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3GPP

Postal address

3GPP support office address

650 Route des Lucioles - Sophia Antipolis
Valbonne - FRANCE
Tel.: +33 4 92 94 42 00 Fax: +33 4 93 65 47 16

Internet

http://www.3gpp.org

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Foreword

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Introduction

The concept of Precoded EGPRS2 – PC EGPRS2 - DL has recently shown to provide significant performance gains for EGPRS2. Gains are expected both in interference and sensitivity limited scenarios, allowing significant increase to both data capacity and spectral efficiency for EGPRS2.

Performing precoding of EGPRS2 modulated data is expected to increase robustness of the system significantly while keeping the spectral properties of the signal intact.

1 Scope

The present document contains the results from the 3GPP study item on Signal Precoding Enhancements for EGPRS2 DL, SPEED.

The following is covered by the study:

- Objectives for the study
- Common assumptions to be used in the evaluation
- Technical work and performance evaluations based on the objectives

2 References

The following documents contain provisions which, through reference in this text, constitute provisions of the present document.

- References are either specific (identified by date of publication, edition number, version number, etc.) or non-specific.
- For a specific reference, subsequent revisions do not apply.
- For a non-specific reference, the latest version applies. In the case of a reference to a 3GPP document (including a GSM document), a non-specific reference implicitly refers to the latest version of that document *in the same Release as the present document*.

[1] 3GPP TR 21.905: "Vocabulary for 3GPP Specifications".

[2] GP-101088 “New SI proposal: Signal Precoding Enhancements for EGPRS2 DL”, source Telefon AB LM Ericsson, ST-Ericsson SA, Telecom Italia S.p.A, Vodafone Group Plc, China Mobile Com. Corp. ZTE Corporation, Huawei Technologies Co. Ltd.

[3] 3GPP TS45.003, v8.3.0 “Channel coding”

[4] 3GPP TS45.005, v9.3.0 “Radio transmission and reception”

3 Definitions and abbreviations

3.1 Definitions

For the purposes of the present document, the terms and definitions given in TR 21.905 [1] and the following apply. A term defined in the present document takes precedence over the definition of the same term, if any, in TR 21.905 [1].

3.3 Abbreviations

For the purposes of the present document, the abbreviations given in TR 21.905 [1] and the following apply. An abbreviation defined in the present document takes precedence over the definition of the same abbreviation, if any, in TR 21.905 [1].)

BMD	Blind Modulation Detection
CP	Cyclic Prefix
DAS	Downlink level A modulation and coding scheme
DBS	Downlink level B modulation and coding scheme
DFT	Discrete Fourier Transform
HSR	Higher Symbol Rate

MCS	Modulation and coding scheme
NSR	Normal Symbol Rate
PAR	Peak-to-Average Ratio
PC EGPRS2	Precoded EGPRS2
SPEED	Signal Precoding Enhancements for EGPRS2 DL
TSC	Training Sequence Code
USF	Uplink State Flag

4 Objectives

4.1 Performance Objectives

4.1.1 Improved EGPRS2 throughput

The introduction of Precoded EGPRS2, PC EGPRS2, shall significantly improve data throughput performance as compared to realistic EGPRS2 performance.

4.2 Compatibility Objectives

4.2.1 Spectral properties

PC EGPRS2 shall obey the current spectral requirements on spectrum due to modulation and wideband noise and on switching transients of EGPRS2 DL[4].

4.2.2 Impact to legacy service

The impact of PC EGPRS2 on GSM speech codecs, GPRS, EGPRS and EGPRS2 shall be kept at a minimum. Impact on cell reselection performance of mobile stations should be avoided by operation of PC EGPRS2 on the BCCH carrier. Impacts from PAN and USF multiplexing on PC-EGPRS2 and legacy user throughput should be minimized.

4.2.3 Implementation impact to base stations

The introduction of Precoded EGPRS2 in the base station transmitter should change BTS hardware as little as possible.

It is recommended for each candidate techinque to state compliance with Radix size used in LTE up to 5 (e.g. 2, 3, 5) in the complexity evaluation in order for the BTS to have an efficient implementation.

Note: SPEED TR will still open for other candidate techniques which are not fulfilling such criterion.

4.2.4 Implementation impact to mobile station

The introduction of Precoded EGPRS2 in the mobile station receiver should change MS hardware as little as possible. Both impact to stand-alone PC-EGPRS2 platforms and combined EGPRS2 and PC-EGPRS2 platforms shall be considered.

It is recommended for each candidate techinque to state compliance with Radix size used in LTE up to 5 (e.g. 2, 3, 5) in the complexity evaluation in order for the UE to have an efficient implementation.

Note: SPEED TR will still open for other candidate techniques which are not fulfilling such criterion.

5 Common assumptions for the evaluation

In the following sub-clauses common assumptions for the performance evaluation are listed

5.1 Overall design principles

5.1.1 Precoder module

The Inverse Discrete Fourier Transform shall be used as precoder module.

5.1.2 Burst length

The length of the active part of the burst (including current tail symbols) of EGPRS2 shall be kept for PC EGPRS2, i.e.

- EGPRS2-A: 148 NSR symbols
- EGPRS2-B: 177 HSR symbols

5.1.3 Cyclic prefix

One cyclic prefix length, defined in duration of microseconds, is required. When applied on PCE2-A and PCE2-B this duration must be translated to CP lengths defined by normal and higher symbol rate symbols respectively.

5.1.4 Positioning of training symbols

To enable easier detection of the precoded bursts only a single positioning vector of training symbols onto the burst for the respective EGPRS2 level will be applied, i.e. one for PC EGPRS2-A and one for PC EGPRS2-B.

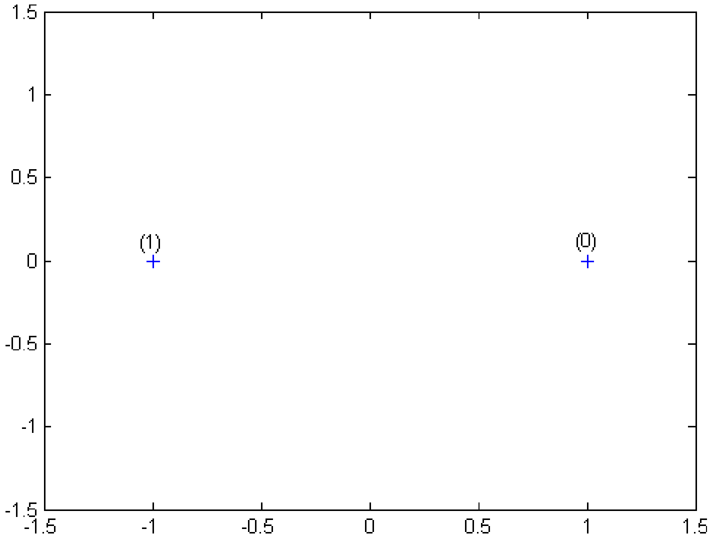
5.1.5 Modulation and coding scheme

All channel coding definitions of EGPRS2 shall be kept intact with regards to payload size, channel coding and interleaving, except for the highest MCS of each set, i.e. DAS-10/11/12 and DBS-10/11/12 for EGPRS2-A and EGPRS2-B, where only the payload size is required to be kept.

5.1.5.1 Modulation scheme

In addition to the modulation schemes defined in [4] for EGPRS2, BPSK and 64QAM can be used for the mapping of bits onto modulation symbols in the burst.

The definition of the additional modulation schemes are shown in Figure 5.1.



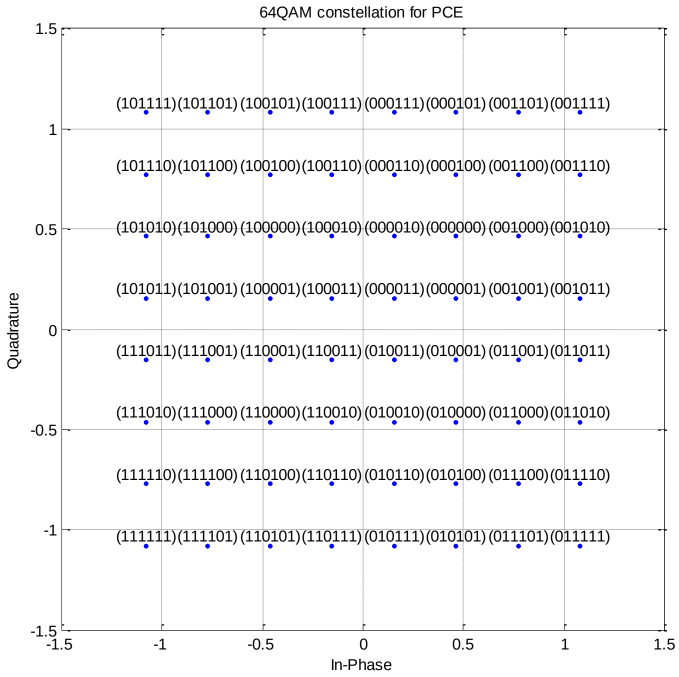


Figure 5.1 Modulation scheme definitions of BPSK and 64QAM.

5.1.5.2 Channel coding

For the re-definition of channel coding of DAS-10/11/12/DBS-10/11/12 the definitions in [1], sub-clause 5.1a.1.3 shall be used as baseline.

5.2 Link parameters

The following link level parameters are to be used as base line for the evaluation. Additional simulations assumptions may be used when deemed necessary.

Table 5.1 Link level simulation assumptions.

Parameter	Value
Link direction	Downlink
Frequency band	900MHz, 1900 MHz
Interference modulation	GMSK, Precoded modulation schemes
MCSs	DAS-5 - DAS-9, DAS-10b/11b/12b DBS-5 - DBS-9, DBS-10b/11b/12b.
Impairments	Typical TX and RX impairments.

Table 5.2 Link level simulation scenarios.

Propagation Conditions	Test Scenario			
	Sensitivity Limited	Co. channel interference	Adj. channel interference	DTS-2
Static	X			
TU3noFH		X	X	
TU3iFH		X	X	
TU50noFH	X	X	X	X
HT100noFH	X			
RA250noFH	X	X	X	

5.3 Evaluation output

When comparing EGPRS2 and PC EGPRS2 performance, as stated in subclause 8, each of the assumptions and test scenarios proposed in section 5.2 shall be considered. Ideal link adaptation throughput envelopes shall be used as baseline in the evaluation.

When evaluating the items of subclause 6, a subset of the listed assumptions in subclause 5.2, deemed as sufficient for the given evaluation, shall be used.

To enable comparison of simulations provided by different companies and/or between different proposals, details on receiver assumptions relevant to both complexity and performance shall be provided together with simulated results. Such details should at minimum include RX-filter bandwidth and TX and RX impairments used.

Absolute performance figures are to be provided when comparing PC EGPRS2 and EGPRS2 performance.

5a Concept description

In this sub clause an overall concept description of each proposed candidate techniques is given.

The description is based on the design used in the analysis on *Impact to base station and mobile station* (Section 6.6) and in the section on *Comparison of PC EGPRS2 with EGPRS2 performance* (Section 8).

The TR can also contain other design alternatives within the same candidate technique. However, these have not been used for the final proposal of each respective candidate technique.

5a.1 Single Block Precoded EGPRS2 – SBPCE2

Single Block Precoded EGPRS2 introduces an IDFT precoder module in the transmitter chain to form a single OFDM symbol modulated upon the GSM carrier frequency. Since only a precoder module is inserted in the transmitter chain the useful part of the burst is still pulse shaped as in legacy operation, keeping the spectral properties of the signal.

The IDFT is applied to both payload and TSC symbols in the burst. Conventional tail symbols used for up/down ramp and providing known states for the termination of the trellis search are discarded, and, instead these symbols are used to add a cyclic prefix to the transmitted OFDM symbol to keep the orthogonality of the transmitted sub carriers, taking the added time dispersive nature of Tx/Rx filters and channel propagation into account.

By applying the precoder module in the transmitter the Peak-to-Average (PAR) of the signal is significantly increased, and stochastic in nature, compared to the deterministic PAR characteristics of EGPRS2. In order to reduce the PAR without adding clipping distortion to the signal, a rotation based PAR reduction is applied where the TSC symbols and half of the data symbols are rotated. A predetermined set of rotation angles are used and for each burst the most suitable rotation angle is chosen. Rotation based PAR reduction is only applied to DBS-11b/12b since for lower MCSs the SINR operating range of the MCSs increases the detection error of the rotation and thus degrades performance, see subclause 6.4.1.

The number of total symbols in the burst has been kept, to the farthest extent possible, similar to the legacy EGPRS2 burst format while still finding a DFT size suitable for efficient FFT operations. The DFT sizes used for level A and B is 144 and 162 respectively to enable FFT Radix size of two and three, see subclause 6.2.1.1.

Since the tail symbols are discarded, and with an IDFT size of 144 for level A, a ramp function is applied to the signal, as decribed in Section 6.2.7.1, to comply with the normal burst duration and preserve spectrum due to switching transients, see subclause 6.2.7.1.

For level B the IDFT size of 162 will reduce the OFDM symbol duration and thus a longer CP has been added compared to the one used for level A, see 6.2.2.1.

Mixed modulations in the burst have been investigated as a means to keep the bit energy constant over the burst, despite suppression from the Tx filter on the edge carriers. It was found that the gains were depending on scenario investigated and thus it has not in general been applied to the design. However, with the chosen DFT size of 162 for level B a certain amount of modulation mix has been applied together with a reduced TSC length, compared to the 32 symbols currently used for EGPRS2-B, see subclause 6.2.5.1 and 6.2.1.1.

To improve performance of the higest MCSs in each respective EGPRS2 set, 64QAM modulation is used for MCSs (DAS-10b/11b/12b and DBS-10b/11b/12b). Further, to improve coderate, the number of TSC/pilot symbols has been reduced to 16 for these MCSs, see subclause 6.1.1.1.4.

The channel coding definitions, of the MCSs in terms of SF, USF, PAN, Payload and Header have been kept intact for all MCSs except for the three highest in each set where the rate of the channel coding for the payload have been reduced, while keeping the payload size. Also, rate matching and interleaving have been optimized for the highest MCSs.

The burst mapping has been modified with the TSC symbols evenly spread over sub carriers not significantly suppressed by the transmit pulse shape while for sub carriers attenuated at the carrier bandwidth edge, consecutive tranining/pilot symbols are used to improve channel estimation and equialization.

In order for the receiver to blindly detect the modulation used in the burst, a predefined set of circular shifts of the TSC symbol vector is performed, see 6.3.1.

To keep the performance of USF and the relative performance of SF/Header to data intact, the currently defined mapping of bits in each half burst is circularly shifted. The shift is different depending on the MCSs used and is performed to map the USF/SF/Header fields on stronger sub channels than what is achieved with the current EGPRS2 mapping. Also, bit swapping is applied to place the SF/Header bits, and the USF bits for 8PSK, on strong bit positions in the burst (for higher order modulations the USF bits are mapped to full symbols, as is done for EGPRS2), see subclause 6.2.4.1.

The design of Single Block EGPRS2 described in this sub clause, is captured in table 5a.1-1 and table 5a.1-2 for SBPCE2-A and SBPCE2-B respectively.

Table 5a.1-1. MCS design of SCPCE2-A.

MCS	DAS-5	DAS-6	DAS-7	DAS-8	DAS-9	DAS-10b	DAS-11b	DAS-12b
Modulation	8PSK			16QAM		64QAM		
Bit rate [kbps/TS]	22.4	27.2	32.8	44.8	54.4	65.6	81.6	98.4
Sub-carrier spacing	1.88 kHz							
CP length	6 NSR symbols [22.2 µs]							
DFT size	144							
Radix sizes	2,3							
TSC symbols	26					16		
Payload symbols	118					128		
Padded symbols	0							
Rotation based PAR reduction	No	No	No	No	No	No	No	No
Mixed modulation	No	No	No	No	No	No	No	No
Burst definition/limitation by predefined ramp	Yes	Yes	Yes	Yes	Yes	Yes	Yes	Yes
Circ. shift of burst mapping	50			45		70	75	50
Code rate (w/o PAN)	0.37	0.45	0.54	0.56	0.68	0.47	0.59	0.70
# puncturing patterns	2	2	2	2	3	2	2	3
Swap parameter in rate matching [%]	5	0	0	0	0	5	0	0
Interleaver type	$^{(1)}$			1		1	1	
Interleaver parameter 'a'	$^{(1)}$			199		641	117	
NOTE1: Interleaver of MCS-5, see 3GPP TS45.003								

Table 5a.1-2. MCS design of SCPCE2-B.

MCS	DBS-5	DBS-6	DBS-7	DBS-8	DBS-9	DBS-10b	DBS-11b	DBS-12b
Modulation(1)	QPSK		16QAM			64QAM		
Bit rate [kbps/TS]	22.4	29.6	44.8	59.2	67.2	88.8	108.8	118.4
Sub-carrier spacing	2.01 kHz							
CP length	15 HSR symbols [46.2 µs]							
DFT size	162							
Radix sizes	2,3							
TSC symbols	26					16		
Payload symbols	136					146		
Padded symbols	0							
Rotation based PAR reduction	No	No	No	No	No	No	Yes	Yes
Mixed modulation (each burst half)	QPSK: 66 8PSK: 2		16QAM: 64 32QAM: 4			No	No	No
Burst definition/limitation by predefined ramp	Yes	Yes	Yes	Yes	Yes	Yes	Yes	Yes
Circ. shift of burst mapping	45		55		65	40	20	0
Code rate (w/o PAN)	0.49	0.63	0.47	0.60	0.71	0.56	0.70	0.75
# puncturing patterns	2	2	2	2	3	2	3	3
Swap parameter in rate matching [%]	5	15	0	10	0	0	0	5
Interleaver type	1		1		1	1	1	
Interleaver parameter 'a'	173		325		283	318	595	
NOTE1: Main modulation								

* Main modulation

5a.2 Padded Higher Order Modulation– Padded HOM

Padded HOM introduces a Symbol swapping and an IDFT precoder module in the transmitter chain to form a single OFDM symbol modulated upon the GSM carrier frequency. The pulse shaping and the spectral properties of the signal are kept intact for legacy EGPRS2.

Symbol swapping module swaps half burst symbol's position to ensure the performance of USF and Header in a simple way, see subclause 6.2.4.2.

One level higher order modulation is utilized for each coding scheme so that only a portion of symbols are needed to carry the same amount of payload. The padded symbols are introduced and could be arranged at the edges of the signal bandwidth to adapt to the intact total symbol numbers for the burst. The padded symbols have different choices. Zero-padded symbols are used for the final design evaluation. For the re-designed MCS DAS-10b/11b/12b v1/v2 and DBS-10b/11b/12b v1/v2, no padded symbols are ultilized.

In order to achieve high efficient FFT/IFFT implementation and comply with radix size used in LTE up to 5, the DFT length used for level A and B is 144 and 162 respectively, see subclause 6.2.1.2.

Conventional tail symbols are discarded and instead these symbols are used to add a cyclic prefix. With the DFT length of 144 for level A, some change in ramping is applied to the signal, see subclause 6.2.7.1.

For level B the DFT length of 162 will reduce the OFDM symbol duration and thus a longer CP has been added compared to the one used for level A, see 6.2.2.2.

Different TS lengths with corresponding number of padded symbols have been investigated to improve the performance for Padded HOM. The number of TS symbols has been reduced to 17/18 for level-A and level-B MCSs except the three higest MCSs in each respective EGPRS2 set. The position of the TS symbols is evenly spread over the whole burst, see subclause 6.2.3.2.

In order for the receiver to blindly detect the modulation used in the burst a predefined set of circular shifts of the TS symbol vector is performed, see 6.3.1.

To improve performance of the highest MCSs in each respective EGPRS2 set, 64QAM modulation is used for MCSs (DAS-10b/11b/12b and DBS-10b/11b/12b). To improve the coderate, the number of TS symbols has been further reduced to 14/13 for these MCSs, see subclause 6.1.1.2.

The channel coding definitions of the MCSs in terms of SF, USF, PAN, Payload and Header have been kept intact for all MCSs except for the three highest in each set where the coderate of the channel coding for the payload have been reduced, while keeping the payload size intact.Furthermore, the puncturing and interleaving parameters of the three highest MCSs in each set have been re-searched, see subclause 6.1.1.2.

To optimize the performance of USF and the relative performance of SF/Header to data intact the currently defined mapping of bits to the burst is circularly shifted and bit swapping is applied for the three highest MCSs in each respective EGPRS2 set, depending on the MCSs used, see subclause 6.1.1.2.

The design of Padded HOM described in this sub clause, is captured in table 5a.2-1 and table 5a.2-2 for Padded HOM-A and Padded HOM-B respectively.

Table 5a.2-1. MCS design of Padded HOM-A.

MCS	DAS-5	DAS-6	DAS-7	DAS-8	DAS-9	DAS-10b	DAS-11b	DAS-12b
Modulation	16QAM			32QAM		64QAM		
Bit rate [kbps/TS]	22.4	27.2	32.8	44.8	54.4	65.6	81.6	98.4
Sub-carrier spacing	1.88 kHz							
CP length	6 NSR symbols [22.2 µs]							
DFT size	144							
Radix sizes	2,3							
TSC symbols	17					14		
Payload symbols	87			93		130		
Padded symbols	40			34		0		
Rotation based PAR reduction	No							
Mixed modulation	No							
Burst definition/limitation by predefined ramp	Yes	Yes	Yes	Yes	Yes	Yes	Yes	Yes
Circ. shift of burst mapping	0			0		20	50	0
Code rate (w/o PAN)	0.37	0.45	0.54	0.56	0.68	0.46	0.58	0.69
# puncturing patterns	2	2	2	2	3	2	2	3
puncturing parameters swap	swap = 0.05	swap = 0	swap = 0	swap = 0	swap = 0	swap = 0	swap = 0	swap = 0
Interleaving parameters a	Refer to 45.003 MCS-5			Type 1 a =199	Type 1 a =199	Type 1 a = 87	Type 1 a = 103	Type 1 a = 643

Table 5a.2-2. MCS design of Padded HOM-B.

MCS	DBS-5	DBS-6	DBS-7	DBS-8	DBS-9	DBS-10b	DBS-11b	DBS-12b
Modulation	8PSK		32QAM			64QAM		
Bit rate [kbps/TS]	22.4	29.6	44.8	59.2	67.2	88.8	108.8	118.4
Sub-carrier spacing	2.01 kHz							
CP length	15 HSR symbols [46.2 µs]							
DFT size	162							
Radix sizes	2,3							
TSC symbols	18					13		
Payload symbols	92		111			149		
Padded symbols	52		33			0		
Rotation based PAR reduction	No							
Mixed modulation (each burst half)	No							
Burst definition/limitation by predefined ramp	Yes	Yes	Yes	Yes	Yes	Yes	Yes	Yes
Circ. shift of burst mapping	0		0			0	0	0
Code rate (w/o PAN)	0.49	0.63	0.47	0.60	0.71	0.55	0.68	0.74
# puncturing patterns	2	2	2	2	3	2	3	3
puncturing parameters **swap**	swap = 0.05	swap = 0.15	swap = 0	swap = 0.1	swap = 0	swap = 0	swap = 0	swap = 0
interleaving parameters **a**	Type 1 a = 173	Type 1 a = 173	Type 1 a = 325	Type 1 a = 325	Type 1 a = 283	Type 1 a = 107	Type 1 a = 107	Type 1 a =107

6 Design and Evaluation

6.1 Re-design of DAS-10b/11b/12b/DBS-10b/11b/12b

A re-definition of DAS-10/11/12 and DBS-10/11/12 for EGPRS2-A and EGPRS2-B respectively is foreseen in order to optimize performance. The payload size of each MCS, i.e. 2x82, 3x68, 3x82 bytes and 3x74, 4x68 and 4x74 bytes for DAS-10/11/12 and DBS-10/11/12 respectively are to be kept.

This sub-clause evaluates modifications in the MCS design and its impact on performance.

6.1.1 Burst formatting and channel coding

In the sub clauses below a re-design of DAS-10/11/12 and DBS-10/11/12 is proposed. These new redesigned MCSs have been denoted DAS-10b/11b/12b and DBS-10b/11b/12b in this technical report.

Two designs of DAS-12b and DBS-12b are presented, and denoted v1 and v2. The difference between the two designs is the number of TSC symbols used. DAS-12b and DBS-12b v1 is used in this document, unless otherwise stated, while DAS-12b v2 and DBS-12b v2 is used in the final performace evalution in Sub clause 8.

6.1.1.1 Burst formatting for Single Block PCE2

6.1.1.1.1 Introduction

This sub-clause proposes a new burst formatting design for DAS-10b/11b/12b and DBS-10b/11b/12b for Single Block PCE2 in terms of

- Mapping of bits onto modulation symbols and
- Burst mapping of Header and Data fields.

For the mapping of bits onto modulation symbols the objective is to find the best mixed mode modulation combination for the studied MCSs in terms of optimal throughput performance.

For the burst mapping of Header and Data fields the objective is to derive

- header bit swapping, which swaps header bits at weak positions with data bits at strong positions and
- burst bit shifting, which circularly shifts each half burst so that part of, or all header bits end up at higher SNR region,

to achieve robust header and incremental redundancy performance.

The evaluation in section 6.1.1.1.2 and 6.1.1.1.3 was for simplicity only performed for DAS-12b and DBS-12b, as it was shown in [6.1-9] that the derived designs are robust enough to be applied also to DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2.

6.1.1.1.2 Evaluation Method

Several best MMM candidates are first derived, in terms of providing as robust performance as possible at lower SNR region, which will later lead to robust data performance when header bit swapping and shifting is used. This evaluation is done by placing all header/USF/SF bits at high SNR region, and focusing therefore on the data BLER performance.

The header bit swapping and shifting are then evaluated together with the best MMM candidates. The best burst formatting for DAS-12b and DBS-12b is then selected as the one that keeps the relative performance between header and data BLER (using non-precoded DAS-12 (data BLER@50% to avoid error floor and headerBLER@1%) and DBS-11(data BLER@10% and headerBLER@1%) as reference), meanwhile achieves as low data BLER as possible.

For more information please refer to [6.1-2].

6.1.1.1.3 Simulation Assumptions and Results

6.1.1.1.3.1 Simulation Assumptions

The selection of the optimal burst formatting, including MMM, header bit swapping and shifting is based on performance simulated in a sensitivity limited scenario given a TU50nFH propagation environment. The data and header BLER are then verified in other propagation models and interference scenarios.

A detailed list of simulation assumptions are presented in table 6.1-1.

Table 6.1-1 Simulation assumptions

Parameter	Value
MCSs	DAS-12b, DBS-12b
TSC placement	According to [6.1-3]
Burst length	According to [6.1-4]
CP length	PCE2A: 6 PCE2B: 9
RX BW	PCE2A: 240kHz PCE2B: 275kHz
ICI Suppression	No
Backoff	No
Channel propagation	TU50nFH, HT100nFH
Interference	AWGN, CO, DTS-2 modified
Frequency band	900 MHz
Puncturing	Patterns generated with swap set to 0.
Interleaver	Interleaver type 1 with a set to 97.
Frames	5000
Tx/Rx impairments - Phase noise - I/Q gain imbalance - I/Q phase imbalance - DC offset - Frequency error	Ericsson typical TX/RX impairments: 0.8 / 1.2 [degrees (RMS)] 0.1 / 0.2 [dB] 0.2 / 2.0 [degrees] -45 / -40 [dBc] - / 25 [Hz]

6.1.1.1.3.2 Simulation Results

Evaluation according to 6.1.1.2 shows that the best burst formatting is:

- For DAS-12b: MMM pattern containing 56 64QAM symbols, one 32QAM and one 8PSK symbol, with header bit swapping and a right shift of 50 bits;
- for DBS-12b: no MMM, with header bit swapping and no shifting.

Simulation results are listed in table 6.1.1-2 and table 6.1.1-3.

Table 6.1-2: Absolute Performance with/without optimal burst format.

MCS	Intf. Scen.	Data BLER @ 10%					
		No bitswap/shifting & no MMM		With bitswap(and shifting), no MMM		With bitswap(and shifting) & Opt MMM	
		TU50nFH	HT100nFH	TU50nFH	HT100nFH	Tu50nFH	HT100nFH
DAS-12b	CO	26,5	27,3*	26,9	29,5*	26,4	27,5*
	DTS-2	28,1	28,8*	28,9	31,5*	28,2	29,5*
	AWGN	27,7	28,7*	28,8	31,6*	27,9	29,3*
DBS-12b	CO	26,6		27,1		27,1	
	DTS-2	30,0		31,2		31,2	
	AWGN	28,4		30,0		30,0	

*Data bler@30% for HT100nFH;

MCS	Intf. Scen.	Header BLER @ 1%					
		No bitswap/shifting & no MMM		With bitswap and Shifting, no MMM		With bitswap(and shifting & Opt MMM	
		TU50nFH	HT100nFH	TU50nFH	HT100nFH	Tu50nFH	HT100nFH
DAS-12b	CO	17,4	21	13,3	13,0	13,1	12,8
	DTS-2	24,7	27,4	15,0	14	15,2	14
	AWGN	26,5	28,3	12,4	12,0	12,5	11,8
DBS-12b	CO	18,5		14,6		14,6	
	DTS-2	31,9		17,4		17,4	
	AWGN	31,2		14,3		14,3	

Table 6.1-3: Relative performance using optimal burst formatting, comparing with DAS-12b/DBS-12b no bitswap & no MMM.

MCS	Intf. Scen.	Data improvements(dB) With bitswap(and shifting) & Opt MMM		Header improvements(dB) With bitswap(and shifting) & Opt MMM	
		TU50nFH	HT100nFH	TU50nFH	HT100nFH
DAS-12b	CO	0,1	-0,2	4,3	8,2
	DTS-2	-0,1	-0,7	9,5	13,4
	AWGN	-0,2	-0,6	14,0	16,5
DBS-12B	CO	-0,5		3,9	
	DTS-2	-1,2		14,5	
	AWGN	-1,6		16,9	

It can be seen that the use of header bit swap and burst shift significantly improves the header performance of DAS-12b and DBS-12b, up to 17dB, meanwhile, the data performance has degraded moderately, in almost all the simulated scenarios. It should also be noticed that, for DAS-12b, the use of MMM decreases the degradation in data performance due to header bit swapping and shifting.

6.1.1.1a TSC symbol position for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 for SBPCE2

6.1.1.1a.1 Introduction

In EGPRS2-A and EGPRS2-B the TSC lengths were chosen to secures accurate receiver channel estimation to facilitate synchronization, ISI suppression and equalization. As SBPCE2 introduces a new air interface, where e.g. orthogonality between adjacent symbols is granted by the introduction of a cyclic prefix, it is not necessarily true that the EGPRS2 TSC lengths are optimal for SBPCE2. In the following an empirical method is described that is designated to investigate the optimal TSC length for SBCPE2-A DAS-10b/11b/12b v2 and Low Complexity SBPCE2-B DBS-10b/11b/12b v2. For detailed information see [6.1-9].

The method is based on a search for a TSC vector *TSC(T,cn,sp)* of length *T* and defined by a number *cn* of symbols located in the centre of the burst and a number *T-cn* of symbols evenly distributed over the remains of the burst, starting at position *sp*, as illustrated in Figure 6.2-4.

The task is to find the best TSC vector, offering optimum performance for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2, among the set of TSC positing vectors *TSC(T,cn,sp)* defined by:

$$T \in \{10,12,14,16,18,20,22,24,26\}$$

$$cn \in \{0,1,2,3,4,5\}$$

$$sp = \{1,3,7\}$$

6.1.1.1a.2 Evaluation Method

To find the best TSC vector the Data and Header BLER performance for each TSC positing vector *TSC(T,cn,sp)* was evaluated for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 given TU50noFH propagation conditions and;

- Sensitivity limited,
- Co-channel interference limited and
- Adjacent channel interference limited scenarios.

During the evaluations simulation settings according to Table 6.1-4 was used. For each evaluated TSC length new puncturing patterns and interleaver schemes were derived, using a best guess value on interleaver parameter *a* and the puncturing *swap* parameter set to 0. Header definitions were in accordance with section 6.1.1.1.

Table 6.1-4. Simulation settings.

Simulation parameters	
Frequency band	900
Tx filter	LinGMSK
Rx filter BW	SBPCE2-A: 280 kHz LCSBPCE2-B: 340 kHz
Blind Detection of modulation	On
MCSs	SBPCE2-A DAS10b/11b/12b v2 LC SBPCE2-B DBS10b/11b/12b v2
Frames	20000
MMM	No
Tx/Rx impairments - Phase noise [degrees (RMS)] - I/Q gain imbalance [dB] - I/Q phase imbalance [deegrees] - DC offset [dB] - PA model - Frequency error [Hz]	Tx/Rx 0.8/1.2 0.1/0.2 0.2/2.0 -45/-40 On [8-8]/ - -/25
TSC placement	Reference according to subclause 6.2.3.1, else according to *TSC(T,cn,sp).*
TSC length	10-26 symbols
Puncturing	Patterns P1, P2 and P3, if applicable generated with swap set to 0.
Interleaver	Interleaver type 1 with a set to 97.
PAR reduction method	On with hard clipping and soft clipping
Achieved PAR	6 dB
CP length	SBPCE2-A: 6 LCSBPCE2-B: 15

The achieved performance for each MCS and scenario was compared with the reference performance achieved when utilizing the optimum TSC positioning vector found in subclause 6.2.3.1 and Table 6.2-5a. The relative performance at 10% Data BLER was used as metric to identify the best TSC vector *TSC(T,cn,sp)* for each MCS. Table 6.1-4a shows the results for the best TSC vector *TSC(16,4,7)* for DAS-10b/11b/12b v2. Gains are observed in all scenarios. The same was true for DBS-10b/11b/12b v2 where *TSC(16,4,7)* gave best performance, and gains on all scenarios except for DBS-10b when exposed to adjacent interference.

Table 6.1-4a. Relative data performance improvement for DAS-10b/11b/12b v2 in TU50nFH propagation conditions.

TSC(T,cn,sp)	Es/N0 [dB]	C/ICO [dB]	C/IADJ [dB]	Average [dB]
DAS-12b v2				
TSC(16,4,7)	2.2	2.7	2.9	2.6
DAS-11b				
TSC(16,4,7)	0.9	1.4	0.9	1.1
DAS-10b				
TSC(16,4,7)	0.6	0.8	0	0.5

The improvement in Data BLER, seen in Table 6.1-4a, is achieved by a decreased Data code rate. As the Header block code is unaffected by the choice of *TSC(T,cn,sp)* it must be verified that the improved data performance is not achieved at the expense of an unacceptable degradation in header performance. In Table 6.1-4b it is shown that the relative header performance degradation for DAS-10b/11b/12b v2 is limited. The same is true for DBS-10b/11b/12b v2 header performance.

Table 6.1-4b. Relative header performance improvement for DAS-10b/11b/12b v2 in TU50nFH propagation conditions.

TSC(T,cn,sp)	Es/N0 [dB]	C/ICO [dB]	C/IADJ [dB]	Average [dB]
DAS-12b v2				
TSC(16,4,7)	-1.0	-0.8	-0.7	-0.9
DAS-11b				
TSC(16,4,7)	-1.0	-0.8	-0.7	-0.9
DAS-10b				
TSC(16,4,7)	-1.0	-0.7	-1.0	-0.9

6.1.1.1a.3 Verification and Conclusion

To secure that the performance gains were not limited to TU50nFH propagation conditions the choice of TSC vectors *TSC(16,4,7)* for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 was verified in the following scenarios:

- Es/N0, TU50nFH,
- Es/N0, HT100nFH,
- Es/N0, RA250nFH,
- C/I_{CO}, TU3iFH and
- C/I_{ADJ}, TU3iFH

Table 6.1-4c summarizes, in similarity to Table 6.1-4a, the relative performance improvement at 10% Data BLER for each of the evaluated MCSs in the listed set of interference and sensitivity limited scenarios. Gains are observed in all scenarios except for DBS-10b in the C/I_{ADJ}, TU3iFH evaluation. This confirms the conclusion that *TSC(16,4,7)* was the best choice for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2.

Table 6.1-4c. Performance improvement defined at 10% Data BLER.

	Es/N0, TU50nFH	Es/N0, HT100nFH	Es/N0, RA250nFH	C/ICO, TU3iFH	C/IADJ, TU3iFH
DAS-12b v2	2.2	*	**	2.4	2.3
DAS-11b	0.9	1.7	**	1.4	0.7
DAS-10b	0.6	0.5	3.5	0.8	0
DBS-12b v2	1.2	**	**	1.6	0.7
DBS-11b	0.7	**	**	1.5	0.3
DBS-10b	0.2	1.5	**	1.0	-0.9

* Reference performance never reaches Data BLER of 10% or below.

** Reference and TSC(16,4,7) performance never reaches Data BLER of 10% or below.

6.1.1.1b Puncturing and Interleaver design for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 for SBPCE2

6.1.1.1b.1 Introduction

Sub-clause 6.1.1.1.4 presents a proposal to reduce the TSC length and decrease the code rate for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 to achieve throughput gains at high SINR levels. During the evaluations performed [6.1-9] to derive the results presented in sub-clause 6.1.1.1.4 the optimization of puncturing patterns and interleaver schemes were left for further study. In this sub-clause the evaluation is made complete as a search for optimal puncturing patterns and interleaver schemes is performed. For detailed information see [6.1-10].

6.1.1.1b.2 Evaluation Method

The EGPRS2 puncturing pattern and interleaver scheme design are based on the *swap* and *a* design parameters. To derive optimal *swap* and *a* parameters for each of DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 the method outlined in [6.1-7] was followed, i.e. to

- Simulate incremental redundancy Data BLER performance for a range of *swap* values, given an initial best guess of the *a* parameter value.
- Select the best *swap* value and simulate first transmission Data BLER performance over a range of feasible *a* values for the chosen interleaver type.

In the evaluation Interleaver type 1 was used since it is typically used for EGPRS2 MCS with low code rate and secures a good spread of bits over the entire radio block as well as within each burst. It restricts *a* to take on a value less than the number of data bits per burst, N_b, and prohibits *a* to have factors common to N_b.

Table 6.1-4d lists the code rate for each of the targeted MCSs along with the applied interleaver type and the required number of puncturing patterns.

Table 6.1-4d. Code rate, interleaver type and number of puncturing patterns.

	DAS-10b	DAS-11b	DAS-12b	DBS-10b	DBS-11b	DBS-12b
Code rate	0.47	0.59	0.70	0.56	0.70	0.75
Interleaver type	1	1	1	1	1	1
Puncturing patterns	2	2	3	2	3	3

The simulations were performed for TU50nFH and TU3nFH channels which are believed to correspond to typical propagation conditions, while covering scenarios where high as well as low fading diversity can be expected. Table 6.1-4e presents a detailed summary of the simulation assumptions.

Table 6.1-4e. Simulation settings.

Simulation parameters

Channel	TU50, TU3
Frequency hopping	No
Frequency band	900
Tx filter	LinGMSK
Rx filter BW	SBPCE2-A: 280 kHz SBPCE2-B: 340 kHz
MMM	No
Blind Detection of modulation	On
MCSs	SBPCE2-A DAS10b/11b/12b v2 LC SBPCE2-B DBS10b/11b/12b v2
Frames	80000
Tx/Rx impairments	Tx/Rx
- Phase noise [degrees (RMS)]	0.8/1.2
- I/Q gain imbalance [dB]	0.1/0.2
- I/Q phase imbalance [deegrees]	0.2/2.0
- DC offset [dB]	-45/-40
- PA model	On [8-8]/ -
- Frequency error [Hz]	-/25
TSC placement and length	*TSC(16,4,7)*
Swap	0, 5, 10, 15, 20
Interleaver	Interleaver type 1
a	DAS-10b: 1-705, DAS-11/12b: 1-700 DBS-10b: 1-810, DAS-11/12b: 1-805
PAR reduction method	On with hard clipping and soft clipping
Achieved PAR	6 dB
CP length	SBPCE2-A: 6 NSR symbols SBPCE2-B: 15 HSR symbols

6.1.1.1b.3 Puncturing design

Incremental redundancy performance was evaluated for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2. Figure 6.1 shows the performance for DAS-12b and DBS-12b RV 1, 2 and 3 over the range of selected *swap* values, simulated for TU50nFH and TU3nFH, relative the performance achieved for the best choice of *swap*. Reference [6.1-10] presents similar results for DAS-11b/10b and DBS-11b/10b.

