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TECHNICAL REPORT

**Satellite Earth Stations and Systems (SES);
Technical analysis for the Radio Frequency, Modulation and Coding for Telemetry Command and Ranging (TCR) of Communications Satellites**

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coding, modulation, satellite, telemetry

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Foreword

This Technical Report (TR) has been produced by ETSI Technical Committee Satellite Earth Stations and Systems (SES).

Modal verbs terminology

In the present document "**should**", "**should not**", "**may**", "**need not**", "**will**", "**will not**", "**can**" and "**cannot**" are to be interpreted as described in clause 3.2 of the ETSI Drafting Rules (Verbal forms for the expression of provisions).

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Executive summary

The present document provides the rationale for the revision of the ETSI TCR Standard [i.1].

The need for such revision appeared mainly as a consequence of the evolution of satellites, with new operational frequency bands and configurations like mega-constellations; the progress on spread spectrum technology; the quest for more flexibility in frequency planning and operations on geostationary telecommunication satellite fleets; and novel demands concerning the accommodation of hosted payloads and their segregated operations in those satellites.

Therefore, the existing standard was revised in the following areas: frequency plan, operational phases, hosted payload management application, mega-constellation application, spread spectrum modulation, phase and frequency modulation and finally, coding and interleaving.

The revision has borrowed from the experience acquired by suppliers, operators and space agencies as well as from standards produced in other relevant standardization fora like the European Cooperation for Space Standardization (ECSS) or the Consultative Committee for Space Data Systems (CCSDS).

In summary, the present document provides a sufficient justification of the revision with pointers to annexes and relevant references for those readers seeking further detail.

Introduction

The European Telecommunication Standards Institute (ETSI) established a telemetry, command and ranging (TCR) standard ETSI EN 301 926 [i.2] in 2002.

In recent years telecommunication satellite operators have shown a renewed interest in spread spectrum systems. Their improved performance under interference over the classical frequency modulation (FM) or phase modulation (PM) could ease operations. In mission phases like orbit drift the TCR can be subject to interference while crossing the equatorial orbital arc. With the emergence of electric orbit raising strategies, this phase has actually become much longer making resistance to interference even more relevant.

In addition, hosted payloads are emerging as an attractive business proposition for telecommunication satellite operators. The capability to have direct telecommand (TC) and telemetry (TM) communications with segregated radiofrequency (RF) carriers and avionics could off-load to some extent hosted payload operations from satellite operations. In addition, it could limit the interface impact between the satellite and the hosted payload respective ground and space segments.

Meanwhile the technology of transponders/receivers as well as ground modems have evolved to support spread spectrum modulation as well as frequency flexibility on FM/PM. For instance, the ability to acquire and track a spread spectrum signal under high dynamics is no longer considered an issue, in contrast to the times when the first version of the ETSI TCR standard was published. Such capability could simplify satellite TT&C sub-systems by eliminating the need for dual-mode transponders (FM/PM and spread spectrum).

Moreover, mega-constellations for telecommunication missions are currently being developed. To accommodate a very large number of new TCR carriers on existing bands, spread spectrum modulation could offer an efficient solution.

In consideration of all the above, ETSI initiated a work item to revise the standard in 2015. The goal was to match the revised standard with the current and expected capabilities of transponders and ground modems for future telecommunication missions, not only geostationary. ETSI EN 301 926 [i.1] revision has been published in 2017. The present document, therefore, provides a description and justification of this revision.

Readers are encouraged to take into account that the present document builds upon and complements ETSI TR 101 956 [i.3]. ETSI EN 301 926 (V1.3.1) [i.1] has not questioned the existing modulation trade-offs carried out for the definition of the first issue of the standard.

Furthermore, it does not question the concept of Collocated Equivalent Capacity (CEC). However, it is recognized that the addition of channel coding and interleaving will impact CEC by allowing to enhance capacity with respect to the first version of the standard.

Following this introduction, the present document is organized as follows.

Clause 1 outlines the scope of the standard revision.

Clause 2 provides relevant informative references that can assist readers seeking a more detailed understanding of some modifications.

Clause 3 recalls the terms and abbreviations employed throughout the document.

Clause 4 discusses the modifications impacting frequency planning and operational scenarios.

Clause 5 provides a detailed discussion of the key modifications affecting spread spectrum modulation like the extension of the Pseudo-noise code family and others.

Clause 6 addresses key modifications affecting non-spread modulations.

Clause 7 introduces coding and interleaving options added to the standard.

Clause 8 gives a conclusion.

Finally, annexes A and B complement the main body of the document addressing detailed aspects, annex C provides information for future possible work.

1 Scope

The present document provides the rationale for the revision of the ETSI TCR Standard ETSI EN 301 926 [i.1] in the following areas:

- frequency plan;
- operational phases;
- hosted payload management application;
- mega-constellation application;
- spread spectrum modulation;
- phase and frequency modulation; and
- coding and interleaving.

2 References

2.1 Normative references

Normative references are not applicable in the present document.

2.2 Informative references

References are either specific (identified by date of publication and/or edition number or version number) or non-specific. For specific references, only the cited version applies. For non-specific references, the latest version of the referenced document (including any amendments) applies.

NOTE: While any hyperlinks included in this clause were valid at the time of publication, ETSI cannot guarantee their long term validity.

The following referenced documents are not necessary for the application of the present document but they assist the user with regard to a particular subject area.

[i.1] ETSI EN 301 926 (V1.3.1) (10-2017): "Satellite Earth Stations and Systems (SES); Radio Frequency and Modulation Standard for Telemetry, Command and Ranging (TCR) of Communications Satellites".

[i.2] ETSI EN 301 926 (V1.2.1) (06-2002): "Satellite Earth Stations and Systems (SES); Radio Frequency and Modulation Standard for Telemetry, Command and Ranging (TCR) of Geostationary Communications Satellites".

[i.3] ETSI TR 101 956: "Satellite Earth Stations and Systems (SES); Technical analysis of Spread Spectrum Solutions for Telemetry Command and Ranging (TCR) of Geostationary Communications Satellites".

[i.4] CCSDS 231.0-B-x: "TC Synchronization and Channel Coding".

[i.5] CCSDS 131.0-B-x: "TM Synchronization and Channel Coding".

NOTE: CCSDS standards always include the issue number on their numbering system; the parameter 'x' on references [i.4] and [i.5] is understood as the highest published number and therefore latest issue of the standard.

[i.6] IEEE Transactions on Information Theory: "Optimal Binary Sequences for Spread Spectrum Multiplexing", R. Gold, vol. IT-13, no. 1, pp. 619-621, 1967.

[i.7] NASA Publication Contract NAS 5-22546: "TDRSS Telecommunication System PN Code Analysis final Report Addendum", R. Gold, Sept. 1977.

[i.8] Space Network Interoperability Group: "Space Network Interoperable PN Code Libraries", Revision 1, Sept. 1998.

[i.9] CCSDS 230.1-G-x: "TC Synchronization and Channel Coding - Summary of Concept and Rationale".

[i.10] CCSDS 130.1-G-x: "TM Synchronization and Channel Coding - Summary of Concept and Rationale".

NOTE: CCSDS reports always include the issue number on their numbering system; the parameter 'x' on references [i.9] and [i.10] is understood as the highest published number and therefore latest issue of the standard.

[i.11] R. L. Miller, L. J. Deutsch, and S. A. Butman: "On the Error Statistics of Viterbi Decoding and the Performance of Concatenated Codes". JPL Publication 81-9. Pasadena, California: JPL, September 1, 1981.

[i.12] L. Deutsch, F. Pollara, and L. Swanson: "Effects of NRZ-M Modulation on Convolutional Codes Performance". TDA Progress Report 42-77, January-March 1984 (May 15, 1984): 33-40.

[i.13] Space Network Users' Guide (SNUG). Revision 10. 450-SNUG. Greenbelt, Maryland: NASA Goddard Space Flight Center, August 2012.

[i.14] I. Aguilar Sánchez et al.: "The Navigation and Communication Systems for the Automated Transfer Vehicle", proceedings of the IEEE 49th Vehicular Technology Conference, Vol. 2, pp. 1187-1192, 1999.

[i.15] G. Lesthievent et al.: "Concatenating the convolutional (7,1/2) code with the BCH in TED mode with CRC for improved TC link in the CNES Myriad satellites family", Paper SLS-NGU-10-CNES01, CCSDS Next Generation Uplink Working Group, London (UK), October 2010.

[i.16] CCSDS 231.1-O-1: "Short Block Length LDPC Codes for TC Synchronization and Channel Coding".

[i.17] ECSS-E-ST-50-05C Rev. 2: "Space Engineering - Radio frequency and modulation", European Cooperation for Space Standardization, 4 October 2011.

[i.18] NIST: "Advanced Encryption Standard (AES)", Federal Information Processing Standard Publication 197, United States, November 26, 2001.

