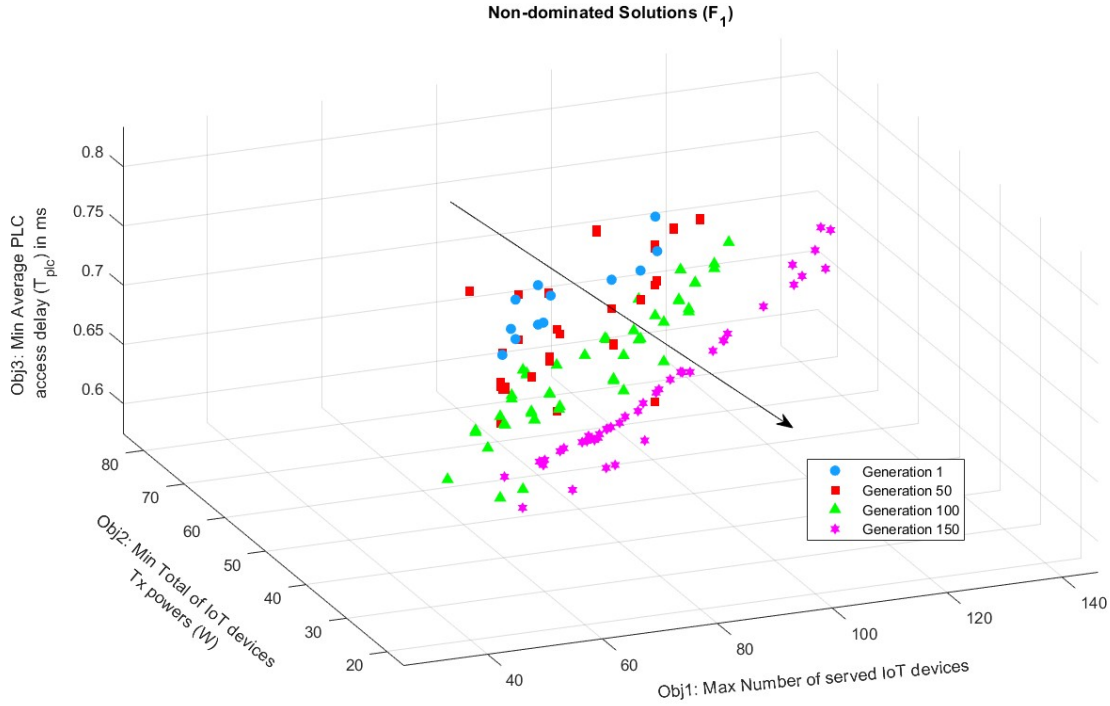


Figure 5.6: 3D Plot of the Pareto-front for Problem **P1** solution for different generations

For a single floor within the building with a $1 \text{ km} \times 1 \text{ km}$ square area, we simulate a network with an average RU density of 20 RUs/km^2 , each equipped with 12 RSCs to serve IoT devices with an average density of $40,000 \text{ devices/km}^2$. Figure 5.6 illustrates the system's progression towards the optimal Pareto-front as more generations are simulated, demonstrating the improvement in the front-frontier/first-frontier. To Further examine Figure 5.6 it is

expanded into multiple 2D-plots. As shown in Figure 5.7a, with each passing generation, the number of satisfied IoT devices increases while maintaining minimal total transmit power from these devices. This ensures extended battery life and reduced energy consumption per device, despite the growing number of satisfied IoT devices. Additionally, Figure 5.7b highlights a decrease in average delay over the PLC link while ensuring that data from all IoT devices is successfully delivered to the DU via the RUs and the PLC based front-haul. This comprehensive simulation underscores the system's capability to balance multiple objectives, enhancing both efficiency and performance in IoT device management.

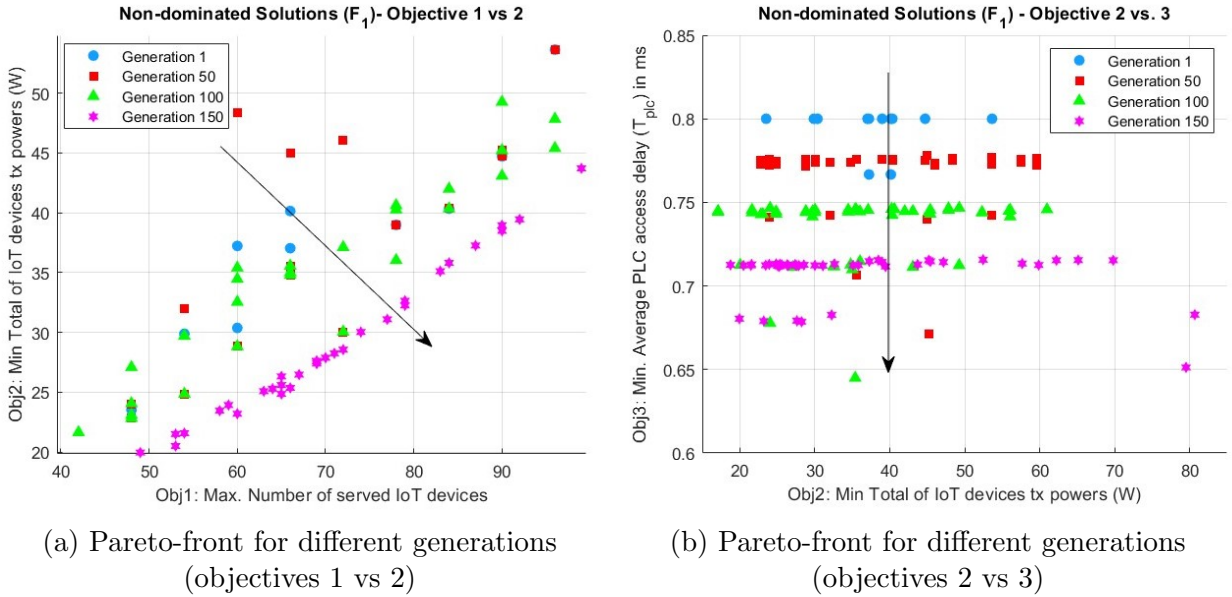


Figure 5.7: 2D Plots of Pareto- front for Problem **P1** solution

To further evaluate our system's performance, a comprehensive comparison is conducted regarding the total number of IoT devices served and the average transmit power per device. We shall compare our system with the main baseline scheme, the Max-RSS device-RU association scheme, where each device is associated with the RU to which it has the best channel conditions; however, resource assignment and power allocation are performed randomly [135]. Our evaluation involves comparing the outcomes of our system across multiple generations verses multiple single runs using the Max-RSS scheme. This comparison will highlight the advantages of our optimization approach in terms of enhancing device-RU associations, optimizing resource allocation, and minimizing power consumption, thereby

demonstrating the efficacy of our system in achieving superior performance metrics in IoT network management.