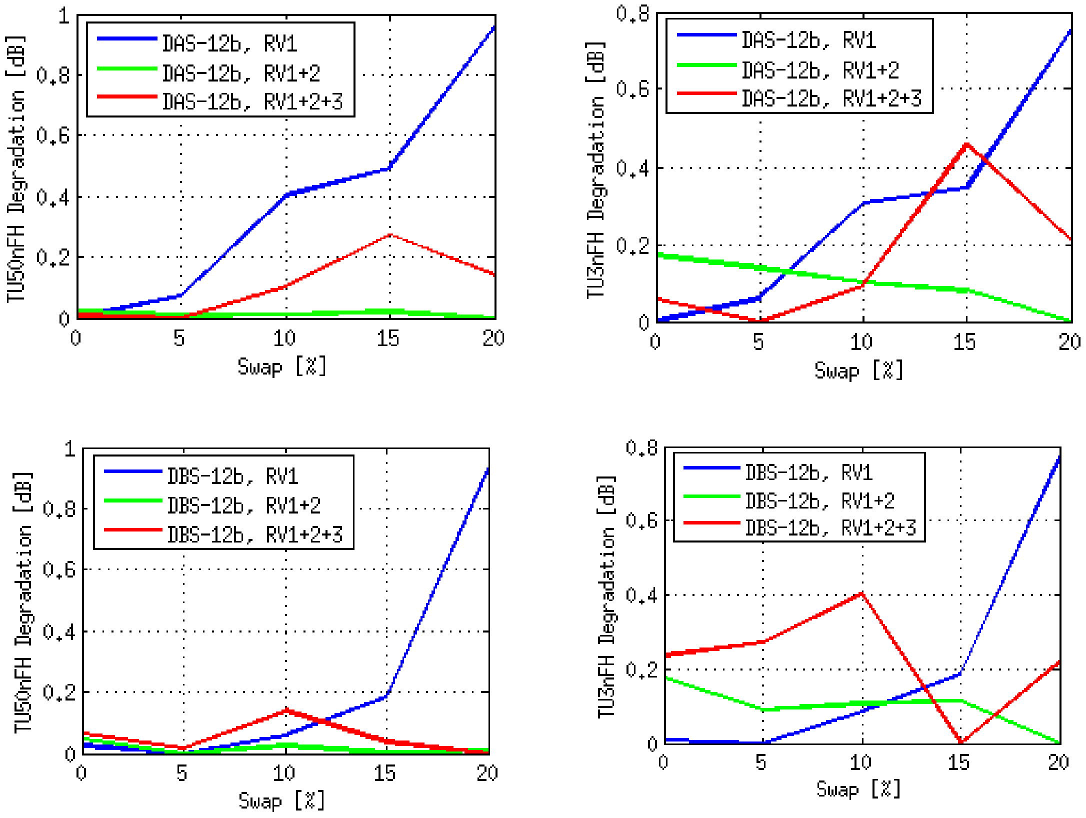


Figure 6.1 DAS-12b v2 (upper) and DBS-12b v2 (lower) evaluation of puncturing *swap* parameter for Tu50nFH (left) and TU3nFH (right) propagation conditions.

To optimize incremental redundancy performance, and minimize the total degradation computed over the simulated scenarios, it was concluded that swap equal to 0% shall be used for DAS-11b/12b v2 and DBS-10b/11b while swap equal 5% is the best choice for DAS-10b and DBS-12b v2. Table 6.1-4f summarizes the results.

Table 6.1-4f. Derived *swap* values for each studied MCS.

MCS	DAS-10b	DAS-11b	DAS-12b v2	DBS-10b	DBS-11b	DBS-12b v2
Swap	5	0	0	0	0	5

6.1.1.1b.4 Interleaver design

The first transmission Data BLER performance was, for each studied MCS, evaluated over a range of a values as defined in Table 6.1-4e, at a fixed Es/N0 corresponding to approximately 10% Data BLER. The puncturing *swap* parameter was set in accordance to the findings in sub-clause 6.1.1.1.5.3.

Figure 6.2 below shows the performance for DAS-12b v2 and DBS-12b v2 over the entire range of a values, simulated for TU50nFH and TU3nFH, relative the performance achieved for the best choice of a. As the resulting performance from the TU50nFH and TU3nFH simulations were strongly correlated an average value of the relative performance was selected as metric in the search for the optimal choice of a. Reference [6.1-10] presents similar results for DAS-11b/10b and DBS-11b/10b.

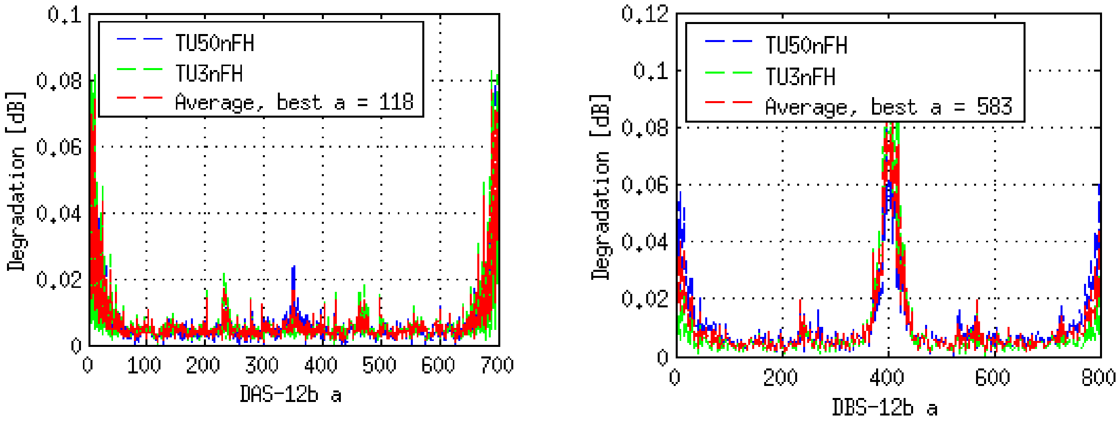


Figure 6.2 DAS-12b v2 (left) and DBS-12b v2 (right) evaluation of Interleaver type 1 a parameter for TU50nFH and TU3nFH propagation conditions.

The optimal *a* value for each MCS was identified as the value offering the lowest metric, i.e. the best average performance over the simulated scenarios. As DAS-11b and DAS-12b v2 have same data block size they can for simplicity be equipped with identical interleaver schemes. The same principle applies to DBS-11b and DBS-12b v2. A summary of the best *a* value, given the condition that DAS-11b and DAS-12b v2 as well as DBS-11b and DBS-12b v2 shall have identical interleavers, is presented in Table 6.1-4g.

Table 6.1-4g. Summary of derived Interleaver type 1 a values.

MCS	DAS-10b	DAS-11b	DAS-12b v2	DBS-10b	DBS-11b	DBS-12b v2
a	641	117	117	318	595	595

6.1.1.1b.5 Conclusion

This section presents an investigation of interleaver and puncturing design for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 based on EGPRS2 Interleaver type 1 a and puncturing swap parameters. Optimal values for the studied parameters have been found through a search over a range of feasible a and selected swap values given TU50nFH and TU3nFH propagation conditions in a sensitivity limited scenario. Table 6.1-4f and 6.1-4g summarizes the findings.

The results from these investigations show that the SBPCE2 performance is robust over a range of puncturing swap and interleaver a parameter values. It is thus concluded that the results from earlier investigations, performed for SBPCE2, where swap has been set to 0% and arbitrary a values have been used, can be considered as robust.

6.1.1.2 Coding parameters and burst formatting for Padded HOM

6.1.1.2.1 Design assumptions

- 64QAM modulation is used over the whole burst for the new MCSs DAS-10b/11b/12b and DBS-10b/11b/12b.
- The same Header, USF and SF definitions for DAS-10/11/12 and DBS-10/11/12 respectively are to be kept.
- The length of the Training Sequence in the burst could be different from DAS-10/11/12 and DBS-10/11/12.
- The mother code and rate matching algorithms are kept the same as EGPRS2 described in TS 45.003 chapter 5.1a.1.3. The interleaver Type1 defined in TS45.003 5.1a.2.1 is used in this study.

Therefore the coding design of the puncturing and interleaving algorithms is limited to determine the rate matching parameter *swap* and the interleaver parameter *a*.

6.1.1.2.2 Design Method

A set of different TS lengths and corresponding best choice of rate matching and interleaver parameters are first derived.

The RV1 Data BLER performances with different TS lengths under the selected coding parameters are evaluated in different propagations and interference scenarios. After comprehensive comparison the TS length 12 symbols for DAS-12b and 13 symbols for DBS-12b are selected.

Burst shift and bit swapper according to [6.1-8] are utilized in searching the best burst format. Based on the performance evaluations, it is shown that the best burst format for DAS-12b is header bit swapping and shifting of 88 bits. And the best format for DBS-12b is header bit swapping and shifting of 54 bits. This version of re-design MCSs could be denoted v1 or no version indication is given, unless otherwise stated.

For more information please refer to [6.1-6].

For Low complexity Padded HOM, the re-designed MCSs have been denoted v2. The TS length 14 and 13 symbols are proposed for DAS-10b/11b/12b v2 and DBS-10b/11b/12b v2 respectively.

6.1.1.2.3 Simulation Assumptions and Results

6.1.1.2.3.1 Simulation Assumptions

The simulation assumptions are listed in table 6.1-5. The burst formats and coding parameters of the new channel coding are depicted in table 6.1-6 and table 6.1-7. Only the simulation results of DAS/DBS-12b are provided in the following section.

Table 6.1-5 Simulation assumptions

Parameter	Value
	DAS/DBS-12b for Padded HOM
IDFT length	PCE2A: 140 PCE2B:168
CP length	PCE2A: 8 PCE2B: 9
TSC length	PCE2A: 12 PCE2B: 13
TSC placement	According to [6.1-5]
RX BW	PCE2A:270kHz PCE2B:325kHz
ICI Suppression	No
Backoff	No
Channel propagation	TU3iFH, TU50noFH
Interference	AWGN, CO (GMSK), ADJ (GMSK)

Frequency band	900 MHz
Frames	5000
PAPR reduction	No
Tx/Rx impairments	No

Table 6.1-6 Burst format for re-designed MCSs for Padded HOM

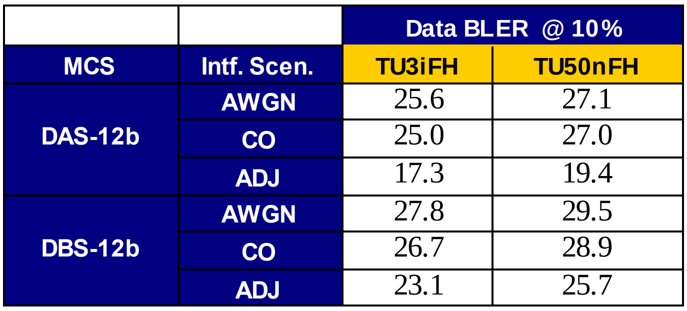
MCS	Bit swapper	Burst shift bits
DAS-12b	on	88
DBS-12b	on	54
DAS-10b v2	on	20
DAS-11b v2	on	50
DAS-12b v2	on	0
DBS-10b v2	on	0
DBS-11b v2	on	0
DBS-12b v2	on	0

Table 6.1-7 Rate matching and interleaving parameters for re-designed MCSs for Padded HOM

MCS	*swap*	*a*
DAS-12b	0%	622
DBS-12b	0%	98
DAS-10b v2	0%	87
DAS-11b v2	0%	103
DAS-12b v2	0%	643
DBS-10b v2	0%	107
DBS-11b v2	0%	107
DBS-12b v2	0%	107

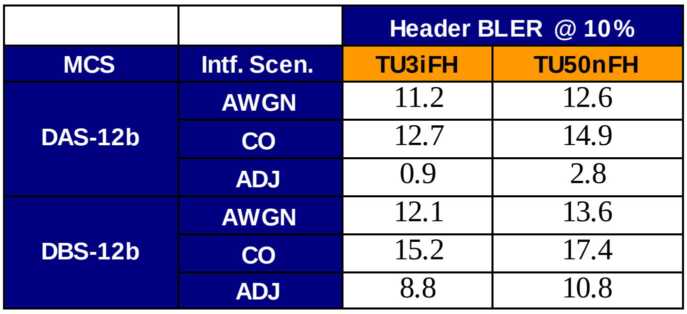
6.1.1.2.3.2 Simulation Results

Table 6.1-8 Data BLER performance of DAS/DBS-12b for Padded HOM



		Data BLER @ 10%	
MCS	Intf. Scen.	TU3iFH	TU50nFH
DAS-12b	AWGN	25.6	27.1
	CO	25.0	27.0
	ADJ	17.3	19.4
DBS-12b	AWGN	27.8	29.5
	CO	26.7	28.9
	ADJ	23.1	25.7

Table 6.1-e Header BLER performance of DAS/DBS-12b for Padded HOM



		Header BLER @ 10%	
MCS	Intf. Scen.	TU3iFH	TU50nFH
DAS-12b	AWGN	11.2	12.6
	CO	12.7	14.9
	ADJ	0.9	2.8
DBS-12b	AWGN	12.1	13.6
	CO	15.2	17.4
	ADJ	8.8	10.8

6.1.2 References

[6.1-1] GP-100918, “Precoded EGPRS2 Downlink”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#46

[6.1-2] GP-101852, “DAS-12b and DBS-12b Burst Formatting”, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN#48

[6.1-3] GP-101350, “Training symbol placements in Precoded EGPRS2 DL”, source Telefon AB LM Ericsson, ST-Ericsson. GERAN#47

[6.1-4] GP-101349, “Aspects of burst formatting of Precoded EGPRS2 DL”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.1-5] GP-101770, "Burst format of Improved Precoded EGPRS2 DL", source Huawei Technologies Co., Ltd., GERAN #48

[6.1-6] GP-111131, "MCS re-design for DAS-12 and DBS-12", source Huawei Technologies Co., Ltd., GERAN #51

[6.1-7] GP-101348, "DAS-12 and DBS-12 puncturing and interleaving", source Telefon AB LM Ericsson, ST-Ericsson SA, TSG GERAN #47

[6.1-8] GP-101850, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA, TSG GERAN #48

[6.1-9] GP-120162, “TSC optimization for SBPCE2”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#53

[6.1-10] GP-120160, “SBPCE2 Puncturing and Interleaving”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#53

6.2 Burst format

6.2.1 DFT length

The length of the Discrete Fourier Transform - DFT – has a direct relation to the computational complexity and also the sub-carrier spacing, which will have impact on performance.

This sub-clause evaluates different DFT sizes, both in terms of computational complexity and performance.

6.2.1.1 Choice of DFT lengths for Single Block PCE2

At the normal symbol rate, a reasonable choice of N, the DFT size, is $N=142=116+26=2\times 71$, since this length yields the same payload and training symbols as EGPRS/EGPRS2-A. Unfortunately N is not a highly composite number since 71 is a prime number. The choice $N=144=2\times 2\times 2\times 2\times 3\times 3$ is proposed since FFT's of size 144 are typically 3 to 10 times faster than FFT's of size 142, depending on implementation details and hardware capabilities. This also gives two extra symbols that can be used for training or other purposes. Although the sub-carrier spacing for $N=144$ is a little smaller than for $N=142$, the difference in performance between the two FFT lengths is negligible.

At the higher symbol rate, a reasonable choice of N is $N=169=138+31=13\times 13$, which gives the same payload and training symbols as EGPRS2-B. A more convenient value from the computational point of view is $N=168=2\times 2\times 2\times 3\times 7$. This last value is obtained by the elimination of one training symbol. Although the sub-carrier spacing for $N=168$ is a little larger than for $N=169$, and even though one training symbol is removed, the difference in performance between the two DFT lengths is negligible. A further reduction in computational complexity is achieved by choosing an FFT size of $N=162=2\times 3\times 3\times 3\times 3$.

This design is in the following sub clauses and in the remainder of the TR referred to as 'Low complexity SBPCE2B', see [6.2-10] for more details. If Low complexity SBPCE2B is not explicitly referred to SBPCE2B is assumed to be based on the design with an FFT size of 168.

As the Low complexity SBPCE2B FFT size is reduced, the modulation order needs to be increased, as shown in table 6.2-0, in order for the Low complexity SBPCE2B radio block to maintain its payload capacity.

Table 6.2-0. Modulation distribution for Low complexity SBPCE2B.

DBS-5/6	DBS-7/8/9	DBS-10/11	DBS-12b
$66_{QPSK}+2_{8PSK}$	$64_{16QAM}+4_{32QAM}$	$63_{32QAM}+5_{64QAM}$	68_{64QAM}

Table 6.2-0a lists the TSC length and payload size of the investigated SBPCE2B burst formats. To accommodate a total burst size of 162 the number of TSC symbols has been reduced to 26 symbols for DBS-5-9 resulting in a payload size of 136 symbols. The Low complexity SBPCE2B TSC length of 26 symbols for DBS-5-9 enables a re-use of the TSC set defined for EGPRS2-A. DBS-10b/11b/12b allows more flexibility in the TSC length which has been investigated in-depth in section 6.1.1.1.4.

Table 6.2-0a. Payload and TSC size for SBPCE2B and Low complexity SBPCE2B.

TSC length	#Payload/half burst length	DFT size	Radix size
30	69	168	2,3,7
26	68	162	2,3

Table 6.2-1 summarizes the lengths of the DFT required by the investigated SBPCE2 burst formats.

Table 6.2-1 DFT length for PCE2

Symbol rate	DFT Length	Radix size
Normal	144	2,3
Higher	168	2,3,7
Higher, low complexity	162	2,3

6.2.1.1.1 Performance Evaluation of DFT lengths

Figures 6.2-1 and 6.2-2 show that a DFT of size 144 for normal symbol rate and 168 for higher symbol rate can be adopted without link performance losses with respect to the sizes 142 and 169 respectively.

Simulation assumptions include TX and RX impairments from [6.2-1] TSC positioning based on [6.2-3] and using an RX bandwidth of 240 kHz and 275 kHz for PCE2-A and PCE2-B respectively.

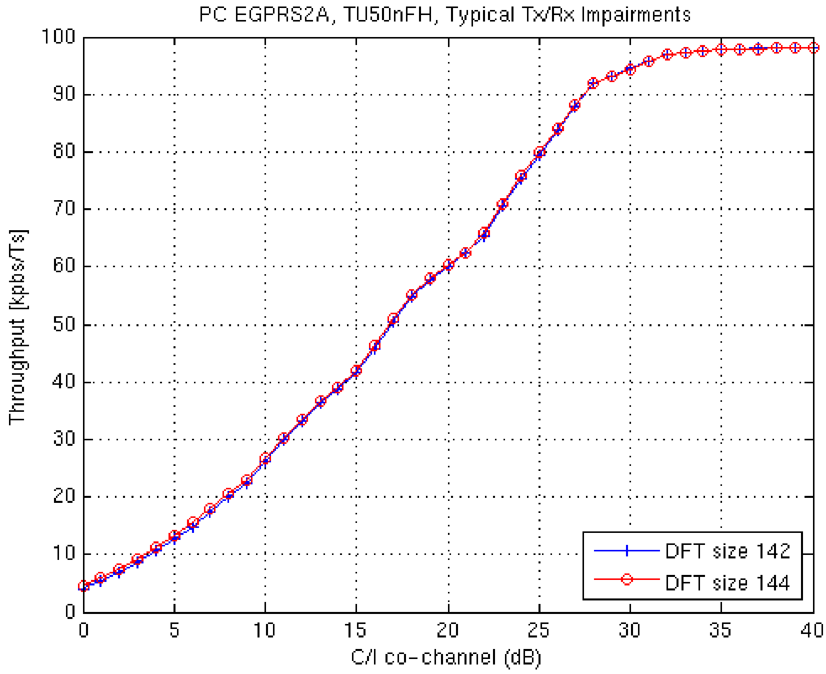


Figure 6.2-1 Performance of PC EGPRS2-A with different DFT lengths

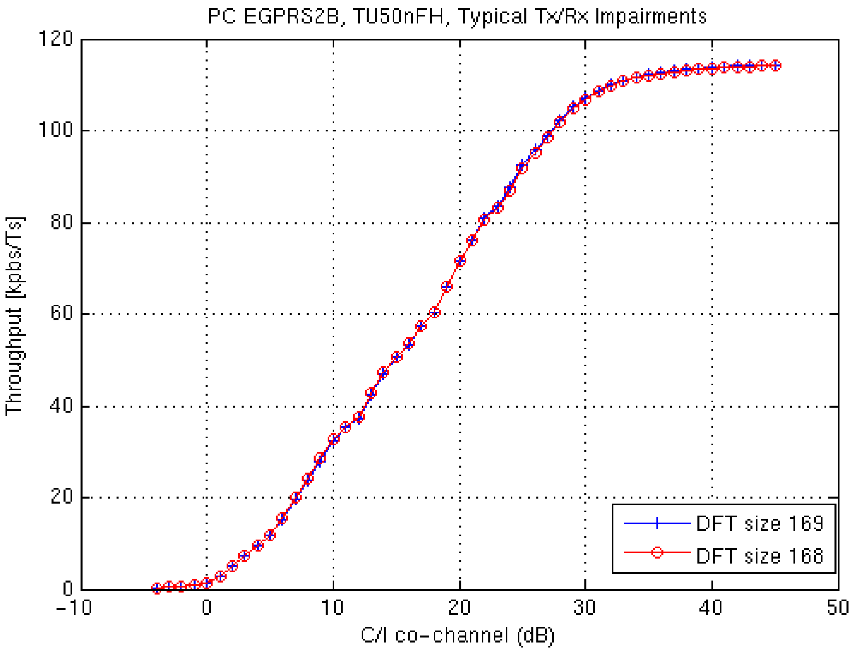


Figure 6.2-2 Performance of PC EGPRS2-B with different DFT lengths.

The performance impact on reducing the size of the FFT from 168 to 162 by utilizing a shorter TSC and increasing the modulation order is listed in table 6.2-1 for different propagation conditions and interference/sensitivity limited scenarios. It can be seen that the performance difference between SBPCE2B and Low complexity SBPCE2B is typically 0.5 dB or less for most of the investigated scenarios.

Table 6.2-1a. Performance comparison between SBPCE2B (FFT size 168) and Low complexity SBPCE2B (FFT size 162)

MCS	Performance difference between SBPCE2B (FFT size 168) and Low complexity SBPCE2B (FFT size 162) @ DataBLER 10% (Db)					
	TU50nFH, Sensitivity	TU50nFH, CO	TU50nFH, DTS-2	TU50nFH, ADJ	HT100nFH, Sensitivity	RA250nFH, Sensitivity
DBS-5	+0.5	+0.5	+0.5	+0.7	+0.6	+0.4
DBS-6	+0.4	+0.5	+0.5	+0.5	+0.8	+0.3
DBS-7	+0.3	+0.3	+0.6	+0.6	+0.5	+0.5
DBS-8	+0.3	+0.4	+0.4	+0.5	+0.5	+0.5
DBS-9	+0.3	+0.4	+0.4	+0.3	+0.5	+0.8*
DBS-10	+0.3	+0.5	+0.4	+0.3	+0.9	
DBS-11	+0.4	+0.4	+0.4	+0.5		
DBS-12b	+0.7	+0.1	+0.5	+1.2		

6.2.1.2 DFT length for Padded HOM

6.2.1.2.1 Principle of DFT length choices

For the normal case, the DFT size is determined primarily by the number of user data symbols, the number of training symbols and some special symbols (such as padding symbols at the edge of the signal BW). Considering the efficient of DFT algorithm and the impact on other parts, the final DFT length should be carefully designed.

At the normal symbol rate, a common choice of N, the DFT size, is N = 142 = 116+26=2*71. Unfortunately N is not a highly composite number since 71 is a prime number, which will result in high algorithm complexity both in BTS and MS. So is the N=169=138+31=13*13 at the higher symbol rate.

A good way to achieve highly efficient DFT is to find a number which is highly composite number with the small prime numbers, such as 2,3,5,7. At the same time, the change of the DFT size would have little impact on the functionality of other parts of burst and the performance. The DFT sizes, N = 140 = 2*2*5*7 and N = 144 =2*2*2*2*3*3 are both candidates at the normal symbol rate. Compared to DFT size N = 142 the increased DFT size N = 144 will result in even narrower sub-carrier bandwidth which may bring some frequency offset sensitivity issues. Since the total duration of the burst is kept intact and it is determined by the guard period, data symbols, CP, training symbols and other special symbols (if exists), if keeping the guard period, the CP length will be shortened because of the increased DFT size. It is not expected especially for the large delay spread channel. At the higher symbol rate, the DFT size N=168=2*2*2*3*7 is an appropriate highly composite number close to 169.

According to above considerations, a reasonable choice of DFT size N =140=2*2*5*7 is recommended for NSR and N =168=2*2*2*3*7 for HSR, shown in Table 6.2-2. This DFT size can be easily obtained by removing 2(for NSR) or 1(for HSR) padding symbols in Padded HOM (reference to the section 6.2.5.2).

In order to further reduce computational complexity of FFT/IFFT implementation and comply with radix size used in LTE up to 5, alternative DFT sizes of N=144 for NSR and N=162=2*3*3*3*3 for HSR are also proposed for Padded HOM.

In the following sub clauses and in the remainder of the TR this alternative DFT sizes design is referred to as ‘Low complexity Padded HOM’. If Low complexity Padded HOM is not explicitly stated, Padded HOM is assumed to be based on the design with DFT sizes of 140 and 168.

Table 6.2-2 DFT sizes for PCE2

Symbol rate	DFT size
NSR	140
HSR	168
NSR, low complexity	144
HSR, low complexity	162

6.2.1.2.2 Performance evaluations for the DFT length

Figure 6.2-3 shows that the DFT size of 140 (for NSR) and 168 for HSR can be adopted without link performance losses compared to the DFT sizes 142 and 169 respectively. The simulation assumptions are depicted in Table 6.2.1.2-4.

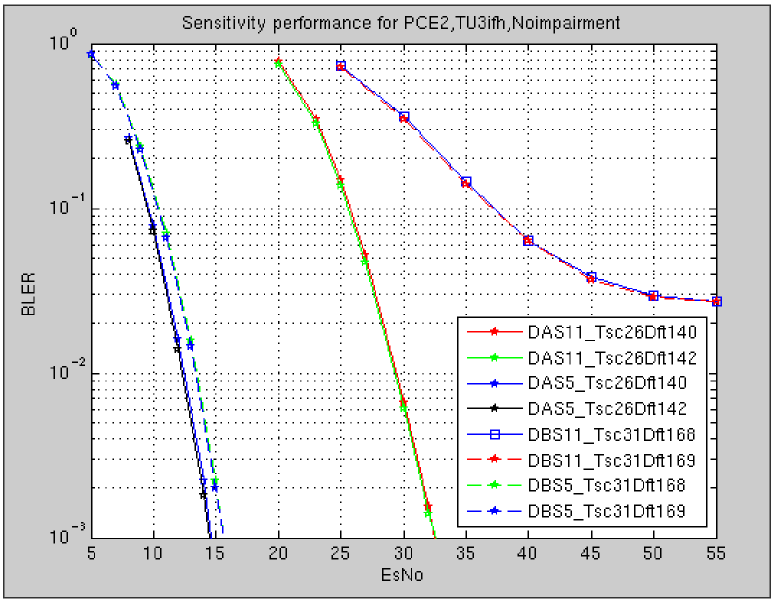


Figure 6.2-3 Sensitivity performance with different DFT sizes

6.2.2 CP length

A cyclic prefix is needed to mitigate the impact from channel time dispersion, on PC EGPRS2 performance.

This sub-clause evaluates different CP lengths.

6.2.2.1 Choice of CP length for Single Block PCE2

The CP introduces overhead, but also helps reduce the complexity of the demodulator. In order to keep both the duration of the PCE2 time slot, the length of the DFT and maintain the guard period as in EGPRS2, the following values are proposed.

Table 6.2-3 CP length for PCE.

Symbol rate	CP duration [µs]
Normal	6*48/13
Higher	9*40/13
Higher, low complexity	15*40/13

The proposed duration of the CP covers the time dispersion due to the Tx/Rx filters, as well as the time dispersion found in the most common propagation environments (except for Hilly Terrain).

6.2.2.2 Choice of CP length for Padded HOM

In order to keep the duration of the burst intact and maintain the guard period, the CP length will be increased from 6 symbols to 8 symbols when the DFT size is decreased from 142 to 140, as described in section 6.2.1.2.1. The CP length defined in duration of microseconds, given in Table 6.2-4, is very close for NSR and HSR (with only a difference of 1.8us).

For Low complexity Padded HOM, the CP length of 6 symbols and 15 symbols for NSR and HSR are proposed.

Table 6.2-4 CP length for PCE2

Symbol rate	CP length [µs]
NSR	8*3.69=29.52
HSR	9*3.08=27.72
NSR, low complexity	6*3.69=22.14
HSR, low complexity	15*3.08=46.2

6.2.3 TSC symbol position

This sub-clause evaluates different positioning of the training symbols for improved performance.

6.2.3.1 TSC symbol position for Single Block PCE2

In Precoded EGPRS2 [6.2-1], the channel estimate is performed based on the training symbols, which are spread over the transmitted burst. To achieve good quality of the channel estimate and increased throughput, the placement of the training symbols should be carefully designed.

In the following sub-clauses two design criteria are presented and evaluated to find best the placement of the training symbols,

6.2.3.1.1 Minimum Mean Square Error (MMSE) criterion

Under the MMSE criterion, a good training symbol placement tsc_ind_{opt} is the one that minimizes the mean square error of the channel estimate:

$$tsc\_ind_{opt} = \underset{tsc\_ind}{\arg\min}\{E(\sum_{i=1}^{N}\left\|\lambda_i - \hat{\lambda}_i\right\|^2)\}$$

where $\hat{\lambda}_i$ is the channel frequency response estimate for the i -th symbol.

Given Additive White Gaussian Noise (AWGN) and flat fading channel, the optimal training symbol placement based on MMSE criterion are those that are equally spaced. The training symbol indices can be calculated by:

$$tsc\_ind_{uniform}(sp) = sp + \left\lfloor \frac{L}{Ntr}[0:1:Ntr-1] \right\rfloor,$$

with L the DFT size, Ntr the training sequence length, and sp the index of the first training symbol.

6.2.3.1.2 Balance Signal-to-Noise Ratio (SNR) criterion

Due to the low-pass characteristic of a GSM/EDGE channel, the instantaneous SNR for symbols transmitted at the edge of the frequency band in a PCE2 burst will always be lower than that for those transmitted in the middle of the frequency band. By placing training symbols at different positions, the instantaneous SNR can be adjusted.

Consider the signal tap model for PCE2:

$$Y_i = \hat{\lambda}_i S_i + \underbrace{(\lambda_i - \hat{\lambda}_i)S_i + N_i}_{Noise'_i},\ i = 1,\dots,N,$$

where the noise term consists of both the noise (and interference) N_i, and the channel estimation error $\hat{\lambda}_i$. Expectation of the instantaneous SNR is calculated as:

$$E(SNR) = E\{\frac{\left\|\hat{\lambda}_i S_i\right\|^2}{\left\|(\lambda_i - \hat{\lambda}_i)S_i + N_i\right\|^2}\} = \frac{\left\|\hat{\lambda}_i\right\|^2}{\left\|\lambda_i - \hat{\lambda}_i\right\|^2 + \left\|N_i\right\|^2} = \frac{\left\|\hat{\lambda}_i\right\|^2}{MSE(\lambda_i) + \left\|N_i\right\|^2}$$

with $MSE(\lambda_i)$ contributing to the SNR calculation. By modifying the training symbol placement, and hence the MSE, the instantaneous SNR over the burst can be adjusted towards a more balanced form. Please refer to [6.2-3] for illustration of the SNR balancing effect.

To make the design feasible, the following constraints are used:

- The training symbols are placed symmetrically over the part consisting of payload and training symbols;
- In the center of the burst, the training symbols are consecutively placed to achieve a lower MSE in this part (the weak taps), thus increase the SNR for this part;
- In other parts of the burst, the training symbols are evenly placed.

A simplified implementation of training symbol placement, denoting as $tsc\_ind(sp, cn)$ is calculated as follows. The Matlab operators ":" and $fliplr$ will be used.

$$tsc\_ind_{left\_concentrated}(sp, cn) = [L/2 - cn/2 + 1 : 1 : L/2] \; ;$$

$$tsc\_ind_{left\_uniform}(sp, cn) = \left\lfloor sp + \frac{L/2 - cn/2 - sp + 2}{(Ntr - cn)/2}[0 : 1 : (Ntr - cn)/2 - 1 :] \right\rfloor ;$$

$$tsc\_ind_{left}(sp, cn) = \lfloor tsc\_ind_{left\_uniform} \quad tsc\_ind_{left\_concentrated} \rfloor ;$$

$$tsc\_ind_{right}(sp, cn) = L + 1 - flip\_left\_right(tsc\_ind_{left});$$

$$tsc\_ind(sp, cn) = \lfloor tsc\_ind_{left} \quad tsc\_ind_{right} \rfloor .$$

with sp denotes the index of the first training symbol, and cn the number of training symbols concentrated in the middle of the burst.

Figure 6.2-4 illustrates the concentrated training symbol placement according to the above procedure.

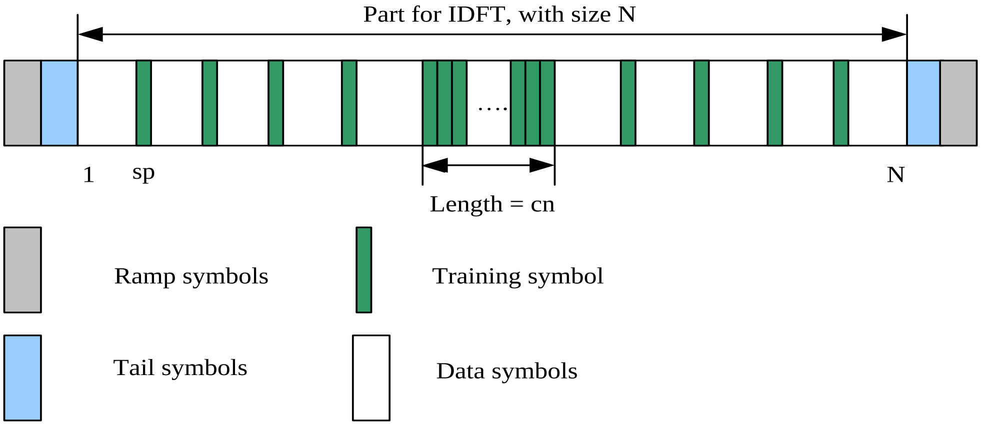


Figure 6.2-4: Burst with training symbols intercalated.

6.2.3.1.3 Evaluation of training symbol placement

In this section, performances with different training symbol placements based on design criteria in 6.2.3.1 are evaluated. The design parameters are tuned and a recommendation of the training symbol placement is made.

6.2.3.1.3.1 Simulation Assumptions

A detailed list of simulation assumptions are presented in table 6.2-5.

Table 6.2-5. Simulation assumptions.

Parameter	Value
MCSs	DAS5-DAS11, and DAS-12b, DBS7-DBS11, and DBS-12b
Burst length	According to [6.2-4]
CP length	PCE2A: 6 PCE2B: 9
RX BW	PCE2A: 240kHz PCE2B: 275kHz
Channel propagation	TU50nFH, TU3iFH, Static
Interference	AWGN, CO, DTS-2
Frequency band	900 MHz
Frames	5000
Impairments - Phase noise - I/Q gain imbalance - I/Q phase imbalance - DC offset - Frequency error	Ericsson typical TX/RX impairments: 0.8 / 1.0 [degrees (RMS)] 0.1 / 0.2 [dB] 0.2 / 1.5 [degrees] -45 / -40 [dBc] - / 25 [Hz]

The training symbol placements evaluated are the uniform placement based on MMSE criterion, $tsc\_ind_{uniform}(4)$, and the concentrated placements based on balancing SNR criterion, $tsc\_ind(sp, cn)$, with $sp = [1, \cdots, 7], cn = 2 * [1, \cdots, 10]$.

The data BLER is used as an indication for the overall impact.

6.2.3.1.3.2 Simulation Results and conclusions

The evaluation process involves:

- Tuning the concentrated training symbol placement, including the first training symbol, sp, and the number of concentrated training symbols, cn in $tsc\_ind(sp, cn)$, in both sensitivity and interference limited scenarios (single interference and multiple interferences);
- Performance comparison with uniform placement and concentrated placement, in both sensitivity and interference limited scenarios.

An example of performance comparison between uniform and concentrated TSC placement can be seen in Figure 6.2-5.

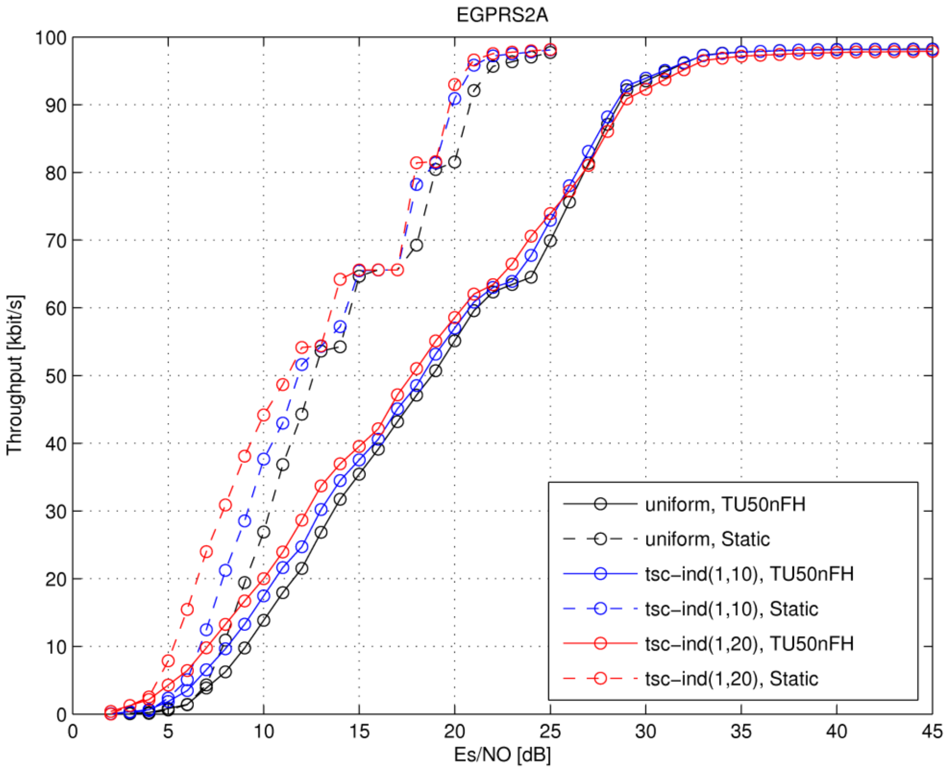


Figure 6.2-5. Performance of different training symbol placements in sensitivity.

It can be seen that the concentrated placement generally outperforms the uniform placement, which results from a balanced SNR. In TU channel, the gain with concentrated placement can be over 1dB for SNR below 24 dB, and in static channel, an overall improvement up to 2dB is achieved. It can also be noticed that, in static channel, the gain increases as more training symbols are concentrated, which is well aligned with the theoretical analysis of a more balanced SNR.

Please refer to [6.2-3] for a detailed information about the evaluation and simulation results.

The following conclusions are reached based on the evaluation:

Both the MMSE criterion and the balancing SNR criterion can be used to reach a reasonable design of the training symbol placement;

- The effectiveness of the balancing SNR criterion is limited by non-AWGN noise, unknown and fading channel, and constraints made for design simplicity;
- For sensitivity, the concentrated placement outperforms the uniform placement (especially at low SNR region). The gain is up to 1dB in TU channel and up to 2dB in a Static channel.
- In single interference case, concentrated placement is outperformed by the uniform placement. The degradation is up to 1.5dB. The gap steadily closes as cn increases;
- In multi-interference case, the concentrated placements generally outperform the uniform placement. At high C/I region, the uniform placement outperforms concentrated placements with a lower cn;
- The placement of the first training symbol, sp, does not affect the performance much in both sensitivity and interference limited scenario. The differences are within 0.2dB for the simulated scenarios.

It is therefore recommend to use a concentrated training symbol placement, with moderate amount of concentrated training symbols, to have non-negligible gain in sensitivity and multi-interference scenario, while only a small degradation in single interference case. For PCE2A, $tsc\_ind(7,10)$ is recommended; for PCE2B, $tsc\_ind(8,8)$ is recommended.

Table 6.2-5a shows the TSC placements used for SBPCE2B and Low complexity SBPCE2B in the TR.

Table 6.2-5a. TSC placements for SBPCE2B and Low complexity SBPCE2B.

Format	TSC placement
SBPCE2	[7 18 30 41 53 64 75:92 103 114 126 137 149 160]
Low complexity SBPCE2	[7 20 33 47 60 73:88 101 114 128 141 154]

6.2.3.2 TSC symbol position for Padded HOM

6.2.3.2.1 Principle of TS symbol position generation

Since the padding symbols are already used to balance the SNR in Padded HOM, a kind of uniform placement of TS symbols are investigated in this section. In order to simplify the description, the burst format shown in this section is given in the format before the Symbol swapping. As illustrated in [6.2-5], symbols before the IDFT are swapped to ensure the performance of USF and Header which are located in the middle of the burst. That means, in the burst format shown here the symbols which are located in the middle of the burst are mapped to the centre of the signal Bandwidth (BW) and the symbols at the both ends of the burst are mapped to the edge of the signal BW.