[i.19] NIST Special Publication 800-38A: "Recommendation for Block Cipher Modes of Operation: Methods and Techniques", United States, December 2001.

3 Definition of terms and abbreviations

3.1 Terms

For the purposes of the present document, the following terms apply:

binary channel: binary communications channel (BPSK has 1 channel, QPSK has 2 channels)

channel symbol rate: rate of binary elements, considered on a single wire, after FEC coding and channel allocation

NOTE: See Figures 2, 3 and 4. This applies only to multi-channel modulations, thus to spread spectrum QPSK modes and not to PM/FM modes.

Co-located Equivalent Capacity (CEC): number of collocated satellites that can be controlled with a perfect power balanced link between the ground and the satellite

Code Division Multiple Access (CDMA): technique for spread-spectrum multiple-access digital communications that creates channels through the use of unique code sequences

Command Link Transmission Unit (CLTU): telecommand protocol data structure providing synchronization for the codeblock and delimiting the beginning of user data

NOTE: See [i.4], section 4 for further details.

data rate: total number of uncoded data bits per second after packet and frame encoding

NOTE: See Figures 1 to 4. This is the data rate used in link budgets in ETSI TR 101 956 [i.3].

Direct Sequence Spread Spectrum (DSSS): form of modulation where a combination of data to be transmitted and a known code sequence (chip sequence) is used to directly modulate a carrier, e.g. by phase shift keying

symbol rate: rate of binary elements, considered on a single wire, after FEC coding

NOTE: See Figures 1 to 4.

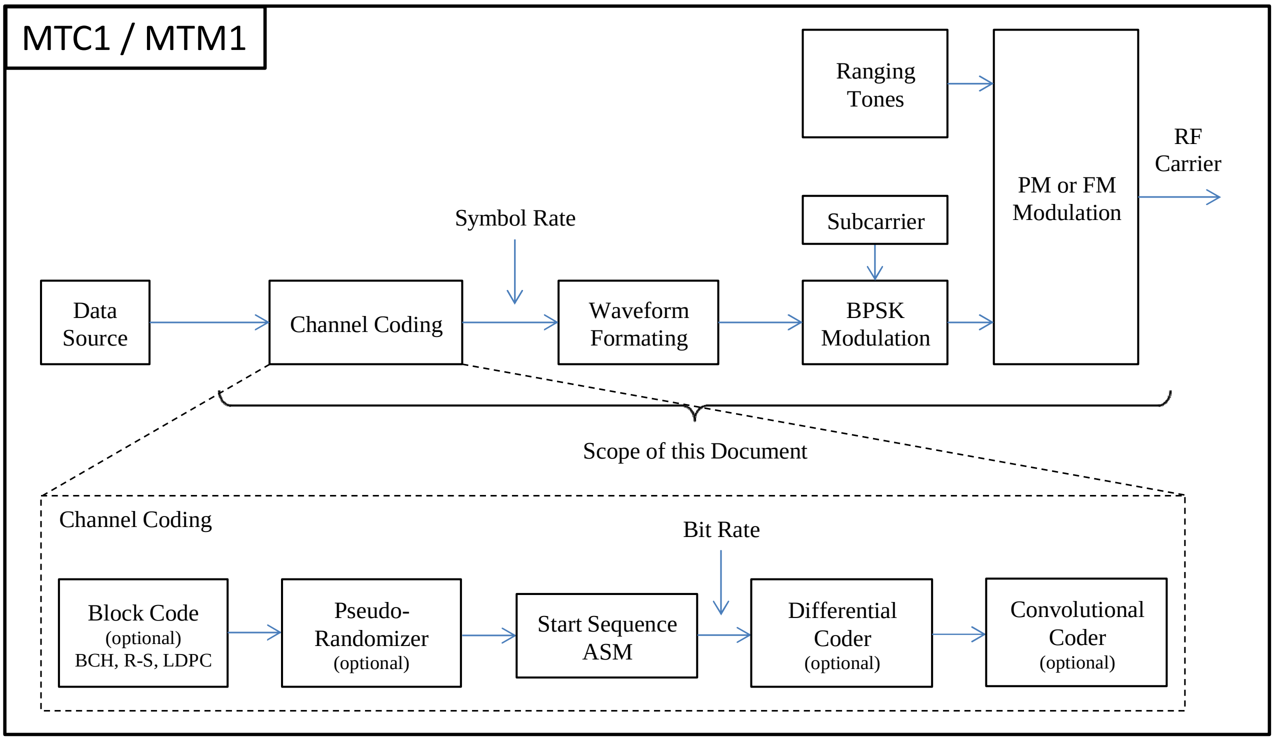

Figure 1: Functional stages of transmit chain for FM/PM modulation (MTC1/MTM1)

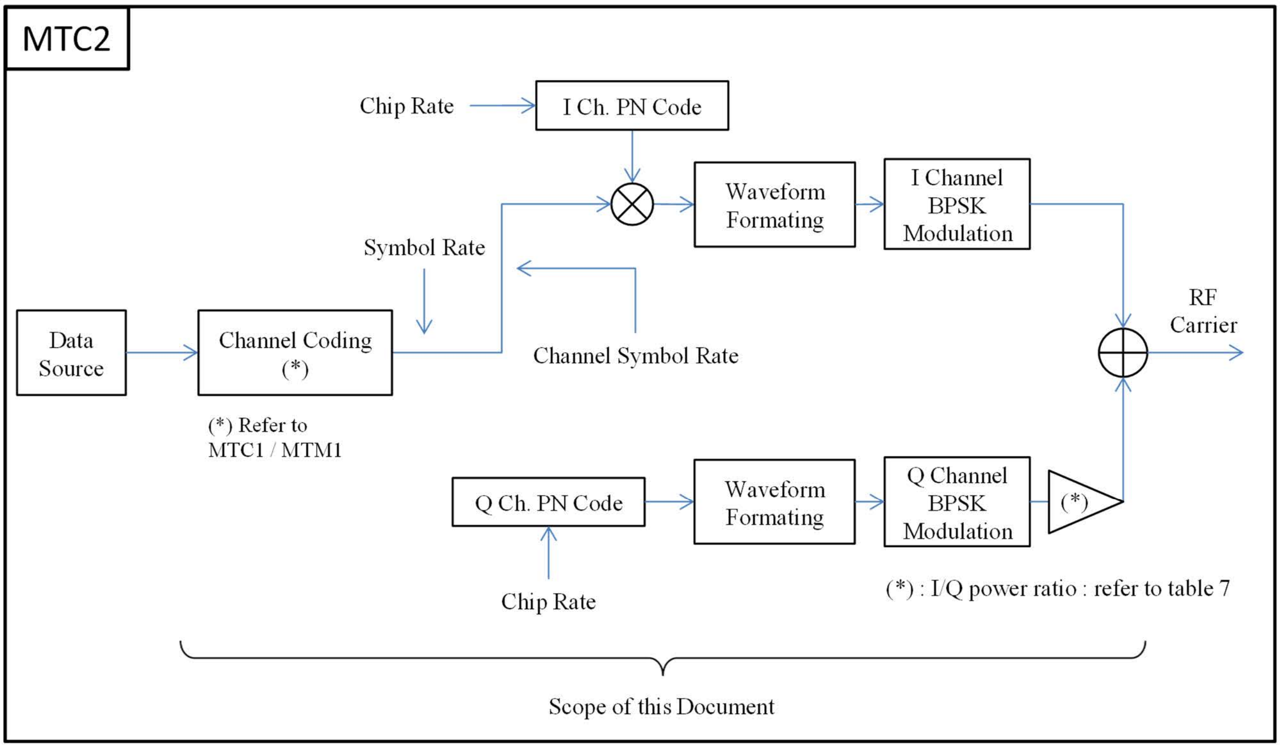

Figure 2: Functional stages of transmit chain for spread spectrum modulation MTC2