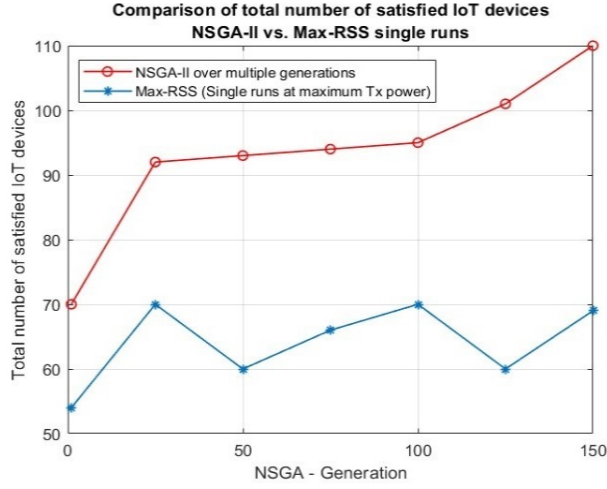
Figure 5.8a depicts the total number of satisfied IoT devices over 150 generations for NSGA-II versus multiple single runs for Max-RSS. The results show a consistent increase in the number of satisfied IoT devices when using NSGA-II based MOO, culminating at around 110 devices by the 150th generation, whereas Max-RSS fluctuates between a total of 60 and 80 IoT devices yielding an average increase in the number of satisfied IoT devices by $\approx 50\%$. This indicates the superior ability of NSGA-II to optimize and increase the number of satisfied devices over successive generations.

Analysing the system outcomes regarding the average PLC access delay (T_{PLC}), Figure 5.8b shows that the Max-RSS scheme maintains a relatively stable delay around $0.77ms$, whereas the NSGA-II scheme exhibits a notable decrease from $0.78ms$ to $0.71ms$ over 150 generations, suggesting decrease in the average PLC delay of about 7% when using NSGA-II while ensuring all data from all IoT devices are delivered to the DU.

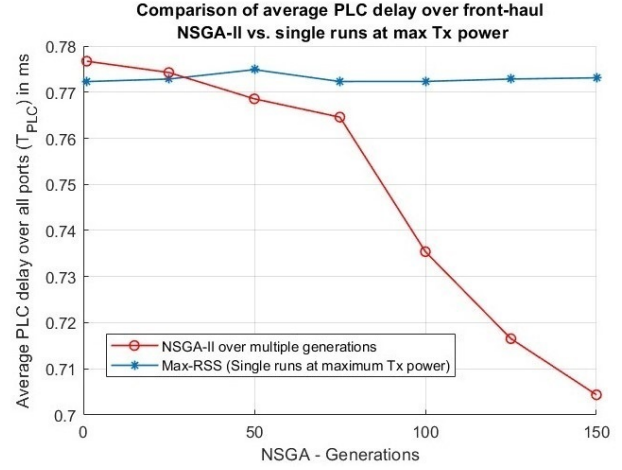
Figure 5.8c compares the total transmit power per device when using NSGA-II scheme across the 150 generations vs. multiple single runs using of Max-RSS scheme. Here, NSGA-II algorithm shows a gradual reduction in total IoT devices transmit power, starting from $48W$ and stabilizing around $35W$ total power, whereas the Max-RSS scheme varies significantly but remains lower than NSGA-II, with peaks around $40W$ and drops to approximately $25W$. Looking at this comparison, at first sight it seems as if the NSGA-II based scheme is performing worse in comparison to Max-RSS scheme. However, we shall not forget that the number of satisfied IoT device is higher when using NSGA-II scheme.

Therefore, it is essential to examine Figure 5.8d, which demonstrates the average transmit power per IoT device. The NSGA-II algorithm shows a clear reduction in transmit power per device, starting from approximately $0.5W$ and decreasing steadily to around $0.35W$ which indicates an average decrease of 30% in device energy consumption. On the other hand, the Max-RSS method keeps a relatively constant power level, highlighting the advantage of NSGA-II in optimizing individual device power consumption. The above system simulation outcomes highlights the efficiency of NSGA-II in minimizing power consumption while in-

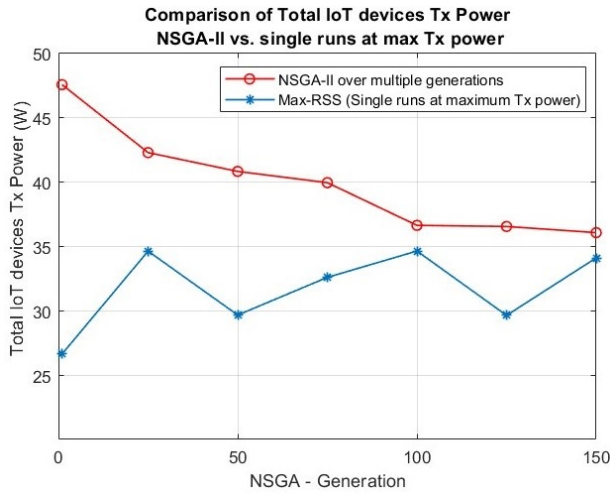
creasing the number of satisfied devices, thereby enhancing overall system performance and energy efficiency.