The exemplary uniformly placed TS symbol position could be achieved as follows:

- Two TS symbols are assigned to each end of the burst;
- Other TS symbols are evenly placed in the whole burst.

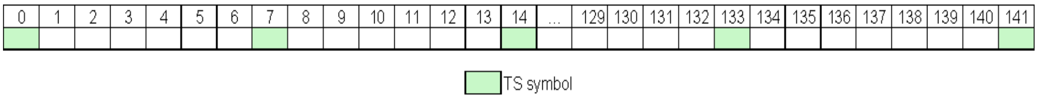


Figure 6.2-6 TS symbol position

For example, as shown in Figure 6.2-6, when the DFT size N=142, two TS symbols are allocated at the symbol index 0 and 141. Other TS symbols are evenly placed in the whole burst according to the TS length. A simple way to implement this can be elaborated below.

The interval of every two TS symbols could be roughly calculated by:

- TS_interval = round((N-2)/(NTS-1)), where N is the DFT size and NTS is the TS length.

Hence, the training symbol indices TS_ind are usually calculated by:

- TS_ind(1) = 0;
- TS_ind(NTS)=N-1;
- TS_ind(k) = TS_ind(k-1)+ TS_interval, for k=2,3,…,NTS-1.

A small adjustment is needed with some exception when TS symbol position index exceeds the bound of the DFT size.

With respect to the TS length, it should also be taken into account for PCE2. As we know, the TS symbols are mainly used to estimate the channel property. The wireless channels change both in time and frequency. The frequency coherence shows how quickly it changes in frequency. Coherence bandwidth is the parameter which expresses the frequency coherence. It gives the range of frequencies over which the channel can be considered "flat". Considering the new characteristics of PCE2, the TS length in EGPRS2 may not be suitable for PCE2. Within the coherence bandwidth, optimized TS length can give some extra symbols for other use, such as more padding symbols at the edge of the signal BW which can bring some performance improvement with little loss on channel estimation.

Table 6.2-6 shows the resulting TS position index with different TS lengths according to the principle described above.

Table 6.2-6 TS symbol position index

level A	N=142	TS13	[0:12:132,141]
		TS17	[0:9:135,141]
		TS21	[0:7:133,141]
		TS26	[0,3,6:6:138,141]
level B	N=169	TS13	[0:14:154,168]
		TS18	[0:10:160,168]
		TS25	[0:7:161,168]
		TS31	[0,3,6:6:162,165,168]
Level A	N=144	TS17	[0:9:135,143]
Level B	N=162	TS18	[0,13:9:148,161]

6.2.3.2.2 Performance evaluations

Some performance evaluations have been carried out on TS symbol position with different TS lengths. The simulation assumptions are shown in Table 6.2-7. All the evaluations in this section are performed in Zero-padded pattern. The pattern description is in section 6.2.5.2.2.

Table 6.2-7 simulation assumptions

Parameter	Value
Coding Schemes	DAS5,DAS11,DAS12; DBS5, DBS11,DBS12
Channel propagation	TU3iFH
Frequency band	900 MHz
Noise/Interference	Sensitivity/CCI, ACI
Interference modulation	GMSK
Tx filter	Lin GMSK
Rx filter	Level A: 270 kHz; Level B: 325 kHz
ICI equalizer	No
Frames	5000
Tx /Rx impairments	No
TS position index	According to Table 6.2-6
Cyclic prefix length	level A: 6 level B: 8

Table 6.2-8 Absolute performance for Data@10% BLER with different TS lengths

Coding schemes	Sensitivity (dB)		CCI (dB)		ACI (dB)	
	TS 17	TS 26	TS 17	TS 26	TS 17	TS 26
DAS-5	9.5	9.5	10	9.7	-6.3	-5.5
DAS-11	25.3	25.6	26.1	26	13.5	14.7
DAS-12	39.6	40				
Coding schemes	**Sensitivity (dB)**		**CCI (dB)**		**ACI (dB)**	
	TS 18	**TS 31**	**TS 18**	**TS 31**	**TS 18**	**TS 31**
DBS-5	10.6	10.4	11.8	11.2	-2.3	-1.1
DBS-11	37.5	37	32	32	26	29
DBS-12	42.2	/				

Table 6.2-9 Performance improvements for Data@10% BLER with different TS lengths

Coding scheme	Sensitivity (dB)	CCI (dB)	ACI (dB)
	TS 17	TS 17	TS 17
DAS-5	0	-0.3	0.8
DAS-11	0.3	-0.1	1.2
DAS-12	0.4		
Coding scheme	**Sensitivity (dB)**	**CCI (dB)**	**ACI (dB)**
	TS 18	**TS 18**	**TS 18**
DBS-5	-0.2	-0.6	1.2
DBS-11	-0.5	0	3
DBS-12			

In these evaluations, the TS symbols with shorter length are simply intercepted from the legacy TS symbols. The detailed figures can be found in [6.2-6].The absolute performance and the relative performance for PCE2-A and PCE2-B with different TS lengths are given in Table 6.2-8 and Table 6.2-9 .The gray cell means no simulation has been done in that scenario. The sign "/" means 10% BLER cannot be reached in that case. The positive value in Table 6.2-9 refers to the performance gain with the shorter TS length compared to the legacy TS length while the negative value means the loss.

It can be seen that the performance of shorter TS length has little impact on the performance in sensitivity and CCI scenario and has about 1 dB gain for DAS-5, DAS-11 and DBS-5 and up to 3 dB gains for DBS-11 in ACI scenario.

6.2.4 Mapping of block fields

The optimal position of Data, Header, USF and SF fields in EGPRS2 and PC EGPRS2 is not necessary identical.

This sub-clause evaluates different placements of the burst fields to i) Ensure similar performance as EGPRS2 USF reception and ii) Ensure incremental redundancy performance of Precoded EGPRS2.

6.2.4.1 Header bit swap and burst shift of Single Block PCE2

6.2.4.1.1 Introduction

In an EGPRS2 radio block there are different information fields that need to be mapped on the information symbols of the bursts: Data, Header, USF, SF (and optionally PAN). The mapping of these information fields are done in a specific manner to guarantee both absolute and relative performance targets.

In Precoded EGPRS2 (PCE2) the characteristics of the burst is changed and thus the mapping of these information fields need to be optimized.

This sub-clause presents results from an investigation performed in [6.2-7] on the use of header bit swap and burst shift for PCE2 Header field to guarantee robust incremental redundancy performance.

6.2.4.1.2 PCE2 Burst characteristics

In EGPRS2 the training sequence code (TSC), is placed in the middle of the burst. In general it can be assumed that the further away a data symbol is from the TSC the weaker the performance gets. In PCE2 the TSC symbols need to be spread out over the burst to optimize the channel estimation, see [6.2-3]. Further, the quality is no longer as dependent on the distance from the TSC symbol but rather characterized by the spectral properties of the TX pulse.

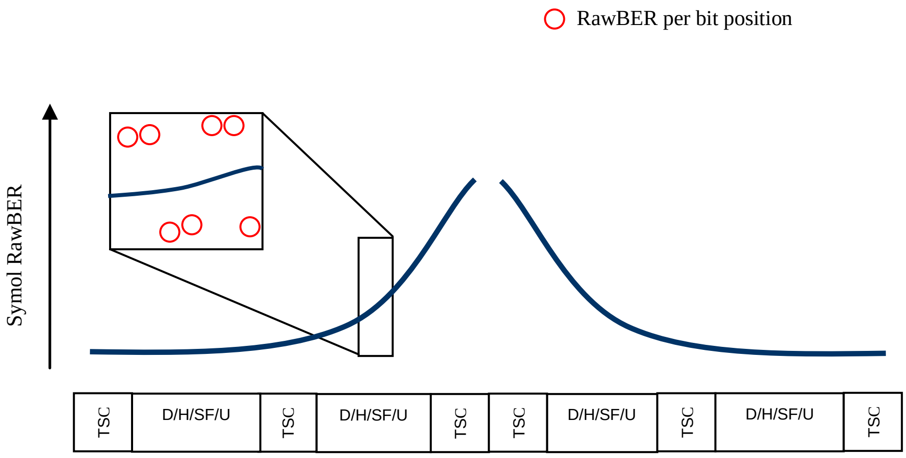


Figure 6.2-10 Burst characteristics of a Precoded 16QAM EGPRS2 burst.

6.2.4.1.3 Burst shift and Bit swapper

The MCS used, and the channel coding definitions needed to decode the data block is signaled in the header. Thus, the header needs to be correctly decoded for the data block to be correctly decoded. Further, if incremental redundancy is used the operative range of the MCS will be wider but still the condition of header decoding is valid.

To enhance header performance and secure a Header BLER performance relative to the data BLER a burst shift can in combination with a bit swap be used on weak header bits. The shift of the bits in the burst is done in a cyclic manner for each half burst as shown in Figure 6.2-11.

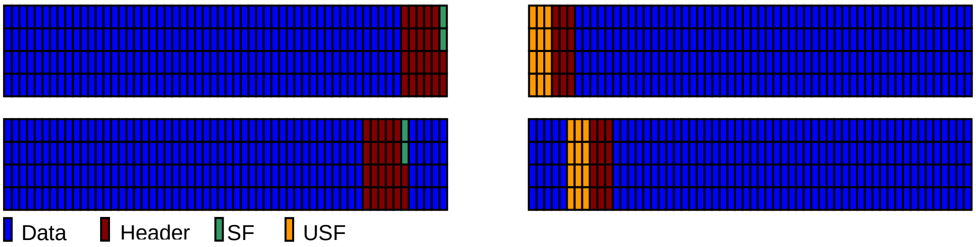


Figure 6.2-11 Burst mapping 16QAM, EGPRS2-A with (bottom) and without (top) a circular shift of 20 bits.

The bit swapper will swap weak header bits to strong bit positions occupied by data. This is illustrated in Figure 6.2-12.

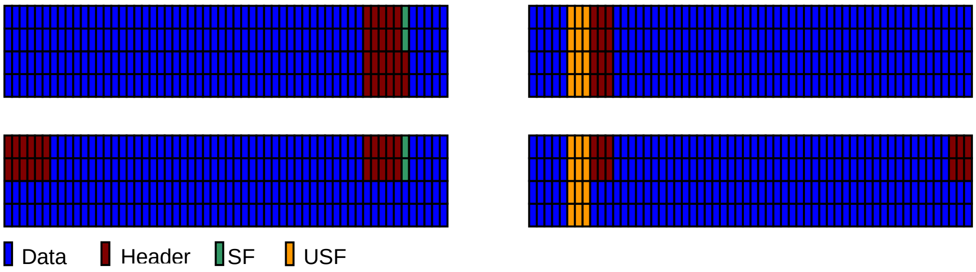


Figure 6.2-12. Burst mapping 16QAM, EGPRS2-A with 20 bits circular shift and with (bottom) and without (top) header bit swap.

In order to evaluate the relative performance needed between the Data and Header BLER the relative performance of EGPRS2 Data and Header BLER has been used as a reference. Since the channel coding definitions are kept intact for PCE2 it is reasonable to assume that similar relative performance figures should be kept.

6.2.4.1.4 Simulation assumptions and Results

Table 6.2-10 summarizes the simulation assumptions used when deriving burst shift and bit swap required to maintain the relative performance of EGPRS2 Header and Data BLER in PCE2.

Table 6.2-10. Simulation assumptions.

Parameter	Value
MCS	DAS-5-12, DBS-5-12
Intf/Sens.	N0
Impairments	[6.2-8]
Frames	5000
Max transmission for IR	5
Channel propagation	TU50nFH
Backoff	No
TSC placement	According to [6.2-3]
Burst length	According to [6.2-4]
CP length	PCE2-A: 6 PCE2-B: 9
RX BW	PCE2-A: 240kHz PCE2-B: 275kHz
ICI suppression	No
MMM	According to [6.2.2]

The reference relative performance between 1% Header BLER and 10% Data BLER for EGPRS2-A and EGPRS2-B is depicted in Table 6.2-11.

Table 6.2-11. Relative performance between 1% Header BLER and 10% Data BLER, EGPRS2-A and EGPRS2-B.

MCS	Header@1% [dB]	Data@10% [dB]	Rel. perf. [dB]
DAS-5	12.0	12.2	0.2
DAS-6		13.5	1.5
DAS-7		15.1	3.1
DAS-8	13.6	19.1	5.5
DAS-9		21.4	7.8
DAS-10	14.7	25.4	10.7
DAS-11	16.3	30.3	14.0
DAS-12	16.3	34.5*	18.2

*Performance at 20% BLER

MCS	Header@1% [dB]	Data@10% [dB]	Rel. perf. [dB]
DBS-5	11.4	11.8	0.4
DBS-6		14.1	2.7
DBS-7	17.4	21.9	4.5
DBS-8		24.4	7.0
DBS-9	18.2	26.1	5.9
DBS-10	20.2	37.5	16.7
DBS-11	21.7	35.9**	14.2
DBS-12		38.9**	17.2

** Performance at 30% BLER

In Table 6.2-12 the swap and shift required to achieve a PCE2 relative performance similar to EGPRS2, shown in Table 6.2-11, is presented. Achieved Header and Data BLER at selected swap and shift levels, is presented for completeness. Swap is not applicable on QPSK modulated MCSes DBS-5 and DBS-6.

Table 6.2-12. Header mapping of PCE2-A and PCE2-B with relative difference between Data and Header.

MCS	Swap	Shift	Header@1% [dB]	Data@10% [dB]	Diff [dB]
DAS-5	1	50	9,8	9,9	0,1
DAS-6	1	50	9,8	11,5	1,7
DAS-7	1	50	9,8	13,3	3,5
DAS-8	1	45	10,5	15,2	4,7
DAS-9	1	45	10,5	18,0	7,5
DAS-10	1	70	10,5	20,1	9,6
DAS-11	1	75	11,5	25,2	13,7

MCS	Swap	Shift	Header@1% [dB]	Data@10% [dB]	Diff [dB]
DBS-5	-	45	10.4	10.5	0.1
DBS-6	-	45	10.4	14.2	3.8
DBS-7	1	55	11.1	14.5	3.4
DBS-8	1	55	11.1	18.2	7.1
DBS-9	1	65	11.2	21.1	9.9
DBS-10	1	40	12.6	24.6	12
DBS-11	1	20	14.3	38.2	23.9

6.2.4.1.5 IR performance

In Figure 6.2-4 and -5 ideal link adaptation envelope curves are shown for both PCE2-A, EGPRS2-A and PCE2-B, given shift and swap parameters according to table 6.2-12. Visible impacts from the header is seen at Es/N0 < 10 dB for both PCE2-A and PCE2-B.

DAS-12 and DBS-12 has been used based on the design given in [6.2-9].

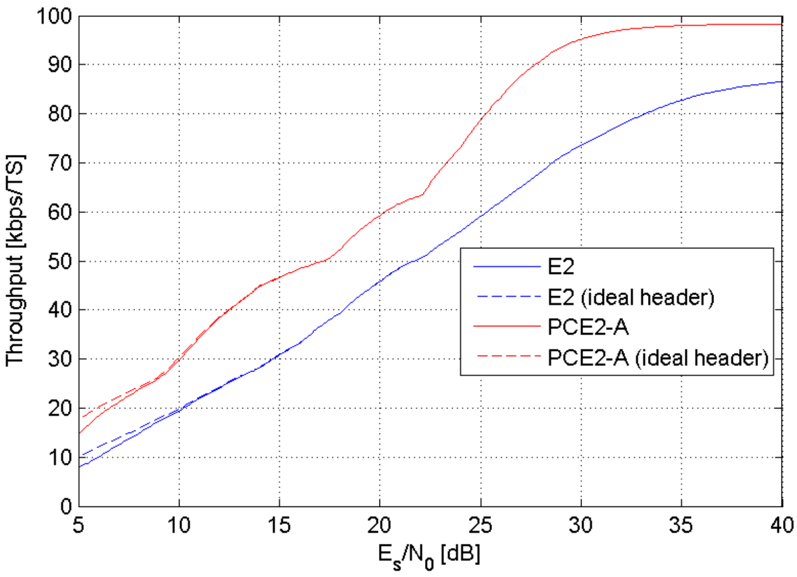


Figure 6.2-13. Ideal LA envelope for E2-A and PCE2-A (DAS-5-12).

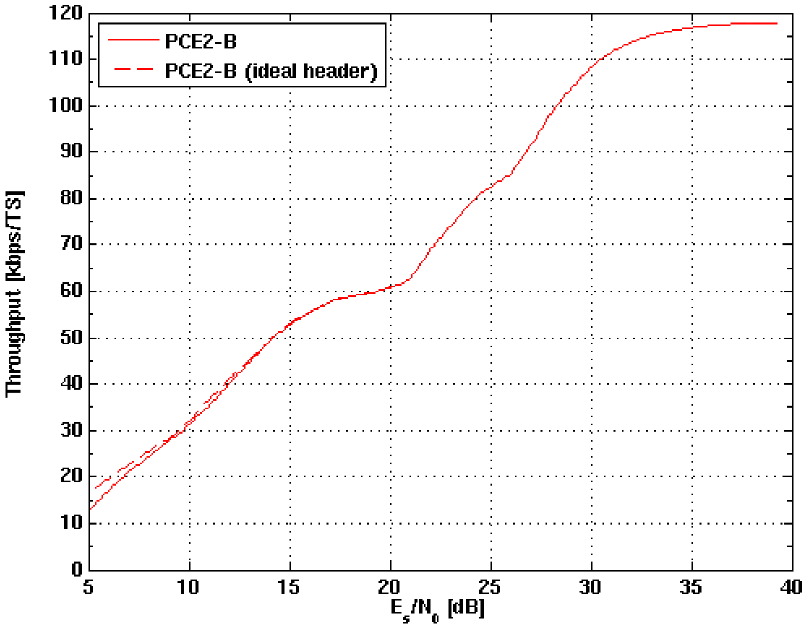


Figure 6.2-14. Ideal LA envelope for PCE2-B (DBS-5-12b).

6.2.4.1.6 Mixed Mode Modulation Results

When applying mixed mode modulation as derived in [6.2-2] the shift can be reduced while still offering similar performance as presented in Table 6.2-12. The performance when mixed mode modulation is active is presented in Table 6.2-13.

Tabel 6.2-13. Header mapping of PCE2-A and PCE2-B, MMM with relative difference between Data and Header.

MCS	Swap	Shift	Header@1% [dB]	Data@10% [dB]	Diff [dB]
DAS-5	1	30	9,8	9,4	-0,4
DAS-6	1	30	9,8	10,8	1,0
DAS-7	1	30	9,8	12,4	2,6
DAS-8	1	30	10,5	15,0	4,5
DAS-9	1	30	10,5	17,5	7,0
DAS-10	1	45	10,5	19,5	9,0
DAS-11	1	50	11,5	24,2	12,7

MCS	Swap	Shift	Header@1% [dB]	Data@10% [dB]	Diff [dB]
DBS-5	-	10	10.4	9.3	-1.1
DBS-6	-	10	10.4	12.3	1.9
DBS-7	1	35	10.9	14.4	3.5
DBS-8	1	35	10.9	17.4	6.5
DBS-9	1	45	11	19.5	8.5
DBS-10	1	35	12.6	23.2	10.6
DBS-11	1	15	14.1	38.2	24.1

6.2.4.1.7 Conclusions

A burst mapping approach for header and data fields has been presented. This approach is based on a swap and shift of header bits that guarantees robust incremental redundancy performance of PCE2-A and PCE2-B. It was further shown that the shift could be reduced if mixed mode modulation is in use.

6.2.4.2 Block mapping for Padded HOM

The transmitter of Padded HOM for PCE2 is illustrated in Figure 6.2-14.

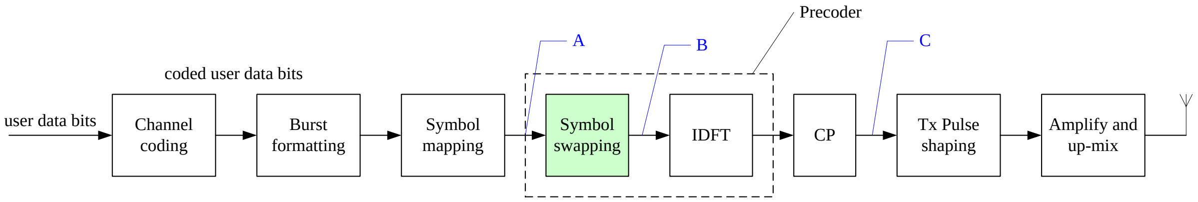


Figure 6.2-14 Transmitter for Precoded EGPRS2

Channel coding module here includes all parts of processing (including mapping on a burst subsection) defined in TS45.003. After channel coding, the order of coded user data bits in each burst is decided. The burst formatting module intercalates Training Sequence (TS) into the coded user data bits. Symbol mapping module maps coded user data bits and TS bits to PSK/QAM symbols. Symbol swapping is a new simple module added to achieve the function illustrated in Figure 6.2-15. Symbols are swapped to ensure the performance of USF and Header which are located in the middle of the burst. The relationship of symbol position in the burst and corresponding frequency characteristics before and after Symbol swapping is depicted in Figure 6.2-15. The middle figure is the position of user symbols in a burst (taking DAS-5 as an example). The top and bottom figures show the frequency characteristics before and after Symbol swapping respectively. It can be observed that the performance of Header and USF will be heavily degraded if the symbols are not swapped. That would affect the total BLER performance in PCE2.

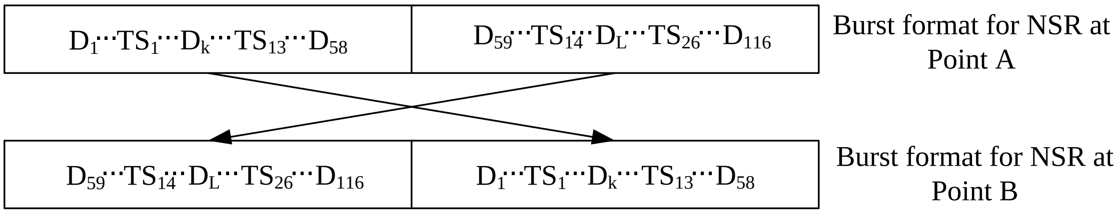


Figure 6.2-15 Symbol swapping

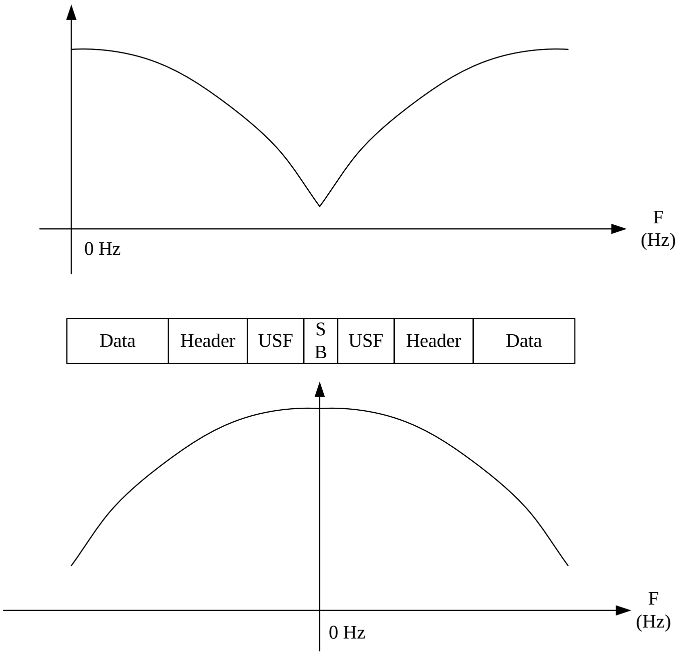


Figure 6.2-16 Frequency characteristics analyses

Thus the block mapping (including Data, Header, USF and SF fields but excluding the TS) before the Symbol swapping for each coding scheme is kept intact with the legacy EGPRS2 with only several padding symbols bits are added at both ends of the burst. The numbers of padding symbols are depicted in Table 6.2-20. For example, the burst format with different padding pattern of DAS-5 in symbols is shown in Figure 6.2-17.

It should be noted that this is the simple way to mapping each block of the coded user data to the burst.

6.2.5 Modulation scheme

The introduction of the precoder module allows for different mapping of the encoded bits onto modulation symbols.

This sub-clause contains evaluation of different mappings of bits to modulation symbols in the burst.

6.2.5.1 Mixed Mode Modulation for Single Block PCE2

6.2.5.1.1 Concept

Precoded EGPRS2 (PCE2) utilizes orthogonal frequency division multiplexing (OFDM) to transmit symbols modulated upon orthogonal sub carriers. It is known [6.2-1] that the low pass characteristics of the GSM pulse shaping filter will attenuate sub carriers located in the proximity of the edges of the signal bandwidth (BW).

To counteract this inherent property of PCE2 lower order modulated (LOM) symbols can be modulated upon the sub carriers at the BW edges. This mixture of modulations adapts the energy per bit to compensate for the decreased energy per symbol caused by the pulse shaping filter.

6.2.5.1.2 Evaluation method

To derive an optimal mixture of modulations the Raw BER performance of a given MCS was simulated at a SINR corresponding to the 10% Data BLER operative point. This was done for a large set of modulation mixtures. A detailed list of simulation assumptions are presented in table 6.2-14. For more information please see [6.2-2]

To study the general applicability of the derived MMM patterns Data and Header BLER performance of each MCS was simulated with and without the derived MMM pattern activated.

6.2.5.1.3 Simulation assumptions and results

The optimal mixture of modulations was derived in a sensitivity limited scenario given a TU50nFH propagation environment. The Data and Header BLER performance of the optimal mixture was then evaluated in

- a sensitivity limited scenario,
- a CO-channel and
- a modified DTS-2 interference scenario.

This evaluation was completed for TU50nFH, HT100nFH and RA250nFH channels. A detailed list of simulation assumptions are presented in table 6.2-14.

Table 6.2-14 Simulation assumptions.

Parameter	Value
Link direction	Downlink
Frequency band	900MHz
Channel model	TU50nFH, HT100nFH, RA250nFH
Simulated scenario	Interference limited Modified DTS-2 CO channel interference Sensitivity limited scenario
Interference modulation	GMSK
MCSs	DAS5-DAS11, DBS5-DBS11.
TSC placement	According to [6.2-3]
Burst length	According to [6.2-4]
CP length	PCE2-A: 6 PCE2-B: 9
RX BW	PCE2-A: 240kHz PCE2-B: 275kHz
ICI equalization	No
Impairments: – Phase noise – I/Q gain imbalance – I/Q phase imbalance – DC offset – Frequency error	Ericsson typical TX/RX impairments: 0.8 / 1.2 [degrees (RMS)] 0.1 / 0.2 [dB] 0.2 / 2.0 [degrees] -45 / -40 [dBc] - / 25 [Hz]

Table 6.2-15 and Table 6.2-16 summarizes the MMM patterns for PCE2-A and PCE2-B providing the best Raw BER performance given the sensitivity limited scenario and TU50noFH propagation model.

Table 6.2-15 MMM patterns derived for PCE2-A.

MCS	64QAM	32QAM	16QAM	8PSK	QPSK	BPSK
DAS-10/11	28	11	13	3	3	0
DAS-8/9	0	24	21	2	11	0
DAS-5/6/7	0	0	28	8	16	6

Table 6.2-16 MMM patterns derived for PCE2-B.

MCS	64QAM	32QAM	16QAM	8PSK	QPSK	BPSK
DBS-10/11	40	6	14	1	8	0
DBS-7/8/9	12	16	20	2	19	0
DBS-5/6	0	0	8	10	25	26

The tables below summarize the relative Data BLER and Header BLER performance improvements when MMM is activated compared to the performance achieved without MMM. If a BLER target could not be reached no value is presented in the corresponding cell.

Table 6.2-17 PCE2-A and 2-B Data and Header BLER improvement in sensitivity limited scenario.

MCS	Data / Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DAS-5	0.2 / 4.9	0.3 / 5.3	0.2 / 5.6
DAS-6	0.3 / 4.9	0.4 / 5.3	0.2 / 5.6
DAS-7	0.4 / 4.9	0.6 / 5.3	0.4 / 5.6
DAS-8	-0.6 / 6.2	-0.5 / 6.5	-1.3 / 6.8
DAS-9	-0.4 / 6.2	-0.5 / 6.5	-5.3 / 6.8
DAS-10	0.0 / 4.6	0.1 / 5.4	
DAS-11	-0.0 / 4.9	-1.2 / 6.1	

MCS	Data/Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DBS-5	0.4 / 5.4	0.6 / 6.3	0.5 / 5.4
DBS-6	0.7 / 5.4	0.7 / 6.3	0.4 / 5.4
DBS-7	0.1 / 3.1	0.1 / 3.6	-0.3 / 3.1
DBS-8	0.5 / 3.1	0.6 / 3.6	-2.0 / 3.1
DBS-9	0.8 / 3.2	0.6 / 3.7	
DBS-10	1.1 / 0.9	1.2 / 0.4	
DBS-11	-1.3 / 0.1		

Table 6.2-18 PCE2-A and 2-B Data and Header BLER improvement in CO channel interference scenario.

MCS	Data / Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DAS-5	-1.0 / 2.6	-0.6 / 3.3	-0.8 / 3.0
DAS-6	-1.0 / 2.6	-0.8 / 3.3	-1.1 / 3.0
DAS-7	-1.2 / 2.6	-0.9 / 3.3	-1.2 / 3.0
DAS-8	-1.6 / 4.0	-1.6 / 4.6	-2.4 / 5.1
DAS-9	-1.8 / 4.0	-1.8 / 4.6	-6.1 / 5.1
DAS-10	-1.2 / 3.6	-1.1 / 4.4	
DAS-11	-1.6 / 3.7	-2.5 / 4.4	

MCS	Data / Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DBS-5	-1.2 / 3.8	-1.0 / 4.7	-1.2 / 4.4
DBS-6	-1.5 / 3.8	-1.3 / 4.7	-1.9 / 4.4
DBS-7	-1.0 / 2.8	-0.8 / 3.0	-1.6 / 3.7
DBS-8	-1.4 / 2.8	-1.1 / 3.0	-4.3 / 3.7
DBS-9	-1.5 / 3.0	-1.3 / 3.7	
DBS-10	-0.7 / 0.8	-0.5 / 0.8	
DBS-11	-4.6 / 0.5		

Table 6.2-19 PCE2-A and 2-B Data and Header BLER improvement in modified DTS-2 channel interference scenario.

MCS	Data / Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DAS-5	-0.5 / 4.9	-0.2 / 5.5	-0.5 / 5.6
DAS-6	-0.6 / 4.9	-0.3 / 5.5	-0.6 / 5.6
DAS-7	-0.6 / 4.9	-0.3 / 5.5	-0.7 / 5.6
DAS-8	-1.3 / 6.1	-1.2 / 6.6	-2.1 / 7.3
DAS-9	-1.3 / 6.1	-1.4 / 6.6	-5.8 / 7.3
DAS-10	-0.7 / 4.8	-0.6 / 5.8	
DAS-11	-1.1 / 5.0	-2.3 / 6.0	

MCS	Data / Header improvements @ 10% / 1% BLER		
	MMM, Opt RawBER		
	TU50nFH	HT100nFH	RA250nFH
DBS-5	0.1 / 6.3	0.2 / 6.9	0.2 / 6.3
DBS-6	0.2 / 6.3	0.4 / 6.9	0.3 / 6.3
DBS-7	-0.1 / 2.9	0.0 / 4.0	-0.4 / 3.5
DBS-8	0.2 / 2.9	0.2 / 4.0	-1.7 / 3.5
DBS-9	0.3 / 3.3	0.5 / 3.8	
DBS-10	0.9 / 0.3	1.3 / 0.8	
DBS-11	-0.9 / 0.0		

6.2.5.1.4 Conclusion

Since the Header is modulated upon sub carriers located at the edges of the signal BW the Header BLER performance is as expected improving from MMM in all studied scenarios.

The relative performance of Data BLER is dependent upon scenario and propagation model. It is seen that PCE2-A Data BLER shows gains in the sensitivity limited scenario and degradation in the interference limited scenarios. PCE2-B Data BLER shows gains both in sensitivity limited and the modified DTS-2 scenario, while degradation is observed in the CO-channel interference scenario.

The reason for larger degradations in the CO-channel scenario is that the spectral characteristics of the interference is identical to the spectral characteristics of the carrier and a more constant C/I per sub carrier is experienced, where a mix modulations does not improve channel capacity.

It can further be seen that the relative gains/losses seen are consistent for different propagation conditions. When larger degradations are observed in especially the RA250nFH, it is due to the target BLER being close to an error floor.

6.2.5.2 Padded HOM

6.2.5.2.1 Concept

It has been discovered that the GSM radio channel typically possesses low pass characteristics. This implies that the SNR varies over the precoded symbols within a burst. Symbols transmitted at the edges of the signal bandwidth (BW) will experience a degraded SNR while symbols in the centre of the BW will benefit from an enhanced SNR. Thus if more user data symbols are centralized in the BW of the burst, the performance could get improved.

One way to achieve this goal is to use higher order modulation for each coding scheme in EGPRS2 Downlink. Currently a defined modulation is selected for each coding scheme in such a way that symbol numbers for the user data just fit for the symbol numbers in the burst. In order to put most user data symbols in the centre of signal BW, one level higher order modulation is used to each coding scheme (e.g. 16QAM instead of 8PSK for DAS-5) so that only a portion of symbols are needed to carry the same amount of user data. Some padding symbols could be arranged at the edges of the signal BW to adapt the total symbol numbers for the burst. This kind of modulation scheme can be called Padded HOM. The detailed information is shown in Table 6.2-20 when keeping TS length 26/31 and DFT size 142/169 for NSR/HSR respectively. Other choices of TS length and DFT sizes are also allowed for Padded HOM. Some example figures are listed in Table 6.2-20a and Table 6.2-20b. A diagram depicts the frequency characteristics for the PCE2 in Figure 6.2-17. The shadow part of the figure is the position that the padding symbols are located.

Table 6.2-20 Relationship of modulations and coding schemes in Padded HOM

Coding scheme	Legacy modulation	Proposed modulation	Legacy user data symbols/Burst	Proposed User.data symbols/Burst	Proposed padding symbols/Burst
DAS-5~7	8PSK	16QAM	116	87	29
DAS-8~9	16QAM	32QAM	116	93	23
DAS-10~12	32QAM	64QAM	116	97	19
DBS-5~6	QPSK	8PSK	138	92	46
DBS-7~9	16QAM	32QAM	138	111	27
DBS-10~12	32QAM	64QAM	138	115	23

Table 6.2-20a The number of padding symbols with DFT size 140/168 for Padded HOM

Coding scheme	DFT size	TS length	Proposed User.data symbols/Burst	Proposed padding symbols/Burst
DAS-5~7	140	17	87	36
DAS-8~9	140	17	93	30
DAS-10~12	140	17	97	26
DBS-5~6	168	18	92	58
DBS-7~9	168	18	111	39
DBS-10~12	168	18	115	35

Table 6.2-20b The number of padding symbols with DFT size 144/162 for Low complexity Padded HOM

Coding scheme	DFT size	TS length	Proposed User.data symbols/Burst	Proposed padding symbols/Burst
DAS-5~7	144	17	87	40
DAS-8~9	144	17	93	34
DAS-10~12	144	17	97	30
DBS-5~6	162	18	92	52
DBS-7~9	162	18	111	33
DBS-10~12	162	18	115	29

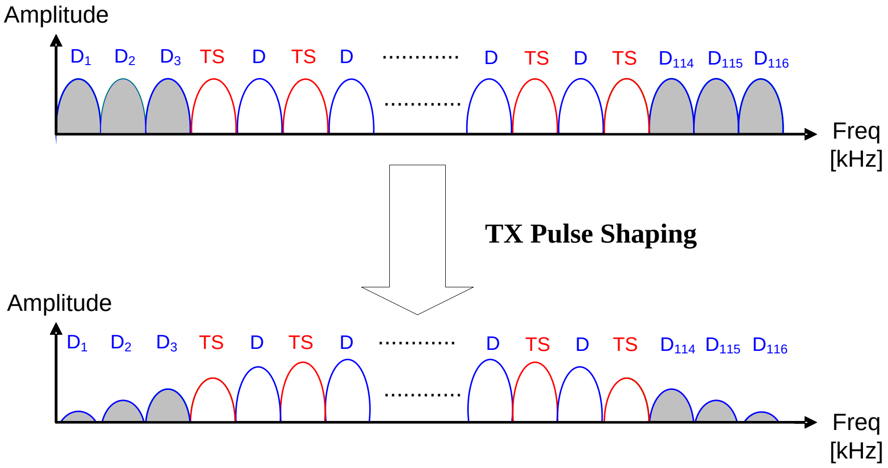


Figure 6-17 Frequency characteristics of Precoded EGPRS2

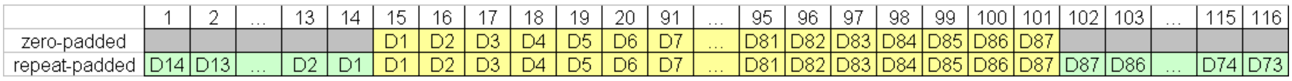
The padding symbols have different choices. They can be random symbols, zero symbols or repeated data symbols.

6.2.5.2.2 Zero-padded and Repeat-padded pattern

As described in 6.2.5.2.1, padding symbols for Padded HOM have different choices. The zero-padded and repeat-padded patterns are two kinds of padding pattern. For example, DAS-5 with these two padding pattern is illustrated in Figure 6.2-18.

The gray symbols shown in zero-padded pattern means the padding symbols at the edge of the signal BW is null. No specific symbols are transmitted there.

In repeat-padded pattern, user's data symbols at both ends are repeated with the reverse order on the sub-carriers at the edge of the signal BW. At the receiver the information of these repeated data symbols could be combined together to get some further performance improvement.



	1	2	…	13	14	15	16	17	18	19	20	91	…	95	96	97	98	99	100	101	102	103	…	115	116
zero-padded	▨	▨	▨	▨	▨	D1	D2	D3	D4	D5	D6	D7	…	D81	D82	D83	D84	D85	D86	D87	▨	▨	▨	▨	▨
repeat-padded	D14	D13	…	D2	D1	D1	D2	D3	D4	D5	D6	D7	…	D81	D82	D83	D84	D85	D86	D87	D87	D86	…	D74	D73

Figure 6.2-18 Zero-padded and Repeat-padded pattern

6.2.5.2.3 Performance evaluations

The simulation assumptions are shown in Table 6.2-7.

Table 6.2-21 Absolute performance and improvements for Data@10% BLER,PCE2-A

Coding scheme	Sensitivity (dB)		Improvements (dB)
	Rp-TS 17	TS 26	
DAS-5	9.2	9.5	0.3
DAS-11	24.7	25.6	0.9
DAS-12	37.8	40	2.2

Table 6.2-22 Absolute performance and improvements for Data@10% BLER,PCE2-B

Coding scheme	Sensitivity (dB)		Improvements (dB)
	Rp-TS 18	TS 31	
DBS-5	10.3	10.4	0.1
DBS-11	33.2	37	3.8
DBS-12	41	/	

The performance evaluation for Zero-padded pattern is in section 6.2.3.2.2.

The simulation results for Repeat-padded pattern are collected in Table 6.2-21 and Table 6.2-22. The detailed performance figures can be found in [6.2-6]. Since the shorter TS length can have more extra padding symbols for repeat. The gains for repeat-padded pattern with shorter TS length are larger than zero-padded pattern. There are about 1 dB gains for DAS-11 and up to nearly 4 dB gains for DBS-11.

It should be noted that even for DAS-12 and DBS-12 repeat-padded pattern also can have un-negligible performance improvement without any change of the channel coding.

6.2.6 Tail Symbols

6.2.6.1 Removal of Tail Symbols for Single Block PCE2

The fixed tail symbols serve several purposes in EGPRS2, such as controlling the amplitude of the signal during ramp up and ramp down, and providing known states for the termination of the trellis search. However, none of these properties are essential for the proper operation of Precoded EGPRS2 in DL. Therefore, it is proposed to remove the tail symbols in order to compensate for the overhead introduced by the CP.

6.2.6.2 Removal of Tail Symbols for Padded HOM

It is proposed to remove all the tail symbols for normal and Low complexity Padded HOM.