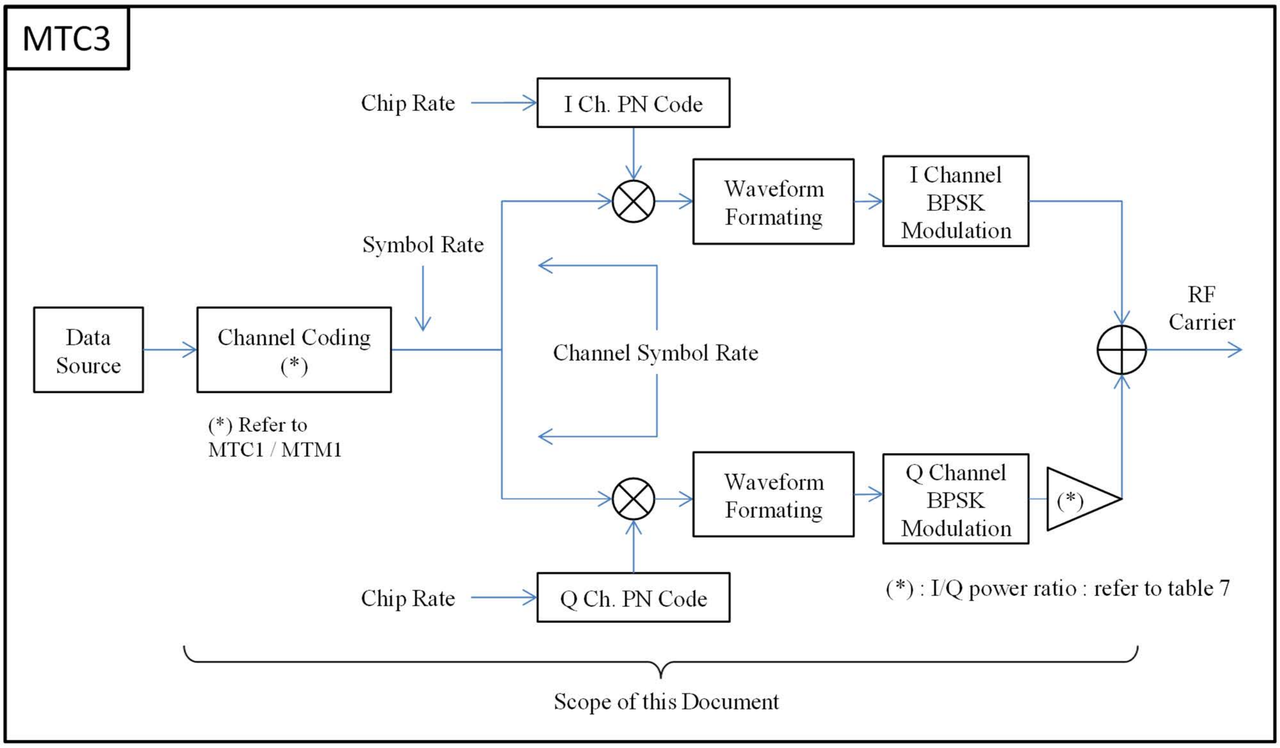

Figure 3: Functional stages of transmission chain for spread spectrum modulation MTC

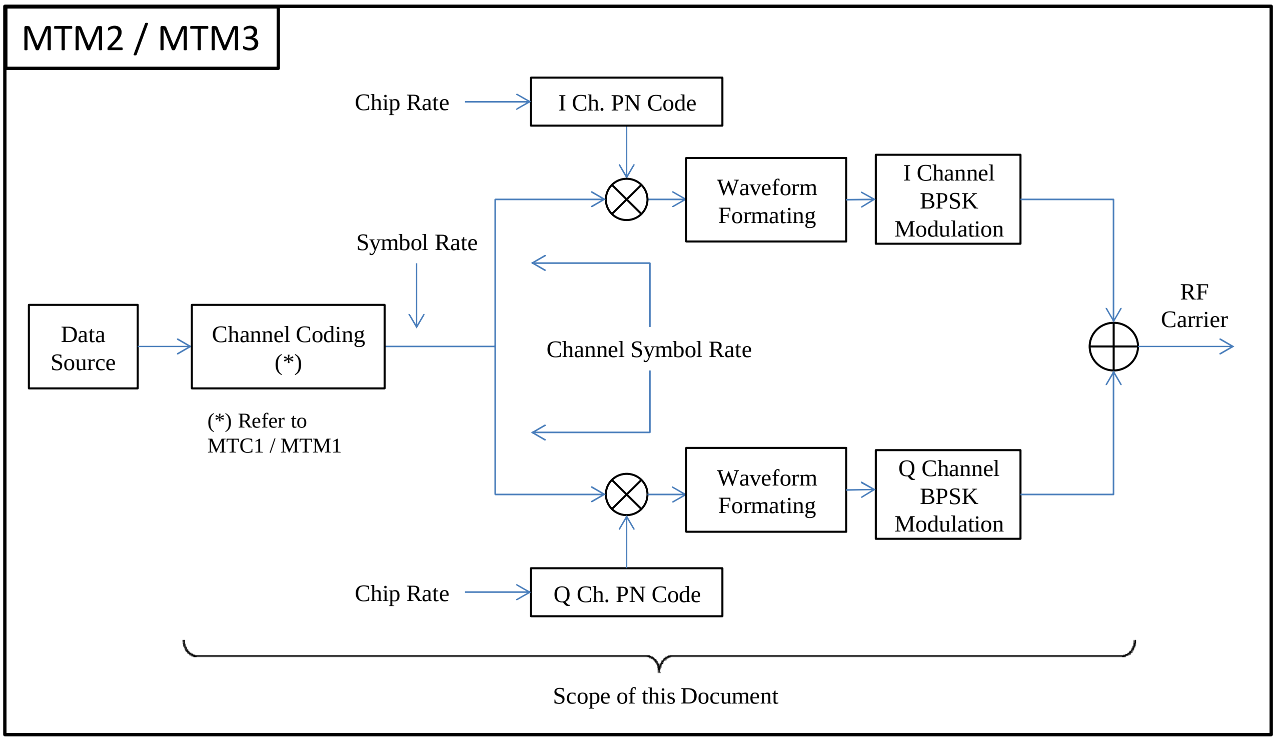

Figure 4: Functional stages of transmission chain for spread spectrum modulation MTM2/MTM3

3.2 Abbreviations

For the purposes of the present document, the following abbreviations apply:

AES	Advanced Encryption Standard
ATV	Automated Transfer Vehicle
BCH	Bose-Chaudhuri-Hocquenghem
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CCSDS	Consultative Committee for Space Data Systems
CDMA	Code Division Multiple Access
CEC	Co-located Equivalent Capacity
CLTU	Command Link Transmission Unit
CMM	Carrier Modulation Modes
CNES	Centre National d'Etudes Spatiales
CRC	Cyclic Redundancy Check
CW	Continuous Wave
DC	Direct Current
DSSS	Direct Sequence Spread Spectrum
ECSS	European Cooperation for Space Standardization
FEC	Forward Error Correction
FM	Frequency Modulation
GSID	Ground-to-Satellite Interface Specification
GSO	Geo-Stationary Orbit
GTO	Geostationary Transfer Orbit
I	In-phase
LDPC	Low Density Parity Check
LEO	Low Earth Orbit
LEOP	Launch and Early Orbit Phase
MAI	Multiple Access Interference
MTC1	TeleCommand Mode 1
MTC2	TeleCommand Mode 2
MTC3	TeleCommand Mode 3
MTM1	TeleMetry Mode 1

MTM2	TeleMetry Mode 2
MTM3	TeleMetry Mode 3
NASA	National Aeronautics and Space Administration (USA)
NIST	National Institute of Standards and Technology
NRZ	Non-Return Zero
NRZ-L	Non Return to Zero-Level
NRZ-M	Non Return to Zero-Mark
NTIA	National Telecommunications Industry Association
PDF	Probability Density Function
PED	Phase Error Detector
PLL	Phase Locked Loop
PLOP	Physical Layer Operating Procedures
PM	Phase Modulation
PN	Pseudo Noise
Q	Quadrature
QPSK	Quaternary Phase Shift Keying
RF	Radio Frequency
RS	Reed-Solomon
SEC	Single Error Correction
SER	Symbol Error Rate
SNR	Signal to Noise Ratio
SSTO	Single-Stage-To-Orbit
TC	TeleCommand
TCR	Telemetry, Command and Ranging
TDRSS	Tracking and Data Relay Satellite System (NASA)
TED	Triple Error Detection
TM	TeleMetry
TT&C	Telemetry, Tracking and Command
VLSI	Very Large Scale Integration

4 Frequency Planning and Operational Scenarios

4.1 Frequency Planning

4.1.1 Frequency Bands

a) C-band: 5 850 MHz to 6 725 MHz uplink, 3 400 MHz to 4 200 MHz downlink;

b) Ku-band: 12 750 MHz to 14 800 MHz and 17 300 MHz to 18 100 MHz uplink, 10 700 MHz to 12 750 MHz downlink;

c) Commercial Ka-band: 27 500 MHz to 30 000 MHz uplink, 17 700 MHz to 20 700 MHz downlink.

It should be noted that these bands are due to prevailing regulations, not physics. Possible usage of the TCR techniques considered in the present document and ETSI EN 301 926 [i.1] in adjacent bands between 1 GHz and 44 GHz may be envisaged.

4.1.2 Frequency Flexibility

Modern command receivers and telemetry transmitters often utilize fractional N phase-locked loop (PLL) synthesizers for frequency generation. It is possible to generate different output frequencies from a single input reference frequency using this technology. The frequency resolution of such synthesizers is very high, in the order of a few Hertz, which is clearly more than needed for communication satellite transceivers.

For practical purposes, a 100 kHz resolution is recommended. This is based on experience from commercial programs, which already use frequency flexibility and also the fact that the resolution does not need to be higher than the downlink frequency stability requirement of ±5 ppm [i.1], clause 5.1.2.