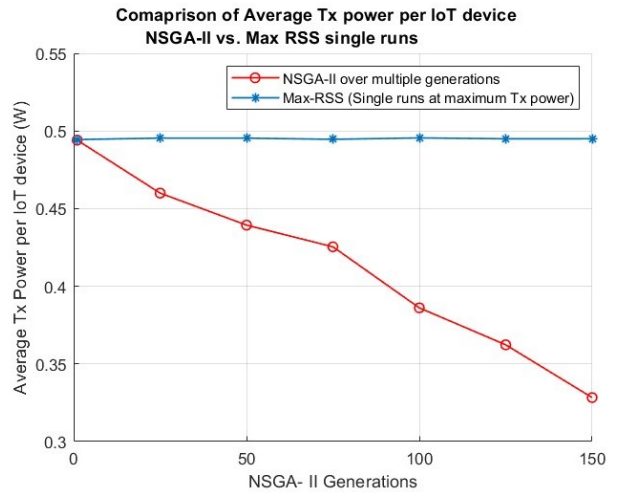
(a) Total number of satisfied IoT devices



(b) Average PLC delay (T_{PLC}) over all ports



(c) Total Transmit Power from all IoT devices



(d) Average transmit Power per IoT devices

Figure 5.8: Comparison of system outcomes between NSGA-II across 150 generations vs. multiple single runs of Max-RSS

5.5 Chapter Summary

This chapter explores the optimization problem for maximizing throughput in a split-RAN system with a PLC front-haul link in a multi-story building. The system consists of RUs, DU, and a PLC front-haul link connecting them. The objective is to maximize the number of satisfied IoT devices while meeting their QoS requirements. The optimization problem is formulated as a mixed-integer non-convex problem, involving multiple objectives: maximizing satisfied devices, minimizing device transmit power, and minimizing PLC access delay. The chapter discusses using an EMO algorithm, specifically NSGA-II, to solve the problem with conflicting objectives in communication systems. The algorithm works by iteratively creating new generations of solutions using tournament selection, single-point binary and simulated binary crossovers, and polynomial mutation operators. The system outcomes provide a Pareto front of non-dominated solutions for the MOO formulated problem. The system simulations, demonstrate the efficacy of the NSGA-II algorithm in improving the performance of IoT networks. Specifically, it enhances device satisfaction, decreases overall and individual device transmit power, and ensures consistent decrease in PLC delays, which outperforms the outcomes when using Max-RSS scheme.

Chapter 6

Conclusion and Future Work

6.1 Conclusion

This thesis investigated the use of indoor low-voltage power lines as the last hop of the C-RAN front-haul that transports TDM based CPRI hyper-frames as well as eCPRI frames. To meet the CPRI and eCPRI requirements on data rate, BER and FER defined under 5G NR, we introduced CPG and eCPG, which serve as agent for creating a virtual CPRI/eCPRI link between the DU and RUs. Simulations have proven that the CPG/eCPG operating the proposed IND technique with spatial multiplexing and HARQ can effectively create a CPRI/eCPRI link over the noisy PLC channel without modification of any existing protocol operation and hardware arrangement (DU & RU).

This research addressed the challenges and potential solutions related to using PLC networks modified with CPG units to support eCPRI-based front-haul in 5G IoT indoor mobile coverage. It emphasized the advantages of PLC technology, compared to optic fibres and coaxial cables, by pointing out the possibility of leveraging existing indoor power lines without requiring alterations to the infrastructure, and of operating without restrictions or licensing requirements. The utilization of low-voltage PLC links for eCPRI-based front-hauling in 5G IoT indoor mobile coverage environments has been investigated, focusing on split C-RAN architecture. The research introduced a novel PLC-based indoor front-haul solution, presented an analytical framework for the RAN e2e delay performance, and derived closed-form expressions for key system parameters.

The meticulous analysis and system-level simulations validate the mathematical models

and explore the intricate dynamics of RU and IoT device densities on e2e delay within the RAN system. Simulations reveal the nuanced relationship between these variables, where increasing IoT device density leads to a rise in concurrent data transmission, reducing air delay until radio resource limitations intervene. However, as the number of RUs increases to serve the same number of devices, interference escalates, diminishing coverage probability and escalating e2e average delay in the system. Despite challenges like interference and resource saturation, the system consistently meets 5G, CPRI and eCPRI benchmarks, showcasing reliability in real-world scenarios. Considerations regarding re-sequencing delay and re-transmission rates underscore further complexities impacting access latency. Overall, this thorough analysis not only affirms model validity but also offers crucial insights for future system design and optimization, ensuring continual enhancement of wireless communication technologies.

This research assessed radio and access network delays in a 5G large-scale indoor IoT network. The scenario focuses on a unique front-haul system that utilizes CPRI/ eCPRI-frames for communication between DU and RU over indoor low-voltage power lines. Employing stochastic geometry, Markov modeling, and queuing theory, a comprehensive mathematical model was developed to analyze various network performance metrics, including scheduling probability, coverage probability, and air interface latency. The interference faced by these devices was precisely specified in the simulation. A mathematical model was formulated to determine the maximum number of IoT devices supported in the PLC front-haul. The study successfully identified and assessed four key factors contributing to access delay: incoming queue delay, transport delay, the impact of impulsive noise events, and re-sequencing delay. The validity of the proposed system and the associated end-to-end latency were evaluated through extensive simulations, demonstrating compliance with 5G, CPRI and eCPRI delay requirements.

This thesis has comprehensively analyzed the system model aimed at maximizing the deployment efficiency of low-mobility, battery-powered IoT devices in a multi-story service facility under a C-RAN architecture. The analysis delved into the complexities of PLC access delay, including queuing delays, transport and propagation delays, re-sequencing challenges,

and the impact of IND. This thorough exploration highlights the intricate nature of C-RAN systems with PLC front-haul, underscoring opportunities for enhancing efficiency and performance through advanced optimization methodologies. Additionally, the chapter explores the optimization problem of maximizing throughput in a split-RAN system with a PLC front-haul link in a multi-story building. It details the formulation of a mixed-integer non-convex problem aimed at simultaneously maximizing the number of satisfied IoT devices, minimizing operational costs, device transmit power, and PLC access delay while meeting QoS requirements. The use of the EMO algorithm, NSGA-II, is discussed for solving this multi-objective optimization problem, employing strategies such as tournament selection, binary and simulated binary crossovers, and polynomial mutation to generate a Pareto front of non-dominated solutions. The results of the system simulation demonstrated the effectiveness of NSGA-II in reducing power consumption while increasing the number of satisfied devices, thus improving overall system performance and energy efficiency.