6.2.7 Generation of the Baseband Signal

6.2.7.1 Pulse Shaping and Ramping for Single Block PCE2

At the normal symbol rate, it is possible to increase the size of the DFT from 142 to 144 without increasing the burst duration, without decreasing the length of the CP and without degradation of the link performance. At the higher symbol rate it is not necessary to enlarge the length of the DFT, but it is necessary to cope with the lack of tail symbols. Both tasks can be accomplished by a slight modification of the usual generation of the baseband signal by means of a linear modulator [6.2-4].

6.2.7.1.1 Pulse Shaping

The length of the DFT is written in the form $(N+\lambda)$, where N is 142 at the normal symbol rate, 168 at the higher symbol rate, and λ is a small integer (equal to 2 at NSR and 0 at HSR).

Define

$$\vec{d} = [d_1, d_2, \ldots, d_K] \overset{def}{=} \left[\underbrace{g_1,\ldots,}_{guard} \underbrace{Z_1^P, Z_2^P, \ldots, Z_{N+L+\lambda}^P}_{CP+payload+pilots}, \underbrace{\ldots, g_\eta}_{guard} \right]$$

to be the $(N+\lambda)$ data and training symbols after precoding and after having appended a CP of L symbols and η guard symbols. The guard symbols are chosen so that the signal generated after pulse shaping has approximately constant amplitude over the guard. For example, if the pulse shaping filter is the linearized GMSK filter, then the guard symbols may be defined by

$$g_m = A_l e^{j\frac{\pi}{2}m},\ 1 \le m \le \xi, \qquad (l \text{ for left})$$

$$g_m = A_r e^{j\frac{\pi}{2}(m-\xi)},\ \xi+1 \le m \le \eta.\ (r \text{ for right})$$

The positive real numbers A_l, A_r are the amplitudes of the left and right guard symbols respectively. These amplitudes, together with a window described below, can be adjusted to control the power of the signal during ramp up or ramp down.

The vector $\vec{d}$ is pulse shaped using the linearized GMSK filter p.

$$y(t) = \sum_n d_n \cdot p(t - nT + \varphi),\ 0 \le t \le \tau + \lambda T,$$

where T is the symbol period (in seconds), φ is the phase, τ is the duration of the burst (in seconds) and $\lambda \le L$ is a non-negative integer. Afterwards, the first λT seconds of the signal $y(t)$ are erased. This yields a new signal

$$y_\lambda(t) = y(t + \lambda T),\ 0 \le t \le \tau.$$

Finally, a new baseband signal $s(t)$ is generated by the multiplication $y_\lambda(t)$ with two ramps. In other words, the desired baseband $s(t)$ signal is given by

$$s(t) = y_\lambda(t) \cdot r_\beta^l(t) \cdot r_\beta^r(t),\ 0 \le t \le \tau,$$

where

$$r_\beta^l(t) = \begin{cases} \frac{1}{2}\left(1 + \cos\left(\pi \frac{(\beta T - t)}{\beta T}\right)\right), & \text{if } 0 \le t \le \beta T \\ 1, & \text{if } t \ge \beta T \end{cases}, \text{ and}$$

$$r_\beta^r(t) = \begin{cases} \frac{1}{2}\left(1 + \cos\left(\pi \frac{(t - \tau + \beta T)}{\beta T}\right)\right), & \text{if } \tau - \beta T \le t \le \tau \\ 1, & \text{if } t \le \tau - \beta T \end{cases}.$$

The parameter β determines the duration of the ramping period.

This pulse shaping method is applied with the parameter values shown in Table 6.2-23 below.

Table 6.2-23 Parameter Values for Pulse Shaping

Parameter	NSR	HSR
λ	2	0
η	8	10
ξ	4	5
β	4	5
T	48/13 μs	40/13 μs
N	142	168
τ	577 μs	577 μs
L	6	9
A_l	1	TBD
A_r	0.5	TBD

6.2.7.1.2 Signal Spectrum

With the values of the parameters shown in Table 6.2-23 the EGPRS2 spectrum mask requirement for normal symbol rate is fulfilled, see Figure 6.2-20.

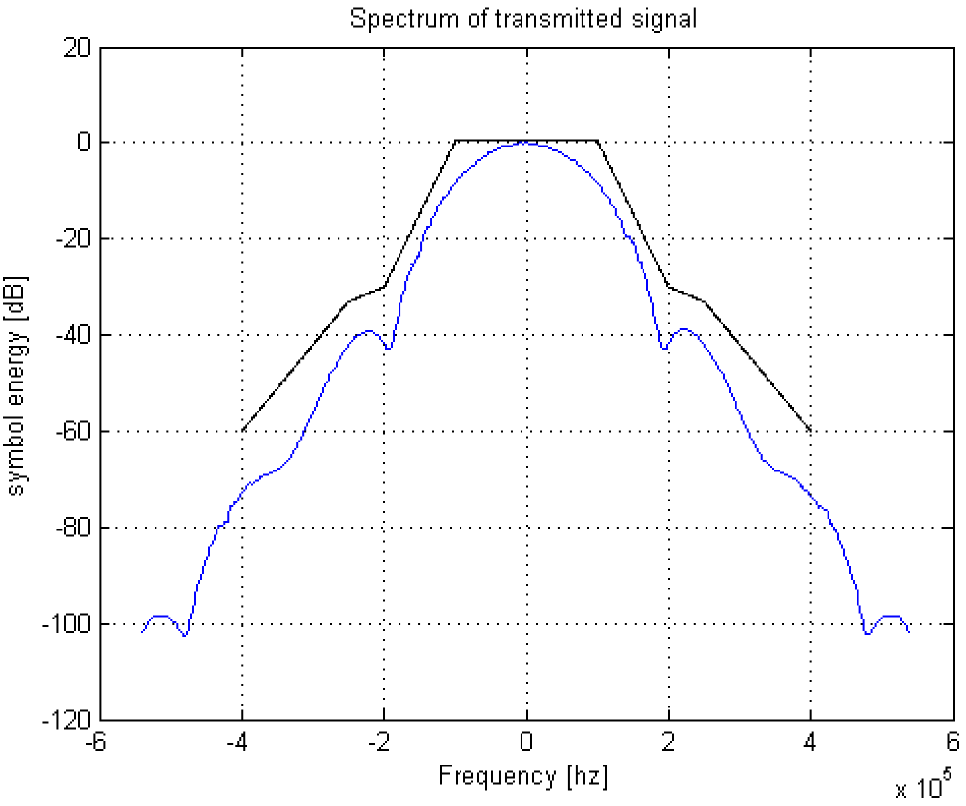


Figure 6.2-20 Simulated spectrum at the Normal Symbol Rate.

6.2.7.2 Pulse Shaping and Ramping for Padded HOM

With the DFT sizes 140/168 for NSR/HSR and 162 for low complexity HSR, CP length is increased while the DFT size is decreased compared with 142/169 to keep the burst duration. Pulse shaping and ramping are kept as the same processing as EGPRS2.

With the DFT size 144 for low complexity NSR, the method proposed in section 6.2.7.1could also be used for Padded HOM.

6.2.8 References

[6.2-1] GP-101088, "WID SPEED", source Telefon AB LM Ericsson, ST-Ericsson SA, Telecom Italia S.p.a, Vodafone Group Plc, China Mobile Com. Corp., ZTE Corporation, Huawei Technologies Co. Ltd. GERAN#46

[6.2-2] GP-101847, "Mixed Mode Modulation", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48

[6.2-3] GP-101350, "Training symbol placements in Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.2-4] GP-101349, "Aspects of burst formatting of Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.2-5] GP-101431, "Improved Precoded EGPRS2 DL (update of GP-101289)", source Huawei Technologies Co., Ltd., TSG GERAN #47.

[6.2-6] GP-101770, "Burst format of Improved Precoded EGPRS2 DL", source Huawei Technologies Co., Ltd., TSG GERAN #48

[6.2-7] GP-101350, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48

[6.2-8] GP-101352, "SPEED Working Assumptions", GERAN#47, source Telefon AB LM Ericsson, ST-Ericsson SA

[6.2-9] GP-101852, "DAS-12b and DBS-12b burst formatting", source Telefon AB LM Ericsson, ST-Ericsson SA

[6.2-10] GP-111182, "Complexity reduction of SBPCE2B", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#40.

6.3 Blind detection

To receive and demodulate a precoded burst, blind modulation detection - BMD - needs to be performed.

This sub-clause contains methods and evaluations of blind detection for MS supporting both EGPRS2 and PC EGPRS2 where both EGPRS2 and PC EGPRS2 need to be blindly detected

NOTE: Impact on BMD of MSs supporting only PC EGPRS2 is considered to be covered by this evaluation.

6.3.1 Blind Detection in for Single Block PCE2

6.3.1.1 Scope of blind detection in SBPCE2

For SBPCE2, the blind detection involves:

- Detect between an EGPRS2 and SBPCE2 user;
- For SBPCE2: Detect between level A and level B;
- Detect the modulation type.

The method is described in the following section. For more information please refer to [6.3-1].

6.3.1.2 Method of blind detection in PCE2

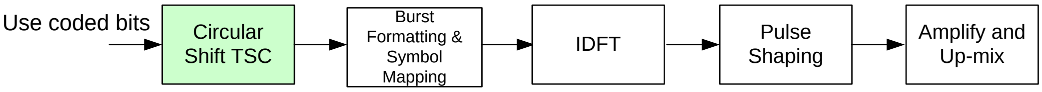
Modulations in PCE2 can be signaled by circularly shifting the TSC with an unique offset for each modulation type. If m denotes the offset for a given modulation type, the m-shifted training sequence S used for burst formatting is obtained by:

$$\mathrm{circshift}(S,m)=\mathrm{circshift}((S_0,S_1,\cdots,S_{Ntr-1}),m)=(S_{Ntr-m},\cdots,S_{Ntr-1},S_0,S_1,\cdots,S_{Ntr-m-1})$$

where circshift denotes the operation of circular shift.

Depending on the signaling method, blind detection at the demodulator can be carried out in a similar way as EGPRS2, with the difference that the training sequence is circularly shifted instead of rotated. Detecting the modulation type is therefore equivalent to finding the offset that gives the best detection metric. The procedure is illustrated in Figure 6.3-1. Blocks added to facilitate BMD are highlighted in green.

Transmitter



Receiver

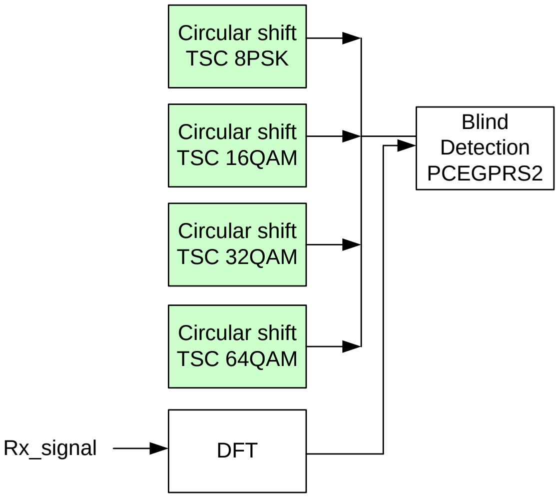


Figure 6.3-1: Blind detection of modulation type based on circular shifted training sequence.

The different modulation signaling methods used in EGPRS2 and PCE2 enable the detection between an EGPRS2 and a PCE2 user. The procedure is illustrated in Figure 6.3-2. In the RX selection block, the user type and the modulation type are detected together as the one that gives the best metric.

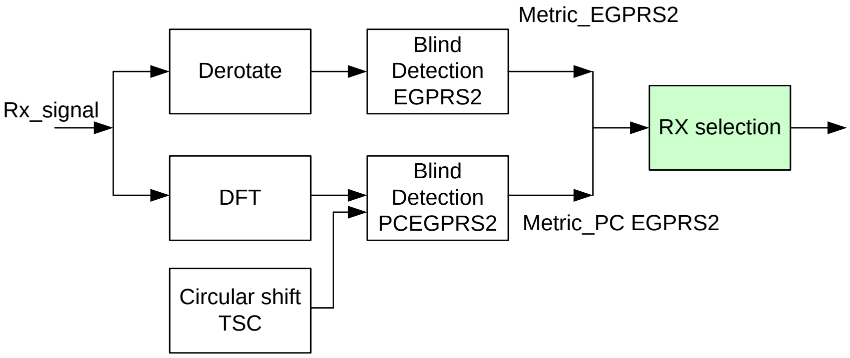


Figure 6.3-2: Blind detection of receiver type based on circular shifted training sequence.

No additional functions are need to detect between level A and level B, as the different symbol rates and different settings (training symbol placement etc. in PCE) are expected to result in a sufficiently higher (or lower) metric for the selection process to perform a successful detection. The only change involved is that the detection must be performed among a larger number of candidates.

6.3.1.3 Computational complexity of blind detection in PCE2

The following aspects contribute to the computational complexity increase of blind detection in PCE2, comparing with that in EGPRS2:

- A Fast Fourier Transform (FFT) has to be performed .
- The blind detection procedure for EGPRS2 also needs to be performed in order to detect between a PCE2 and EGPRS2 user, for equipment supporting both EGPRS2 and SBPCE2;
- Circular shift of training sequence, which adds negligible computations.

With the DFT size 144 (2*2*2*2*3*3) and 168 (2*2*2*3*7) [6.1-4] proposed for PC EGPRS2A and PC EGPRS2B, the FFT calculation can be optimized, thus limiting the increase of computational complexity due to the addition of FFTs.

To summarize, the computational complexity of blind detection in SBPCE2, given method in 6.3.1.2, is roughly same as that for EGPRS2 (twice for equipment supporting both E2 and SBPCE2) plus the computation of the FFT.

6.3.1.4 Simulation Assumptions and Results

6.3.1.4.1 Simulation Assumptions

Table 6.3-1 lists the common simulation assumptions in the evaluation.

Table 6.3-1 Simulation assumptions.

Parameter	Value
MCSs	DAS5-DAS11 and DAS-12b, DBS5-DBS11 and DBS-12b
Channel propagation	TU50nFH
Interference	AWGN, CO, ADJ_PLUS
Frequency band	900 MHz
Frames	5000
TSC placement	According to [6.3-3]
Burst length	According to [6.3-4]
Burst mapping (DAS-5-11 & DBS-5-11)	According to [6.3-5]
Burst mapping (DAS-12b & DBS-12b)	According to [6.3-6]
Soft clipping	Off
Hard clipping	Off
CP length	PCE2-A: 6 PCE2-B: 9
RX BW	PCE2-A: 280kHz PCE2-B: 340kHz
ICI Suppression	No
TX/RX impairments	Ericsson typical TX/RX impairments:
- Phase noise	0.8 / 1.2 [degrees (RMS)]
- I/Q gain imbalance	0.1 / 0.2 [dB]
- I/Q phase imbalance	0.2 / 2.0 [degrees]
- DC offset	-45 / -40 [dBc]
- Frequency error	- / 25 [Hz]

The offsets used for each modulation type are listed in Table 6.3-2. The cross-correlation property of the circular shifted training sequences is verified and no degradation has been observed in CO-channel interference scenario.

Table 6.3-2: Offsets used for circular shift training sequence.

Modulation type	QPSK	8PSK	16QAM	32QAM	64QAM
Offset	3	0	2	5	7

When using DAS-10b/11/12b v2 and DBS-10b/11b/12b v2, introduced in Section 6.1.1, the following modulation types and offsets are proposed.

Table 6.3-3: Offsets used for circular shift training sequence for DAS-10b/11b/12b and DBS-10b/11b/12b.

Modulation type	QPSK	8PSK	16QAM	64QAM DBS-11b/12b	64QAM DAS-10b/11b/12b or DBS-10b
Offset	3	0	2	5	7

In Section 6.3.1.4.2, only results with modulation types according to table 6.3-2 are shown. However, given that the same numbers of offsets are proposed for SBPCE2-B while one offset less is needed for SBPCE2-A, the presented results are valid for both cases.

6.3.1.4.2 Simulation Results

Figure 6.3-3 shows the SBPCE2-A and SBPPCE2-B performance of blind detection based on the circular shifted training sequence, in interference limited and noise limited scenario respectively. It can be seen that the modulation type (QPSK/8PSK/16QAM/32QAM/64QAM), user type (EGPRS2/SBPCE2) and levels (LevelA/LevelB for SBPCE2B) can be correctly detected. The performance difference between ideal blind detection and blind detection based on circular shift training sequence is small for the simulated scenarios, with the largest degradation around 0.3dB for QPSK in sensitivity.

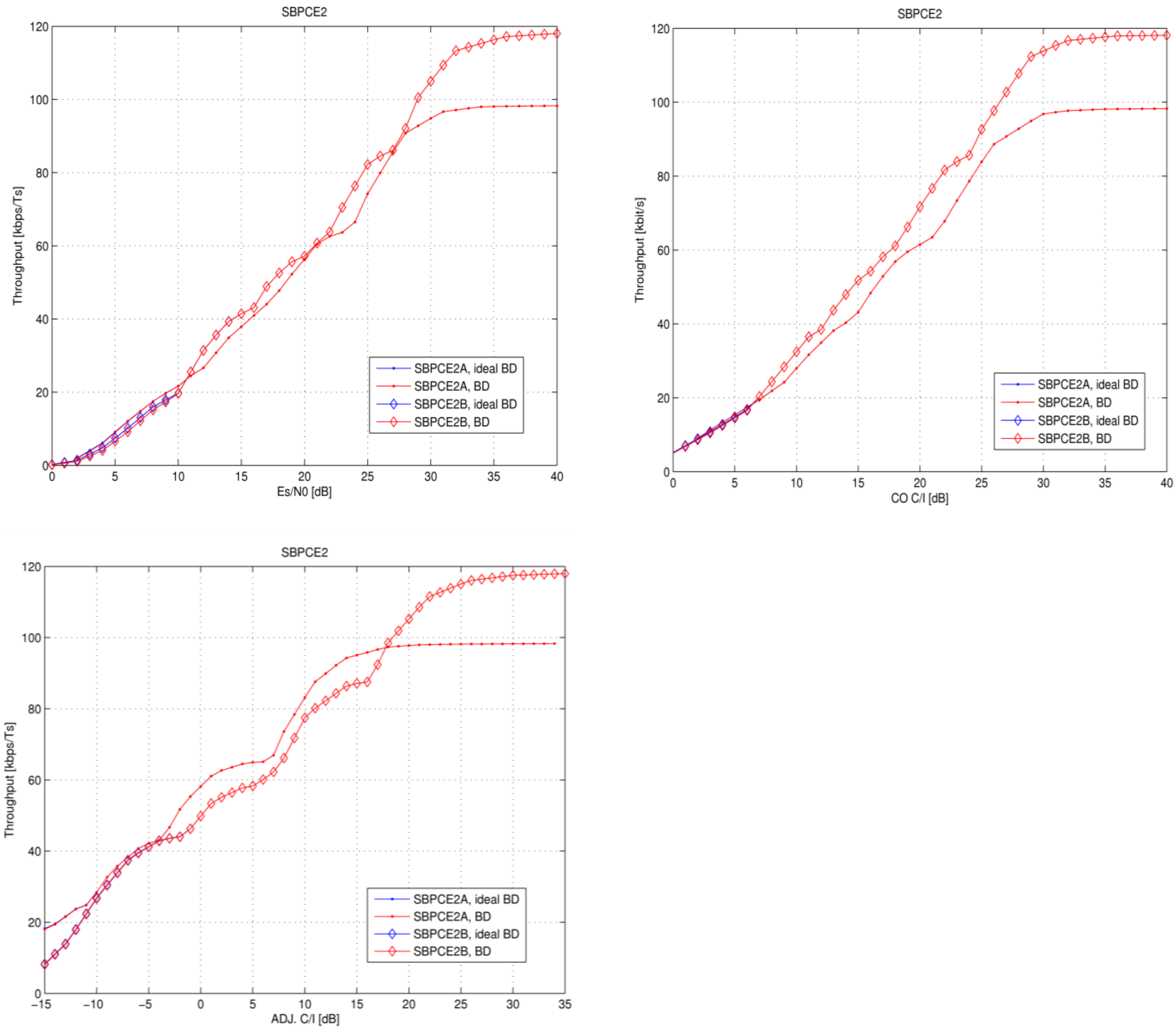


Figure 6.3-3: Performance of BD based on circular shift training sequence, SBPCE2.

6.3.2 References

[6.3-1] GP-101854, "Blind Modulation Detection in PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48.

[6.3-2] GP-101847, "Mixed Mode Modulation", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48.

[6.3-3] GP-101350, "Training symbol placements in Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson. GERAN#47,

[6.3-4] GP-101349, "Aspects of burst formatting of Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47,

[6.3-5] GP-101850, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48

[6.3-6] GP-101852, "DAS-12b and DBS-12b burst formatting", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#48.

6.4 PAR reduction

By applying precoding to EGPRS2 the Peak-to-Average Ratio – PAR – of the signal is increased.

This sub-clause contains methods and evaluations to reduce the PAR of PC EGPRS2 while maintaining the spectral properties of the signal and minimizing impact on link level performance.

6.4.1 PAR reduction for Single Block PCE2

Single Block Precoded EGPRS2 (SBPCE2) possesses an inherently large Peak to Average power Ratio (PAR). This section presents an evaluation of PAR reduction methods for SBPCE2, taking into consideration both the computational complexity of the evaluated methods and their impact on performance.

Three alternative PAR reduction methods for SBPCE2 are briefly introduced in the sub-clause 6.4.1.1. The chosen evaluation method and the results from the PAR reduction exercise, along with complexity estimates of the methods are described in sub-clause 6.4.1.2.

A detailed description on PAR reduction for SBPCE2 can be found in [6.4-1].

6.4.1.1 PAR Reduction Methods

6.4.1.1.1 Soft clipping

Soft clipping targets signal peaks exceeding a configured threshold. Each targeted peak is compressed by a compensation signal as described in [6.4-2]. To maintain the spectrum of the soft clipped signal the compensation signal is filtered through the Linearized GMSK pulse shaping filter.

6.4.1.1.2 Hard clipping

Hard clipping limits the amplitude of all signal peaks exceeding a configured threshold to the level of the threshold. This operation will widen the spectrum of the signal, due to the sharp transitions around the clipped peaks. The spectrum widening limits the PAR level that can be achieved while meeting the spectrum requirements.

6.4.1.1.3 Symbol rotation

In SBPCE2, a signal peak occurs when the sub-carriers transmitted at different frequencies add constructively together. By rotating part of the signal with angles selected from a pre-defined set, the signal PAR characteristics can be altered. At the receiver side blind detection is performed over the rotated training sequence, to detect the used rotation angle. Both the rotation angle selection and blind detection are performed on a burst-by-burst basis.

In the results below the choice of rotation angle is used after upsampling and pulse shaping. Alternative ways to choose rotation angle has been investigated in [6.4-9] with minimal impact to performance (at most 0.2 dB) but a significant reduction in complexity. The rotation angle is then chosen based from a signal without oversampling and a 2 tap pulse shaping filter [1 1].

6.4.1.2 PAR Reduction Evaluation

The PAR reduction methods are evaluated for all SBPCE2-A and SBPCE2-B MCSs given a target PAR of 6dB. For DAS-5 – DAS-7 and DBS-5 – DBS-6 the PAR reduction are also evaluated for a 4dB target.

Sub-clause 6.4.1.2.1 presents the simulation assumptions used through the evaluations. To achieve maximal performance the PAR reduction methods presented in sub-clause 6.4.1.1 were combined, as described in sub-clause 6.4.1.2.2 and 6.4.1.2.3. The sub-clauses also present the impact on performance and spectral characteristics of the signal along with an assessment of the computational complexity of the PAR reduction methods.

6.4.1.2.1 Simulation Assumptions

The PAR reduction evaluations were performed with a PA model, verified towards measurements on Spectrum due to modulation and wideband noise, to secure that the signal spectrum characteristics were assessed correctly in the evaluation. Figure 6.4-1 depicts the simulated spectrum of the basic SBPCE2 signal given different impairment models.

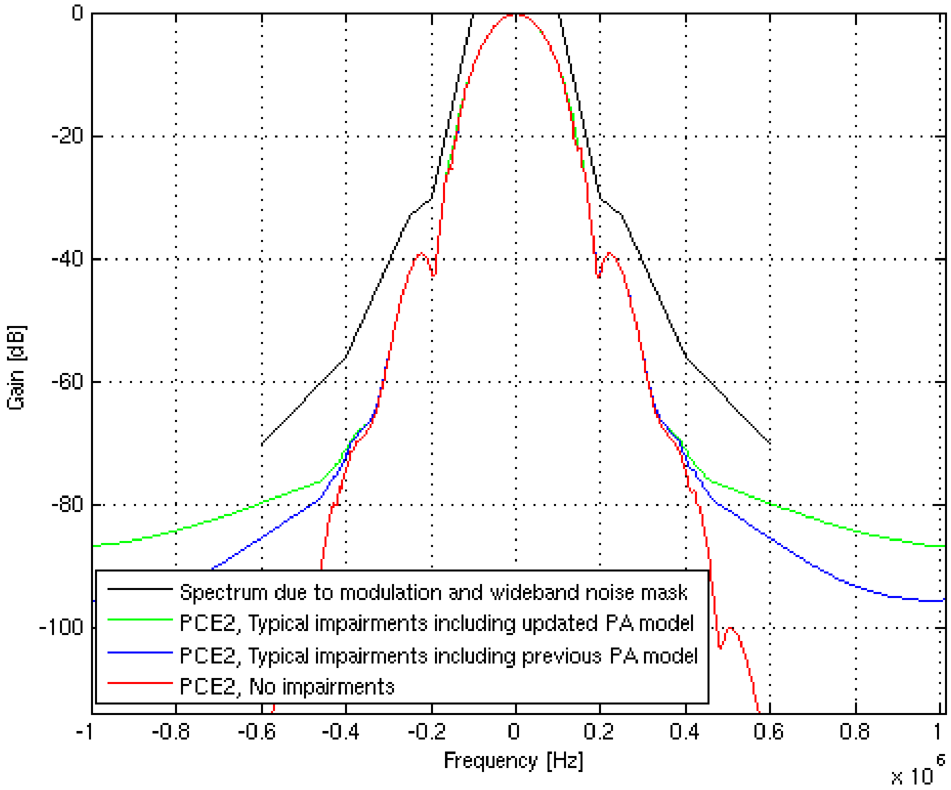


Figure 6.4-1 Spectrum characteristics of SBPCE2.

The simulation settings used through the evaluation are listed in Table 6.4-1.

Table 6.4-1 Simulation settings.

Parameters	Value
MCSs	DAS5-11, DAS-12b, DBS5-11, DBS-12b
Burst mapping	According to [6.4-3] and [6.4-4]
TSC placement	According to [6.4-5]
Burst length	According to [6.4-6]
Mixed Mode Modulation	Not used (except for DAS-12b)
Blind Detection	On (when symbol rotation is used)
CP length	PCE2A: 6 PCE2B: 9
RX BW	PCE2A: 280kHz PCE2B: 340kHz
Channel propagation	TU50nFH
Interference	AWGN
Tx filter	Lin GMSK
Frequency band	900 MHz
Frames	10000
Tx/Rx impairments - Phase noise [degrees (RMS)] - I/Q gain imbalance [dB] - I/Q phase imbalance [deegrees] - DC offset [dB] - PA model - Frequency error [Hz]	Tx/Rx 0.8/1.2 0.1/0.2 0.2/2.0 -45/-40 Yes / - -/25

6.4.1.2.2 Soft Clipping combined with Symbol rotation

Soft clipping and Symbol rotation can be combined to compress the signal to a desired PAR target. Table 6.4-1 lists the number of peaks clipped in order to reach a target PAR, given this approach. The impact on both computational complexity and performance from usage of symbol rotation are presented.

It can be concluded that soft clipping efficiently reduces the PAR. For the 6 dB target, 98% of the bursts reach their target after less than 11 clipped peaks. The impact on performance is small for lower order MCSs, while 32QAM and higher order modulated MCSs are less robust to clipping.

It is beneficial to use symbol rotation for the higher order modulated MCSs in level B, while the performance degrades at lower MCSs due to erroneous blind detection of the rotation angle.

Table 6.4-1 Summary of PAR reduction results.

Target PAR	Level	MCS	#Peaks clipped@98% w/wo rotation	Achieved PAR @99.9%	Degradation @10%DataBLER w/wo rotation
4 dB	Level A	DAS-7	27/27	4.0 dB	0.7/0.6 dB
6 dB		DAS-7	9/9	6.0 dB	0.3/0.1 dB
		DAS-9	10/10	6.0 dB	0.2/0.2 dB
		DAS-11	9/10	6.0 dB	0.7/0.7 dB
		DAS-12b	9/10	6.0 dB	1.5/1.9 dB
4 dB	Level B	DBS-6	27/27	4.0 dB	1.2/0.0 dB
6 dB		DBS-6	9/9	6.0 dB	0.9/0.0 dB
		DBS-9	11/11	6.0 dB	0.2/0.3 dB
		DBS-11	9/10	6.0 dB	0.6/0.9* dB
		DBS-12b	10/10	6.0 dB	2.4/3.6 dB

* Evaluated at 30% DataBLER.

6.4.1.2.3 Soft Clipping combined with Hard clipping and Symbol rotation

To reduce the computational complexity Soft clipping and Symbol rotation can be combined with Hard clipping. For a given PAR target, the soft clipping is performed first and targets only the highest signal peaks. The hard clipping is applied on the remaining peaks. The maximum number of hard clippings is limited to secure that the spectrum are maintained within the 3GPP requirements.

It can be seen in table 6.4-2 that by combining soft clipping with hard clipping, the clipping efficiency is significantly enhanced, while performance is maintained. Again it is beneficial to use symbol rotation for the higher order modulated MCSs in level B, while the performance degrades at lower MCSs due to erroneous blind detection of the rotation angle.

Table 6.4-2 Summary of PAR reduction results.

Target PAR	Level	MCS	#Peaks clipped@98%w/wo rotation	Achieved PAR @99.9%	Degradation @10%DataBLER w/wo rotation
4	Level A	DAS-7	20 / 20	4.0	0.8 / 0.7
6		DAS-7	4 / 4	6.0	0.3 / 0.1
		DAS-9	4 / 4	6.0	0.1 / 0.3
		DAS-11	4 / 4	6.0	0.7 / 0.7
		DAS-12b	4 / 4	6.0	1.4 / 1.6
4	Level B	DBS-6	20 / 20	4.0	1.2 / 0.1
6		DBS-6	4 / 4	6.0	1.0 / 0.1
		DBS-9	4 / 5	6.0	0.2 / 0.3
		DBS-11	4 / 4	6.0	0.6 / 0.9*
		DBS-12b	4 / 4	6.0	2.1 / 3.1

* Evaluated at 30% DataBLER.

For Low complexity SBPCE2 the performance degradation is at most an additional 0.5 dB compared to table 6.4-2. The additional degradation is mainly due to the higher order modulation symbols that are needed to compensate for the smaller FFT size used (compared to SBPCE2B). For more details, see [6.4-8]

As the signal is hard clipped the spectrum of the clipped signal will widen. This is exemplified in the figures below for DAS-7 and DBS-6 given PAR targets of 4dB and 6dB. In the figures it is seen that the spectrum widening is dependent on the number of soft clipped peaks. As the number of soft clipped peaks decrease, the impact from hard clipping on the spectrum characteristics increase and the margin to the Spectrum due to modulation and wideband noise decreases.

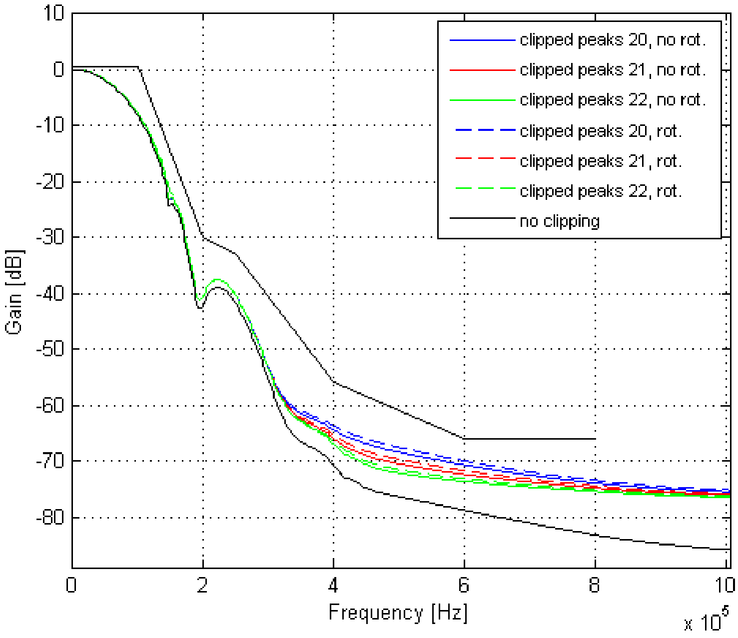


Figure 6.4-2 Spectrum widening of DAS-7 at a PAR target of 4dB due to hard clipping.

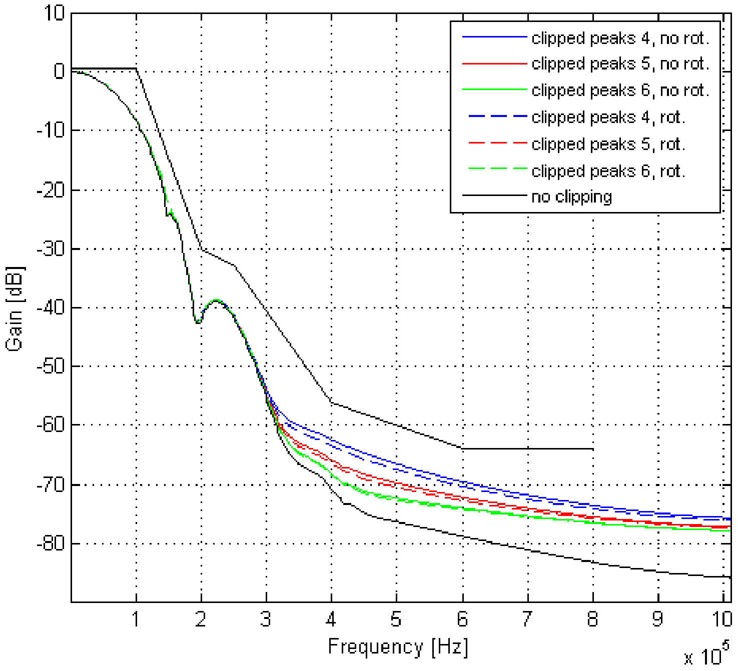


Figure 6.4-3 Spectrum widening of DAS-7 at a PAR target of 6dB due to hard clipping.

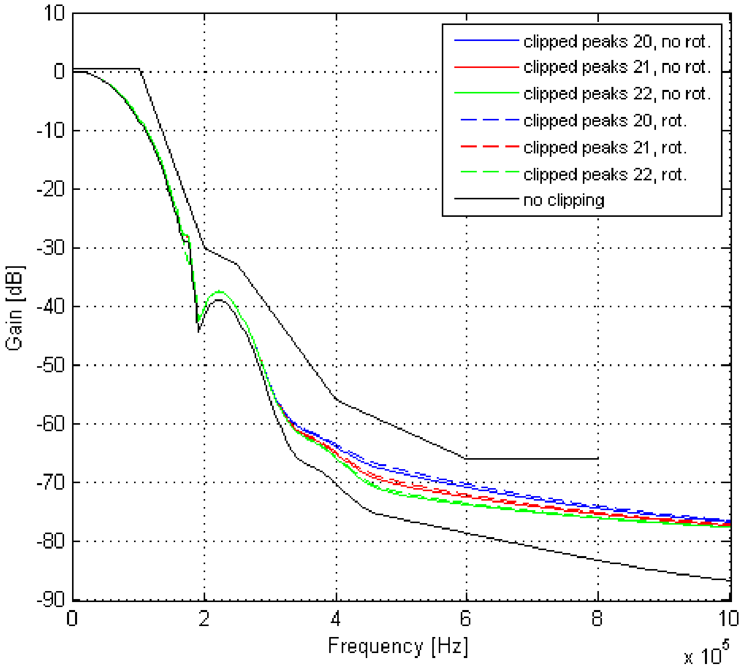


Figure 6.4-4 Spectrum widening of DBS-6 at a PAR target of 4dB due to hard clipping.

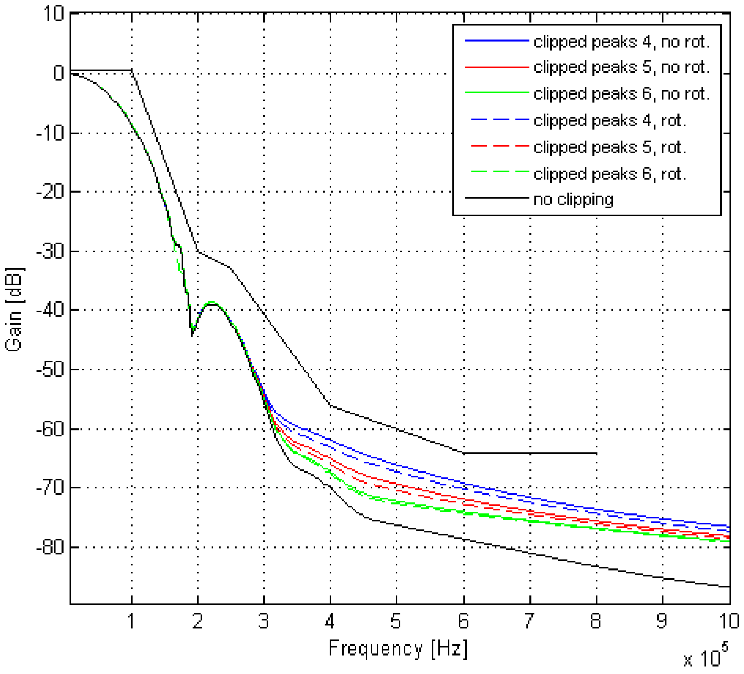


Figure 6.4-5 Spectrum widening of DBS-6 at a PAR target of 6dB due to hard clipping.

6.4.2 PAR Reduction for Padded HOM

Higher peak-to-average power ratio is introduced by the IDFT precoder in PCE2.The clipping based PAR reduction method is performed for Padded HOM.This section presents the techniques and investigates of the impact on the performance and spectrum.

PAR reduction methods for Padded HOM are described in sub-clause 6.4.2.1.

Both the performance degradation and impact on the spectrum are evaluated and described in sub-clause 6.4.2.2.

6.4.2.1 PAR Reduction Method

The same soft and hard clipping algorithms described respectively in sub-clause 6.4.1.1.1 and 6.4.1.1.2 are used for Padded HOM. The evaluations of the clipping based PAR reduction method are given in the following sub-clauses.

6.4.2.2 PAR Reduction Evaluation

In this sub-clause, the impact of PAR reduction with the combined soft and hard clipping on the performance and spectrum are investigated. For a given PAR target, the soft clipping is performed first and targets only the highest peaks. The hard clipping is then applied on the remaining peaks. The number of soft clipping peaks aligns with [6.4-7].

6.4.2.2.1 Simulation Assumption

The simulation assumptions are listed in table 6.4-3.

Table 6.4-3 Simulation settings.

Parameter	Value
Coding Schemes	DAS-7,DAS-9,DAS-11b v2 DBS-6,DBS-9,DBS-11b v2
Channel propagation	TU3iFH
Frequency band	900 MHz
Interference scenario	AWGN
Tx filter	Lin GMSK
Frames	5000
Tx/Rx impairments	Tx/Rx
- Phase noise [degrees (RMS)]	0.8/1.2
- I/Q gain imbalance [dB]	0.1/0.2
- I/Q phase imbalance [deegrees]	0.2/2.0
- DC offset [dB]	-45/-40
- PA model	Yes / -
- Frequency error [Hz]	-/25
Cyclic prefix length	level A: 6 level B: 15
DFT size	level A:144 level B: 162
TS position index	According to [6.4-10]

6.4.2.2.2 Simulation Result

Table 6.4-4 Performance degradation of PAPR reduction.

Target PAR	Level	MCS	Soft clipping	Achieved PAR@99.9%	Degradation @10%DataBLER (dB)
4 dB	Level A	DAS-7	20	4.0 dB	1.1
6 dB		DAS-7	4	6.0 dB	0.2
		DAS-9	4	6.0 dB	0.4
		DAS-11b v2	4	6.0 dB	0.6
4 dB	Level B	DBS-6	20	4.0 dB	1.5
6 dB		DBS-6	4	6.0 dB	0.1
		DBS-9	4	6.0 dB	0.7
		DBS-11b v2	4	6.0 dB	2.1

Table 6.4-4 shows the impact of PAR reduction on the link level performance. For DAS-7, with a target PAR of 4 dB and 6 dB the spectrum requirements could be met with a performance degradation of 1.5 dB (see Figure 6.4-6) and 0.2 dB (see Figure 6.4-7), respectively. For DBS-6, with a target PAR of 4 dB and 6 dB the spectrum requirements could be met with a performance degradation of 1.5 dB (see Figure 6.4-10) and 0.1 dB (see Figure 6.4-11), respectively. In other cases, with a target PAR of 6 dB, the spectrum requirements could be met with a small performance degradation of less than 0.7 dB (except for DBS-11b v2).