4.2 Hosted Payloads

Hosted payloads are emerging as an attractive business proposition for telecommunication satellite operators. When additional "Hosted" payloads are embarked on a given "Host" satellite, they may require telemetry and command functionalities beyond what the host's heritage hardware is designed for. The traditional concept for the command and telemetry of hosted payloads relies on a tight share of the classical TCR and spacecraft platform avionics resources by means of data multiplexing mechanisms and the common spacecraft computer and applications software.

An alternative design solution consists on adding dedicated TCR hardware to the hosted payload. Thus, monitoring and control can be carried out independently of the host, decoupled from host avionics and data handling. The resulting reduction of interface control documentation between host and hosted payload is an additional advantage. It is a use case considered as natural for minimally intrusive and spectrally robust TCR systems. Since the hosted payload TCR use case is in effect a separate payload that just happens to be on the same structure as the host, the same TCR standards apply as for the host.

In addition, such architecture can off-load to some extent hosted payload operations from satellite operations. Thus, higher levels of operational autonomy can be reached and offered to hosted payload operators in contrast to the classical operations concept.

4.3 Operation during Launch and Early Orbit

After successful launch and separation all satellites have a Low Earth Orbit Phase (LEOP) while propulsive manoeuvres are performed to get the spacecraft to their intended final orbit. TCR is also needed in this phase.

The main TCR consequences of LEOP are due to geometry. The spacecraft is moving differently (and generally faster) than in the final orbit, the link ranges may be different and, because one may want to do certain manoeuvres at specific parts of the LEOP orbits, time criticality means that the technical TCR solutions used are expected to be robust and reliable.

Except when operating in a GSO, the satellite has a relative velocity to the earth-fixed ground station. This relative velocity (radial velocity) results in a Doppler shift of all frequency components (carrier, subcarrier, chip rate, symbol rate, ranging tone). The maximum Doppler shift has been specified as 22 ppm in ETSI EN 301 926 [i.2], clause 6.1, and this value has been maintained in ETSI EN 301 926 [i.1], clause 5.1.

As a satellite passes by an earth station (worst case is directly over) the Doppler frequency shift changes from positive to negative, passing through zero at closest approach, and thus resulting in a Doppler Rate. The Doppler Rate is maximal when a satellite passes directly over a ground station, and increases as the altitude decreases. The maximum Doppler Rate has been specified as 1,7 ppm/s in ETSI EN 301 926 [i.2], clause 6.1, and this value has been maintained in ETSI EN 301 926 [i.1], clause 5.1.

These limits of 22 ppm for the Doppler shift and 1.7 ppm/s for the Doppler Rate were determined for a Geosynchronous Transfer Orbit and justified in ETSI EN 301 926 [i.2], annex A.

Annex B provides the results of an analysis in different types of transfer orbits or highly elliptical orbits (GTO, Super Synchronous Transfer Orbit, Molniya orbit).

This analysis confirms the validity of the specified limits for all of these orbits under the specified limitation of a true anomaly between 40° and 320°.

4.4 Operation in Other Orbits

The opportunity to open the standard to future telecommunication missions (e.g. mega-constellations) operating at other orbits than the geo-stationary (GSO) was one of the key goals of the revision effort and has been taken into account to some extent. In contrast to classical GSO telecommunication missions, emerging non-GSO systems often have unique architectures. At first glance standardization appears to be at odds with the motivation behind the designers and operators of such systems, which seems to be ad-hoc design and operation.

Furthermore, without some in-depth knowledge of those systems and their operations concept it is considered impractical to attempt standardization.

In consideration of those arguments, ETSI EN 301 926 [i.1] to address this area has taken a pragmatic approach. It has considered two main common characteristics of these novel systems: orbits other than GSO and a very large numbers of satellites (mega-constellations).

The first characteristic is covered with the revision of signal frequency dynamics to support LEOP.

Annex B provides the results of an analysis in different types of Low Earth Orbit (LEO) circular orbits at altitudes as low as 781 km. This analysis confirms the validity of the 22 ppm limit for the Doppler shift and 1,7 ppm/s limit for the Doppler Rate for these orbits. The analysis has not been conducted for LEO circular orbits with an altitude lower than 780 km.

Thus, transponders/transceivers/receivers compliant with the revised ETSI TCR [ref] should be able to accommodate the expected frequency dynamics of those systems.

For the second characteristic, it is considered attractive but also crucial for these systems to provide a sufficiently large number of unique spread spectrum pseudo-noise (PN) codes so that they can share efficiently and effectively the available spectrum with code division multiple access (CDMA) technique.

5 Spread Spectrum Modulation

5.1 Extension of PN Code Family

In order to fulfil the expected future needs of telecommunication satellite fleets as well as missions embarking hosted payloads, the introduction of more flexibility on spread spectrum modulation was desired. The standard PN code family found on the initial version of the standard defines only 85 different codes. It was considered necessary to extend the current PN code family in order to be able to support both larger and smaller CDMA systems. Therefore, both shorter and longer PN sequence lengths have been accommodated.

In contrast to the existing code family, where different types of codes have be defined, all codes of the extended family are defined as dual channel Gold codes. Such codes are currently used only for MTM3 mode. Figure 5 shows the schematic diagram of the generic dual channel (In-phase/Quadrature) Gold code generator. The feedback taps of Register A and C are the same. The feedback taps of register B and register A (or C) should form a preferred pair according to R. Gold's rule [i.6]. Register B is always initialized with 00…01. The initial settings of Register A and C determine the individual codewords from the code defined by the preferred pair of feedback taps.

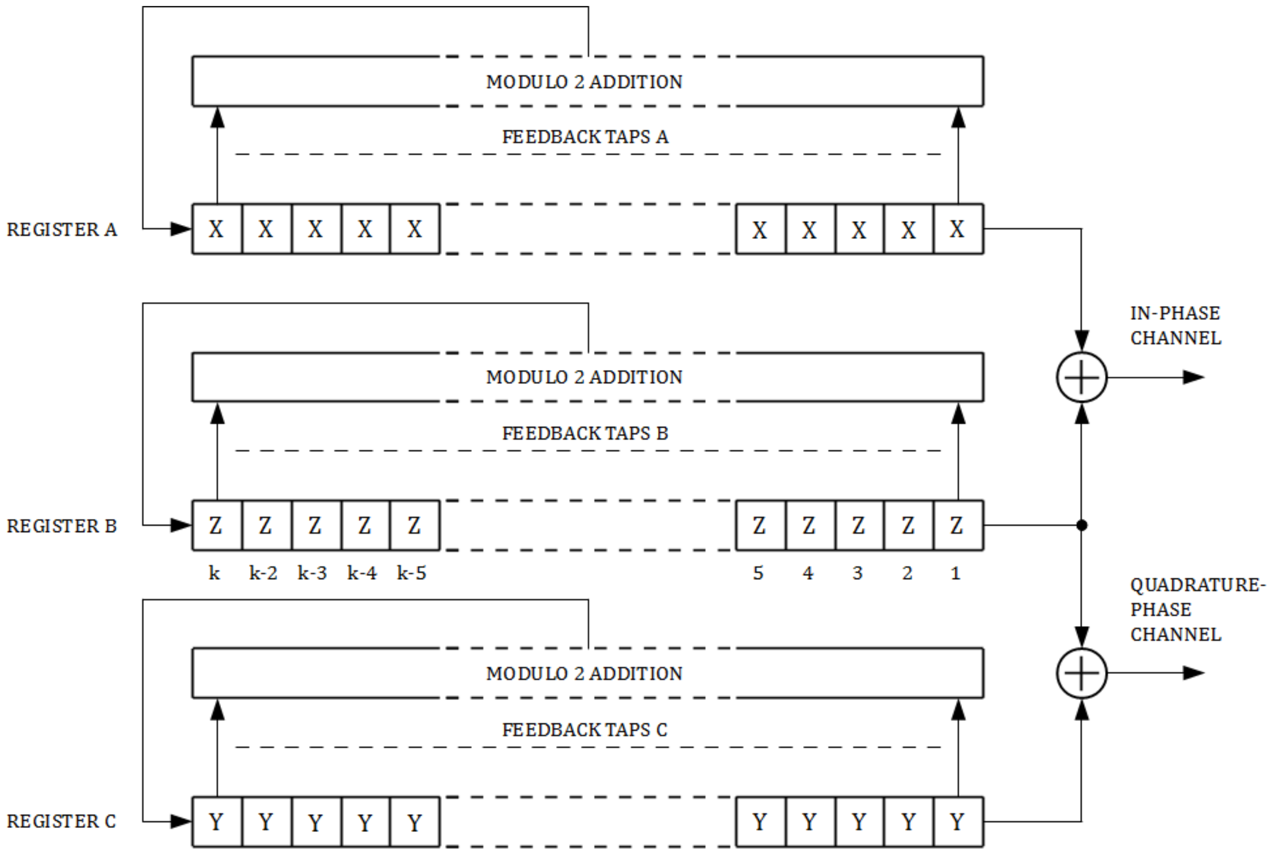

Figure 5: Generic I/Q Gold code generator

The codes are uniquely defined by the shift register length and the feedback taps of register B and register A (or C). Different links in a CDMA system use different codewords of the same code for spreading. These codewords are defined by the initial settings of register A and C (while the initial setting of register B is 00…01 for all codewords).