6.2 Future Work and Research Directions

Below is a list of potential areas for future work based the work in this dissertation:

1. **Changeable IoT Device Density:** The current performance analysis considers a fixed density of IoT devices. Future work can extend this by incorporating dynamic device density into the performance analysis as well as. This would provide valuable insights into how the system behaves under varying load conditions, which is more representative of real-world scenarios with fluctuating numbers of connected devices.
2. **Mid-voltage Power lines for DU-CU Connection:** Investigate the feasibility of using mid-voltage power lines as a mid-haul channel between DU and CU. Full system design including a full PLC based front-haul link. Analyze the impact of this approach on the system's performance, including factors like the number of DUs that can be supported and the overall communication efficiency.
3. **RU Assignment Constraint:** In formulating the goodput MOO problem, in future work, additional constraints shall be considered where RU can only be assigned to

a single phase. This would require modifying the optimization algorithm to account for this limitation. This would provide more practical solutions that consider real-world deployment scenarios where RUs might be restricted to specific phases due to infrastructure or regulatory reasons.

4. **Deep Reinforcement Learning for Optimization:** Explore the application of Deep Reinforcement Learning to solve the optimization problem. Reinforcement learning can be particularly well-suited for problems with complex dynamics and uncertain environments, potentially leading to superior performance compared to the current EMO approach.
5. **Exploration of Non-Orthogonal Multiple Access (NOMA) over the radio and/or access links:** We shall also investigate the potential of NOMA as a method to enhance radio resource efficiency and increase the number of supported devices in our network. NOMA allows multiple users to share the same frequency spectrum, thereby improving spectral efficiency and accommodating more devices. This technique not only boosts the overall capacity but also ensures that resources are utilized more efficiently, addressing the growing demand for connectivity in dense IoT environments. Additionally, ensure the PLC front-haul could support the increase capacity. We aim to achieve a balanced distribution of radio and access resources that meets the specific needs of each device, ensuring robust and reliable network performance. This comprehensive approach aims to maximize network efficiency while adhering to the stringent QoS standards essential for various IoT applications.

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Appendix A

CPG Virtual link flow chart in UL

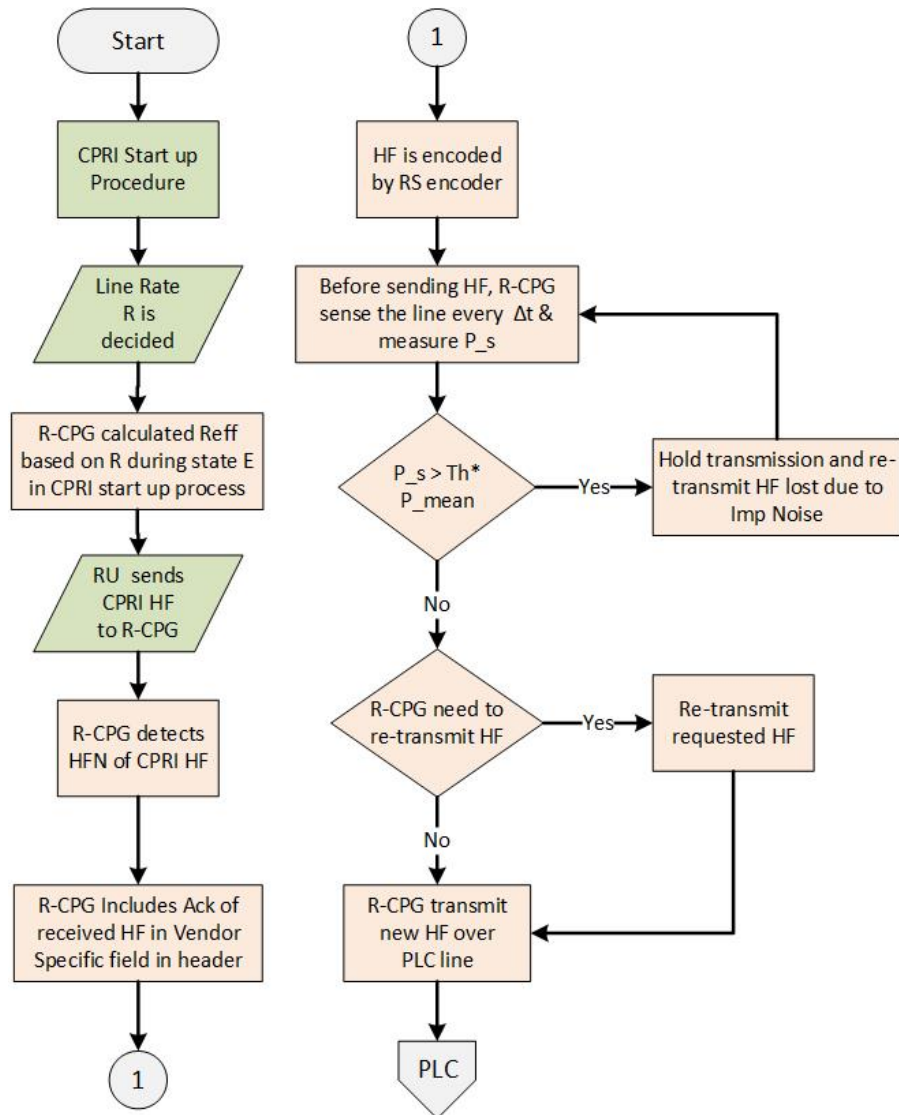


Figure A.1: Flow chart for the PLC based communication system - RU act as transmitter