It can be seen in the figures that the target PAR could be efficiently reached with several times of soft and hard clipping. The spectrum of the clipped signal could be kept within the spectrum mask with acceptable performance degradation.

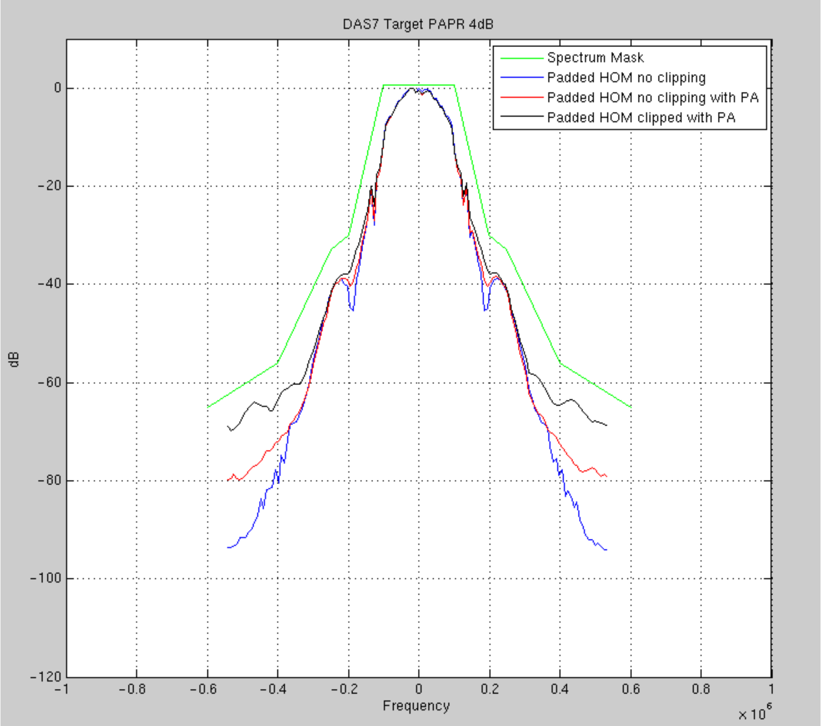


Figure 6.4-6 Spectrum of DAS-7(Target 4dB) due to PAR reduction with soft and hard clipping.

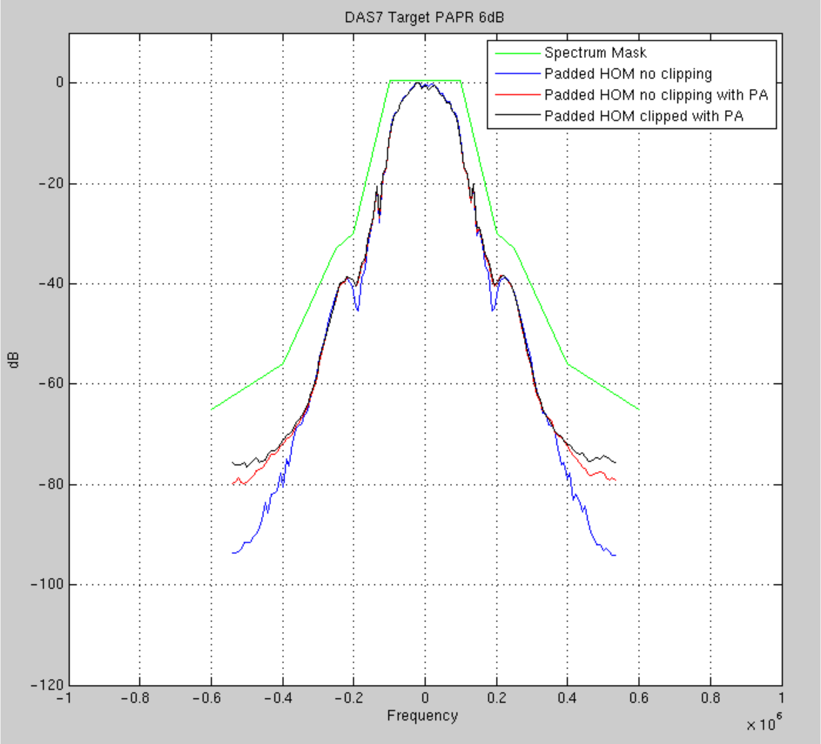


Figure 6.4-7 Spectrum of DAS-7 (Target 6dB) due to PAR reduction with soft and hard clipping.

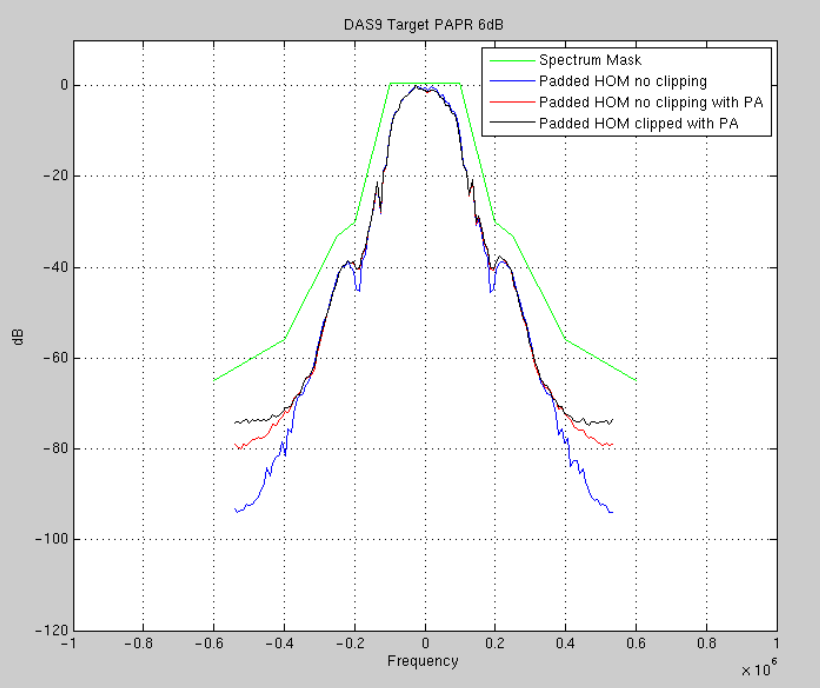


Figure 6.4-8 Spectrum of DAS-9(Target 6dB) due to PAR reduction with soft and hard clipping.

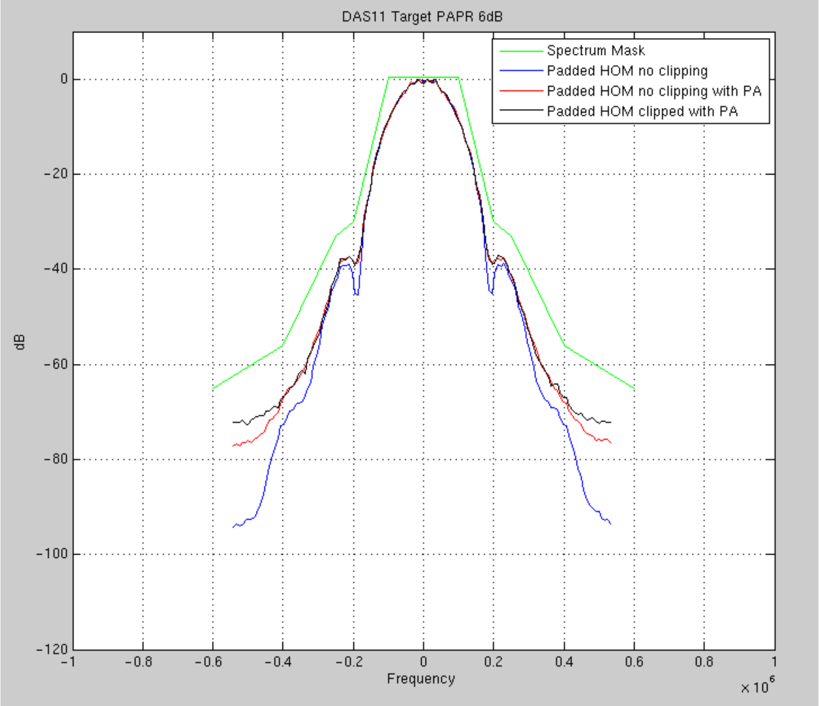


Figure 6.4-9 Spectrum of DAS-11b(Target 6dB) due to PAR reduction with soft and hard clipping.

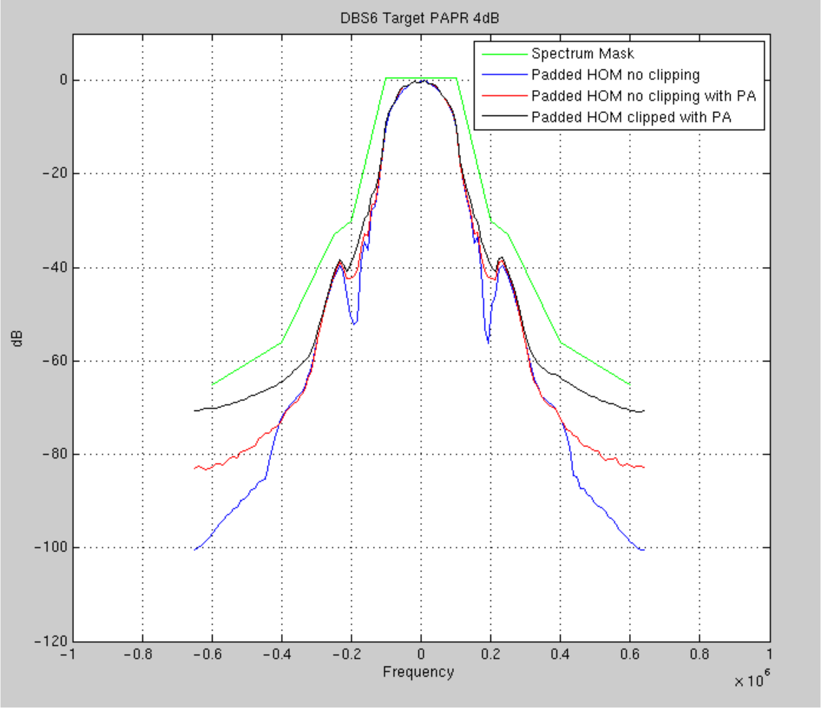


Figure 6.4-10 Spectrum of DBS-6(Target 4dB) due to PAR reduction with soft and hard clipping.

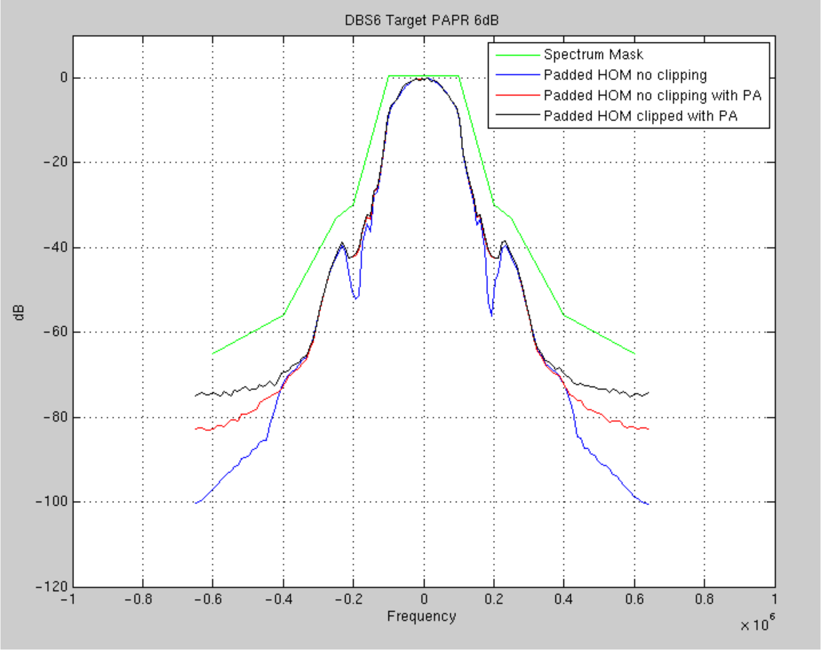


Figure 6.4-11 Spectrum of DBS-6(Target 6dB) due to PAR reduction with soft and hard clipping.

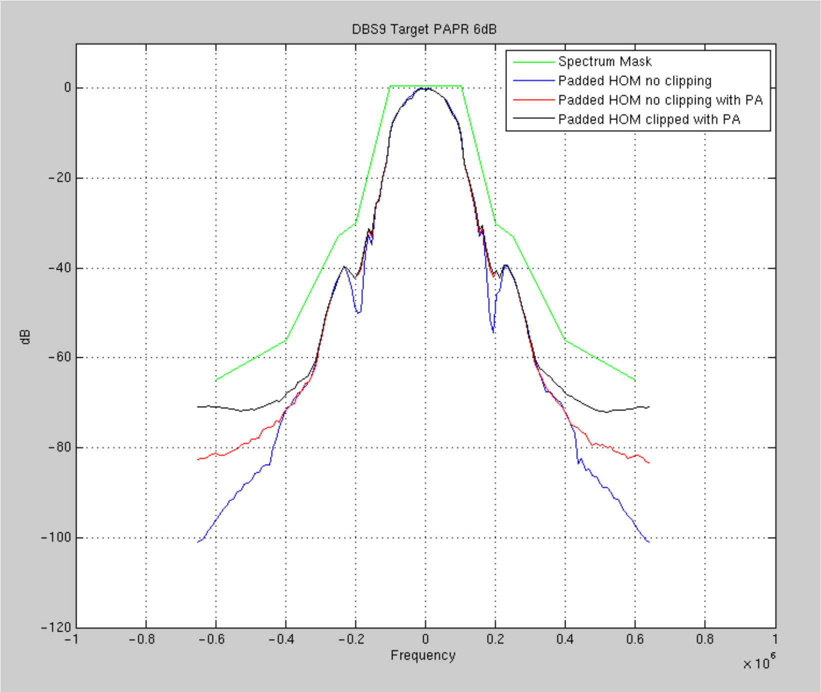


Figure 6.4-12 Spectrum of DBS-9(Target 6dB) due to PAR reduction with soft and hard clipping.

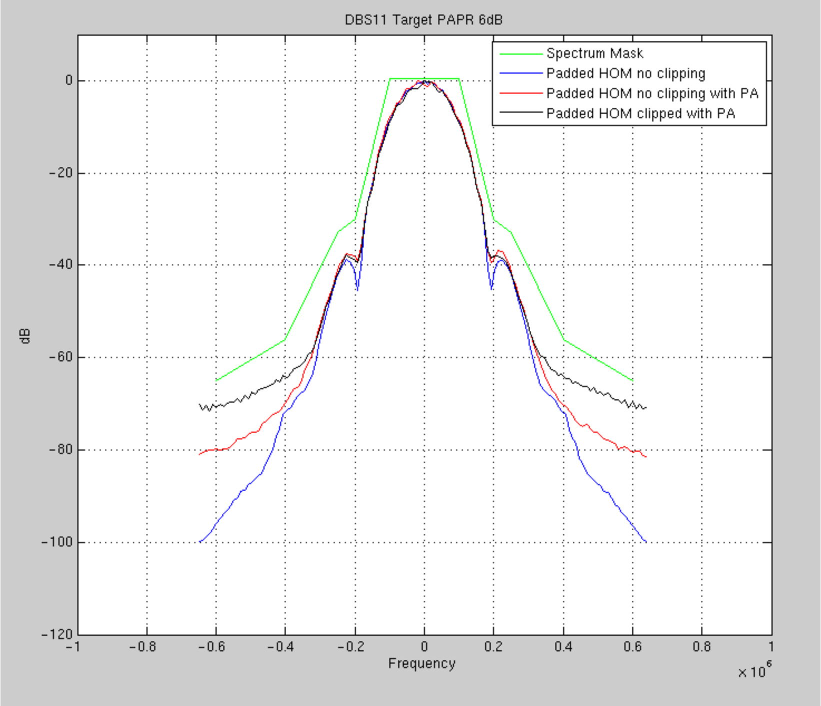


Figure 6.4-13 Spectrum of DBS-11b(Target 6dB) due to PAR reduction with soft and hard clipping.

6.4.3 PAR reduction complexity for Single Block PCE2

The evaluation in sub-clause 6.4.2 indicates that the combined soft and hard clipping algorithm offers the best tradeoff between link level performance and computational complexity. The ambition of this sub-clause is to verify this indication, and make a thorough assessment of the computational complexity demanded by the combined soft and hard clipping algorithm.

6.4.3.1 Soft and Hard clipping algorithms

The soft clipping algorithm applies a compression signal to all signal peaks exceeding a configured threshold. This compression signal c_i is typically generated based on all signal samples s_i in a burst exceeding a configured power threshold th_{pow}, or the corresponding amplitude threshold th_{amp}.

$$c_i = \begin{cases} s_i - th_{amp} \cdot \dfrac{s_i}{\sqrt{s_i s_i^*}}, & if\ s_i s_i^* - th_{pow} > 0 \\ 0, & if\ s_i s_i^* - th_{pow} < 0 \end{cases}$$

Based on the derived compression signal the soft clipped signal cs_i can be calculated as

$$cs_i = s_i - \sum_{k=0}^{L-1} h_k \cdot c_{i-k}$$

where h_k equals the soft clipping filter. In the evaluations presented in sub-clause 6.4.2 the soft clipping filter was set to equal the TX pulse shaping filter, i.e. the Linearized GMSK pulse shaping filter with a length L equal to five symbols.

In the case of Hard clipping the amplitude of all signal samples s_i exceeding the threshold th_{pow}, is set to th_{amp} while conserving the phase of the samples.

$$s_i = \begin{cases} th_{amp} \cdot \dfrac{s_i}{\sqrt{s_i s_i^*}}, if \; s_i s_i^* - th_{pow} > 0 \\ s_i, \qquad\qquad if \; s_i s_i^* - th_{pow} < 0 \end{cases}$$

The above equations involve logical comparison operations as well as arithmetic operations. In sub clause 6.4.3.2 a generic evaluation of the clipping algorithms computational complexity is performed. This evaluation is based on the estimated number of operations that is required to reach a PAR of 4 or 6dB for SBPCE2.

6.4.3.2 Computational complexity

The flow chart of Figure 6.4-6 outlines the main functional blocks in the combined hard and soft clipping algorithm, while estimates the computational complexity associated with each of the functional blocks are listed in table 6.4-3. The estimates are calculated as the number of real arithmetic and logic operations consumed when the transmitter soft clips x peaks, and hard clips y peaks per burst. Each estimate is a function of the length of the useful part of the burst *bl*, the selected oversampling rate *os*, the soft clipping filter length *fl* and the number of soft and hard clipped peaks *x+y*.

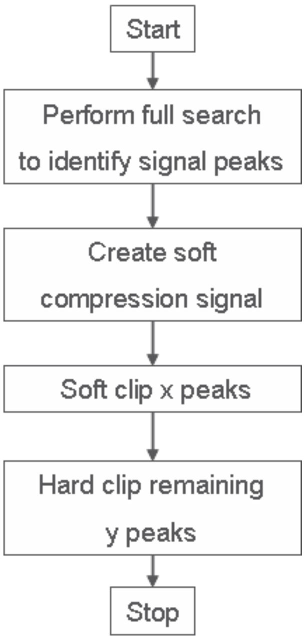


Figure 6.4-6 Block diagram over clipping algorithm.

Table 6.4-3 Clipping computational complexity.

Operation type	Full search	Soft clip signal	Soft clip	Hard clip
Comparisons	$bl \cdot os$			
Additions and Subtractions	$2bl \cdot os$	$3x$	$2x \cdot fl \cdot os$	y
Multiplications	$2bl \cdot os$	$4x$	$2x \cdot fl \cdot os$	$4y$
Divisions		$2x$		$2y$
Square roots		x		y

To get an overview of the total complexity Table 6.4-4 quantifies the expected computational complexity for SBPCE2-A and 2-B with some examples on the total number of operations per burst The number of comparisons, additions, subtractions and multiplications are summed together, as they usually consume the same number of CPU cycles. Divisions and square roots are kept separate since they are considered as more demanding operations to implement.

The examples are based on the assumption that the compression to 6dB PAR requires 4 soft and 7 hard clippings, while 4dB PAR requires 20 soft and 7 hard clippings, as stated in sub-clause 6.4.2. The SBPCE2-B burst length corresponds to the Low Complexity SBPCE2 burst format introduced in sub-clause 6.2.1.1.

Table 6.4-4 Examples of computational complexity.

Level	*os*	*bl*	*x + y*	*fl*	Number of <, +, - and * operations	Number of / and √ operations
EGPRS-2A	5	144	20 + 7	5	5 775	81
	5	144	4 + 7	5	4 063	33
	50	144	20 + 7	5	56 175	81
	50	144	4 + 7	5	40 063	33
EGPRS-2B	5	162	20 + 7	6	6 625	81
	5	162	4 + 7	6	4 593	33
	50	162	20 + 7	6	64 675	81
	50	162	4 + 7	6	45 363	33

The figures presented in table 6.4-4 can be compared to the cost of performing pulse shaping of a burst with the Linearized GMSK filter. Given a filter length *fl* of 6 symbols, a burst length *bl* of 177 symbols and an over sampling factor *os* of 5, the pulse shaping operation requires:

$$2 \cdot bl \cdot fl \cdot os \ additions + 2 \cdot bl \cdot fl \cdot os \ multiplications = 21240 \ operations$$

In the case of an oversampling factor *os* of 50 the number of operations will increase 10 times to 212 400. It can be concluded that in comparison to these estimates the computational complexity required by the SBPCE2 compression algorithm presented in table 6.4-4 is moderate.

6.4.4 References

[6.4-1] GP-119696, "PAR reduction for SBPCE2", source Telefon AB LM Ericsson, ST-Ericsson SA., GERAN#49.

[6.4-2] GP-061690, "Compressed QAM Modulation", source Telefon AB LM Ericsson, GERAN#31.

[6.4-3] GP-101850, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA., GERAN#48.

[6.4-4] GP-101852, "DAS-12b and DBS-12b burst formatting", source Telefon AB LM Ericsson, ST-Ericsson SA., GERAN#48.

[6.4-5] GP-101350, "Training symbol placements in Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson. GERAN#47.

[6.4-6] GP-101349, "Aspects of burst formatting of Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47.

[6.4-7] GP-111180, "PAR reduction complexity of SBPCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#51.

[6.4-8] GP-111183, "Complexity reduction of SBPCE2B", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#51.

[6.4-9] GP-111641, "Complexity reduction for rotation based PAR reduction", sourced Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#52.

[6.4-10] GP-101770, "Burst format of Improved Precoded EGPRS2 DL", source Huawei Technologies Co., Ltd., TSG GERAN #48.

[6.4-11] GP-120918, "Discussion on PAR Reduction for Padded HOM", sourced Huawei Technologies Co., Ltd. GERAN#55.

6.5 Impact on legacy services

PC EGPRS2 introduces a new modulation technique in GERAN, which might have impact on performance of legacy services but also impact on throughput due to segregation of resource in multiplexing scenarios of USF and PAN transmission.

6.5.1 Impact on EGPRS2 performance

6.5.1.1 Impact from Single Block PCE2 interference

6.5.1.1.1 Simulation assumptions and results

Figures 6.5-1 and 6.5-2, originally presented in [6.5-1], show simulated EGPRS2-A and EGPRS2-B throughput performance when exposed to PCE2-A and PCE2-B interference, respectively. The impact on performance is compared with the impact caused by EGPRS2-A and EGPRS2-B interference.

The simulations were performed in a Co-channel interference scenario since this is expected to most accurately reflect the characteristics of the interference and expose possible differences between PCE2 and EGPRS2 interference. In each scenario the interfering MCS is chosen to be identical to the carrier MCS. The detailed simulation assumptions are presented in Table 6.5-1

Table 6.5-1 Simulation assumptions.

Parameter	Value
Link direction	Downlink
MS Type	Non-SAIC
Frequency band	900MHz
Channels	TU3iFH
Interference scenario	Co-channel interference
Interference modulation	PCE2-A, PCE2-B, EGPRS2-A and EGPRS2-B.
MCSs	DAS5-DAS12, DBS5-DBS12
TSC placement	According to [6.5-2]
Burst length	According to [6.5-3]
CP length	PCE2-A: 6 PCE2-B: 9
RX BW	PCE2-A: 240kHz PCE2-B: 275kHz
ICI equalization	No
Impairments: – Phase noise – I/Q gain imbalance – I/Q phase imbalance – DC offset – Frequency error	Tx / Rx 0.8 / 1.0 [degrees (RMS)] 0.1 / 0.2 [dB] 0.2 / 1.5 [degrees] -45 / -40 [dBc] - / 25 [Hz]

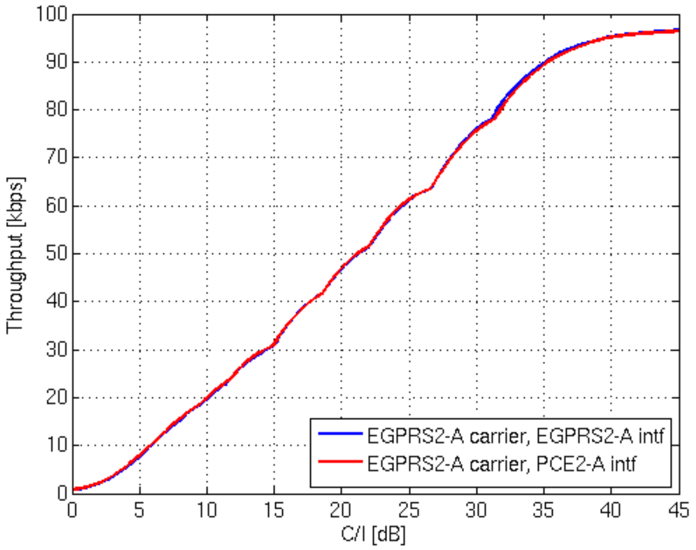


Figure 6.5-1 EGPRS2-A carriers exposed to PCE2-A and EGPRS2-A Co-channel interference.

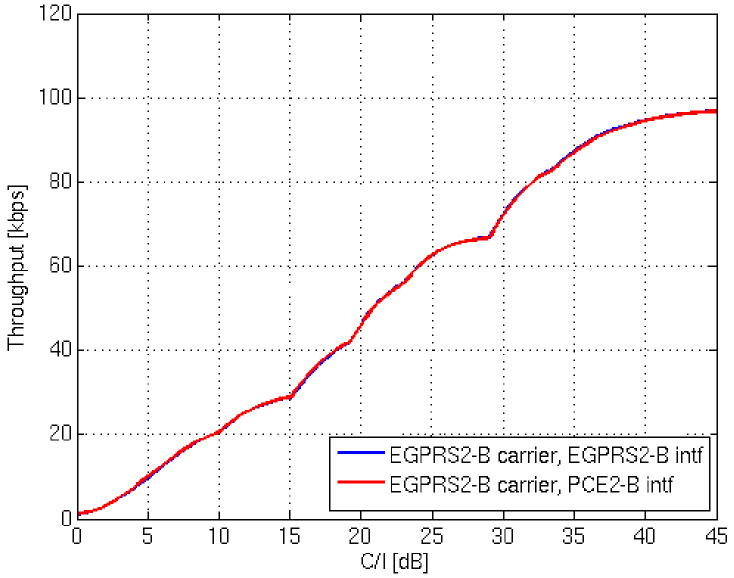


Figure 6.5-2 EGPRS2-B carriers exposed to PCE2-B and EGPRS2-B Co-channel interference.

6.5.1.1.2 Conclusions

Since PCE2 uses the same transmitter pulse shaping filters as EGPRS2 it can be concluded that the difference between PCE2 and EGPRS2 signal characteristics lies within the modulation method of the techniques. PCE2 is a narrow band OFDM based technique and its characteristics can be roughly approximated as pulse shaped Gaussian noise, regardless of transmitted MCS.

To conclude, it is expected that the impact from EGPRS2 and PCE2 interference on EGPRS2 performance is similar. Figure 6.5-1 and 6.5-2 shows that this assumption is valid as practically no difference can be seen between simulated EGPRS2 performances when exposed to PCE2 or EGPRS2 interference.

6.5.1.2 Impact from Padded HOM interference

6.5.1.2.1 Simulation assumptions and results

Figures 6.5-1a/1b and 6.5-2a/2b present simulated EGPRS2 BLER performance when exposed to EGPRS2 and PCE2 interference. The simulations were performed in a Co-channel and Adj-channel interference scenario. In each scenario the interfering modulation is the same as the modulation of wanted signal coding scheme for each type of interference. The detailed simulation assumptions are listed in Table 6.5-1a. The re-designed MCSs DAS-10b/11b/12b v1/v2 and DBS-10b/11b/12b v1/v2 are not considered in this evaluation.

Table 6.5-1a Simulation assumptions.

Parameter	Value
Coding Schemes	DAS-5,DAS-9,DAS-11 DBS-5,DBS-9,DBS-11
Channel propagation	TU3iFH
Frequency band	900 MHz
Interference scenario	CCI,ACI
Interference type	EGPRS2-A, EGPRS2-B, PCE2-A, PCE2-B
Tx filter	Lin GMSK
Frames	5000
Cyclic prefix length	level A: 6 level B: 15
DFT size	level A:144 level B: 162

Impairments:	Tx/ Rx
– Phase noise	0.8/1.2 [degrees (RMS)]
– I/Q gain imbalance	0.1/0.2 [dB]
– I/Q phase imbalance	0.2/2.0 [degrees]
– DC offset	-45/-40 [dBc]
– Frequency error	-/25

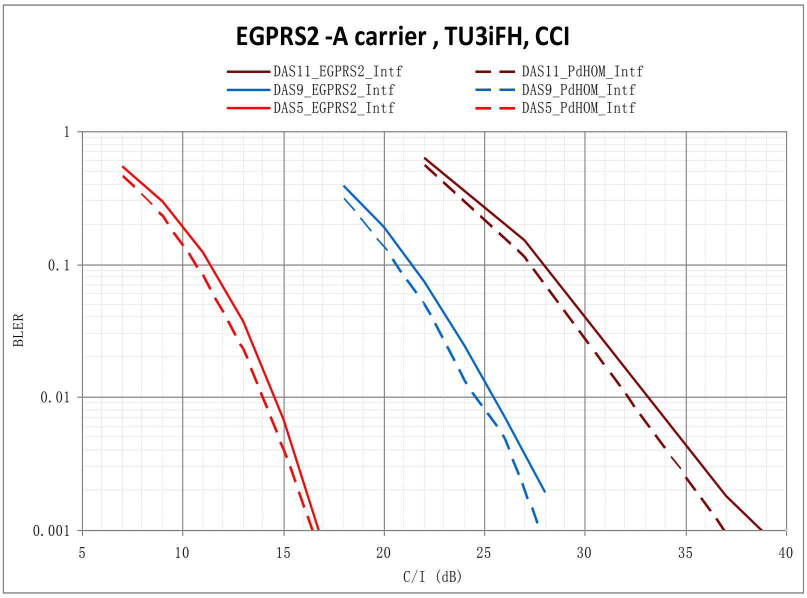


Figure 6.5-1a EGPRS2-A carriers exposed to EGPRS2 and Padded HOM Co-channel interference.

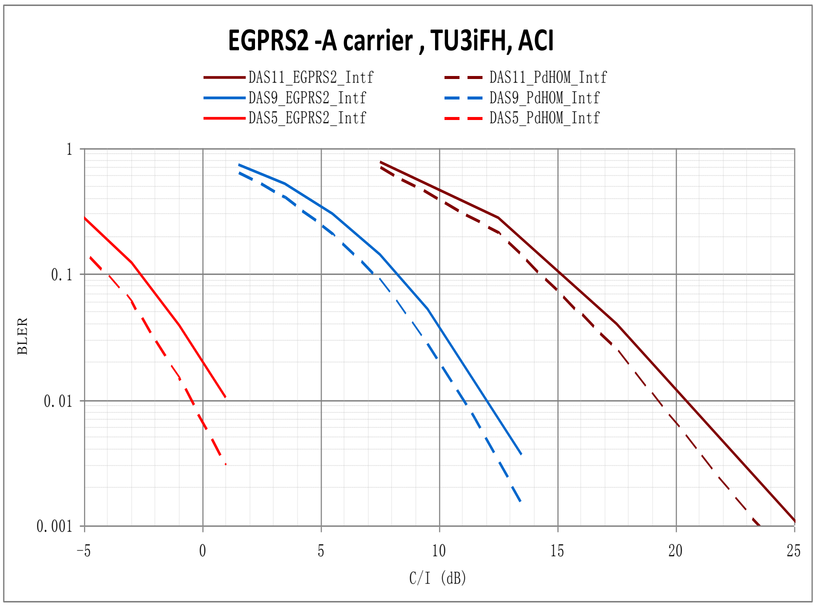


Figure 6.5-1b EGPRS2-A carriers exposed to EGPRS2 and Padded HOM Adj-channel interference.

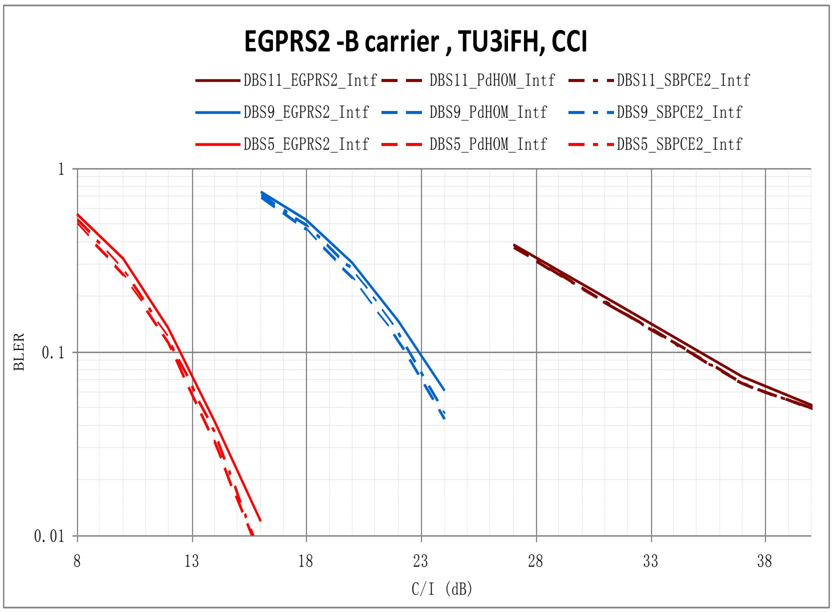


Figure 6.5-2a EGPRS2-B carriers exposed to EGPRS2 and PaddedHOM/SBPCE2 Co-channel interference.

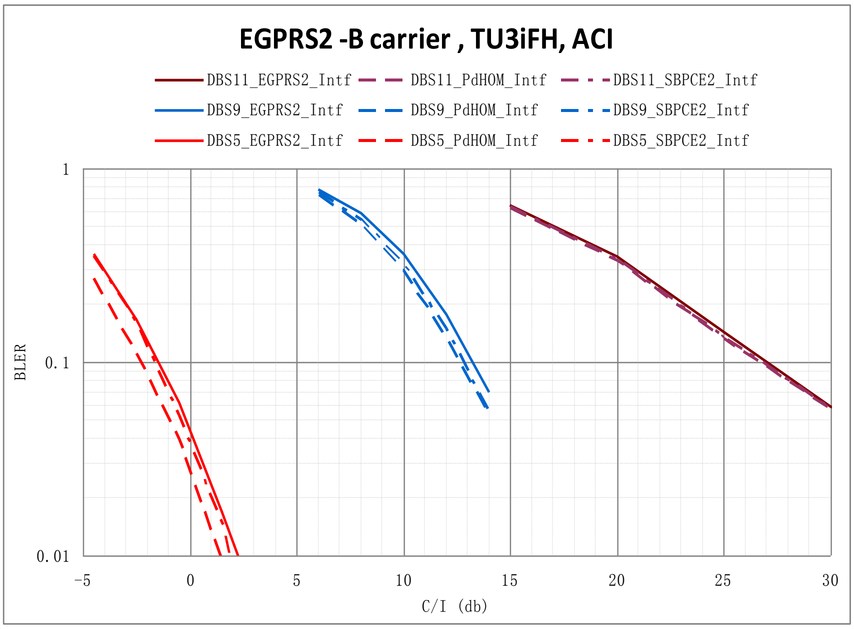


Figure 6.5-2b EGPRS2-B carriers exposed to EGPRS2 and PaddedHOM/SBPCE2 Adj-channel interference.

6.5.1.2.2 Conclusions

It can be seen that there is some performance improvement when the EGPRS2 carrier exposed to Padded HOM interference. The gain is more obvious for lower MCSs especially in adjacent channel interference scenario. Different PCE2 interferences (Padded HOM and SBPCE2 interference) are also taken into account. Padded HOM and SBPCE2 interference have similar impact on the EGPRS2 carrier except a little gain of lower MCSs for Padded HOM interference in simulated scenarios.

It is expected that Padded HOM interference (except re-designed MCSs) has positive impact on EGPRS2 carrier performance compared to EGPRS2 interference. The impacts of EGPRS2 interference and Padded HOM interference for Padded HOM carrier are also evaluated in [6.5-7]. The gain for Padded HOM interference (except re-designed MCSs) is much more obvious for Padded HOM carrier performance.

6.5.2 USF/PAN multiplexing

6.5.2.1 SBPCE2

6.5.2.1.1 Introduction

With the introduction of new modulation techniques there will possibly be multiplexing losses due to incompatibilities of legacy MS not being able to read the precoded USF/PAN.

The USF problem was faced already when EGPRS was introduced in release 99, as 8PSK modulated USF:s cannot be read by GPRS MS. The same problem also occurred in GERAN Evolution in Rel-7 where 16QAM/32QAM modulations and higher symbol rates were introduced.

With Single Block Precoded EGPRS2, SBPCE2, the whole burst is precoded (including TSC and USF) and thus there will be potential losses of throughput in uplink and downlink due to USF multiplexing issues.

With the introduction of Piggy backed Ack/Nacks with the LATRED feature the same type of problem, as for USF, exist for PAN.

In this sub clause simulations are provided on the impact due to USF multiplexing of different MS capabilities.

6.5.2.1.2 USF multiplexing

6.5.2.1.2.1 Simulator description

The simulator used is a dynamic GSM/EDGE traffic simulator with channel management, downlink and uplink scheduling. The traffic model used is file download and file upload with fixed file size (100 kB). MCSs are chosen based on the specified CIR from an ideal LA throughput curve and the MS capabilities. The mix of EDGE and MSs capable of HOM and possibly other modulation techniques is denoted as (PC)E2 penetration. Thus, EGPRS2-A, EGPRS2-B, SBPC EGPRS2-A and SBPC EGPRS2-B are all classified in the "(PC)E2" penetration. Five different MS types have been used in the evaluation, listed in Table 6.5.2.

Table 6.5-2. MS capabilities used in the simulations

MS cap.	TBFs supported
1	EGPRS
2	EGPRS + EGPRS2-A
3	EGPRS + EGPRS2-A + PC EGPRS2-A
4	EGPRS + + PC EGPRS2-A
5	EGPRS + + PC EGPRS2-A + PC EGPRS2-B

The (PC)E2 penetration is swept from 0% to 100 % in steps of 25%. The offered traffic load is specified as 70% of the theoretical cell capacity (kbps/cell) at 100% EDGE penetration and for the given CIR level (for the corresponding MCS which is chosen for that CIR).

Radio link modelling is taken from a CIR distribution from a PS network level simulation for a 3/9 re-use, see [6.5-6], for all users. MCSs are chosen (to maximise throughput) based on the specified CIR and the MS capabilities. Block errors are assumed to be independent.

Simulation parameters are summarised in Table 6.5-3.

Table 6.5-3. Summary of simulation parameters.

Parameter	Value
TRXs per cell	2 (16 timeslots)
Multislot class	Class 12 (i.e., Rx=4, Tx=4, Sum=5) Up to 4+1 for downlink users Up to 3+2 for uplink users
Traffic model	FTP download/upload 100 kB packet size Poisson user arrival process. A user leaves the system when download/upload is completed.
USF granularity	1 or 4 for EDGE MS 1 for (PC)E2 MS
CIR [dB]	Distribution from 3/9 re-use network, see [6.5-6].