Gold Codes can be found for arbitrary shift register lengths n. Strictly speaking, Gold Codes are only defined for such n which are no integer multiple of 4. R. Gold [i.6] showed that, for all other n, preferred pairs of maximum length sequences can be found with 3 level cross-correlation. The 3 levels are -1 and $-1 \pm 2^{(n+2)/2}$, if n even, or -1 and $-1 \pm 2^{(n+1)/2}$, if n odd. Further he described a simple algebraic procedure how to find these preferred pairs. If n is an integer multiple of 4, then it is at least possible to find pairs of maximum length sequences with 5 level cross-correlation. The levels are -1, $-1 \pm 2^{(n+1)/2}$, and $-1 \pm 2^{(n+2)/2}$. That means, the highest level is equal to that for n even but no integer multiple of 4. These sequences can be considered as equivalent to Gold sequences and can be applied in the same way. To our best knowledge no proof is known for that statement, nor a simple algebraic algorithm to find these sequences is known. Suitable sequences can however be found by brute force testing arbitrary pairs. The number of different Gold sequences for a given preferred pair can be shown to be $2^n + 1$ for all n.

The codes are grouped into two sets which are called "short" codes or "long" codes. Short codes are defined for $n = 9, \ldots, 12$; long codes are defined for $n = 15, \ldots, 24$. Usually, short codes and long codes are required for each link. The short codes used for TC and TM may have different length. If the long codes are used for ranging, then same length is mandatory. To ease the initial synchronization of the long code used for TC, its length should be an integer multiple of the short code length. This can be achieved by truncating the length of the long code. Let n be the length of the short code. Then the long code generator consists of shift registers with length $n + m$, which are periodically reset to their initial values after every $2^m(2^n - 1)$ shift.

The dual channel structure of the extended PN code family particularly fits into the needs of the new MTC3 mode. However they can also be applied to the MTC2 mode by using only a single channel.

The longer a code, the more it resembles a noise like signal. Noise like spreading is desirable because de-spreading with noise like sequences turns all types of interferers into white noise, even if the interfering signal is deterministic as for example CW. This means that the link performance only depends on the SNR but not on the particular type of interference. If short Gold codes are used, interference is more deterministic. The average interference is the same as for noise like sequences, but a particular interference can differ significantly from this average. The probability that a particular interference is by a given amount larger than the average decreases with increasing Gold code length.

The price to be paid is that acquisition time increases with code length. The number of different code phases to be tested for acquisition is equal to the code length and hence increases exponentially with the shift register length n.

5.2 Symbol Rate

A customary way to apply BPSK modulation to a DSSS carrier is to multiply the NRZ data sequence with the NRZ chip sequence prior to chip shaping. Two cases can be distinguished here, synchronous and asynchronous modulation. Synchronous modulation means that the bit rate is restricted to values for which the chip rate is exactly an integer multiple of the bit rate. Asynchronous modulation imposes no such restrictions on the data rate and thus is required to obtain high system flexibility. However, asynchronous modulation implies that different bit periods may have different lengths. An example is shown in Figure 6, where the chip rate is 6,5 times the data rate. It can be seen that the first data bit covers 7 chips, while the second bit only covers 6 chips.

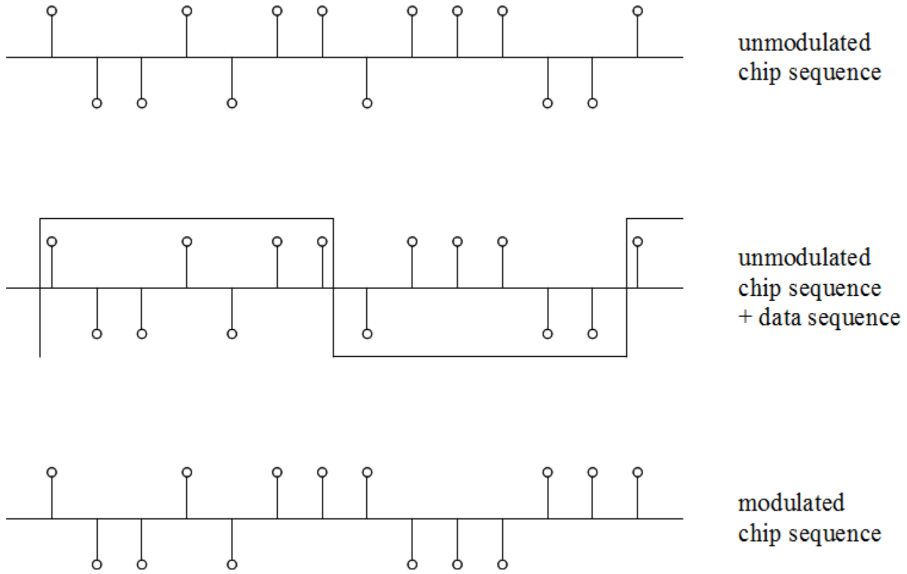

Figure 6: Modulating an NRZ chip sequence with an NRZ data sequence

This can have an impact on jitter and on bit clock recovery. Let the ratio of chip rate and bit rate be $L + p/q$, where L is the greatest integer $< f_{\text{chip}}/f_{\text{bit}}$ and p and q are (relatively prime) integers. Then p out of q bit periods are $L + 1$ chips long and $q - p$ are L chips long. Consider the case $p = 1$. Then $q - 1$ successive bit periods will have length L, followed by only 1 bit period of length $L + 1$. If q is very large, then the clock recovery will adapt to the regular L chips long bit pattern. The intermediate $(L + 1)$-length bit period then introduces a temporary clock shift of exactly one chip period. Since larger clock offsets are not possible, this can be considered as the worst case. The clock shift first means a reduction of the useful detection amplitude equal to $1/L$ and second intersymbol-interference with amplitude $1/L$. For $L = 10$, the first is equivalent to a useful signal power reduction of 0,9 dB; the second means a noise floor at -20 dB.

This shows that asynchronous BPSK modulation enables arbitrary data rates at the expense of some SNR loss. This loss increases with increasing data rate and can be considered as acceptable for data rates below 10 % of the chip rate.

5.3 New MTC3 Mode

The new MTC3 mode is different from the MTC2 mode in that long spread codes are used for spreading the TC uplink signal during tracking. Long spread codes can be considered as noise-like while short codes are treated as deterministic signals. The important difference between random signals and deterministic signal here is, that cross-correlation of random signals changes permanently, while a certain cross-correlation value of deterministic signals can be stable over long times. Therefore, if noise-like spreading is applied to asynchronous CDMA systems, the mutual interference only depends on the average cross-correlation of these signals; if deterministic signals are used instead, the maximum cross-correlation is considered. Thus, the main advantages of random spreading are:

1) Link performance calculation is greatly simplified.

2) CDMA capacity may be significantly improved: if deterministic signals are used, worst case interference is taken into account, which may be significantly larger than average interference of random signals.

The simple idea behind the new MTC3 mode is that, since there is already a long code to be synchronized for ranging, this code could also be used to spread the TC signal. Figure 7 shows the MTC3 mode spreading and modulation scheme.

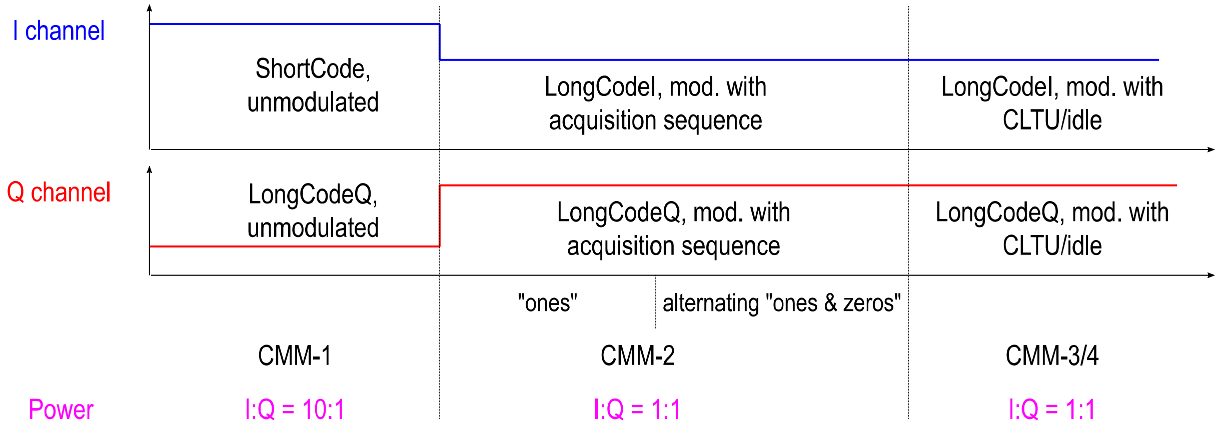

Figure 7: MTC3 mode spreading and modulation scheme

During acquisition a short Gold code is used on In-phase (I) channel and a long Gold code on the Quadrature (Q) channel. The Q-channel power usually is lower than the I-channel power. In Figure 7, a 10:1 ratio is proposed but other ratios may be used as well (see clause 5.4). So far, this is the same as in MTC2 mode. Therefore, acquisition of first the short code and subsequently the long code is the same as in MTC2 mode.