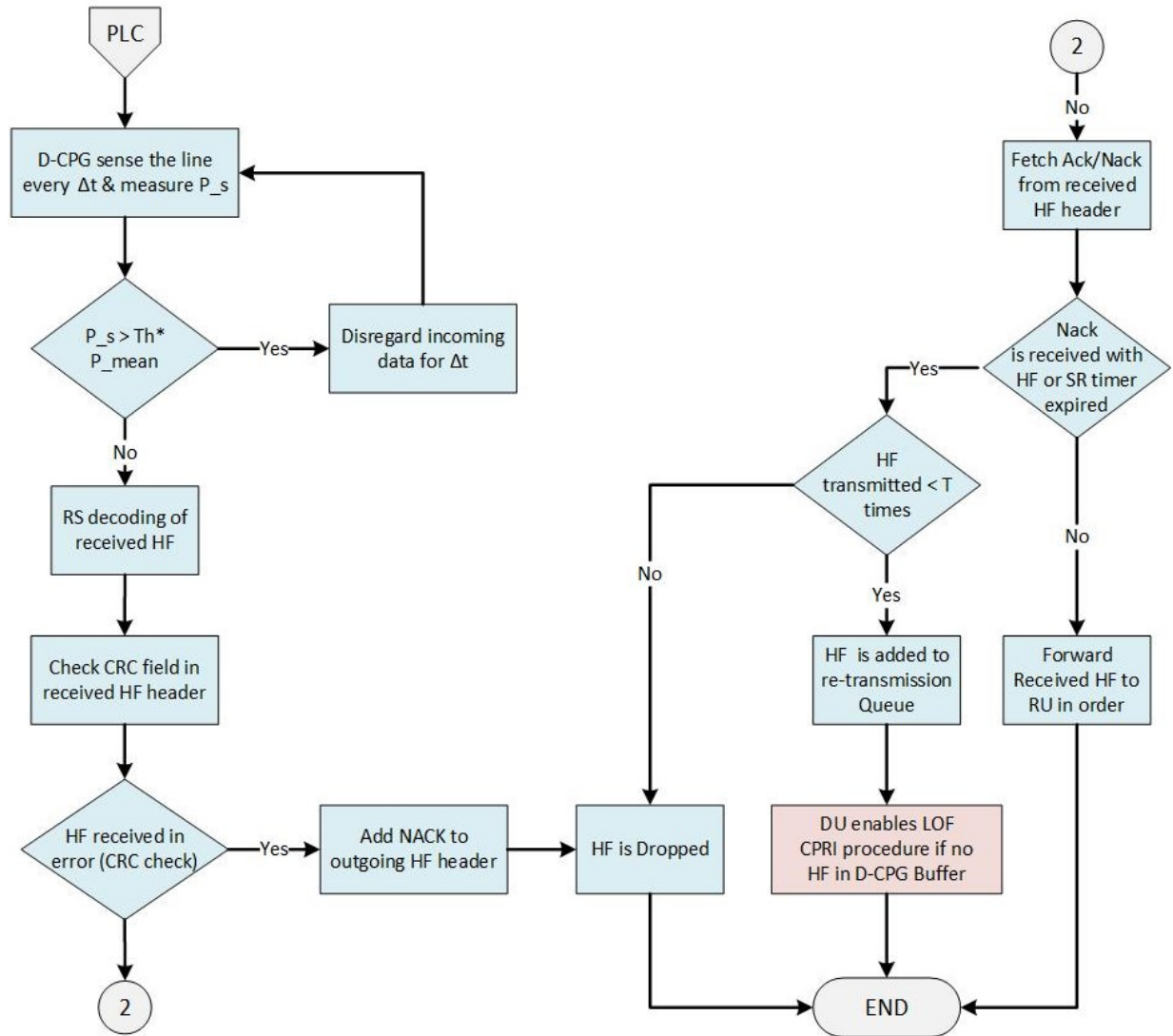


Figure A.2: Flow chart for the PLC based communication system - DU act as receiver

Appendix B

IoT cellular indoor network

B.1 Laplace transform of Inter-cell interference

Devices experience inter-cell interference from other devices served by other RUs within the network and are scheduled on the same sub-channel. Accordingly, an independent thinned PPP ($\Phi_a^{n,t} = \{d_1^{n,t}, d_2^{n,t}, \dots\}$) with density $\lambda_a^{n,t} = \lambda_b$ represents the locations of the inter-cell active device (i.e. in other clusters) interfering on the n^{th} sub-channel during the t^{th} TF. The interference I_a is given by

$$I_a = \sum_{d_i^{n,t} \in \Phi_a^{n,t}} P_a \|d_i^{n,t} - b_i\|^{\epsilon_a \alpha_a} \|d_i^{n,t}\|^{-\alpha_a} g_i \quad (\text{B.1})$$

where g_i is the air interface channel gain experienced by the i^{th} device. $\|d_i^{n,t}\|$ denotes the Euclidean distance between the i^{th} inter-cell interfering device and the origin, and $\|d_i^{n,t} - b_i\|$ is the Euclidean distance between the i^{th} inter-cell interfering device and its serving RU (the i^{th} RU). Note that $\|d_0^{n,t}\|$, $\|d_i^{n,t}\|$ and $\|d_i^{n,t} - b_i\|$ are all random distances as they depend on the locations of the device and the RU.

Let $s = \frac{\tau_a}{P_a} r_t^{\alpha_a(1-\epsilon_a)}$ and $\mathcal{L}_{I_a}\{x\}$ is the Laplace transform of x with respect to inter-cell interference I_a . $\mathcal{L}_{I_a}\{s\}$ is given by [94, 102]

$$\begin{aligned}
\mathcal{L}_{I_a}\{s\} &= \mathbb{E}_{I_a}[\exp\{\{-sI_a\}\}] \\
&= \mathbb{E}_{g_i, s, \|d_i^{n,t} - b_i\|, \|d_i^{n,t}\|} \left[\exp\{-sP_a \cdot \sum_{\|d_i^{n,t}\| \in \Phi_a^{n,t}} \|d_i^{n,t}\|^{-\alpha_a} \|d_i^{n,t} - b_i\|^{\epsilon_a \alpha_a} g_i\} \right] \quad (\text{B.2})
\end{aligned}$$