It is assumed that (PC)E2 MSs can decode the USF of EGPRS blocks as well as (PC)E2 blocks, according to the MS capability in Table 6.5-2. EGPRS MS can only decode the USF of 8PSK modulated blocks

Performance is shown as the relative gain, achieved at a certain (PC)E2 penetration, of the mean user throughput compared to that for 100% EDGE penetration. The download/upload time comprises transmission time, scheduling delays, TBF set-up delays and TCP impact.

One DL MS penetration scenarios have been investigated:

- EGPRS/EGPRS2-A/PC EGPRS2-A using MS capabilities (see 6.5-2) 1, 2, 3 and 4.

The MS mix in uplink is assumed to consist of EGPRS and EGPRS2-A MSs. The EGPRS2-A MS penetration corresponds to the (PC)E2 penetration in each scenario, e.g. DL: [50/25/10/10/5] (see table below) -> UL: [50 (EGPRS) / 50 (25+10+10+5) EGPRS2-A]

Table 6.5-4. Penetration of MS DL capability.

(PC)E2 penetrations MS capability [1/2/3/4/5]				
0	25	50	75	100
[100/0/0/0/0]	[75/12.5/6.25/6.25/0]	[50/25/12.5/12.5/0]	[25/12.5/31.25/31.25/0]	[0/0/50/50/0]

6.5.2.1.2.2 Measures to combat USF multiplexing

In [6.5-5] a modified burst format is shown for Precoded EGPRS2 where part of the payload and the TSC is modulated in time domain to allow for multiplexing with MSs not supporting Precoded EGPRS2. The potential gain with the modified burst format is investigated in the simulations and indicated as 'Mod BF'.

USF granularity is a standardized solution for both EGPRS and EGPRS2, see [6.5-6], where the network assigns (in e.g. PACKET UL ASSIGNMENT) the MS to transmit either one (USF granularity = 1) or four (USF granularity = 4) consecutive radio blocks in the UL at the reception of the assigned USF value. In the following text the term USF granularity is used for the value set to 4.

Both USF granularity and the Modified burst format provide means improve UL scheduling while minimizing restrictions on DL modulation transmissions. The difference is that:

- with USF granularity, restrictions apply on the DL modulation in that the receiving MS shall be able to receive the modulated USF in the first block. In the remaining three blocks no restrictions apply. It should be noted that if the user has less than 4 blocks of data to send in the uplink when receiving a USF with a granularity of 4 in the downlink, the resulting uplink blocks carrying no user data are actually a waste of uplink resources, dependent on the scenario and traffic model investigated.
- with the modified burst format there is an expansion of the DL modulation possible to use, in each block, see [6.5-5]. There is no impact on UL granularity.

6.5.2.1.2.3 Scheduling strategies

Two strategies are investigated:

1. Strategy 1 does not take the USF problem into account. If a downlink block is sent using precoding or higher order modulation, containing a USF to a legacy MS, the legacy MS will not receive the USF and no transmission will occur in the corresponding uplink block. The USF granularity is 1 for all users.

2. Strategy 2 is a simple attempt to reduce the USF problem. If there is a conflict between preferred downlink modulation and USF decoding, the MCS of the downlink block is reduced to accommodate USF reception. Strategy 2 has also been investigated in combination with using USF granularity 4 for EGPRS MSs and with the modified burst format ([6.5-5]) to further reduce multiplexing losses.

6.5.2.1.2.4 Results

In figure 6.5-3 the throughput impact on downlink and uplink is shown at different (PC)E2 MS penetrations using the simulation parameters and traffic model in Table 6.5-3. Both scheduling strategies described in 6.5.2.1.3 have been investigated using both USF granularity 1 and 4 for EGPRS capable MSs for scheduling strategy 2. Note that USF granularity=1 is always used for (PC)E2 MSs. Also, the modified burst format ('Mod BF') has been used both with and without USF granularity 4 of EGPRS MSs.

The large downlink gain seen is a combination of increased transmission bitrate and decreased scheduling delays. The reason for the decreased scheduling delays is the decrease in channel utilization (i.e. that less time is needed to serve the offered traffic).

It can be seen that:

- Scheduling strategy 1, i.e. prioritizing downlink throughput at the expense of not being able to schedule uplink MSs, has a strong negative impact on uplink throughput. This should not be seen as a realistic network implementation but rather serve as the scenario where DL throughput is prioritized.

- Scheduling strategy 2, i.e. prioritizing uplink throughput by choosing an MCS for downlink transmission (even if not optimum) that is compatible with the receiver capability to ensure USF reception, has limited but visible negative impact on downlink throughput if using USF granularity 1. All investigated combinations of USF granularity and modified burst format, with USF scheduling strategy 2, give the same, maximum UL throughput.

- The reduced UL scheduling frequency by using USF granularity 4, see 6.5.2.1.3, for the EGPRS MSs does not seem to have visible impact to the results but minimizes the throughput loss at USF granularity 1 considerably for scheduling strategy 2, almost completely matching scheduling strategy 1.

- For both USF granularity 1 and 4, using the modified burst format does show some additional gains.

- It can be seen that using USF granularity 4 and the modified burst format, in some cases, slightly improves downlink performance compared to scheduling strategy 1 (with USF granularity 1). This gain is believed to be related to the more efficient use of UL resources for scheduling strategy 2.

Further, it should be noted that additional measures to limit the throughput loss such as outlined in [6.5-4] have not been implemented in the simulator.

For more results on other MS penetration scenarios and other network loads please see [6.5-4].

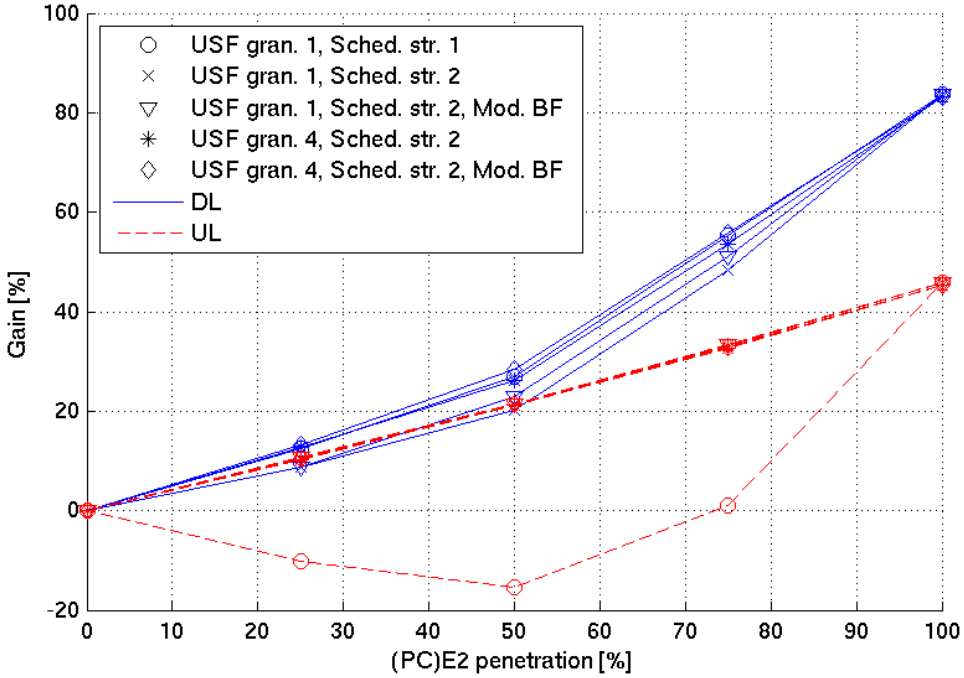


Figure 6.5-3. Throughput impact at different (PC)E2 penetrations and network load levels, 70%.

6.5.2.1.3 PAN multiplexing

The piggy backed Ack/Nack introduced in Rel-7 with the LATRED feature is signaled in the header by the PANI field. The PAN block is accommodated by extra puncturing of the data block and is then interleaved together with the data and bit swapped to ensure good enough performance to allow for PAN scheduling in DL to users not intended for payload reception (similar to the flexibility of USF).

In a SBPCE2 block there is no possibility of the legacy, non Precoded, MS to receive the PAN field which is spread across all four burst of the radio block at positions usually occupied for data.

In similarity to the USF, if a scheduling opportunity for the PAN is lost the Ack/Nack cannot be sent in the block, resulting in larger transfer delay. A number of measures can be taken to alleviate this problem. The mechanisms described for the USF, e.g. lowering of MCS to accommodate PAN reception can be used and/or minimize multiplexing of different MS capabilities in the channel allocation to further minimize impacts of multiplexing. It should also be noted that the higher the multiplexing rate the lower the probability of delay sensitive services being allocated on the channels. Also, low latency applications typically has a relatively low target BLER (low frequency) used for (using PAN functionality) and PAN is typically transmitted per TBF and not on a timeslot resolution as for the USF.

6.5.3 References

[6.5-1] GP-101351, "Precoded EGPRS2 DL – Impact on legacy services", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.5-2] GP-101350, "Training symbol placements in Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.5-3] GP-101349, "Aspects of burst formatting of Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[6.5-4] GP-111185, "SBPCE2 USF/PAN multiplexing", source Telefon AB LM Ericsson ,ST-Ericsson SA. GERAN#51.

[6.5-5] GP-110172, "Burst format for precoded EGPRS2", source Research In Motion UK Ltd. GERAN#49

[6.5-6] GP-101066, "Precoded EGPRS2 Downlink (update of GP-100918)", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#46

[6.5-7] GP-120129, "Impact on EGPRS2 and PCE2 performance for Padded HOM", source Huawei Technologies Co., Ltd. GERAN#53

6.6 Impacts on base station and mobile station

This sub clause evaluates the impact of Precoded EGPRS2 to base station and mobile station implementations.

Considerations of re-usable functionality from LTE, such as FFT size, modulation definitions and demodulation, can be considered in the evaluation.

6.6.1 Impacts on base station

6.6.1.1 Single Block Precoded EGPRS2, SBPCE2

The introduction of PCE2 affects the transmitter chain in the BTS. Table 6.6-1 lists the blocks that are new or need to be modified. The complexity estimates are based on the assumption that SBPCE2 is the technique used in the transmitter chain. Since some of the signal processing algorithms are dependent upon the sampling rate, which is implementation dependent, the complexity estimates are given as a function of the oversampling rate *os* (an integer).

The most computationally demanding block in the modulator is pulse shaping. Its complexity is dependent on the oversampling rate. Thus, it is reasonable to quantify the increase in complexity due to precoding as a percentage of the computational cost of pulse shaping. The overall complexity increase due to precoding is between 43% and 95% for level A, and between 39% and 130% for level B, depending on the oversampling rate. The lowest percentage corresponds to an oversampling rate of 48, while the highest corresponds to an oversampling rate of 4. These estimates include the Inverse Fourier Transform, ramping, soft and hard clipping, and in the case of level B, PAPR reduction by means of training sequence rotation.

For the complexity estimation of SBPCE2-B, Low Complexity SBPCE2-B, LCSBPCE2-B is used.

The addition of clipping is the main contributor to the increase in complexity, followed by the IFFT.

Table 6.6-1 Modifications to the Transmitter Chain

Block	New functionality	Relation to LTE	Comments	Computational Complexity SBPCE2 A	Computational Complexity LCSBPCE2 B
Channel encoder	Similar or identical to EGPRS2. Roughly 20% higher bit rate for DAS-12b/DBS-12b. The encoder comprises interleaving, puncturing and turbo-coding.	Same turbo-encoder	Introduction of 64QAM increases bit rate by 20% for DAS- 12b/ DBS-12b	20% more complex than EGPRS2A encoder for DAS-12b. Same complexity as EGPRS2A for other MCSs.	20% more complex than EGPRS2B encoder for DBS-12b. Same complexity as EGPRS2A for other MCSs.
Symbol mapping	Mapping to 64QAM symbols	Same 64QAM constellation		20% more complex than 32QAM EGPRS2A symbol mapping	20% more complex than 32QAM EGPRS2B symbol mapping
IFFT	Precoder. This is a new block	Radix 2 and 3 are used	It is assumed that a radix 2 butterfly requires 4 real additions, while a radix 3	2304 real additions 384 real multiplications	2916 real additions 864 real multiplications

			butterfly requires 12 real additions and 4 real multiplications		
Ramping	The ramp-up and ramp-down of the modulated signal are different from EGPRS2A	-	Low complexity, only a few symbols need to be multiplied by the ramp function	2*8**os* real multiplications	2*10**os* real multiplications
Soft and Hard clipping	PAPR reduction is a new function	-	The complexity estimate is explained in reference [6.6-1]. The clipping filter length is 5 (LinGMSK filter) Only the worst case (4 dB PAR) is shown	(2*144+2*20*5)* *os* + 3*20 + 7 real additions (2*144+2*20*5)* *os* + 4*20 + 4*7 real multiplications 144* *os* logical operations 54 divisions 27 square roots	(2*162+2*20*5)* *os* + 3*20 + 7 real additions (2*162+2*20*5)* *os* + 4*20 + 4*7 real multiplications 162* *os* logical operations 54 divisions 27 square roots
PAPR reduction by rotation of training sequence by using pulse shaping filter [1 1], see 6.4.1.1.3 [6.6-2]	PAPR reduction is a new function	-	3 possible rotation angles are used for 64QAM modulation in SBPCE2. One extra IFFT's (one for all rotation angles) are needed, plus calculation of PAPR for each angle.	-	2916 + 3*3*187 real additions 864 + 3*6*187 real multiplications 3*187 logical operations
Mixed mode modulation	The possibility of using symbols from different constellations in one burst is new	-	The complexity increase is negligible	-	-

6.6.1.2 Padded Higher Order Modulation, Padded HOM

Table 6.6-1a shows the blocks and computational complexity estimation affected by the introduction of Padded HOM in the BTS transmitter chain. The oversampling rate os (an integer) is used as a parameter when the operations are given.

For the complexity estimation Low complexity Padded HOM is used, where 144 and 162 FFT length are utilized for level-A and level-B estimation.

The increase in complexity due to precoding is quantified as a percentage of the computational cost of pulse shaping. The overall complexity increase due to precoding is between 43% and 95% for level A, and between 35% and 75% for level B, depending on the oversampling rate. The lowest percentage corresponds to an oversampling rate of 48, while the highest corresponds to an oversampling rate of 4. These estimates include the Inverse Fourier Transform, ramping, soft and hard clipping.

Table 6.6-1a Modifications of Padded HOM to the Transmitter Chain

Block	New functionality	Relation to LTE	Comments	Computational Complexity for level-A	Computational Complexity for level-B
Channel encoder	Refer to Table 6.6-1.				
Symbol mapping	Mapping to Higher Order Modulation symbols	Same 64QAM constellation	Except 64QAM, same constellations in EGPRS2 are re-used	Except re-designed coding schemes, the per-symbol complexity is up to 33% higher than EGPRS2A due to the introduction of higher order modulation for each coding scheme, but this is compensated by using a shorter mapping length. The total complexity is at least 2.3% lower than EGPRS2A. For re-designed coding schemes (the worst case limits the complexity), 20% more complex than 32QAM EGPRS2A symbol mapping	Except re-designed coding schemes, the per-symbol complexity is up to 50% higher than EGPRS2A due to the introduction of higher order modulation for each coding scheme, but this is compensated by using a shorter mapping length. The total complexity is at least 2.4% lower than EGPRS2A. For re-designed coding schemes (the worst case limits the complexity), 20% more complex than 32QAM EGPRS2A symbol mapping.
IFFT	Refer to Table 6.6-1.				
Ramping	Refer to Table 6.6-1.				
Soft and Hard clipping	Refer to Table 6.6-1.				

6.6.2 Impacts on mobile station

6.6.2.1 Single Block Precoded EGPRS2, SBPCE2

The introduction of PCE2 affects the receiver chain in the MS. Table 6.6-2 lists the blocks that are new or need to be modified. The complexity estimates are based on the assumption that SBPCE2 is the technique used in the receiver chain. Since some of the signal processing algorithms are dependent upon the length of the synchronization window, which is implementation dependent, the complexity estimates are given as a function of the synchronization function length *win_len* (an integer).

For the complexity estimation of SBPCE2-B, LCSBPCE2-B is used.

Blind detection is the main contributor to the increase in complexity. The demodulation, however is significantly less costly than for EGPRS2.

The complexity of an EGPRS2 receiver can be roughly estimated as follows. It is assumed that a trellis search using a reduced trellis is performed. Moreover, it is assumed that only two survivors are kept at every trellis stage, and that only 2 MLSE taps are used. In other words, a low complexity trellis search is assumed for the EGPRS2 demodulator. The total number of arithmetic operations has been counted, assuming that multiplications, additions and maximum calculations have all the same cost. The total cost is estimated to be 1919 operations per trellis step. For level A this gives a total of 223000 = 1919*116 operations and for level B 265000 = 1919*138 operations. From statistics from actual receivers, it has been estimated that the estimation block, comprising blind detection, synchronization and channel estimation in an EGPRS2 receiver, consumes 30% of the number operations used in the trellis search. Moreover, from receiver statistics it is estimated that blind detection accounts for 20% of the complexity of the estimation block. Based on these estimates, it is possible to calculate the number of operations consumed by the EGPRS2 equalizer and the number of operations consumed by blind detection. The equalization of one 32QAM modulated EGPRS2A burst consumes about 290000 operations, while the equalization of one 32QAM modulated EGPRS2B burst consumes approximately 345000 operations. In addition, blind detection for level A is estimated to consume 13380 operations, while blind detection for level B is estimated to consume 15900 operations. Using these estimates, assuming a synchronization window with length *win_len* = 5 symbols, and using Table 6.6-2 below, it is possible to compare the complexity of PCE2 to EGPRS2. The equalization of one PCE2A burst, for an MS supporting EGPRS, EGPRS2A, PCE2A consumes 103000 operations. The equalization of one PCE2B burst, for an MS supporting EGPRS, EGPRS2A, EGPRS2B, PCE2A, PCE2B, consumes 144000 operations.

Taking into account channel decoding, the total complexity of the equalization and decoding of one 64QAM modulated PCE2A radio block is about 76% as complex as the equalization and decoding of one 32QAM modulated EGPRS2A radio block. Similarly, the equalization and decoding of one 64QAM modulated PCE2B radio block is estimated to be around 82% as complex as the equalization and decoding of one EGPRS2B radio block.

Table 6.6-2 Modifications to the Receiver Chain

Block	New functionality	Relation to LTE	Comments	Computational Complexity SBPCE2-A	Computational Complexity LCSBPCE2-B
Channel decoder	Similar or identical to EGPRS2. Roughly 20% higher bit rate for DAS-12b/DBS-12b. The decoder comprises deinterleaving, depuncturing and turbo-decoding.	Same turbo decoder	Introduction of 64QAM increases bit rate by 20% for DAS-12b/DBS-12b	20% more complex than EGPRS2-A decoder for DAS-12b	20% more complex than EGPRS2-B decoder for DBS-12b.
FFT	Needed for blind detection, synchronization, channel estimation	Radix 2 and 3 are used. It might be possible to re-use implementation or accelerators from LTE	It is assumed that a radix 2 butterfly requires 4 real additions, while a radix 3 butterfly requires 12 real additions and 4 real multiplications	2304 real additions 384 real multiplications	2916 real additions 864 real multiplications
Blind detection for MS supporting EGPRS,	It is necessary to detect whether the signal is	-	The test of each modulation hypothesis (4 in total) is the	Same as blind detection for EGPRS2-A plus	-

PCE2A	precoded.		same as for EGPRS2, plus one FFT for each position in synchronization window.	*win_len* FFT’s	
Blind detection for MS supporting EGPRS, PCE2A, PCE2B and PAPR reduction for 64QAM modulation with training sequence rotation	It is necessary to detect whether the signal is precoded	-	The test of each modulation hypothesis is the same as for EGPRS2, plus one FFT for each position in synchronization window.	Same as blind detection for EGPRS2-A plus *win_len* FFT’s	Same as blind detection for EGPRS2-B plus 2x*win_len* FFT’s
Blind detection for MS supporting EGPRS, EGPRS2A, PCE2A, EGPRS2B, PCE2B and PAPR reduction for 64QAM modulation with training sequence rotation	It is necessary to detect whether the signal is precoded	-	The test of each modulation hypothesis is the same as for EGPRS2, plus one FFT for each position in synchronization window	Twice the complexity of blind detection for EGPRS2-A plus *win_len* FFT’s	Twice the complexity of blind detection EGPRS2-B plus 2x*win_len* FFT’s
Synchronization and channel estimation	Training sequence is modulated in the frequency domain.	-	The FFT calculation can be re-used from the blind detection block.	Same as corresponding functionality for EGPRS2-A plus additional processing for adjacent channel suppression. 53520 + 90000	Same as corresponding functionality for EGPRS2-B plus additional processing for adjacent channel suppression. 63600 + 90000
Demodulation	One tap equalizer	It may be possible to re-use a 16QAM and 64 QAM demodulators from LTE	Efficient methods to compute the soft values exist. For 64QAM a log-max-MAP equalizer requires 40 real additions, 26 real absolute value calculations, 4 arithmetic shifts and 6 real multiplications.	4720 real additions 708 real multiplications 472 arithmetic shifts 3068 real absolute values	5440 real additions 816 real multiplications 544 arithmetic shifts 3536 real absolute values

6.6.2.2 Padded Higher Order Modulation, Padded HOM

Table 6.6-2a shows the blocks and computational complexity estimation affected by the introduction of Padded HOM in the MS receiver chain. The synchronization window length *win_len* = 5 symbols is used when the operations are given.

For the complexity estimation Low complexity Padded HOM is used, where 144 and 162 FFT length are utilized for level-A and level-B estimation.

The operations of legacy 32QAM modulated EGPRS receiver is in reference to the analysis in section 6.6.2.1 Taking into account channel decoding, the total complexity of the equalization and decoding of one 64QAM modulated Padded HOM-A radio block is about 51% as complex as the equalization and decoding of one 32QAM modulated EGPRS2A radio block. Similarly, the equalization and decoding of one 64QAM modulated Padded HOM-B radio block is estimated to be around 53% as complex as the equalization and decoding of one EGPRS2B radio block.

Table 6.6-2a Modifications of Padded HOM to the Receiver Chain

Block	New functionality	Relation to LTE	Comments	Computational Complexity for level-A	Computational Complexity for level-B
Channel decoder	Refer to Table 6.6-2.				
FFT	Refer to Table 6.6-2.				
Blind detection for MS	Refer to Table 6.6-2.				
Synchronization and channel estimation	Training sequence is modulated in the frequency domain.	-	The FFT calculation can be re-used from the blind detection block.	5464 real additions 28932 real multiplications 428 divisions	11276 real additions 36354 real multiplications 454 divisions
Demodulation	One tap equalizer	It may be possible to re-use a 16QAM and 64 QAM demodulators from LTE	Efficient methods to compute the soft values exist. For 64QAM a log-max-MAP equalizer requires 27 real additions, 24 real multiplications, 42 real comparisons and 1 division.	3510 real additions 3120 real multiplications 5460 real comparisons 130 divisions	4023 real additions 3576 real multiplications 6258 real comparisons 149 divisions

6.6.3 References

[6.6-1] GP-111180 "PAR Reduction complexity o SBPCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#51.

[6.6-2] GP-111641, "Complexity reduction for rotation based PAR reduction", sourced Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#52.

7 Comparison of candidate techniques

This sub-clause contains comparison of candidate techniques presented in this TR.

One sub-clause is allocated to each investigation with the sourcing company(ies) of the investigation listed.

7.1 SBPCE2 and Padded HOM (by Telefon AB LM Ericsson and ST-Ericsson SA)

7.1.1 Simulation assumptions

7.1.1.1 Common assumptions

Table 7.1.1. Simulation assumptions

Parameter	Value
Burst length	According to [7.1.2]
RX BW	PCE2A: 280kHz
TSC length	26
ICI Suppression	No
Backoff	No
Channel propagation	TU50nFH
Interference	AWGN, CO (GMSK), ADJ (GMSK)
Frequency band	900 MHz
Frames	20000
Soft clipping	According to [7.1.3]
Target PAR for soft clipping	4 dB (8PSK MCSs) 6 dB (16/32/64 QAM MCSs)
Tx/Rx impairments	Ericsson typical TX/RX impairments:
- Phase noise	0.8 / 1.2 [degrees (RMS)]
- I/Q gain imbalance	0.1 / 0.2 [dB]
- I/Q phase imbalance	0.2 / 2.0 [degrees]
- DC offset	-45 / -40 [dBc]
- Frequency error	- / 25 [Hz]
- PA model	Yes / -

7.1.1.2 SBPCE2 specific assumptions

Table 7.1.2. SBPCE2 specific assumptions

Parameter	Value
MCS	DAS-5-11, DAS-12b
CP length	6
TSC placement	According to

7.1.1.3 Padded HOM specific assumptions

It should be noted that to facilitate the evaluation some of the design parameters for Padded HOM has not been strictly followed. These simplifications are not expected to have any impact on the results, or the conclusions drawn. Simulation assumptions used and modifications done are presented in this section.

Table 7.1.3. Padded HOM specific assumptions

Parameter	Value
MCS	DAS-5-12
CP length	6
TSC placement	According to [7.1.5]

It should be noted that the CP length is not chosen according to [7.1.5] but considering the evaluation done in [7.1.6] where a CP of 5 symbols was seen sufficient for the propagation channel used in this paper. Thus, choosing a longer CP length will not have impact on performance.

The TSC placement has been used as defined in [7.1.5] since this is a design parameter believed to have impact on the results considering the different numbers and allocation of the sub carriers between SBPCE2 and Padded HOM.

In the Padded HOM proposal the burst format is not symmetric nor is the number of bits per radio block equal to the bits carried by the HOM symbols. In this evaluation both burst symmetry and burst sizes that fit the block size have been chosen. The changes to the design are minimal and should not have impact on performance. Table 7.1.4 summarizes the differences.

Additional features such as Repeat pattern ([7.1.5]) or a shorter TSC ([7.1.5]) have also been investigated for Padded HOM previous investigations. Repeat pattern together with shorter TSC was only investigated in sensitivity with 0.3 and 0.9 dB gains seen (in [7.1.5]) for DAS-5 and DAS-11 respectively. For DAS-12 2.2 dB gains were observed.

Table 7.1.4. Difference in burst design between Padded HOM [7.1.7] and Padded HOM used in this evaluation.

		Padded HOM [7.1.7]		Padded HOM in current doc	
MCS	Block payload size	Modulation mix*	Block payload size	Modulation mix*	Block payload size
DAS-5-7	1392	[0,0,0,87,0,0,29]	1392	[0,2,0,86,0,0,28]	1392
DAS-8-9	1856	[0,0,0,0,93,0,23]	1860	[0,2,0,0,94,0,22]	1856
DAS-10-12	2320	[0,0,0,0,0,97,19]	2328	[0,2,0,0,0,96,18]	2320

* Modulation mix : [BPSK,QPSK,8PSK,16QAM,32QAM,64QAM,padding]

7.1.2 Performance plots

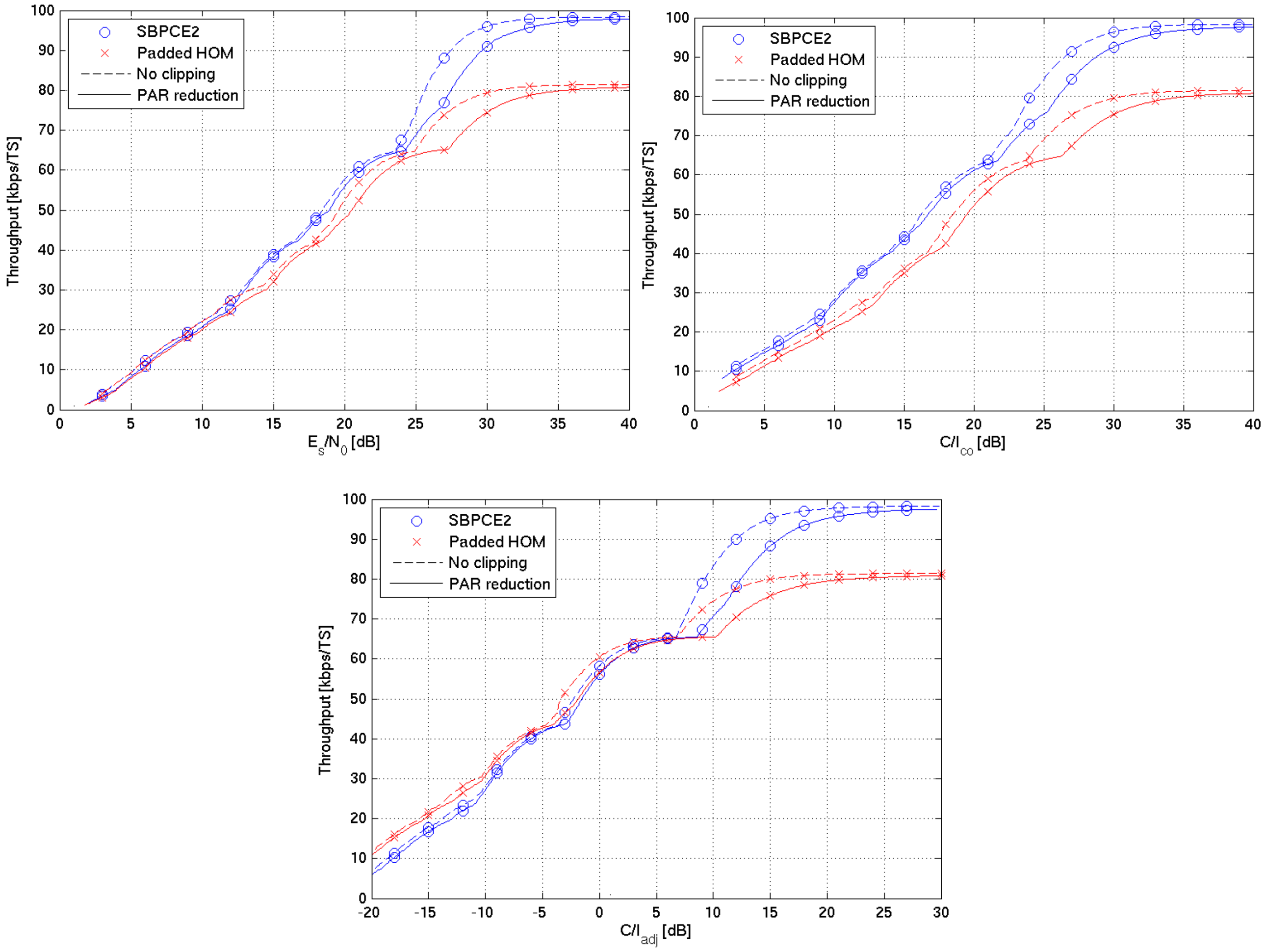


Figure 7.1.1. SBPCE2 and Padded HOM. Sensitivity (top left), Co (top right), Adj (lower left).

It should be noted that the header performance of both techniques were aligned within 1.5 dB, see [7.1.1]. Considering the findings in **Error! Reference source not found.** where a header BLER improvement of 15 dB caused a data BLER degradation of at most 1 dB these differences are expected to have minimal impact on the data BLER performance comparison in this investigation. For more details on the performance evaluation please see [7.1.7].

7.1.3 Discussion

It has been seen that SBPCE2 outperforms Padded HOM in both sensitivity and co-channel interference with the most significant differences for co-channel interference where gains of 2-5 dB can be seen for performance with reduced PAR. For adj-channel interference Padded HOM is superior for most MCSs due to the narrower spectrum used for payload allocation. For the highest MCSs however there is still a significant degradation compared to SBPCE2.

In all scenarios investigated DAS-12 with reduced PAR for Padded HOM did not reach BLER levels lower than 50% which makes the MCSs not useful for link adaptation. Significant degradations when reducing the PAR are seen due to the high modulation order used and the high code rate of this MCSs.

7.1.4 References

[7.1.1] GP-110929 "Comparison: Padded HOM and SBPCE2", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#49.

[7.1.2] GP-101349, "Aspects of burst formatting of Precoded EGPRS2 DL", source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[7.1.3] GP-061690, “Compressed QAM Modulation”, source Telefon AB LM Ericsson. GERAN#31.

[7.1.4] GP-101350, “Training symbol placements in Precoded EGPRS2 DL”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[7.1.5] GP-101770, “Burst format of Improved Precoded EGPRS2 DL”, source Huawei Technologies Co., Ltd., TSG GERAN #48

[7.1.6] GP-110201, “Evaluation of modified burst format of Precoded EGPRS2”, source Telefon AB LM Ericsson, ST-Ericsson SA

[7.1.7] GP-101431, “Improved Precoded EGPRS2 DL (update of GP-101289)”, source Huawei Technologies Co. Ltd.

[7.1.8] GP-101850,”Burst mapping of PCE2”, source Telefon AB LM Ericsson, ST-Ericssson SA. GERAN#48.

7.2 SBPCE2 and Padded HOM (by Huawei Technologies Co., Ltd)

7.2.1 Comparison of legacy coding schemes

7.2.1.1 Simulation assumptions

The common and specific simulation assumptions utilized in these two candidate techniques are listed in Table 7.2.1.

Table 7.2.1 Simulation assumptions

Parameter	Value	
	SBPCE2	Padded HOM
Burst length	PCE2A: 142 PCE2B:168	
RX BW	PCE2A:270kHz PCE2B:325kHz	
ICI Suppression	No	
Backoff	No	
Channel propagation	TU3iFH	
Interference	AWGN, CO (GMSK), ADJ (GMSK)	
Frequency band	900 MHz	
Frames	5000	
PAPR reduction	No	
Tx/Rx impairments	No	
CP length	PCE2A: 6 PCE2B: 9	
MCS	DAS-5-DAS-11 DBS-5~DBS-11	
TSC length	PCE2A: 26 PCE2B: 30	PCE2A: 17 PCE2B: 18
TSC placement	According to [7.1.4]	According to [7.2.2]

7.2.1.2 Simulation results

In this section ideal LA throughput envelope curves of SBPCE2 and Padded HOM both in level-A and level-B are depicted in Figure 2.1.1 and Figure 2.1.2. The detail figures please refer to [7.2.5].

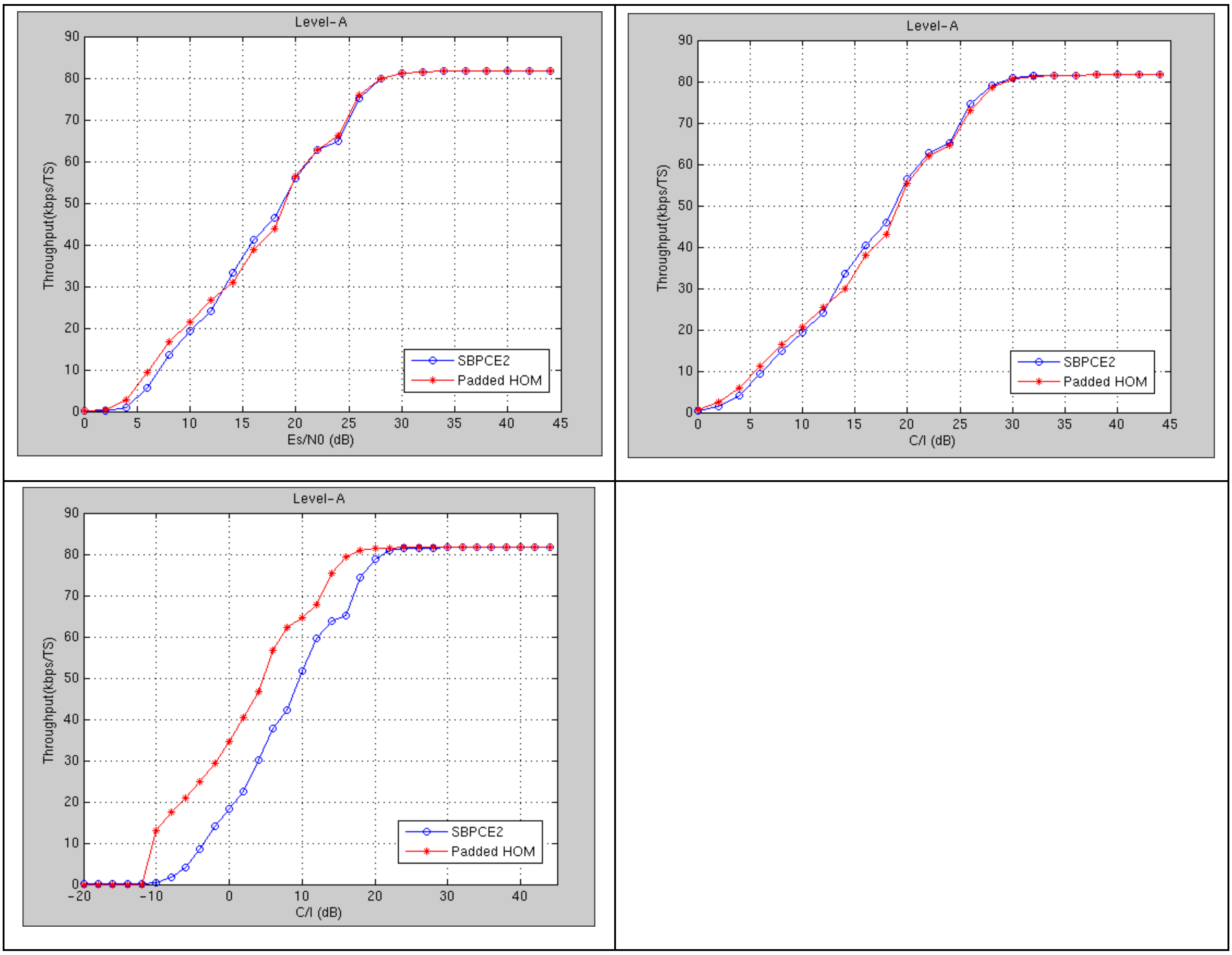


Figure 7.2.1 SBPCE2 and Padded HOM, lever-A, Sensitivity (top left), Co (top right), Adj (lower left).

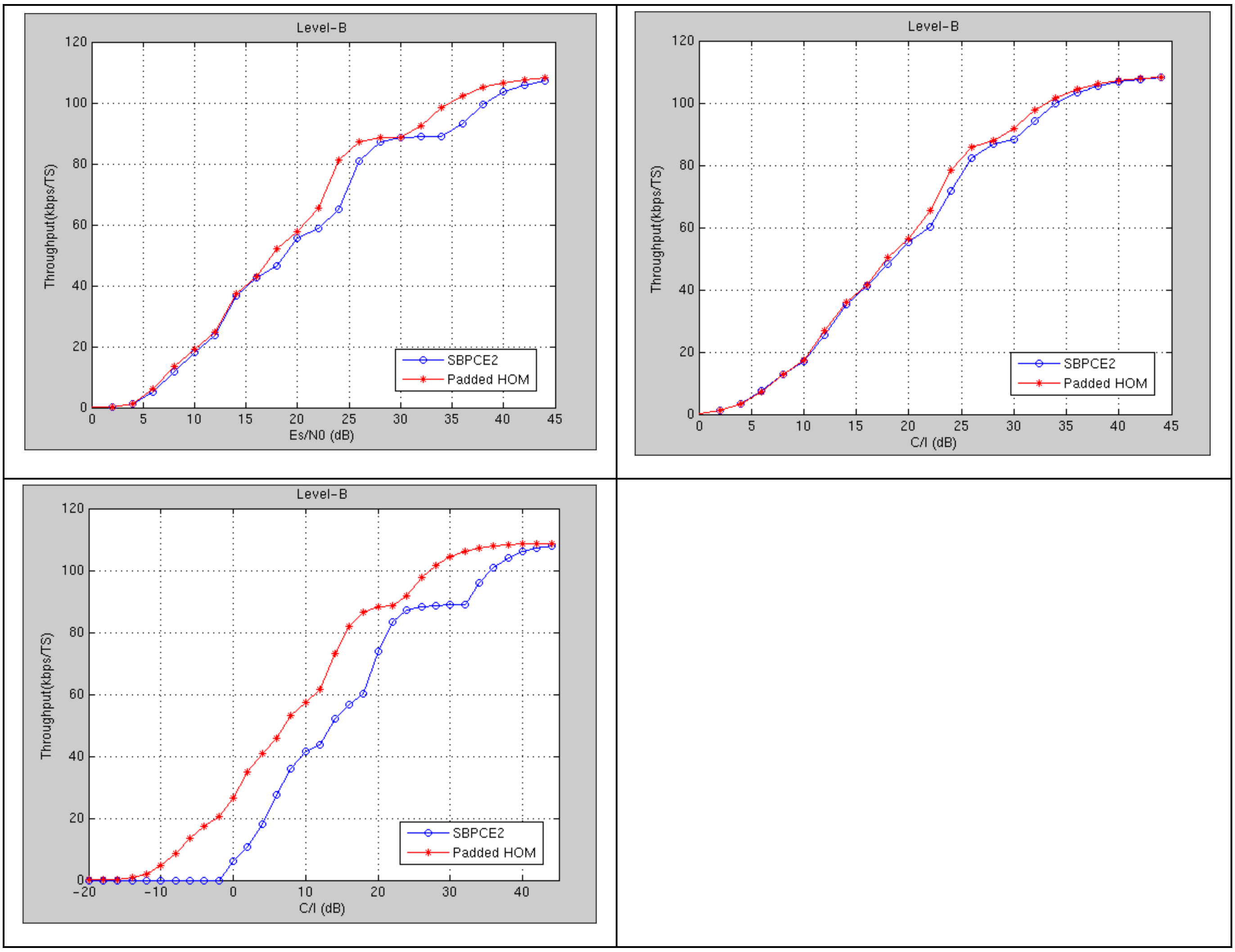


Figure 7.2.2 SBPCE2 and Padded HOM, level-B, Sensitivity (top left), Co (top right), Adj (lower left).