After having acquired the long code on Q-channel, the short code on I-channel is replaced by another long code which is different from the one on Q-channel but epoch-synchronized to it. Therefore no further acquisition is necessary. At the same time the I/Q power ratio is changed to 1. Data modulation is BPSK by synchronously modulating I and Q channel with the same data.

Figure 8 shows the block diagram of the carrier recovery in MTC3 mode. The received signal is de-spread with the local short code in the upper branch as well as with the local long code in the lower branch. The de-spread signals are subsequently integrated and dumped (I&D). During acquisition, the short code is transmitted; therefore only switch S_1 is closed so that only the upper signal is fed to the phase error detector (PED).

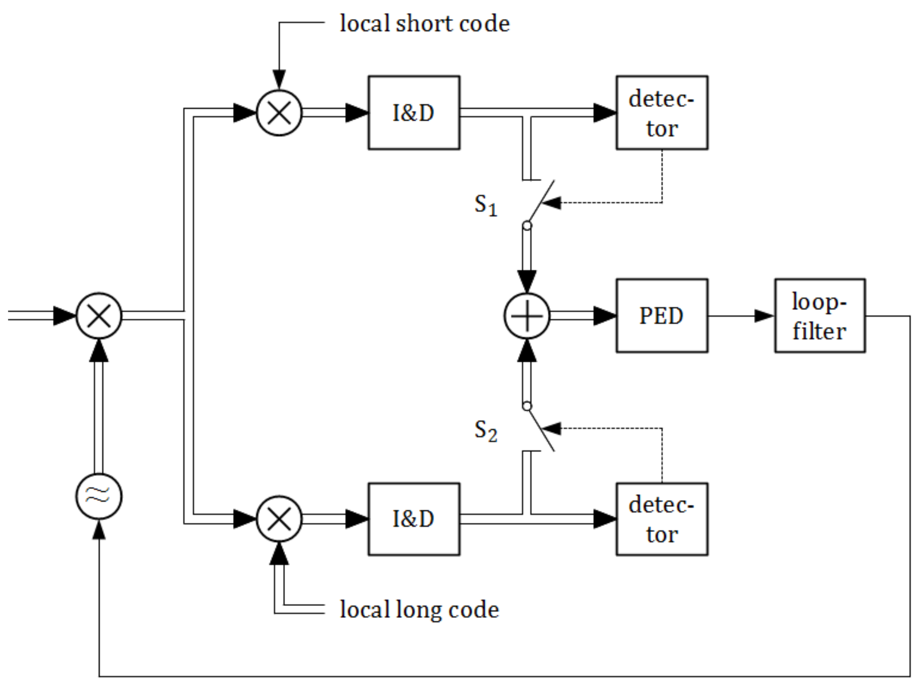

Figure 8: Carrier recovery in MTC3 mode

Since the receiver does not know when the short code is replaced by the long code, after acquisition switch S_2 is closed, too, and the sum of the signals of both branches is forwarded to the PED. Since actually only the short code or the long code is transmitted, only one of the branches delivers a useful signal while the other produces just noise. This reduces the link performance by 3 dB, which is acceptable if no data is transmitted in this phase. This is recognized by the modification of the carrier modulation mode 2 (CMM-2) of the physical layer operation procedure (PLOP) as shown in Figure 7 and described in [i.1], clause 6.2.

Which code actually is transmitted can be detected by simply observing the output power of the I&Ds. During acquisition the output of the short code I&D should be much larger than that of the long code I&D. After code change the ratio of both changes. As soon as the output of the long code I&D is larger, switch S_1 can be opened.

The MTC3 mode preferably uses a dual channel (I/Q) long Gold code for tracking. The Q part of this code is also transmitted during acquisition. In addition a single channel short Gold code is necessary for acquisition. For this either channel of the dual channel generator after Figure 6 can be used. In MTC2 mode a maximal length code is transmitted on the Q-channel for ranging purposes. As shown in Figure 9, the MTC3 tracking codes could also be derived from the ranging code of the standard PN code family.

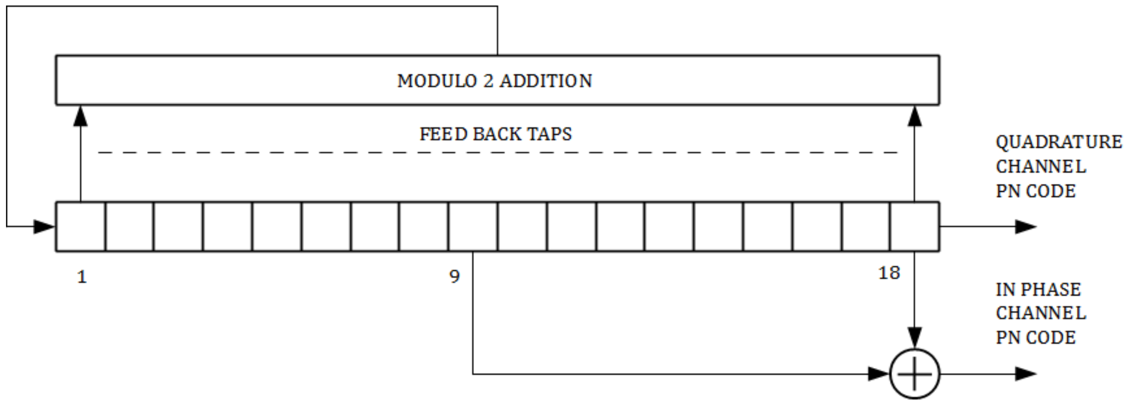

Figure 9: Maximal length code generator for producing the MTC3 tracking codes

In principle the code change can be executed at any time. It is not required that it is done at the epoch of the long and/or short code. Once the receiver is synchronized to the long code on Q it is also synchronized to the long code on I.

It is however important to note that if the code change is performed at the long code epoch, it is possible to switch to completely different long codes for tracking. In particular much longer codes could be used. That means that long acquisition codes and long tracking codes with different lengths can coexist. After having acquired the long acquisition code, the receiver starts detecting the long tracking code at the next epoch of the long acquisition code. Assume first that the acquisition codes are transmitted. Then the detection of the long tracking code fails so a new attempt is started with the next epoch of the long acquisition code. If now the transmit codes are changed, then it is important that the receiver is able to detect the code change within the next acquisition code period. If it fails, it starts detection at the next long acquisition code epoch again and, therefore, loses track if the tracking code length is different from the acquisition code length.

5.4 In-phase to Quadrature Power Ratio

ETSI EN 301 926 (V1.2.1) [i.2] established fixed values for the In-phase to Quadrature (I/Q) ratio, which is the power ratio between I and Q components of the QPSK modulation.

For uplink the value was 10:1. Such value strongly favoured power allocation to the I component, which carried data modulation with a short PN sequence in MTC2 mode, in order to ease short PN sequence acquisition. The consequence was a 10 dB lower power for the Q component, which carried the long PN sequence with no data modulation (ranging component).

However, complete signal acquisition requires the synchronization of the long PN sequence as well. Thus, a relatively lower signal power makes long PN code sequence acquisition last longer or being more complex, depending on the required acquisition time.

In recognition of the technical trade-off between complete signal acquisition time and receiver implementation complexity, the fixed 10:1 I/Q power ratio value has been removed. Instead, the user can select any value between 10:1 and 1:1. This ratio can have an optimized value depending on the long PN sequence detection algorithm strategy and the expected demodulator acquisition time. Note that this I/Q ratio flexibility is also applicable to the new MTC3 mode during acquisition phase (see clause 5.3).

For downlink the value was, and still is, 1:1. This means the same power is allocated to both I and Q components of the QPSK signal. It is important to recall that in non-coherent mode (MTM3) both I and Q components employ short PN sequences whereas in coherent mode (MTM2) both employ long PN sequences (ranging). Furthermore, in contrast to the uplink both I and Q components are always data modulated. Therefore, there has been no reason to change the power balance between I and Q.