Let $y_t = \|d_i^{n,t} - b_i\|$ denote the distance between an arbitrary device located at $d_i^{n,t}$ and its serving RU located at b_i , and $x_t = \|d_i^{n,t}\|$ denotes the distance between the device located at $d_i^{n,t}$ and using the n^{th} sub-channel to the origin and during the t^{th} TF. All the euclidean distances r_t , y_t , and x_t are random distances because they depend on where the device and RU are placed. Using Campbell's theorem on (B.2) and given that the distances r_t , y_t and x_t all follow Rayleigh distributions with parameters λ_b , λ_a and $\lambda_d^{n,t}$ respectively, the Laplace transform of the interference can be expressed as [94, 102]

$$\begin{aligned}
\mathcal{L}_{I_a}\{s\} &= \mathbb{E}_{I_a}(\exp\{-sI_a\}) \\
&= \mathbb{E}_{g_i, y_t, x_t} \left[\exp\{-sP_a \cdot \sum_{x_t \in \Phi_a^{n,t}} y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a} g_i\} \right] \\
&= \mathbb{E}_{g_i, y_t, x_t} \left[\prod_{x_t \in \Phi_a^{n,t}} \exp\{-sP_a y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a} g_i\} \right] \\
&= \mathbb{E}_{x_t} \left[\prod_{x_t \in \Phi_a^{n,t}} \mathbb{E}_{y_t, g_i} \exp\{-sP_a y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a} g_i\} \right] \quad (\text{B.3}) \\
&= \mathbb{E}_{x_t} \left[\prod_{x_t \in \Phi_a^{n,t}} \mathbb{E}_{y_t} \left[\frac{1}{1 + sP_a y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} \right] \right] \\
&= \mathbb{E}_{x_t} \left[\prod_{x_t \in \Phi_a^{n,t}} \left[1 - \int_0^\infty \frac{1}{1 + sP_a y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} dy_t \right] \right] \\
&\approx \exp \left(-2\pi\lambda_d^{n,t} \int_{x_t > 0} \left[1 - \int_0^\infty \frac{2\pi\lambda_b y e^{(-\pi\lambda_b y_t^2)}}{1 + \tau_a r_t^{(1-\epsilon_a)\alpha_a} y_t^{\epsilon_a \alpha_a} x_t^{-\alpha_a}} dy_t \right] x_t dx_t \right)
\end{aligned}$$

B.2 Average active IoT devices in one RU Voronoi

In a cellular network with RUs located as a PPP in 2-D, the RU cell's Voronoi area ($\mathcal{V}_a$) is a random variable that can be approximated using a gamma distribution [98]

$$f_{\mathcal{V}_a}(v_a) = \frac{(c_v \cdot \lambda_b)^{c_v}}{\Gamma(c_v)} v_a^{c_v-1} e^{-v_a \cdot c_v \cdot \lambda_b} \quad (\text{B.4})$$

Let the number of active devices associated with RU_0 during the t^{th} TR is denoted by $M_{a_0}^t$. It is a random variable (RV) with probability mass function (PMF) calculated by approximating the probability density function (PDF), over a Voronoi cell of area $\mathcal{V}_a$. As the IoT devices are located as PPP, the number of active devices in an area v_a is Poisson distributed i.e. [97, 98]

$$\mathbb{P}[M_{a_0}^t = m \mid v_a] = \frac{(\lambda_a \cdot v_a)^m \cdot e^{-\lambda_a \cdot v_a}}{m!}, \quad m = 0, 1, 2, \dots \quad (\text{B.5})$$

Thus $\mathbb{P}[M_{a_0}^t = m]$ is calculated as

$$\begin{aligned} \mathbb{P}[M_{a_0}^t = m] &= \int_0^\infty \mathbb{P}[M_{a_0}^t = m \mid v_a] \cdot f_{\mathcal{V}_a}(v_a) \, dv_a \\ &= \int_0^\infty \frac{(\lambda_a \cdot v_a)^m \cdot e^{-\lambda_a \cdot v_a}}{m!} \cdot \frac{(c_v \cdot \lambda_b)^{c_v}}{\Gamma(c_v)} v_a^{c_v-1} e^{-v_a \cdot c_v \cdot \lambda_b} \\ &= \frac{(c_v \cdot \lambda_b)^{c_v} \lambda_a^m}{m! \Gamma(c_v)} \int_0^\infty v_a^{m+c_v-1} \cdot e^{-(\lambda_a + c_v \cdot \lambda_b) v_a} \\ &= \frac{(c_v \cdot \lambda_b)^{c_v} \lambda_a^m \Gamma(m+c_v)}{\Gamma(c_v) (\lambda_a + c_v \cdot \lambda_b)^{m+c_v} m!}, \quad 0 \leq m < \infty \end{aligned} \quad (\text{B.6})$$

where $c_v = (3d+1)/2 = 7/2$ is a constant defined for the Voronoi tessellation in $\mathbb{R}^2$ with dimensionality $d = 2$ and $\Gamma(c_v) = \int_0^\infty t^{c_v-1} e^{-t} dt$. Thus, the average number of active devices would be calculated as [97, 98]. Thus, the average number of active devices would be calculated as

$$\begin{aligned}
\overline{M}_a^t &= \mathbb{E}_m[M_a^t] \\
&\approx \sum_{m=0}^{\infty} \frac{m(c_v \cdot \lambda_b)^{c_v} \lambda_a^m \Gamma(m + c_v)}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{m+c_v} m!} \\
&= \frac{(c_v \cdot \lambda_b)^{c_v}}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{c_v}} \sum_{m=0}^{\infty} \frac{m \Gamma(m + c_v) \zeta^m}{m!}, \quad \text{where, } \zeta = \frac{\lambda_a}{\lambda_a + c_v \cdot \lambda_b} \\
&= \frac{(c_v \cdot \lambda_b)^{c_v}}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{c_v}} \int_{t=0}^{\infty} \sum_{m=0}^{\infty} \frac{m t^m t^{(c_v-1)} \zeta^m e^{-t}}{m!} dt, \quad \text{as, } \Gamma(a + c) = \int_0^{\infty} t^{a+c-1} e^{-t} dt \\
&= \kappa \int_{t=0}^{\infty} \sum_{m=0}^{\infty} \frac{\zeta (\zeta t)^{m-1}}{(m-1)!} t^{c_v} e^{-t} dt, \quad \text{where, } \kappa = \frac{(c_v \cdot \lambda_b)^{c_v}}{\Gamma(c_v)(\lambda_a + c_v \cdot \lambda_b)^{c_v}} \\
&= \kappa \int_{t=0}^{\infty} \zeta e^{\zeta t} t^{c_v} e^{-t} dt, \quad \text{as, } \sum_{n=0}^{\infty} x^n / n! = e^x \\
&= \kappa \zeta \int_{u=0}^{\infty} \frac{u^{c_v} e^{-u}}{(1 - \zeta)^{c_v+1}} du, \quad \text{where, } u = (1 - \zeta)t \\
&= \frac{\kappa \zeta \Gamma(c_v + 1)}{(1 - \zeta)^{c_v+1}} = \left[\frac{\lambda_a \Gamma(c_v + 1)}{\lambda_b c_v \Gamma(c_v)} \right]
\end{aligned}
\tag{B.7}$$