It can be seen that Padded HOM outperforms SBPCE2 in all scenarios investigated at level-B. SBPCE2 has slightly gains in the middle range of EsN0/CIR in sensitivity and CCI scenarios at level-A. In ACI scenario Padded HOM is superior for all modulation types.

7.2.2 Comparison of redesigned coding schemes

7.2.1.1 Simulation assumptions

The new channel coding DAS/DBS-12b proposed for SBPCE2 in [7.2.1] [7.2.6] and DAS/DBS-12b proposed for Padded HOM in [7.2.4] are evaluated with the same receiver. The specific and common simulation assumptions are listed in table7.2.2. The burst formats and coding parameters of the new channel coding are depicted in table 7.2.3 and table7.2.4.

The IDFT length of DAS-12b for SBPCE2 is not strictly followed. This change of the design is minimal and should not have impact on performance. The MMM is not considered in this evaluation.

Table 7.2.2 Simulation assumptions

Parameter	Value	
	DAS/DBS-12b for SBPCE2	DAS/DBS-12b for Padded HOM
IDFT length	PCE2A: 142 PCE2B:168	PCE2A: 140 PCE2B:168
CP length	PCE2A: 6 PCE2B: 9	PCE2A: 8 PCE2B: 9
TSC length	PCE2A: 26 PCE2B: 30	PCE2A: 12 PCE2B: 13
TSC placement	According to [7.1.4]	According to [7.2.2]
RX BW	PCE2A:270kHz PCE2B:325kHz	
ICI Suppression	No	
Backoff	No	
Channel propagation	TU3iFH, TU50noFH	
Interference	AWGN, CO (GMSK), ADJ (GMSK)	
Frequency band	900 MHz	
Frames	5000	
PAPR reduction	No	
Tx/Rx impairments	No	

Table 7.2.3 Burst format for DAS/DBS-12b

MCS	Bit swapper	Burst shift bits
DAS-12b for SBPCE2	on	50
DBS-12b for SBPCE2	on	0
DAS-12b for Padded HOM	on	88
DBS-12b for	on	54

Padded HOM		

Table 7.2.4 Rate matching and interleaving parameters for DAS/DBS-12b

MCS	*swap*	*a*
DAS-12b for SBPCE2	10%	103
DBS-12b for SBPCE2	0%	691
DAS-12b for Padded HOM	0%	622
DBS-12b for Padded HOM	0%	98

7.2.1.2 Simulation results

Table 7.2.5 Data BLER performance of DAS/DBS-12b for SBPCE2 and Padded HOM

		Data BLER @ 10% [dB]			
MCS	Intf. Scen.	DAS/DBS-12b for SBPCE2		DAS/DBS-12b for Padded HOM	
		TU3iFH	TU50nFH	TU3iFH	TU50nFH
DAS-12	AWGN	27.1	29.3	25.6	27.1
	CO	27.1	29.5	25.0	27.0
	ADJ	18.4	20.9	17.3	19.4
DBS-12	AWGN	29.8	34.0	27.8	29.5
	CO	29.2	34.2	26.7	28.9
	ADJ	24.8	29.3	23.1	25.7

Table 7.2.6 Header BLER performance of DAS/DBS-12b for SBPCE2 and Padded HOM

		Header BLER @ 10% [dB]			
MCS	Intf. Scen.	DAS/DBS-12b for SBPCE2		DAS/DBS-12b for Padded HOM	
		TU3iFH	TU50nFH	TU3iFH	TU50nFH
DAS-12	AWGN	11.1	12.6	11.2	12.6
	CO	12.9	15.0	12.7	14.9
	ADJ	2.1	3.6	0.9	2.8
DBS-12	AWGN	12.9	14.5	12.1	13.6
	CO	16.0	18.3	15.2	17.4
	ADJ	8.4	10.5	8.8	10.8

Table 7.2.7 Data and Header BLER performance improvement of DAS/DBS-12b for Padded HOM

		DAS/DBS-12b for SBPCE2 Vs Padded HOM			
MCS	Intf. Scen.	Data improvements		Header improvements	
		TU3iFH	TU50nFH	TU3iFH	TU50nFH

DAS-12	**AWGN**	1.5	2.2	-0.1	0.0
	CO	2.1	2.5	0.2	0.1
	ADJ	1.1	1.5	1.2	0.8
DBS-12	**AWGN**	2.1	4.5	0.8	0.9
	CO	2.5	5.3	0.8	0.9
	ADJ	1.7	3.6	-0.4	-0.3

The evaluation results are tabulated from table7.2.5 to table 7.2.7. It is noted that in all scenarios investigated the data performance of DAS-12b and DBS-12b for Padded HOM outperforms DAS-12b and DBS-12b for SBPCE2. The gain is as much as 1.2 dB to 5.3 dB. The header performance of DAS-12b and DBS-12b for Padded HOM is also better than DAS-12b and DBS-12b for SBPCE2 in most scenarios except for several degradations less than 0.5dB.

7.2.3 Discussion

The PAR reduction is not considered in this study for the time being. Based on the observations in the [7.1.1], the impact of the PAR reduction between the two candidate techniques is less than 0.6dB (except DAS/DBS-11) by comparing the no-clipping with the clipping values in all scenarios investigated. Therefore the impact of PAR reduction on the performace of these two candidates is very limited in low MCSs.

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7.2.4 References

[7.2.1] GP-101348, "DAS-12 and DBS-12 puncturing and interleaving", source Telefon AB LM Ericsson, ST-Ericsson SA, TSG GERAN #47

[7.2.2] GP-101770, "Burst format of Improved Precoded EGPRS2 DL", source Huawei Technologies Co., Ltd., GERAN #48

[7.2.3] GP-101850, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA, TSG GERAN #48

[7.2.4] GP-111131, "MCS re-design for DAS-12 and DBS-12", source Huawei Technologies Co., Ltd., GERAN #51.

[7.2.5] GP-111132 “Performance comparison: Padded HOM and SBPCE2”, source Huawei Technologies Co., Ltd., GERAN#51.

[7.2.6] GP-101852, “DAS-12b and DBS-12b Burst Formatting”, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN#48.

8 Comparison of PC EGPRS2 with EGPRS2 performance

The performance objective of the SPEED study item, presented in sub-clause 4.2, is to significantly improve data throughput performance when introducing PC EGPRS2 as compared to realistic EGPRS2 performance.

To determine if the performance objective has been fulfilled the following sub-clauses present realistic PC EGPRS2 and EGPRS2 throughput performance. To enable a fair comparison between the candidate PC EGPRS2 techniques and EGPRS2 each performance evaluation shall present a set of simulation parameters, common to all evaluations.

Common assumptions to all evaluations further assume:

- PC EGPRS2 specific:
  - Inclusion of blind modulation detection.
- EGPRS2 specific:

  - PAR, see [8.10]:
    - NSR, 8PSK: 3.2 dB
    - NSR, 16QAM: 4.7 dB
    - NSR, 32QAM: 5.1 dB
    - HSR, QPSK 3.4 dB
    - HSR, 16QAM 5.3 dB
    - HSR, 32QAM 5.7 dB
- EGPRS2 and PC EGPRS2:
  - C/I and Es/N0 is defined by the measured power of the carrier and the interferer/thermal noise , measured over the useful part of the burst (including payload symbols and training sequence symbols), see Section 4 in [8.11], with the exception for DTS-2 where the C/I is defined to the strongest Co-channel interferer (C/I1), see Annex L in [8.11] .
  - If the backoff used is assumed to be equal to or greater than the PAR of the signal the PA model can be omitted from the set of Tx impairments used, if this is not believed to impact performance. PAR is in this context referring to a deterministic maximum of the signal, i.e. for PC EGPRS2 hard clipping or similar functionality is needed to ensure the limited signal dynamics.
  - Realistic Tx and Rx impairments.
  - Backoff included in sensitivity limited evaluations.
  - Header performance included in the evaluation and header design for PC EGPRS2 evaluated in 6.2.4 by incremental redundancy or other means.
  - Inclusions of AFC functionality for EGPRS2. For PC EGPRS2 the use of AFC is TBD.

The interference modulation shall be according to the interferer modulation definitions specified in TS45.005 for EGPRS2.

8.1 Simulation assumptions

Table 8.1-1 presents the common simulation parameter settings used in the evaluation of the various PC EGPRS2 techniques as well as in the EGPRS2 evaluations.

Table 8.1-1 EGPRS2 simulation assumptions.

Parameters	**EGPRS-2A (ST-Ericsson SA)**	**EGPRS-2A & B (Renesas Mobile)**
Frequency band	900 / 1800 MHz	900 / 1800 MHz
Tx filter	LinGMSK	LinGMSK
Rx filter BW	260 kHz	EGPRS2A: 270 kHz EGPRS2B: 325 kHz
Blind Detection of modulation	On	On
MCSs	DAS5-12	MCS1-4 DAS5-12 DBS5-12
Backoff	3.2 dB 8-PSK 4.7 dB for 16QAM 5.1 dB for 32QAM	0 dB for GMSK 3.4 dB for QPSK 3.2 dB for 8PSK 4.7/5.3 dB for 16QAM(NSR/HSR) 5.1/5.7 dB for 32QAM(NSR/HSR)
Frames	20000 for Tu3nFH; 10000 otherwise	25000 for Tu3nFH; 10000 otherwise
Tx/Rx impairments	Tx/Rx	Tx/Rx
- Phase noise [degrees (RMS)]	0.8/1.2	0.8 / 1.2
- I/Q gain imbalance [dB]	0.1/0.2	0.1 / 0.2
- I/Q phase imbalance [deegrees]	0.2/2.0	0.2 / 2.0
- DC offset [dB]	-45/-40	-45 / -40
- PA model	Off / Off	- / -
- Frequency error [Hz]	-/25	- / AFC enabled

Table 8.1-1a Precoded EGPRS2 simulation assumptions.

	SBPCE2 (Ericsson AB)	**SBPCE2 (Huawei)**	**Padded HOM (Huawei)**
Frequency band	900 / 1800 MHz	900 MHz	900 MHz
Tx filter	LinGMSK	LinGMSK	LinGMSK
Rx filter BW	SBPCE2-A: 280 kHz LCSBPCE2-B: 340 kHz	level-A: 270 kHz level-B: 325 kHz	level-A: 270 kHz level-B: 325 kHz
Blind Detection of modulation	On [8-3]	On [8.1]	On [8.1]
MCSs	MCS1-4 (non-precoded), DAS5-9, DAS-10b/11b/12b v2 [8-4] and [8.8] DBS5-9, DBS-10b/11b/12b v2 [8-5] and [8.8]	DAS5-9, DAS10b-12b v2 DBS5-9, DBS10b-12b v2	DAS5-9, DAS10b-12b v2 DBS5-9, DBS10b-12b v2
Backoff	0 dB: GMSK 4 dB: QPSK/8PSK 6 dB: 16QAM/32QAM/64QAM	4 dB: QPSK/8PSK 6 dB: 16QAM /32QAM /64QAM	4 dB: QPSK/8PSK 6 dB: 16QAM /32QAM /64QAM
Frames	20000 for Tu3nFH; 10000 otherwise	20000 for Tu3nFH; 5000 otherwise	20000 for Tu3nFH; 5000 otherwise
Tx/Rx impairments	Tx/Rx	Tx/Rx	Tx/Rx
- Phase noise [degrees (RMS)]	0.8/1.2	0.8/1.2	0.8/1.2
- I/Q gain imbalance [dB]	0.1/0.2	0.1/0.2	0.1/0.2
- I/Q phase imbalance [deegrees]	0.2/2.0	0.2/2.0	0.2/2.0
- DC offset [dB]	-45/-40	-45/-40	-45/-40
- PA model	On [8-7]/ -	Off / -	Off / -
- Frequency error [Hz]	-/25	-/25	-/25
Burst mapping	[8-6]	According to [8.5] and [8.13],	According to [8.12]
TSC placement	SBPCE2-A: According to [8-6] LCSBPCE2-B: According to [8-5]	According to [8.6] and [8.8]	According to [8.12]
TSC length	26	DAS5-9, DBS5-9: 26 DAS10b-12b,DBS10b-12b: 16	DAS5-9: 17 DAS10b-12b: 14 DBS5-9: 18 DBS10b-12b: 13
Burst length	SBPCE2-A: According to [8-7] LCSBPCE2-B: According to [8-5]	level-A: 144 level-B: 162	level-A: 144 level-B: 162
Mixed Mode Modulation	SBPCE2-A: Not used LCSBPCE2-B: According to [8-5]	level-A: Not used level-B: According to [8.3]	No
PAR reduction method	According to [8-7] and [8.9], with hard clipping and soft clipping	hard clipping and soft clipping, According to [8.7], rotation based PAR reduction is not considered	hard clipping and soft clipping, According to [8.7]
Achieved PAR	4 dB: QPSK/8PSK 6 dB: 16QAM/64QAM	4 dB: QPSK/8PSK 6 dB: 16QAM/32QAM/64QAM	4 dB: QPSK/8PSK 6 dB: 16QAM/32QAM/64QAM
CP length	SBPCE2-A: 6 LCSBPCE2-B: 15	level-A: 6 level-B: 15	level-A: 6 level-B: 15
ICI Equalization	No	No	No

*Zero-padded pattern is utilized for Padded HOM unless otherwise stated.

8.2 EGPRS2-A and PC EGPRS2-A Low band performance

8.2.1 Sensitivity limited performance

Table 8.2-1 presents the EGPRS2-A and PC EGPRS2-A Low band sensitivity limited performance in tabulated form for each evaluated technique. The performance figures in table 8.2-1 are defined as throughput in kbps/TS. The backoff presented in table 8.2-1 shall be taken into account when performing the evaluation. This shall be manifested as a shift of the Es/N0 values.

Table 8.2-1 Sensitivity limited performance

	Es/N0 [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
Static						
	10	23.7	20.7	24.8	3.1	6.6
	15	32.8	32.8	32.8	31.3	32.4
	20	62.3	54.3	54.4	62.7	64.7
	25	81.6	80.9	81.6	82.2	89.3
	30	98.4	98.3	98.4	98.3	98.4
	35	98.4	98.4	98.4	98.3	98.4
	40	98.4	98.4	98.4	98.4	98.4
TU50nFH						
	10	12.1*	7.1	12.8*	4.1	5.5
	15	23.2	20.6	24.9	18	18.9
	20	34.4	31.4	38.0	32.9	30.3
	25	57.3	46.5	50.9	57.3	58.5
	30	77.5	61.0	68.1	76.8	77.9
	35	94.9	76.9	80.6	91.8	93.9
	40	98.1	87.2	87.4	97.4	97.5
HT100nFH						
	10	11.2*	1.5	12.1*	1.6	3
	15	21.6	13.9	22.3	17.8	19.4
	20	32.4	25.6	32.1	30.7	31.2
	25	49.5	38.6	42.6	46.6	47.5
	30	66.4	46.9	49.9	60.3	63.1
	35	79.8	51.5	57.2	63.8	65.9
	40	82.8	52.5	62.9	64.5	70.6
RA250nFH						
	10	11.3*	8.4	12.7	2.5	3.6
	15	23.3	23.3	27.9	18.3	19.1
	20	31.1	31.7	31.2	28.1	28.4
	25	42.1	45.0	45.8	38.8	35.5
	30	55.8	58.9	55.4	48.2	49.5
	35	60.4	66.1	61.4	53.5	54
	40	61.7	75.0	66.1	55	55.2

* The throughput is achieved by non-precoded MCS1-4.

Figures 8.2-1 to 8.2-4 depict the results in table 8.2-1.

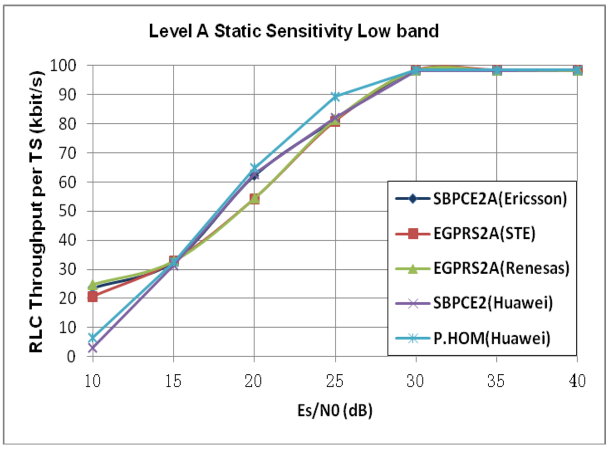

Figure 8.2-1: Static channel

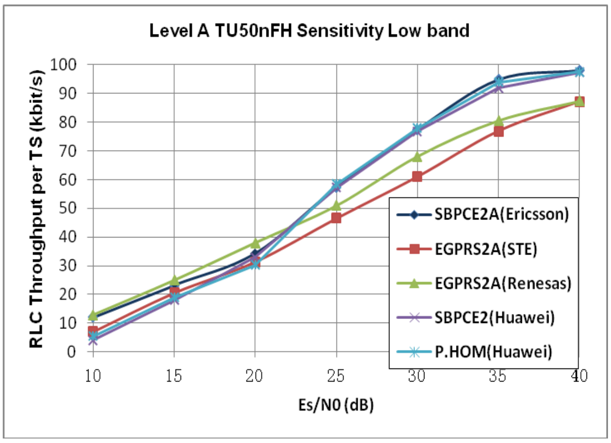

Figure 8.2-2: TU50noFH channel

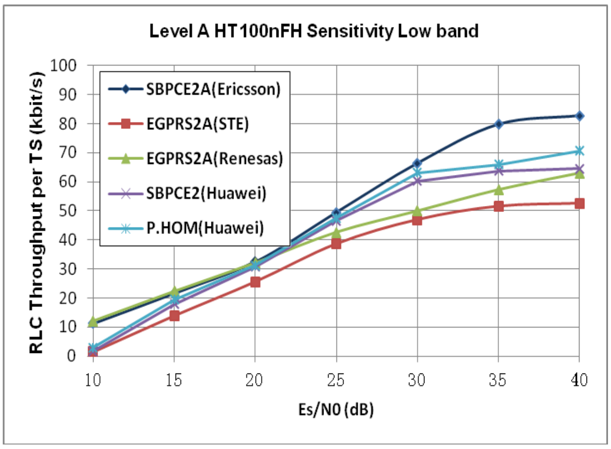

Figure 8.2-3: HT100noFH channel

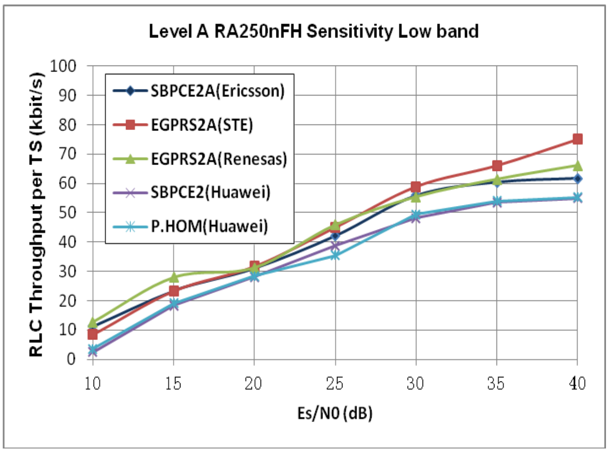

Figure 8.2-4: RA250noFH channel

8.2.2 Co channel interference performance

Table 8.2-2 presents the EGPRS2-A and PC EGPRS2-A Low band Co channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.2-2 are defined as throughput in kbps/TS.

Table 8.2-2 Co channel interference performance

	C/I [dB]	SBPCE2-A (Ericsson)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
Tu3noFH						
	10	25.8	18.4	23.0	19.2	19.1
	15	46.4	31.8	39.3	38.7	40.2
	20	66.3	46.7	59.3	59.8	60.8
	25	86.3	70.8	81.2	78.5	79.5
	30	95.6	88.2	93.2	93.3	93.8
	35	98.0	96.5	96.9	97.3	97.5
	40	98.3	98,1	97.9	98	98.2
Tu3iFH						
	10	26.3	19.9	27.1	21.6	21.8
	15	52.8	34.3	43.7	41.6	42.3
	20	74.6	50.1	61.8	65.6	67.1
	25	93.7	65.8	78.7	84.5	84.9
	30	98.1	83.7	92.2	97.4	97.4
	35	98.4	95.9	97.2	98.3	98.3
	40	98.4	97.9	98.3	98.4	98.4
RA250nFH						
	10	24.2	20.7	23.2	20.4	20.1
	15	39.5	35.8	37.4	30.2	28.7
	20	53.1	50.2	49.3	39.8	41.1
	25	59.7	60.9	60.0	50.1	51.2
	30	61.6	68.3	63.7	53.8	54.5
	35	62.0	77.3	67.4	55	55.3
	40	62.3	79.7	69.3	55.2	55.7

Figures 8.2-5 to 8.2-7 depict the results in table 8.2-2.

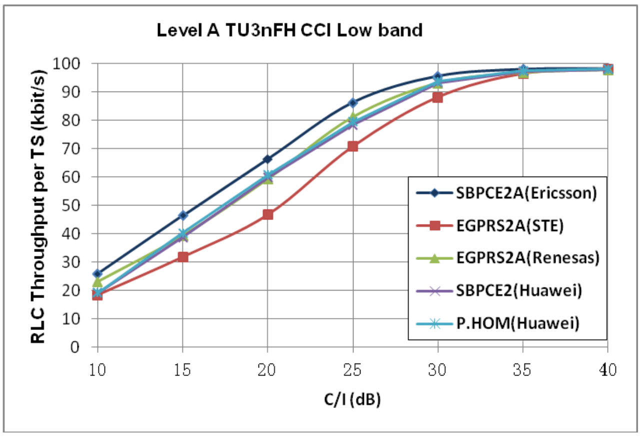

Figure 8.2-5: TU3noFH channel

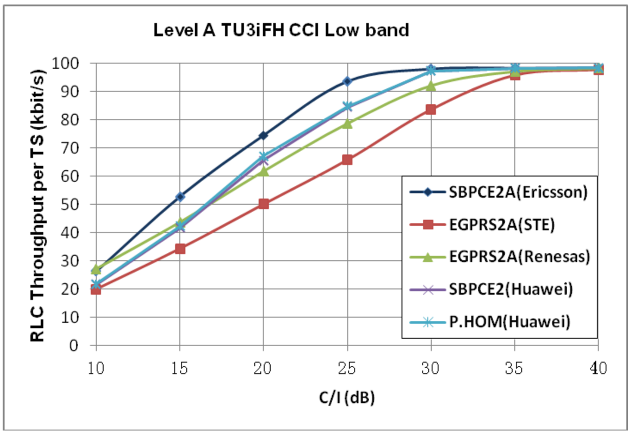

Figure 8.2-6: TU3idFH channel

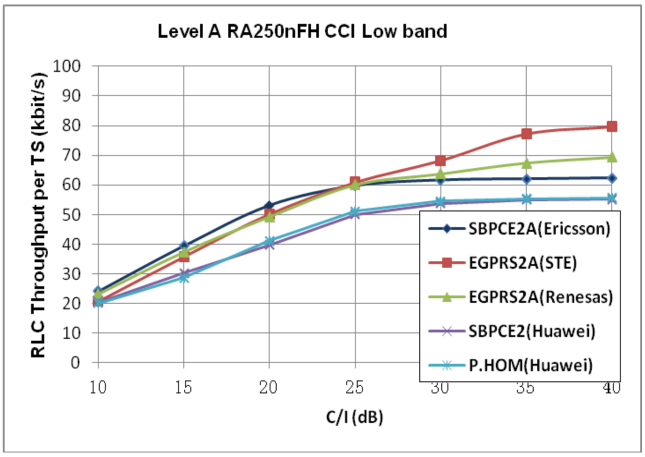

Figure 8.2-7: RA250noFH channel

8.2.3 Adj channel interference performance

Table 8.2-3 presents the EGPRS2-A and PC EGPRS2-A Low band Adj channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.2-3 are defined as throughput in kbps/TS.

Table 8.2-3 Adj channel interference performance

	C/I [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
TU3noFH						
	-10	60.2	12.0	34.7	2.2	14.1
	-5	67.5	22.1	47.9	9.3	22.9
	0	76.8	31.6	67.7	18.2	34.6
	5	82.2	45.2	84.3	41.6	46.4
	10	91.2	57.0	92.4	58.4	62.4
	15	95.0	78.4	95.9	74.5	77.1
	20	97.4	90.3	97.1	88.8	91.7
	25	98.2	95.2	97.8	95.7	96.8
	30	98.4	97.0	98.0	97.7	98
TU3iFH						
	-10	64.5	12.8	39.4	0.6	15.3
	-5	75.2	24.5	52.9	7.5	25.5
	0	81.1	32.0	70.7	19.6	38
	5	91.1	46.4	83.2	42.1	50.4
	10	97.0	55.0	93.6	63.6	67.5
	15	98.1	72.4	96.6	79.4	82.6
	20	98.4	88.5	97.8	95.3	96.9
	25	98.4	94.6	98.2	98	98.3
	30	98.4	96.9	98.3	98.3	98.3
RA250nFH						
	-10	37.6	13.1	31.8	0.7	14.2
	-5	41.0	23.8	45.3	8.3	22.7
	0	45.1	31.2	51.3	19.6	29.7
	5	50.2	40.2	56.7	30.1	34.9
	10	53.2	43.6	60.3	40.5	43.9
	15	56.4	57.2	61.8	50.4	52
	20	58.8	64.6	63.4	53.8	54.5
	25	60.1	72.7	66.0	54.8	55.4
	30	60.7	76.5	68.7	55.2	55.8

Figures 8.2-8 to 8.2-10 depict the results in table 8.2-3.

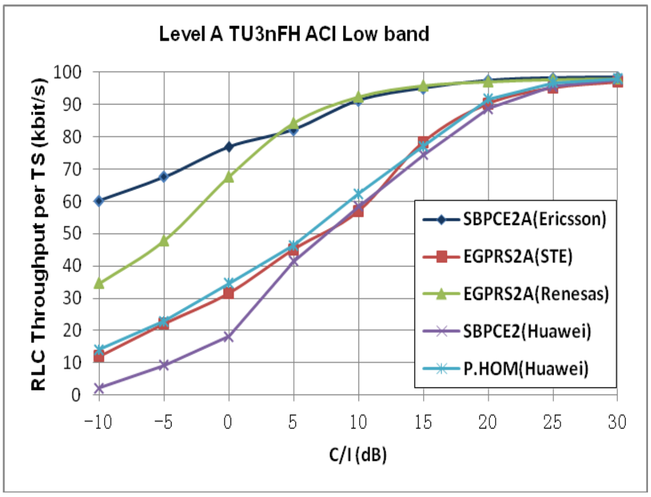

Figure 8.2-8: TU3noFH channel

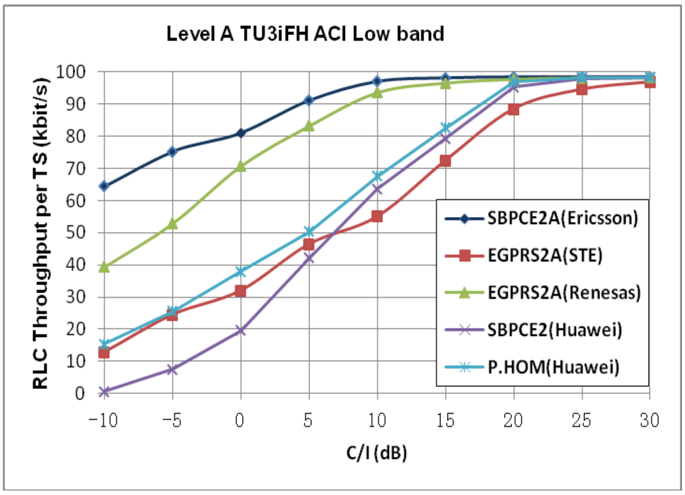

Figure 8.2-9: TU3idFH channel

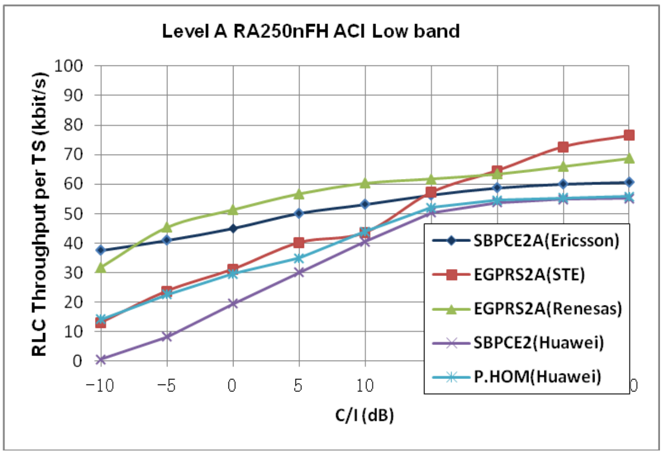

Figure 8.2-10: RA250noFH channel

8.2.4 DTS-2 performance

Table 8.2-4 presents the EGPRS2-A and PC EGPRS2-A Low band Co channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.2-4 are defined as throughput in kbps/TS. In the C/I definition I equals the power of the strongest interferer.

Table 8.2-4 DTS-2 performance

	C/I [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
TU50noFH						
	10	20.9	14.0	21.5	17.5	18.6
	15	37.3	26.0	37.8	31.2	32.1
	20	59.1	38.8	54.5	59.2	59.8
	25	78.3	55.5	72.9	77.9	77.9
	30	95.6	70.6	80.5	92.2	93.6
	35	97.9	82.6	88.2	97.2	97.5
	40	98.3	88.1	90.5	97.7	97.9

Figure 8.2-11 depicts the results in table 8.2-4.

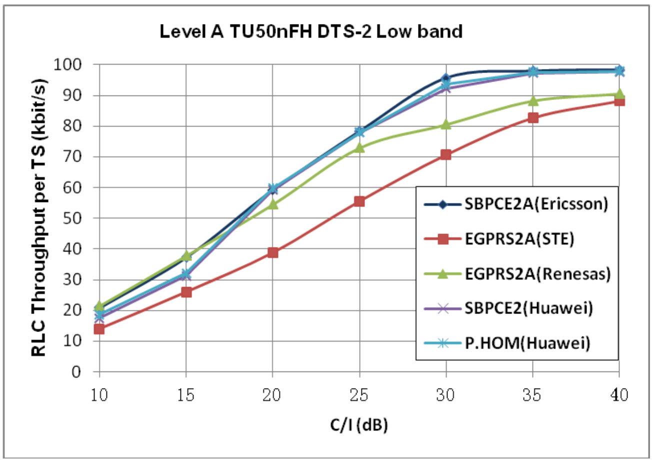

Figure 8.2-11: TU50noFH channel

8.3 EGPRS2-B and PC EGPRS2-B Low band performance

8.3.1 Sensitivity limited performance

Table 8.3-1 presents the EGPRS2-B and PC EGPRS2-B Low band sensitivity limited performance in tabulated form for each evaluated technique. The performance figures in table 8.3-1 are defined as throughput in kbps/TS. The backoff presented in table 8.2-1 shall be taken into account when performing the evaluation. This shall be manifested as a shift of the Es/N0 values.

Table 8.3-1 Sensitivity limited performance

	Es/N0 [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
Static					
	10	20.1	29.2	15.1	3.2
	15	39.2	29.6	33.4	28.2
	20	58.4	64.3	59.1	58.7
	25	88	88.8	78.3	89
	30	113.9	108.8	104.4	107.9
	35	118.4	118.4	115.5	117.3
	40	118.4	118.4	117.8	117.8
Tu50noFH					
	10	12.1*	13.0	4.1	3.8
	15	19.6	25.1	16.9	17.3
	20	37.8	29.5	36.3	34.3
	25	54.5	52.8	51.8	50.8
	30	82.3	76.1	74.8	80.5
	35	103.2	94.2	87.4	91.3
	40	115.7	105.1	101.1	103.9
HT100noFH					
	10	11.2*	12.1*	1.5	3.7
	15	16.8	20.6	14.6	15.8
	20	34.1	28.9	23.2	28.5
	25	51.2	39.1	38.1	43.2
	30	58.6	50.8	42.1	55.3
	35	86.8	55.7	42.9	57.9
	40	90.1	57.5	43.2	59.6
RA250noFH					
	10	11.3*	14.2	4.5	3.7
	15	19.2	25.2	18.4	15.6
	20	35.2	30.5	32.1	28.6
	25	53.6	50.6	42	41.1
	30	58.5	62.0	45.4	43.7
	35	60.2	82.3	48.5	47.4
	40	68.3	89.3	49.5	48.4

* The throughput is achieved by non-precoded MCS1-4.

Figures 8.3-1 to 8.3-4 depict the results in table 8.3-1.

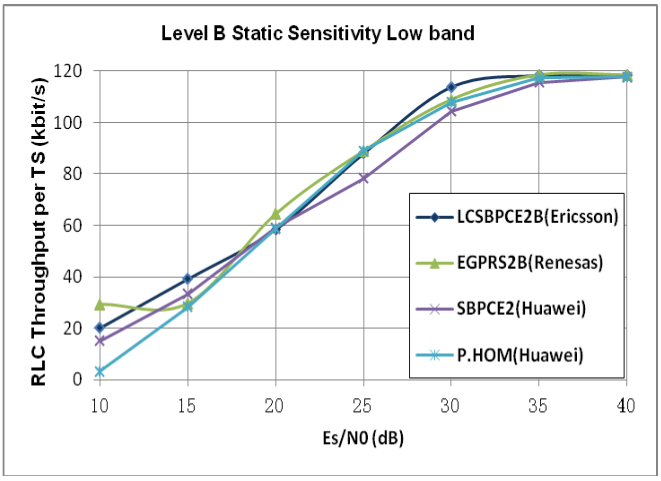

Figure 8.3-1: Static channel

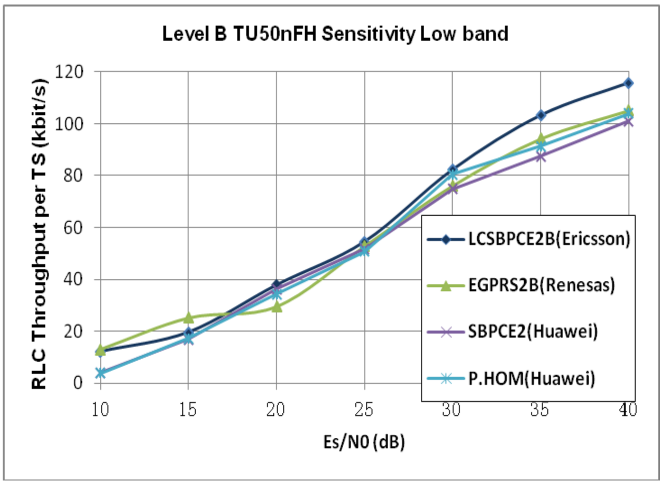

Figure 8.3-2: TU50noFH channel

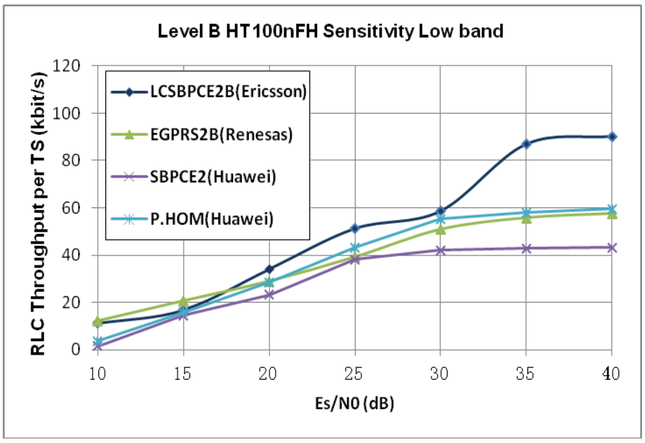

Figure 8.3-3: HT100noFH channel

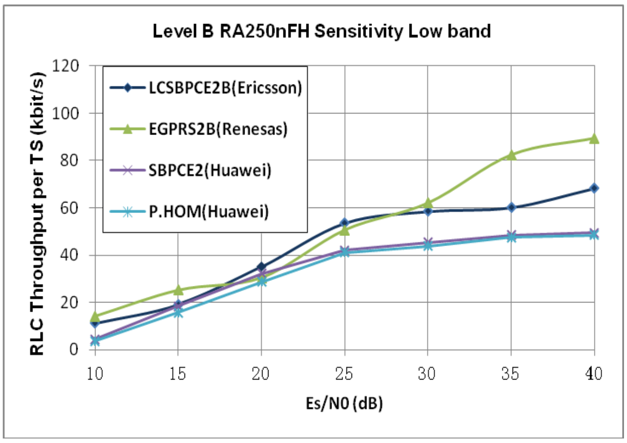

Figure 8.3-4: RA250noFH channel

8.3.2 Co channel interference performance

Table 8.3-2 presents the EGPRS2-B and PC EGPRS2-B Low band Co channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.3-2 are defined as throughput in kbps/TS.

Table 8.3-2 Co channel interference performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
TU3noFH					
	10	26.2	25.8	20.8	18.1
	15	43	45.2	36.4	34.6
	20	67.7	70.6	54.8	57.6
	25	92.3	96.7	78.8	80.4
	30	110.5	112.3	98.3	99.5
	35	117	117.3	108.7	111.1
	40	118.2	117.9	113.4	114.6
TU3iFH					
	10	27.7	25.5	19	18.8
	15	47	49.7	39.5	38.9
	20	76.6	75.8	56.1	56
	25	101.8	97.0	85.1	85.3
	30	116.6	113.2	103.1	104.4
	35	118.3	117.8	111.6	115.1
	40	118.4	118.3	115.1	117.2
RA250noFH					
	10	23.1	23.9	19.3	18.7
	15	40.6	43.0	35.4	33.1
	20	52.9	59.2	42.3	41.7
	25	61.4	78.8	46.8	45.1
	30	69.6	89.1	49.1	47.9
	35	72.1	92.7	49.7	48.6
	40	72.8	93.6	49.9	48.9

Figures 8.3-5 to 8.3-7 depict the results in table 8.3-2.

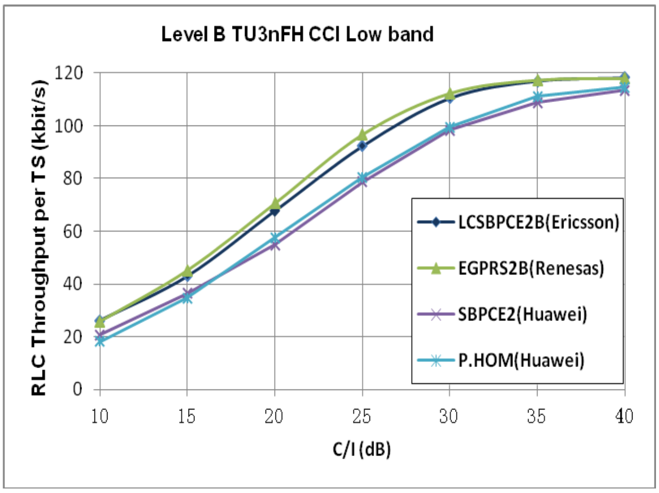

Figure 8.3-5: TU3noFH channel

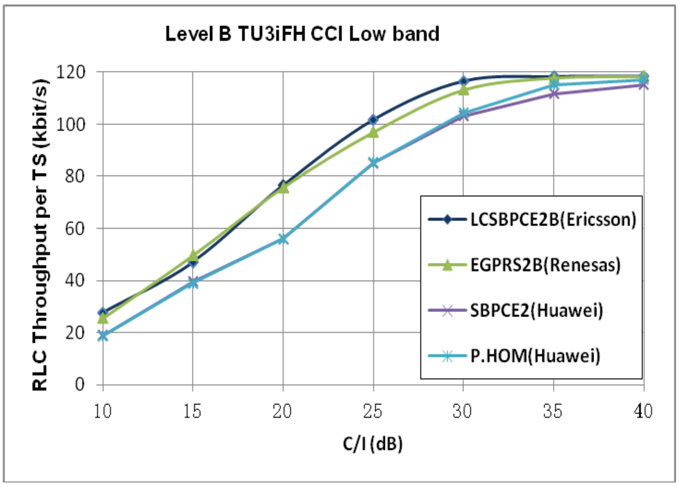

Figure 8.3-6: TU3idFH channel

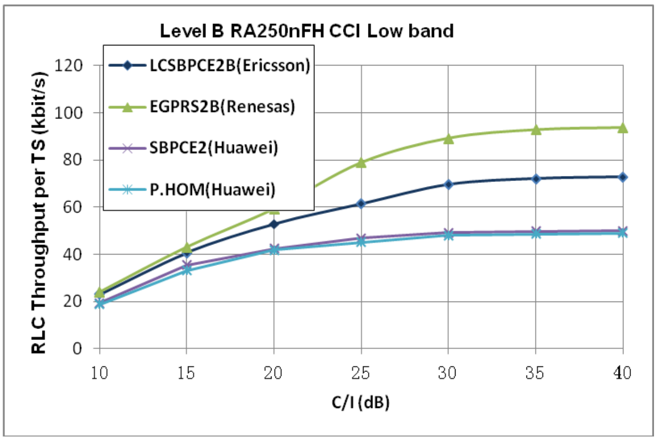

Figure 8.3-7: RA250noFH channel

8.3.3 Adj channel interference performance

Table 8.3-3 presents the EGPRS2-B and PC EGPRS2-B Low band Adj channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.3-3 are defined as throughput in kbps/TS.