5.5 Out-of-Band Emission and Discrete Spurious

ETSI EN 301 926 (V1.2.1) [i.2] did not include this type of requirements on transmitted signals. The revision issue [i.1] includes the following requirements:

- minimum residual power in the modulated carrier for non-spread modulations;
- maximum level of discrete lines in the unmodulated transmitted RF signal spectrum;
- limits to the generation of spurious due to modulation process (discrete lines) near the carrier frequency and in the occupied bandwidth; and
- out-of-band relative power spectral density emission mask to be respected due to modulation process.

The European Cooperation for Space Standardization (ECSS) has established a well-respected Radio frequency and modulation standard [i.17]. The above requirements were therefore extracted from such standard. The following clauses were used as requirements sources:

- clause 6.1.11 for the residual carrier, out-of-band emission and discrete spectral lines of phase modulation with residual carriers;
- clause 6.2.7 for the carrier suppression, out-of-band emission and discrete spectral lines of suppressed carrier modulation; and
- clause 6.3 for the spectral roll-off (out-of-band emission mask) for both non-spread and spread modulations.

Some adaptation to the frequency bands and modulations specified in ETSI EN 301 926 [i.1] was necessary. In particular the following two adaptations were performed:

- clause 6.1.11, sub-clause d): the requirement was extrapolated and generalized to any relevant carrier frequency;
- clause 6.1.11, sub-clause f) and clause 6.2.7, sub-clause d), which both in turn refer to clause 6.3; the ECSS out-of-band emission mask ([i.17], Figure 6-5) was replaced with the United States National Telecommunications Industry Association (NTIA) out-of-band emission mask, which is referenced and shown at [i.13], section D.4.1. This mask was considered more reasonable and compatible with the spread spectrum modulations specified in ETSI EN 301 926 [i.1].

5.6 Physical Layer Operations Procedure

Physical Layer Operations Procedures (PLOPs) define how Command Link Transmission Units (CLTUs) and additional data structures, the Acquisition Sequence and the Idle Sequence, are used with different states of channel modulation. PLOPs for controlling the behaviour of the Physical Layer for CCSDS TC RF and Modulation Standards are described in [i.9], section 5.

The Carrier Modulation Modes (CMMs) consist of different states of data modulation upon the RF carrier which creates the physical telecommand channel. The CMMs employed for CCSDS TC RF and Modulation Standards have been adopted for their application to the ETSI Spread Spectrum modulations defined in ETSI EN 301 926 [i.1] with an adaptation of the definition of "carrier" according to the particular spread spectrum mode:

- For MTC2 a "carrier" is replaced by a "spread spectrum carrier" (see ETSI EN 301 926 [i.1], Table 10). A spread spectrum carrier is simply a carrier which is modulated with a PN sequence in accordance to ETSI EN 301 926 [i.1], Table 7, Mode MTC2 column.
- For MTC3 two types of "spread spectrum carriers" are defined (see ETSI EN 301 926 [i.1], Table 11). Spread spectrum carrier I employs a "short" PN sequence whereas spread spectrum carrier II employs a "long" PN sequence, in accordance to ETSI EN 301 926 [i.1], Table 7 Mode MTC3 column. A visual description of MTC3 PLOP and CMMs is given in clause 5.3.

6 Non-spread Modulation

6.1 Uplink Phase Modulation

The use of a large FM frequency deviation on the Uplink (typically ± 400 kHz) implies a minimum Rx frequency separation between collocated satellites of 1 MHz or more to avoid interference.

Considering that the subcarrier frequency of the TC signal in the MTC1 mode is 8 kHz or 16 kHz, PM modulation of the TC signal occupies a frequency bandwidth of less than 4 times the subcarrier frequency, hence less than 64 kHz for a 16 kHz subcarrier. This allows using a much smaller frequency separation between uplinks for collocated satellites.

Furthermore, most satellite platforms using FM uplink modulation prohibit the use of simultaneous Commanding and Ranging in their Ground-to-Satellite Interface Specifications (GSID). Few satellite platforms specify two modes of operation, with a deviation of ± 400 kHz if a single signal is transmitted (TC or Ranging), and a deviation of ± 200 kHz each when transmitting TC and Ranging simultaneously. Using PM modulation for uplink allows simultaneous Commanding and Ranging, the GSID of such platforms specifying usually a deviation of 1 rad for TC and 0,7 rad for Ranging. This offers added flexibility in the management of the Ranging measurement scheduling as it becomes independent of the TC scheduling.

6.2 Miscellaneous

The ranges for subcarrier frequencies and symbol rates (ETSI EN 301 926 [i.1], clause 4.2) have been extended in order to offer more flexibility for satellites with more demanding TCR requirements.

7 Coding and Interleaving

7.1 General

ETSI EN 301 926 (V1.2.1) [i.2] included an example of a possible forward error correction scheme for uplinks in annex G. With the revision of the standard it was considered valuable to expand the scope of the standard so as to include forward error correction schemes for both uplink and downlink.

The CCSDS has defined a number of synchronization and channel coding standards, which are widely used on space missions and, therefore, of interest for the TCR of telecommunication missions. Hence, ETSI EN 301 926 (V1.3.1) [i.1] has incorporated some of those CCSDS standards.

7.2 Uplink

7.2.1 Forward Error Correction

ETSI EN 301 926 (V1.3.1) [i.1] provides the following options:

1) BCH (63,56) as per [i.4].

2) Concatenated Convolutional (inner code) rate ½ as per [i.5], section 3 and BCH (63,56) as per [i.4].

3) Low-Density Parity-Check (LDPC) as per [i.4].

The BCH scheme is a classic systematic block code standardized long ago by CCSDS. It has been widely adopted by space missions from space agencies but also from commercial operators. It allows to modulate the size of Telecommands by employing small codeblocks (64 bits). These codeblocks consist of 56 bits of information, 7 parity bits and 1 filler bit. The codeblocks are delimited with a 16-bit Start Sequence and a 64-bit Tail Sequence. The aggregation of these delimiters and the codeblocks is called a Command Link Transmission Unit (CLTU).

The CCSDS Standard [i.4] provides two options: single error correction (SEC) and triple error detection (TED). The latter sacrifices error correction capability in exchange of better integrity error detection. More details of its properties as well as performance data can be found in [i.9], section 9.

The second coding scheme relies on the concatenation of two codes: the BCH block code (outer code) and a serial convolutional code (inner code). Each of the two codes have been standardized by CCSDS. Nevertheless, such concatenation has not been standardized by CCSDS. For this reason it deserves further discussion in the present document. Thus, the following paragraphs describe this option and some of its particulars.

The BCH block code is the one described in option 1) above. The serial convolutional code is the one specified by CCSDS for TM as per [i.5], section 3. This is a rate ½ convolutional code with constraint length equal to 7, developed in the 1970s and delivering at that time an excellent compromise between performance and complexity. One interesting property is that the code is *transparent*. This means that at steady state symbol polarity inversion at the input is followed with symbol polarity inversion at the output. This property is particularly useful for BPSK modulations, where symbol polarity ambiguity can manifest on receiver demodulator lock condition. The presence of known headers allows to solve the ambiguity at data handling. Further details on this code can be found in [i.5], section 3 and [i.10], section 4.

This concatenated coding option enhances link performance with respect to just BCH. It is mainly driven by its simple implementation considering the current logical and physical partition between TT&C and data handling subsystems of telecommand protocol functions on-board the spacecraft. Current spacecraft computers directly interface with telecommand receivers. The latter deliver a symbol stream that is processed by the telecommand protocol processor typically siting on a board of the spacecraft computer. Thus, the serial convolutional decoding function can be implemented in the transponder (receiver section) without impacting the well-established telecommand protocol processor. This makes its application relatively straightforward in comparison to more sophisticated coding options like the new Low Density Parity Check code described further down.

However, adopters of this coding option should be aware of the interplay between the serial convolutional code and the BCH block code. It is well known that a Viterbi convolutional decoder can produce errors in bursts [i.11]. Error bursts could impair CLTU synchronization and/or BCH decoding. CLTU synchronizers can tolerate a very limited number of symbol error on the Start Sequence pattern. If the burst(s) exceed that number, the CLTU will not be properly synchronized causing a missed detection. Assuming that burst(s) did not damage the Start Sequence, unfortunately the BCH block code has no capability to correct error bursts if they occur within a codeword. In SEC mode the BCH can only tolerate and correct a single error or detect two errors in a codeword. In TED mode it cannot correct any error but it can detect up to three errors per codeword. The presence of two or more consecutive symbols in error at the output of the Viterbi decoder will either produce a codeword rejection in the BCH block decoder or in worst case lead to undetected codeword error, which ultimately could lead to undetected telecommand error.