B.3 Average achievable bit rate by a device in coverage

A device can only transmit to its serving RU if it is within the coverage area. Consequently, the effective bit rate is dependent on coverage. Using Shannon's channel capacity formula, it is possible to calculate the average achievable bit rate of a device when it is in coverage.

$$\begin{aligned}
\mathcal{F}_a^{n,t} &= \mathbb{E} [\omega_a \log_2(1 + \Theta_0^{n,t})] \\
&= \int_{\zeta>0} \mathbb{P}(\omega_a \log_2(1 + \Theta_0^{n,t}) > \zeta) d\zeta \\
&= \int_{\zeta>0} \mathbb{P}(\Theta_0^{n,t} > 2^{\frac{\zeta}{\omega_a}} - 1) d\zeta \\
&= \int_{\zeta>0} \mathbb{P}(\Theta_0^{n,t} > \beta) d\zeta, \quad \beta = 2^{\frac{\zeta}{\omega_a}} - 1 \\
&= \int_{\zeta>0} \mathbb{P}\left(\frac{P_a \|d_0^{n,t}\|^{(\epsilon_a-1)\alpha_a} g_{00}}{I_a + \sigma_a^2} > \beta\right) d\zeta \\
&= \int_{\zeta>0} \mathbb{P}\left(g_{00} > \frac{\beta}{P_a} (I_a + \sigma_a^2) r_t^{(1-\epsilon_a)\alpha_a}\right) d\zeta \\
&= \int_{\zeta>0} \mathcal{P}_s \cdot \mathbb{E}_{R_t}[C_0^{n,t}(\sigma_a, \tau_a, \lambda_b, \alpha_a, \epsilon_a | R_t)] d\zeta \\
&= \int_{\zeta>0} \mathcal{P}_s \cdot \int_{r_t>0} 2\pi \lambda_b r_t e^{-\pi \lambda_b r_t^2} / C_0^{n,t}(\sigma_a, \tau_a, \lambda_b, \alpha_a, \epsilon_a, r_t) dr_t d\zeta \\
&\approx \int_{\zeta>0} 2\pi \lambda_b \cdot \mathcal{P}_s \cdot \int_{r_t>0} r_t \cdot \exp\left\{-\pi \lambda_b r_t^2\right\} \cdot \exp\left\{-\frac{\beta}{P_a} r_t^{\alpha_a(1-\epsilon_a)} \sigma_a^2\right\} \cdot \mathcal{L}_{I_a}\left\{\frac{\beta}{P_a} r_t^{\alpha_a(1-\epsilon_a)}\right\} d\zeta
\end{aligned} \tag{B.8}$$

Appendix C

PGF of CPG incoming queue length

Let the steady-state of the incoming queue length at the RU-eCPG be denoted $q_{x,y}^h$ when the PLC channel state is (y) and the source state is (x) , then it can be presented as

$$\begin{aligned}
 q_{x,y}^h = & \sum_{u=0}^{\min(N_{src}, h+1)} \left(c_{y,y} \cdot a_{u,x} \cdot r_y \cdot q_{u,y}[h-u+1] + c_{\bar{y},y} \cdot a_{u,x} \cdot r_{\bar{y}} \cdot q_{u,\bar{y}}[h-u+1] \right) \\
 & + \sum_{u=0}^{\min(N_{src}, h)} \left(c_{y,y} \cdot a_{u,x} \cdot v_y \cdot q_{u,y}[h-u] + c_{\bar{y},y} \cdot a_{u,x} \cdot v_{\bar{y}} \cdot q_{u,\bar{y}}[h-u] \right) \\
 & + \sum_{u=0}^{\min(N_{src}, 1)} \left(c_{y,y} \cdot a_{u,x} \cdot r_y \cdot q_{u,y}[1-u] + c_{\bar{y},y} \cdot a_{u,x} \cdot r_{\bar{y}} \cdot q_{u,\bar{y}}[1-u] \right) \\
 & + c_{y,y} \cdot a_{0,x} \cdot q_{0,y}^0 + c_{\bar{y},y} \cdot a_{0,x} \cdot q_{0,\bar{y}}^0
 \end{aligned} \tag{C.1}$$

Based on eq. (C.1), the PGF $Q_{x,y}(z)$ of the CPG incoming queue length $q_{x,y}^h$ when the PLC channel is in state y and the source is in state x is expressed as:

$$\begin{aligned}
Q_{x,y}(z) = & \sum_{h=1}^{\infty} \sum_{m=0}^{\min(N_{src}, h+1)} \left(c_{y,y} a_{m,x} r_y q_{m,y}[h-m+1] + c_{\bar{y},y} a_{m,x} r_{\bar{y}} q_{m,\bar{y}}[h-m+1] \right) . z^h \\
& + \sum_{h=1}^{\infty} \sum_{m=0}^{\min(N_{src}, h)} \left(c_{y,y} a_{m,x} v_y q_{m,y}[h-m] + c_{\bar{y},y} a_{m,x} v_{\bar{y}} q_{m,\bar{y}}[h-m] \right) . z^h \\
& + \sum_{m=0}^1 \left(c_{y,y} a_{m,x} r_y q_{m,y}[1-m] + c_{\bar{y},y} a_{m,x} r_{\bar{y}} q_{m,\bar{y}}[1-m] \right) \\
& + c_{y,y} a_{0,x} q_{0,y}^0 + c_{\bar{y},y} a_{0,x} q_{0,\bar{y}}^0
\end{aligned} \tag{C.2}$$