Table 8.3-3 Adj channel interference performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
Tu3noFH					
	-10	41.4	26.3	0.2	6.2
	-5	53.3	40.8	3.1	14.4
	0	74.2	58.8	10.9	24.6
	5	84.2	83.1	24.6	38.2
	10	89.5	101.5	37.9	52.2
	15	102.3	112.6	61.7	70.2
	20	109.1	116.3	81.3	84.6
	25	114.6	117.4	90.6	99
	30	116.9	117.9	103.9	107.7
Tu3iFH					
	-10	44.4	28.3	0.04	3.9
	-5	56.9	40.0	1.2	14.9
	0	83.5	63.5	8.6	23.9
	5	88.4	84.2	22.2	41.1
	10	101.2	102.7	41.1	56.4
	15	108.2	113.1	62.3	74.6
	20	116.5	116.6	86.4	88
	25	118.2	117.9	95.9	104.9
	30	118.4	118.2	107.5	113.5
RA250noFH					
	-10	37.0	27.3	0.06	4.5
	-5	41.5	35.9	1.7	16.3
	0	43.3	50.6	11.8	22.5
	5	48.5	60.2	22.9	37.3
	10	52.1	71.8	38.7	42.4
	15	54.1	81.5	42	45.4
	20	55.2	87.2	47.2	47.9
	25	61.4	90.5	49.1	48.7
	30	67.7	92.0	49.2	48.9

Figures 8.3-8 to 8.3-10 depict the results in table 8.3-3.

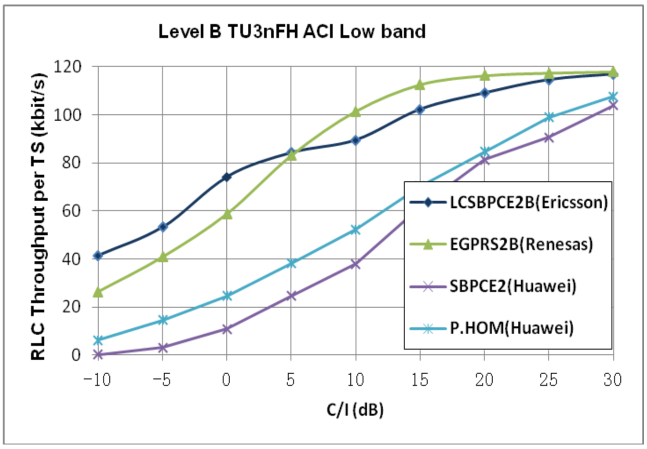

Figure 8.3-8: TU3noFH channel

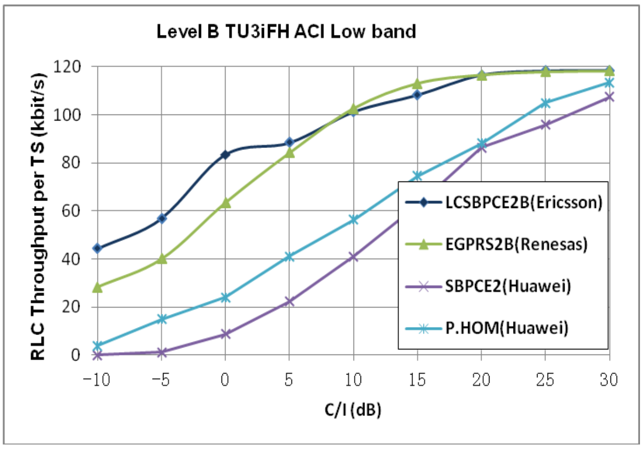

Figure 8.3-9: TU3idFH channel

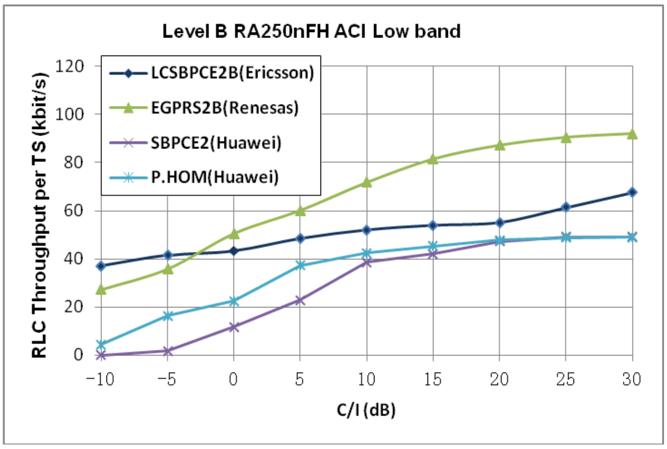

Figure 8.3-10: RA250noFH channel

8.3.4 DTS-2 performance

Table 8.3-4 presents the EGPRS2-B and PC EGPRS2-B Low band DTS-2 channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.3-4 are defined as throughput in kbps/TS. In the C/I definition I equals the power of the strongest interferer.

Table 8.3-4 DTS-2 performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)	SBPCE2 (Huawei)	Padded HOM (Huawei)
Tu50noFH					
	10	16.5	21.2	14.1	15.9
	15	36.3	38.5	29.5	32.2
	20	52.8	60.6	46.8	50.1
	25	80.5	84.5	73.6	77
	30	102.2	100.4	87.8	88.1
	35	115.1	109.0	101.6	103.3
	40	117.8	111.6	105.6	106.1

Figure 8.3-11 depicts the results in table 8.3-4.

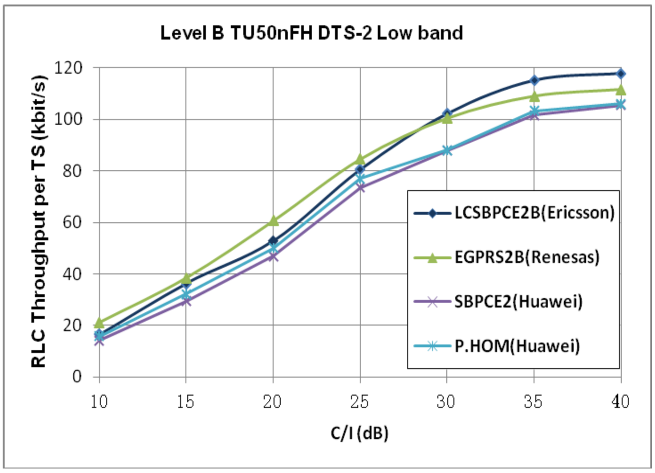

Figure 8.3-11: TU50noFH channel

8.4 EGPRS2-A and PC EGPRS2-A High band performance

8.4.1 Sensitivity limited performance

Table 8.4-1 presents the EGPRS2-A and PC EGPRS2-A High band sensitivity limited performance in tabulated form for each evaluated technique. The performance figures in table 8.3-1 are defined as throughput in kbps/TS. The backoff presented in table 8.2-1 shall be taken into account when performing the evaluation. This shall be manifested as a shift of the Es/N0 values.

Table 8.4-1 Sensitivity limited performance

	Es/N0 [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)
TU50noFH				
	10	12*	2.6	13.0*
	15	24.2	18.1	22.8
	20	34.7	30.7	32.2
	25	57.5	41.6	49.8
	30	75.2	55.5	61.4
	35	85.2	64.1	73.4
	40	93.2	72.6	79.3
HT100nFH				
	10	10.7*	0.5	11.5*
	15	20.3	9.8	18.7
	20	31	21.0	28.2
	25	42.9	29.7	36.4
	30	54.3	39.8	40.8
	35	60.4	42.4	44.1
	40	62.0	43.0	46.0

* The throughput is achieved by non-precoded MCS1-4.

Figures 8.4-1 to 8.4-2 depict the results in table 8.4-1.

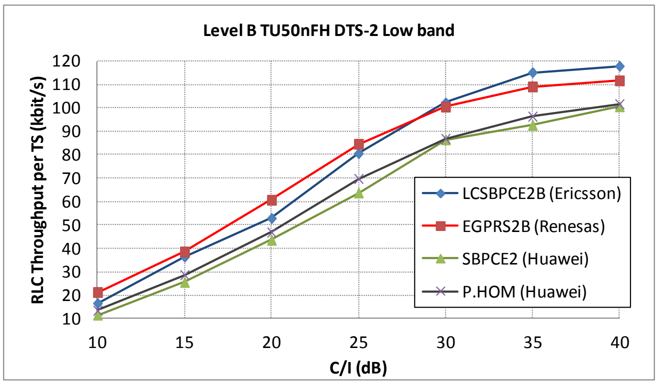

Figure 8.4-1: TU50noFH channel

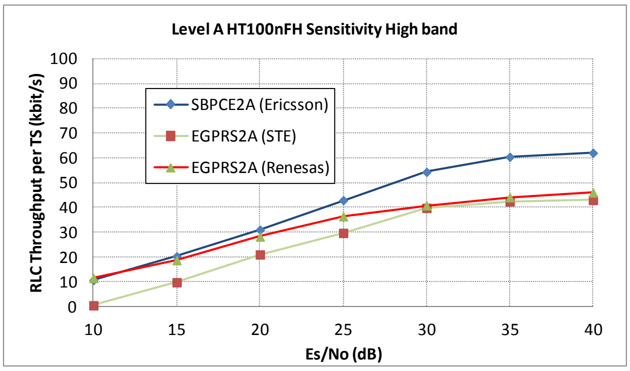

Figure 8.4-2: HT100noFH channel

8.4.2 Co channel interference performance

Table 8.4-2 presents the EGPRS2-A and PC EGPRS2-A High band Co channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.4-2 are defined as throughput in kbps/TS.

Table 8.4-2 Co channel interference performance

	C/I [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)
Tu50noFH				
	10	25.7	19.7	26.1
	15	48.3	32.9	41.8
	20	67.4	46.5	58.5
	25	80.1	61.4	70.9
	30	91.9	69.0	78.4
	35	94.6	89.4	80.3
	40	95.2	92.2	80.7

Figure 8.4-3 depicts the results in table 8.4-2.

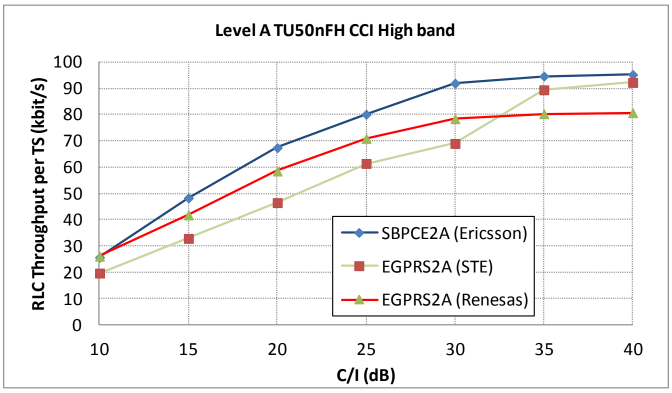

Figure 8.4-3: TU50noFH channel

8.4.3 Adj channel interference performance

Table 8.4-3 presents the EGPRS2-A and PC EGPRS2-A High band Adj channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.4-3 are defined as throughput in kbps/TS.

Table 8.4-3 Adj channel interference performance

	C/I [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)
TU50nFH				
	-10	57.5	12.1	36.2
	-5	63.5	23.2	49.8
	0	67.7	31.6	57.2
	5	75.8	40.8	64.3
	10	79.0	50.5	75.0
	15	80.6	57.9	78.2
	20	88.7	63.1	79.7
	25	92.5	70.9	80.4
	30	94.0	74.3	80.7

Figure 8.4-4 depicts the results in table 8.4-3.

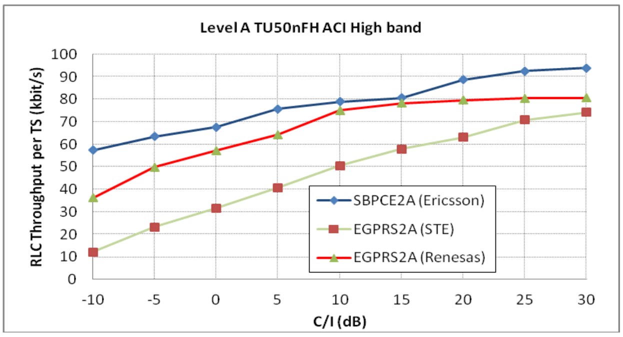

Figure 8.4-4: TU3noFH channel

8.4.4 DTS-2 performance

Table 8.4-4 presents the EGPRS2-A and PC EGPRS2-A High band DTS-2 interference performance in tabulated form for each evaluated technique. The performance figures in table 8.4-4 are defined as throughput in kbps/TS. In the C/I definition I equals the power of the strongest interferer.

Table 8.4-4 DTS-2 performance

	C/I [dB]	SBPCE2-A (Ericsson AB)	EGPRS-2A (ST-Ericsson SA)	EGPRS-2A (Renesas Mobile)
TU50nFH				
	10	21.6	14.1	22.3
	15	38.6	26.7	38.8
	20	60.2	39.5	52.3
	25	77.1	53.1	64.9
	30	86.3	63.6	76.8
	35	92.8	74.1	80.0
	40	94.7	76.4	80.6

Figure 8.4-5 depicts the results in table 8.4-4.

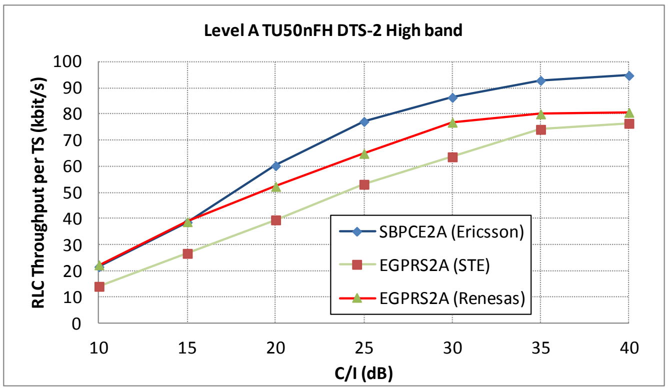

Figure 8.4-5: TU50noFH channel

8.5 EGPRS2-B and PC EGPRS2-B High band performance

8.5.1 Sensitivity limited performance

Table 8.5-1 presents the EGPRS2-B and PC EGPRS2-B High band sensitivity limited performance in tabulated form for each evaluated technique. The performance figures in table 8.5-1 are defined as throughput in kbps/TS. The backoff presented in table 8.2-1 shall be taken into account when performing the evaluation. This shall be manifested as a shift of the Es/N0 values

Table 8.5-1 Sensitivity limited performance

	Es/N0 [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)
TU50nFH			
	10	12*	13.0*
	15	19.8	25.4
	20	39	29.5
	25	54.9	52.9
	30	79.7	66.5
	35	88.2	88.3
	40	104.9	98.2
HT100nFH			
	10	10.7*	11.5*
	15	16	20.1
	20	30.4	28.2
	25	42.5	35.8
	30	55.6	44.3
	35	60.5	50.4
	40	68.7	52.4

* The throughput is achieved by non-precoded MCS1-4

Figures 8.5-1 to 8.5-2 depict the results in table 8.5-1.

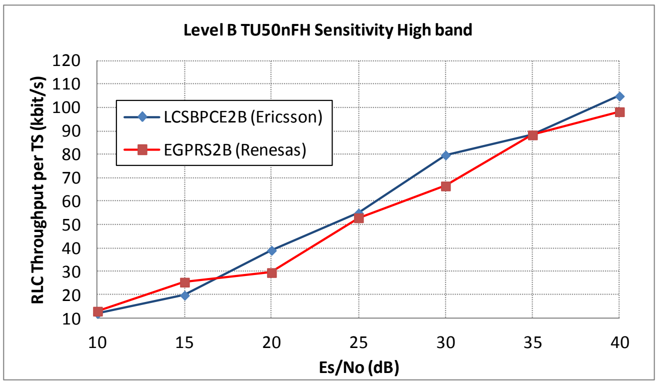

Figure 8.5-1: TU50noFH channel

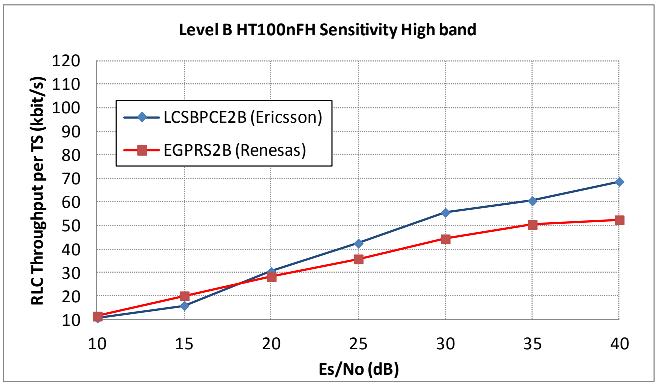

Figure 8.5-2: HT100noFH channel

8.5.2 Co channel interference performance

Table 8.5-2 presents the EGPRS2-B and PC EGPRS2-B High band Co channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.5-2 are defined as throughput in kbps/TS.

Table 8.5-2 Co channel interference performance

Table 8.5-2 Co channel interference performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)
TU50nFH			
	10	26.7	24.7
	15	45.0	46.9
	20	69.5	64.4
	25	88.0	85.5
	30	104.8	97.9
	35	109.6	101.7
	40	111.8	103.9

Figure 8.5-3 depicts the results in table 8.5-2.

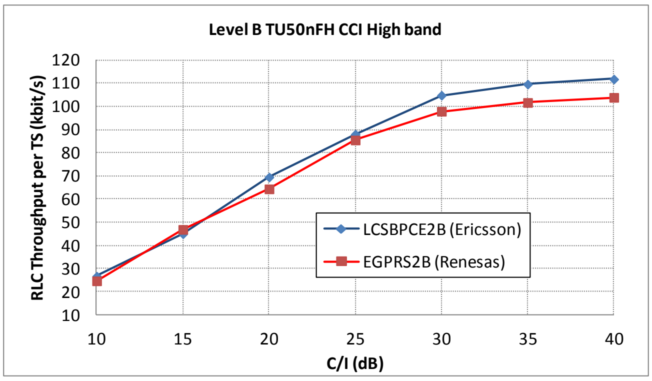

Figure 8.5-3: TU50noFH channel

8.5.3 Adj channel interference performance

Table 8.5-3 presents the EGPRS2-B and PC EGPRS2-B High band Adj channel interference performance in tabulated form for each evaluated technique. The performance figures in table 8.5-3 are defined as throughput in kbps/TS.

Table 8.5-3 Adj channel interference performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)
TU50nFH			
	-10	43.6	28.2
	-5	44.7	38.3
	0	59.7	58.4
	5	78.8	66.4
	10	85.7	83.8
	15	87.6	92.7
	20	94.1	97.2
	25	101.2	99.8
	30	105.2	101.4

Figure 8.5-4 depicts the results in table 8.5-3.

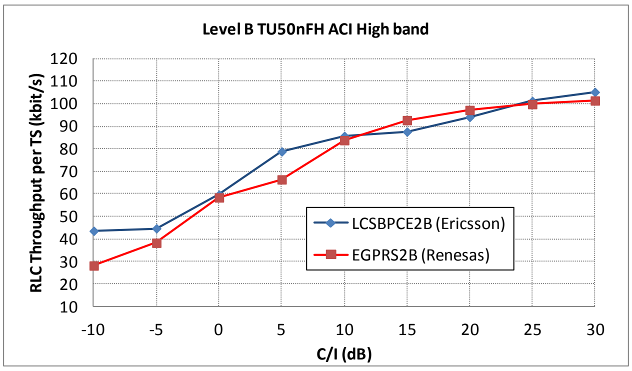

Figure 8.5-4: TU50noFH channel

8.5.4 DTS-2 performance

Table 8.5-4 presents the EGPRS2-B and PC EGPRS2-B High band DTS-2 interference performance in tabulated form for each evaluated technique. The performance figures in table 8.5-4 are defined as throughput in kbps/TS. In the C/I definition I equals the power of the strongest interferer.

Table 8.5-4 DTS-2 performance

	C/I [dB]	LCSBPCE2-B (Ericsson AB)	EGPRS-2B (Renesas Mobile)
TU50nFH			
	10	16.7	21.3
	15	36.9	39.8
	20	52.9	60.6
	25	77.0	81.1
	30	91.8	94.7
	35	104.8	100.5
	40	108.4	103.0

Figure 8.5-5 depicts the results in table 8.5-4.

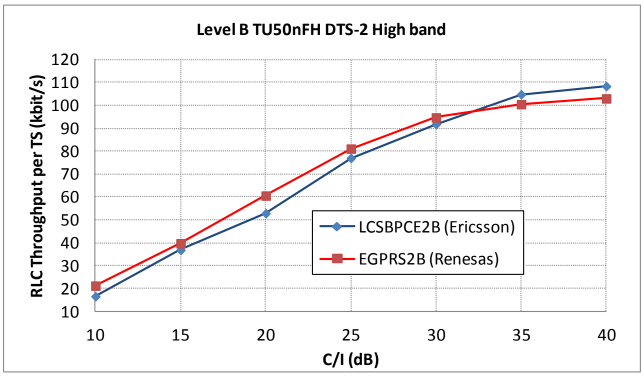

Figure 8.5-5: TU50noFH channel

8.6 References

[8.1] GP-110694, “Blind Modulation Detection of SBPCE2”, source Telefon AB LM Ericsson, ST- Ericsson SA. GERAN#50

[8.2] GP-101852, “DAS-12b and DBS-12b burst formatting”, source Telefon AB LM Ericsson, ST- Ericsson SA. GERAN#48

[8.3] GP-111183, “Complexity reduction of SBPCE2B”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#51

[8.4] GP-101857 ‘pCR45.860 – PCE2 burst formatting’, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN #48

[8.5] GP-101349, “Aspects of Burst Formatting for PC EGPRS2 DL”, source Telefon AB LM Ericson, ST-Ericsson SA, GERAN#47

[8.6] GP-101350, “Training symbol placements in Precoded EGPRS2 DL”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#47

[8.7] GP-110696, “PAR reduction of SBPCE2”, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN#50

[8.8] GP-111642, “TSC Optimization for SBPCE2”, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN #52

[8.9] GP-111641, “Complexity reduction for rotation based PAR reduction”, source Telefon AB LM Ericsson, ST-Ericsson SA, GERAN #52

[8.10] GP-071712, “Tail symbol sequences and Time Masks for EGPRS2”, source Telefon AB LM Ericsson. GERAN#36

[8.11] 3GPP TS 45.005, “Radio transmission and reception”

[8.12] GP-101770, "Burst format of Improved Precoded EGPRS2 DL", source Huawei Technologies Co., Ltd., GERAN #48

[8.13] GP-101850, "Burst mapping of PCE2", source Telefon AB LM Ericsson, ST-Ericsson SA, TSG GERAN #48

9 Summary and conclusions

9.1 Compliance with objectives

In the table below the compliance to the objectives set by the study is summarized for SBPCE2 and Padded HOM.

As the performance presented in Section 8 is extensive, a condensed performance comparison is presented and discussed. The methodology used to derive the figures in the table below is provided in Annex A.

Table 9.1-1. Conclusion summary.

Objectives	**Candidate technique**	
	SBPCE2	**Padded HOM**
Performance objectives		
Improved EGPRS2 throughput *The introduction of Precoded EGPRS2, PC EGPRS2, shall significantly improve data throughput performance as compared to realistic EGPRS2 performance*	Average throughput gains for faded channels, except RA250nFH, unless otherwise stated. Level A: i) when functional blocks of the receiver has been aligned (EGPRS2 and SBPCE2): Sens: 16/27/37/54%. CCI: 28/35/37% DTS-2: 41/44% ACI: 27/36/44%* For RA250nFH: -4/0/1 % * Note that receiver blocks are not aligned in this scenario ii) when functional blocks of the reicever has not been aligned: Sens: 3/12/12/18 % CCI: 9/15/15 % DTS-2: 8/12 % ACI: 0/2/9 % For RA250nFH: -9/-5/2 % iii) Compared to TIGHTER requirements (link level gains [dB]): CCI: 2.6/2.7/2.9/3.0/3.0/… 4.9/5.2/5.6/9.0/10.3 dB ACI: 6.5/6.5/8.1/8.9/10.0/10.0/10.1/… 11.5/16.5/17.5/18/22.0/23.4 dB For RA250nFH: -4.2/+1.5/+4.2/+4.2/+6.6/+8.9 dB	Padded HOM and SBPCE2 share several commonalities, and the design is almost identical for the three highest MCSs in each EGPRS2 set, which are generally used to the largest extent in the LA curves. Based on this observation it is expected that Padded HOM have the same inherent performance as SBPCE2 for DAS-10/11/12 and DBS-10/11/12, and that differences seen between SBPCE2 and Padded HOM performance for these MCSs are due to differences in receiver implementations between the companies contributing to the SPEED study item. For lower MCSs Padded HOM has a design that has positive impact on ACI. Again the differences seen between SBPCE2 and Padded HOM performance for these MCSs are believed to be justified by the differences in receiver implementations.

	Level B: NOTE: Performance with functional blocks of the receiver aligned does not exist: i) with no knowledge of the EGPRS2-B receiver complexity: Sens: +7/9/9/18/20/20%. CCI: -3/1/2/4/10/11% DTS-2: -6/-7/18/20% ACI: -6/-1/0/0/5/14% RA250nFH: -30/-18/-15/-8/-2/9% ii) Compared to TIGHTER requirements (link level gains [dB]): CCI: 1.5/1.9/2.6/2.8/2.8/… 2.8/3.5/4.4/5.5/7.9/8.0 dB For RA250nFH -5.2/+1.4 dB The gains are achieved including typical Tx (no PA modeled for EGPRS2 ref.)/Rx impairments, including impact to blind modulation detection by detecting the circular shift of the TSC and PAR reduction, and without ICI equalization. Also sensitivity figures are compensated by the PAR of the signal.	

Compatibility objectives		
Spectral properties *PC EGPRS2 shall obey the current spectral requirements on spectrum due to modulation and wideband noise and on switching transients of EGPRS2 DL, see 3GPP TS45.005*	Compliant with PAR reduction of 4 dB or 6 dB depending on modulation used when using soft and hard clipping and pre-defined ramp up and ramp down of the burst.	The current spectrum requirement could be obeyed by the combined soft and hard clipping.
Impact on **Legacy services** *The impact of PC EGPRS2 on GSM speech codecs, GPRS, EGPRS and EGPRS2 shall be kept at a minimum.*	No impact has been seen on legacy services when subject to SBPCE2 interference	No impact or small improvement has been seen on legacy services when subject to Padded HOM interference
Cell reselection *Impact on cell reselection performance of mobile stations should be avoided by operation of PC EGPRS2 on the BCCH carrier.*	With rotation based PAR reduction, soft clipping and hard clipping, the achieved PAR is on par with current average power decrease requirement on BCCH.	With soft clipping and hard clipping, the achieved PAR is on par with current average power decrease requirement on BCCH.
USF/PAN multiplexing *Impacts from PAN and USF multiplexing on PC-EGPRS2 and legacy user throughput should be minimized.”*	No or little impact seen on throughput if USF granularity = 4 is used by the network in all multiplexing scenarios investigated.	Conclusion from SBPCE2 is expected to hold also for Padded HOM.
Implementation impact to base station *The introduction of Precoded EGPRS2 in the base station transmitter should change BTS hardware as little as possible*	Overall computational complexity: EGPRS2-A: +43 – +95 % EGPRS2-B: +39 – +130 % +35 - +75%* * without rotation based PAR reduction used by the BTS	Overall computational complexity: EGPRS2-A: +43 – +95 % EGPRS2-B: +35 – +75 %
Implementation impact to mobile station *The introduction of Precoded EGPRS2 in the mobile station receiver should change MS hardware as little as possible. Both impact to stand-alone PC-EGPRS2 platforms and combined EGPRS2 and PC-EGPRS2 platforms shall be considered*	Overall computational complexity: EGPRS2-A: -50/-25%* EGPRS2-B: -40/-20%* * without/with functional block in receiver for ACI suppression	Overall computational complexity: EGPRS2-A: -50% EGPRS2-B: -50%

🟩	Compliant
🟧	Not compliant
🟨	Unclear / FFS
🟦	Expected to be fulfilled

9.2 Conclusions

During the SPEED feasibility study two candidate techniques, Single Block Precoded EGPRS2 – SBPCE2, and Padded Higher Order Modulation – Padded HOM, have been proposed and evaluated against the objectives of the study to significantly improve throughput compared to realistic EGPRS2 performance, while keeping negative impact to the spectral properties, cell reselection, USF/PAN multiplexing to a minimum, and avoiding hardware impact to both base station and mobile station.

Both techniques share several commonalities and the design is almost identical for the three highest MCSs in each EGPRS2 set, which are generally used to the largest extent in the LA curves. For lower MCSs, Padded HOM uses less sub carriers, with positive impact on ACI, while the SBPCE2 design keeps the number of sub carrier the same for all MCSs. The performance difference in absolute performance between the two candidate techniques seen in the TR, see [9.1-2], and [9.1-3], is expected to be related to the receiver design rather than to the difference in the design of the physical layer. As seen in the performance set provided in [9.1-2], small differences are seen between the candidate techniques given an evaluation by the same company.

Given the reasoning above it is expected that Padded HOM fulfils the objectives of the study. The candidate technique that has shown compliance to all objectives is Single Block Precoded EGPRS2.

The throughput gains of the techniques have been evaluated by ideal link adaptation throughput curves on link level in all currently specified scenarios in [9.1-4].

SBPCE2 has shown to give average throughput gains, based on calculations using C/I distributions from network simulations, with realistic performance (i.e. with PAR reduction for SBPCE2, typical Tx/Rx impairments modeled, and impact to Blind detection of modulation taken into account) of:

It should be noted that when functional blocks of the receiver are not aligned or a complexity estimate of the receivers is not available it is not clear if the observed performance differences seen below for Level A and Level B are due to different receiver optimization or the different modulation techniques.

Level A

When functional blocks of the receiver has been aligned/not been aligned between EGPRS2 and SBPCE2:

Sensitivity: +16-54% / +3-18%
CCI: +28-37% / +9-15%
DTS-2: +41-44% / +8-12%
ACI*: +27-44% / +2-9%
* Functional receiver blocks not aligned.

At RA 250 km/h SBPCE2 is usually inferior or on par with EGPRS2 performance.

Level B

When functional blocks of the receiver have not been aligned between EGPRS2 and SBPCE2:

Sensitivity: +7-+20%
CCI: -3 - +11%
DTS-2: -6 – +20%
ACI: -6 - +14%

At RA 250 km/h SBPCE2 is usually inferior or on par with EGPRS2 performance. A significant degradation is seen for ACI case with a degradation of at most 30%.

TIGHTER

Compared to TIGHTER performance, see [9.1-4], SBPCE2 show link level gains [dB] of:

Level A:	
- CCI	3-10 dB
- ACI	6-23 dB
Level B:	
- CCI	1.5-8 dB

NOTE: Low Complexity SBPCE2-B, LC SBPCE2-B, has been used in the evaluation of SBPCE2 for level B.

9.3 References

[9.1-1] GP-120286, “Reference EGPRS2 performance”, Renesas. GERAN#53.

[9.1-2] GP-120134, “Reference performance for Padded HOM and SBPCE2”, Huawei Technologies Co. Ltd. GERAN#53.

[9.1-3] GP-120280, “Update of SBPCE2 Reference performance”, source Telefon AB LM Ericsson, ST-Ericsson SA. GERAN#53.

[9.1-4] 3GPP TS45.005 v. 10.2.0, “Radio transmission and reception”.

Annex A: Throughput comparison

The throughput comparison presented in this Annex is based on the performance of Section 8 in the TR. The performance has been used in the conclusion of the study in Section 9 to state compliance to the performance objective. A SINR probability distribution derived from a 3/9 frequency reuse network simulation has been used to get an estimation of the gain in CCI, Sensitivity and the DTS-2 scneario. For ACI a shift of 6 dB for the distribution has been applied.

In A.2 an example of the methodology used is shown.

In A.3 the tabulated results are presented.

A.1 SINR-distribution

Table A.1. SINR distribution used in CCI, Sensitivity and DTS-2 scenario.

SINR	3/9
10	0,04
15	0,18
20	0,29
25	0,23
30	0,13
35	0,07
40	0,06

Table A.2. SINR distribution used in ACI scenario.

SINR	3/9
-10	0
-5	0,0035
0	0,005
5	0,05
10	0,22
15	0,29
20	0,2
25	0,11
30	0,11

A.2 Average throughput gain calculation

The calculation of average throughput gain used in A.3 is shown below from Table 8.2-2, TU3iFH between SBPCE2-A(Ericsson) and EGPRS2(ST-Ericsson SA).

A weighted average of the gain based on Table A.1 is calculated as input to the tables in A.3:

$32*0.04 + 54*0.18 + 49*0.29 + 42*0.23 + 17*0.13 + 3*0.07 + 1*0.06 = 37.35\ \% \approx 37\ \%$

Table A.2-1

C/I	SBPCE2-A (Ericsson)	EGPRS2-A (ST-Ericsson SA)	Gains [%]
10	26.3	19.9	32
15	52.8	34.3	54
20	74.6	50.1	49
25	93.7	65.8	42
30	98.1	83.7	17
35	98.4	95.9	3
40	98.4	97.9	1

A.3 Average throughput gain

A.3.1 SBPCE2-A

Table A.3-1. Throughput gain [%]. SBPCE2-A vs. EGPRS2-A,

Sensitivity.

EGPRS2-A reference	Low band				High band	
	Static	TU50nFH	HT100nFH	RA250nFH	TU50nFH	HT100nFH
[A.1]	5	16	37	-4	27	54
[A.3]	4	3	12	-5	12	18

Table A.3-2. Throughput gain [%]. SBPCE2-A vs. EGPRS2-A,

CCI.

EGPRS2-A reference	Low band			High band
	TU3nFH	TU3iFH	RA250nFH	Tu50nFH
[A.1]	28	37	0	35
[A.3]	9	15	2	15

Table A.3-3. Throughput gain [%] of SBPCE2-A vs. EGPRS2-A,

ACI.

EGPRS2-A reference	Low band			High band
	TU3nFH	TU3iFH	RA250nFH	TU50nFH
[A.1]	27	36	1	44
[A.3]	0	2	-9	9

Table A.3-4. Throughput gain [%] of SBPCE2-A vs. EGPRS2-A.

DTS-2.

EGPRS2-A reference	Low band	High band
	TU50nFH	TU50nFH
[A.1]	41	44
[A.3]	8	12

A.3.2 SBPCE2-B

Table A.3-5. Throughput gain [%]. SBPCE2-B vs. EGPRS2-B,

Sensitivity.

EGPRS2-B reference	Low band				High band	
	Static	TU50nFH	HT100nFH	RA250nFH	TU50nFH	HT100nFH
[A.3]	2	7	18	-2	9	9
[A.2]	22	20	*	9	20	*
* No performance provided in [A.2]						

Table A.3-6. Throughput gain [%]. SBPCE2-B vs. EGPRS2-B,

CCI.

EGPRS2-B reference	Low band			High band
	TU3nFH	TU3iFH	RA250nFH	Tu50nFH
[A.3]	-3	1	-15	4
[A.4]	2	10	-8	11

Table A.3-7. Throughput gain [%] of SBPCE2-B vs. EGPRS2-B,

ACI.

EGPRS2-B reference	Low band			High band
	TU3nFH	TU3iFH	RA250nFH	TU50nFH
[A.3]	-6	-1	-30	0
[A.4]	0	5	-18	14

Table A.3-8. Throughput gain [%] of SBPCE2-B vs. EGPRS2-B,

DTS-2.

EGPRS2-B reference	Low band	High band
	TU50nFH	TU50nFH
[A.3]	-6	-7
[A.4]	20	18

A.4 References

[A.1] GP-111751, "CR 45.860-0008 EGPRS2 reference performance (Rel-11)", source ST-Ericsson SA. GERAN#52

[A.2] GP-111739, "Inclusion of EGPRS2 reference receiver performance", source Renesas Mobile Europe Ltd

[A.3] GP-120286, "Reference EGPRS2 performance", Renesas. GERAN#53

Annex B:
Change history

Change history							
Date	**TSG #**	**TSG Doc.**	**CR**	**Rev**	**Subject/Comment**	**Old**	**New**
2011-09	GERAN#51	GP-111472			Presented at plenary for approval	1.0.0	11.0.0
2011-11	GERAN#52	GP-111791	0001	1	Evaluating SPEED candidate techniques with Radix size used in LTE	11.0.0	11.1.0
2011-11	GERAN#52	GP-111792	0003	1	Introduction of new MCSs for Padded HOM	11.0.0	11.1.0
2011-11	GERAN#52	GP-111793	0004	1	Update of SBPCE2 reference performance	11.0.0	11.1.0
2011-11	GERAN#52	GP-111862	0005	2	TSC optimization for SPBCE2	11.0.0	11.1.0
2011-11	GERAN#52	GP-111645	0006		Update of complexity estimate and inclusion of complexity reduction for rotation based PAR reduction	11.0.0	11.1.0
2011-11	GERAN#52	GP-111758	0008	1	EGPRS2A reference performance	11.0.0	11.1.0
2011-11	GERAN#52	GP-111864	0010	2	Clarification for simulation assumptions for final evaluation of SPEED	11.0.0	11.1.0
2012-03	GERAN#53	GP-120393	0009	3	Conclusion of SI "Signal Precoding enhancements for EGPRS2 DL" (SPEED)	11.1.0	11.2.0
2012-03	GERAN#53	GP-120286	0011	1	Inclusion of EGPRS2 reference receiver performance	11.1.0	11.2.0
2012-03	GERAN#53	GP-120291	0012	1	Impact on legacy servies for Padded HOM	11.1.0	11.2.0
2012-03	GERAN#53	GP-120131	0013		Low complexity Padded HOM	11.1.0	11.2.0
2012-03	GERAN#53	GP-120132	0014		Impacts on base station and mobile station for Padded HOM	11.1.0	11.2.0
2012-03	GERAN#53	GP-120287	0015	1	Padded HOM concept description	11.1.0	11.2.0
2012-03	GERAN#53	GP-120290	0016	1	Reference performance for Padded HOM and SBPCE2	11.1.0	11.2.0
2012-03	GERAN#53	GP-120292	0017	1	Single Block Precoded EGPRS2 – concept description	11.1.0	11.2.0
2012-03	GERAN#53	GP-120293	0018	1	SBPCE2 Puncturing and interleaving	11.1.0	11.2.0
2012-03	GERAN#53	GP-120294	0019	1	TSC optimization for SBPCE2	11.1.0	11.2.0
2012-03	GERAN#53	GP-120295	0020	1	Update of impact to MS and BTS from SBPCE2	11.1.0	11.2.0
2012-03	GERAN#53	GP-120280	0021	1	Updated reference performance for SBPCE2	11.1.0	11.2.0
2012-03	GERAN#53	GP-120392	0022	1	Addition of performance graphics for PC-EGPRS2 candidates and EGPRS2 reference	11.1.0	11.2.0
2012-05	GERAN#54	GP-120594	0023		Correction of CR implementation from GERAN#53	11.2.0	11.3.0
2012-08	GERAN#55	GP-120916	0024		Update of Reference Performance for Padded HOM and SBPCE2	11.3.0	11.4.0
2012-11	GERAN#56	GP-121252	0026		PAR Reduction for Padded HOM	11.4.0	11.5.0