Using the relations:

$$\begin{aligned}
\sum_{h=1}^{\infty} \sum_{m=0}^{\min(N_{src}, h+1)} a_{m,x} q_{m,y}[h-m+1] . z^h = & \sum_{m=2}^{N_{src}} \sum_{h=m-1}^{\infty} a_{m,x} q_{m,y}[h-m+1] . z^h \\
& + \sum_{m=0}^1 \sum_{h=1}^{\infty} a_{m,x} q_{m,y}[h-m+1] . z^h
\end{aligned} \tag{C.3}$$

and

$$\sum_{h=1}^{\infty} \sum_{m=0}^{\min(N_{src}, h)} a_{m,x} q_{m,y}[h-m] . z^h = \sum_{m=0}^{N_{src}} \sum_{h=m}^{\infty} a_{m,x} q_{m,y}[h-m] . z^h - a_{0,x} q_{0,y}^0 \tag{C.4}$$

Equation (C.2) can be simplified as:

$$\begin{aligned}
Q_{x,y}(z) = & \sum_{m=2}^{N_{src}} \sum_{h=m-1}^{\infty} c_{y,y} a_{m,x} r_y q_{m,y} [h-m+1] \cdot z^h + \sum_{m=0}^1 \sum_{h=0}^{\infty} c_{y,y} a_{m,x} r_y q_{m,y} [h-m+1] \cdot z^h \\
& + \sum_{m=2}^{N_{src}} \sum_{h=m-1}^{\infty} c_{\bar{y},y} a_{m,x} r_{\bar{y}} q_{m,\bar{y}} [h-m+1] \cdot z^h + \sum_{m=0}^1 \sum_{h=0}^{\infty} c_{\bar{y},y} a_{m,x} r_{\bar{y}} q_{m,\bar{y}} [h-m+1] \cdot z^h \\
& + \sum_{m=0}^{N_{src}} \sum_{h=m}^{\infty} c_{y,y} a_{m,x} v_y q_{m,y} [h-m] \cdot z^h - c_{y,y} a_{0,x} v_y q_{0,y}^0 \\
& + \sum_{m=0}^{N_{src}} \sum_{h=m}^{\infty} c_{\bar{y},y} a_{m,x} v_{\bar{y}} q_{m,\bar{y}} [h-m] \cdot z^h - c_{\bar{y},y} a_{0,x} v_{\bar{y}} q_{0,\bar{y}}^0 \\
& + c_{y,y} a_{0,x} q_{0,y}^0 + c_{\bar{y},y} a_{0,x} q_{0,\bar{y}}^0
\end{aligned} \tag{C.5}$$

Using the definition of PGF as $Q_{x,y}(z) = \sum_{h=0}^{\infty} q_{x,y}[h] \cdot z^h$, (C.5) can be further simplified to:

$$\begin{aligned}
Q_{x,y}(z) = & c_{y,y} a_{0,x} r_y q_{0,y}^0 + c_{\bar{y},y} a_{0,x} r_{\bar{y}} q_{0,\bar{y}}^0 + \sum_{m=2}^{N_{src}} c_{y,y} a_{m,x} r_y Q_{m,y}(z) z^{m-1} \\
& + \sum_{m=2}^{N_{src}} c_{\bar{y},y} a_{m,x} r_{\bar{y}} Q_{m,\bar{y}}(z) z^{m-1} + \sum_{m=0}^{N_{src}} c_{y,y} a_{m,x} v_y Q_{m,y}(z) z^m \\
& + \sum_{m=0}^{N_{src}} c_{\bar{y},y} a_{m,x} v_{\bar{y}} Q_{m,\bar{y}}(z) z^m + c_{y,y} a_{0,x} r_y z^{-1} (Q_{0,y}(z) - q_{0,y}^0) \\
& + c_{y,y} a_{1,x} r_y Q_{1,y}(z) + c_{\bar{y},y} a_{0,x} r_{\bar{y}} z^{-1} (Q_{0,\bar{y}}(z) - q_{0,\bar{y}}^0) + c_{\bar{y},y} a_{1,x} r_{\bar{y}} Q_{1,\bar{y}}(z)
\end{aligned} \tag{C.6}$$

Rearranging the previous equation using $\xi_y(z) = v_y + r_y z^{-1}$ and $\xi_{\bar{y}}(z) = v_{\bar{y}} + r_{\bar{y}} z^{-1}$, then we obtain $Q_{x,y}(z)$ as:

$$\begin{aligned}
Q_{x,y}(z) = & c_{y,y}a_{0,x}(1 - \xi_y(z))q_{0,y}^0 + c_{\bar{y},y}a_{0,x}(1 - \xi_{\bar{y}}(z))q_{0,\bar{y}}^0 + c_{y,y}\xi_y(z) \cdot \sum_{m=0}^{N_{src}} a_{m,x}z^m Q_{m,y}(z) \\
& + c_{\bar{y},y}\xi_{\bar{y}}(z) \sum_{m=0}^{N_{src}} a_{m,x}z^m Q_{m,\bar{y}}(z)
\end{aligned} \tag{C.7}$$

Therefore, the PGF can be expressed as:

$$\begin{aligned}
\mathbf{Q}(z) = & \sum_{m=0}^{\infty} [\mathbf{A}^\dagger \text{diag}[z^x] \otimes \mathbf{C}^\dagger \mathbf{E}(z)]^{m+1} \cdot [\mathbf{I} \otimes [\mathbf{E}(z)^{-1} - \mathbf{I}]] \mathbf{Q}_0 \\
\text{where, } \mathbf{Q}(z) \triangleq & [Q_{0,0}(z), Q_{0,1}(z), \dots, Q_{N_{src},1}(z)]^\dagger, \\
\mathbf{Q}_m(z) \triangleq & [Q_{m,0}(z), Q_{m,1}(z)]^\dagger, \quad m = 0, 1, \dots, N_{src}, \\
\mathbf{E}(z) \triangleq & \text{diag}[\xi_0(z), \xi_1(z)], \quad \xi_y(z) \triangleq v_y + r_y z^{-1}, \\
\mathbf{Q}_0 \triangleq & [q_{0,0}^0, q_{0,1}^0, 0, 0, \dots, 0]^\dagger, \\
\text{diag}[z^x] \triangleq & \text{diag}[1, z, z^2, \dots, z^{N_{src}}], \\
\mathbf{A} \text{ is the transition probability matrix for HF arrival at RU,} \\
\mathbf{C} \text{ is the transition probability matrix for PLC channel state, and} \\
\uparrow \text{ is the matrix transpose}
\end{aligned} \tag{C.8}$$