To mitigate the effect that symbol polarity ambiguity introduced by BPSK demodulators can have on symbol synchronization and consequently on convolutional decoding performance, a symbol conversion from NRZ-L to NRZ-M before the convolutional encoding (at the sending end) and after the convolutional decoding (at the receiving end) is advisable [i.12]. Furthermore, this symbol conversion provides an additional benefit. Under low signal-to-noise ratio the BPSK demodulator may introduce symbol polarity jumps as a result of carrier loop cycle slipping. These jumps will in turn produce error bursts. At the expense of a minor increase of symbol energy such symbol conversion reduces the length of error bursts when compared to NRZ-L. With the latter the occurrence of a cycle slip may introduce a consecutive string of symbol errors until the next cycle slip or the new symbol polarity ambiguity resolution. With NRZ-M the symbol polarity change due to a cycle slip may lead to a single symbol error.

As a further measure to mitigate the undesirable effects on undetected error performance the user has at TC frame level the option to insert a frame Cyclic Redundancy Check (CRC), which will reduce the likelihood of undetected frame error.

Thus, it is important to carefully consider the BCH and convolutional codes interplay, which is also influenced by the selection of SEC or TED mode, and the presence or absence of a frame CRC to properly set the maximum acceptable symbol error rate (SER) at the output of the convolutional decoder and consequently the frame error rate. To be on the safe side SER should be lower than 10^{-5}.

To balance the cautionary view given by the precedent paragraphs it is worth mentioning that such concatenation has been, and continues to be, used by users of NASA TDRSS. This concatenation offers a simple way to enhance link budget performance on noise-constrained forward data relay satellite links [i.13]. One European example of such use was the Automated Transfer Vehicle (ATV) TC link [i.14]. Another European example of such concatenation but on direct-to-ground links was given by the CNES Myriad micro-satellites family, where the additional coding gain allowed to reduce the size and cost of the uplink station [i.15].

Finally, the LDPC code provides users with the coding option with the highest performance (coding gain) at the price of higher implementation complexity. The CCSDS work for the definition of the LDPC codes has been mainly driven by NASA's need to retire costly 70-meter Deep Space Network antennas but at the same time maintain or even improve power efficiency (link performance) to accommodate increasingly higher data rates with smaller antennas. Hopefully the exploitation of more modern and powerful on-board signal processing technologies has enabled the application of higher performance channel codes with more complex decoders providing several dBs of link budget improvement with respect to BCH.

The two proposed LDPC codes belong to the *binary* family of LDPC codes. They are specified with codeword lengths $(n = 128, k = 64)$ and $(n = 512, k = 256)$, so that both have code rate $r = k/n = 1/2$. They provide a substantial performance improvement with respect to BCH without incurring in the high implementation complexity brought by the slightly higher performance *non-binary* LDPC codes.

Furthermore, these two LDPC codes are systematic. They are transparent, so phase ambiguities may be resolved by using frame markers (start sequence), or by another means after decoding.

At the time of writing the present document the CCSDS is working on the update of the Informational Report [i.9] that will detail the characteristics and performance of the two proposed LDPC codes. In the meantime readers can refer to [i.16], which is an Experimental Specification describing and comparing a number of binary and non-binary LDPC codes. Among the studied binary LDPC codes one can find the two finally selected by CCSDS for the standard [i.4].

7.2.2 Pseudo-randomization

Pseudo-randomization as specified in [i.4], section 5 ensures sufficient symbol transition density on the modulated carrier and, therefore, eases and maintains symbol synchronization process at the TC receiver. A detailed description of pseudo-randomization characteristics, related procedures as well as recommended pseudo-random sequence generation is provided in [i.9], section 7.

7.3 Downlink

7.3.1 Forward Error Correction

ETSI EN 301 926 (V1.3.1) [i.1] provides the following options:

1) Convolutional coding rate 1/2 as per [i.5], section 3.

2) Reed-Solomon coding and interleaving as per [i.5], section 4.

3) Turbo Coding as per [i.5], section 6.

4) LDPC as per [i.5], section 7.

5) Concatenated Convolutional (inner code) rate 1/2 as per [i.5], section 3 and Reed-Solomon coding and interleaving (outer code) as per [i.5], section 4.

The serial convolutional code is a rate ½ convolutional code with constraint length equal to 7, developed in the 1970s and delivering at that time an excellent compromise between performance and complexity. One interesting property is that the code is transparent. This means that at steady state symbol polarity inversion at the input is followed with symbol polarity inversion at the output. This property is particularly useful for BPSK modulations, where symbol polarity ambiguity can manifest on receiver demodulator lock condition. The presence of known headers allows to solve the ambiguity at data handling. Further details on this code as well as its performance can be found in [i.5], section 3 and [i.10], section 4.

The Reed-Solomon code is a linear block code. The block length n of an RS code is $q–1$, with $q = 2^J$ being the alphabet size of the symbols. RS codes with k information symbols and block length n have a minimum distance $d = n–k + 1$. The error probability is an exponentially decreasing function of the block length, and the decoding complexity is proportional to a small power of $n–k$.

The Reed-Solomon code is also a transparent code. If the channel symbols have been inverted somewhere along the line, the decoders will still operate. The result will be the complement of the original data (except, usually, for the codeword in which the inversion occurs). However, this property is lost if virtual 'zero' fill is used. Therefore, symbol polarity ambiguity needs to be resolved before RS decoding.

Two RS codes are recommended by CCSDS, both having codeword size $n = 255$ symbols and symbol size $J = 8$ bits or alphabet size $2^J = 256$. The first code has information block size $k = 223$, minimum distance $d = 33$, and can correct $E = 16$ errors. The second code has $k = 239$, $d = 17$, and can correct $E = 8$ errors. Further details on these codes, their interleaving as well as their performance can be found in [i.5], section 4 and [i.10], section 5.

Turbo codes were introduced in 1993 as a new class of concatenated codes, capable to achieve near-Shannon limit error correction performance with reasonable decoding complexity. It was found that good Turbo codes can come within approximately -0,8 dB of the theoretical limit at a BER of 10^{-6}. To achieve highest performance Turbo codes use large block lengths and correspondingly large interleavers. Further details on these codes and their performance can be found in [i.5], section 6 and [i.10], section 7.

Low Density Parity Check codes were rediscovered at mid-1990s. Initially invented by R. Gallager in 1961 they were forgotten for more than 30 years given the implied complexity of the decoder, an unsurmountable barrier at that time. Decades of VLSI development has enabled their practical implementation. They provide excellent performance and offer a different solution space with respect to Turbo codes for the trade-off between power and spectral efficiencies. Further details on these codes and their performance can be found in [i.5], section 7 and [i.10], section 8.

Finally, concatenated convolutional and Reed-Solomon provides for a classical solution in many space missions. By combining a serial code with a block code, performance can be improved with a reasonable amount of additional complexity. The reader is referred to additional information on each separate code component (see the coding options listed earlier in this clause and corresponding references) as well as to specifics of the concatenation process in [i.10], section 6.

7.3.2 Pseudo-randomization

Pseudo-randomization as specified in [i.5], section 10 ensures sufficient symbol transition density on the modulated carrier and, therefore, eases and maintains symbol synchronization process at the TM receiver. The downlink pseudo-randomization is not the same as the uplink pseudo-randomization. The basic principle is the same but the pseudo-random bit pattern and its underlying generator polynomial are different. A detailed discussion and description of pseudo-randomization benefits, characteristics, related procedures as well as recommended pseudo-random sequence generation is provided in [i.10], section 9.2.

8 Conclusion

The present document provides a sufficient justification for the revision of the ETSI TCR Standard [i.1]. When necessary pointers to more detailed analyses (e.g. annexes) or more detailed sources of information (references) are provided.

The revision effort has added flexibility in a number of areas such as spread spectrum modulation with the expansion of the family of PN codes and corresponding library that can be used, the symbol rate ranges for both non-spread and spread modulations and other.

Ka-band frequency bands have been specified. Furthermore, PM modulation has been introduced for up-links. Corresponding requirements have been borrowed from ECSS. Channel coding and interleaving has been specified by referring and adopting in most cases the relevant CCSDS options for TC and TM.

To conclude the present document, it is worth mentioning that during the study of PN code family expansion, novel concepts like PN codes produced with cryptographic sequence generators (counter-mode block-cipher stream) were explored given their potential, but not adopted for the revision (see annex C). Perhaps their study will be resumed on a future revision of the present document.

Annex A: Generation and Validation of the Extended PN Code Library

A.1 Dual Channel Gold Codes

As already mentioned in clause 5.1 above, a new comprehensive spread code library, based on dual channel Gold codes, has been created. The codes are specified by the shift register feedback taps (see Figure 5). To minimize multiple access interference (MAI) different users in a CDMA ensemble use different codewords from the same code for spreading. These codewords are determined by the initial settings of register A and C, while the initial setting of register B is fixed to 00…01.

Not every pair of initial settings is a valid pair. There are two additional rules to be considered:

1) If the Q channel sequence is delayed 1/2 chip as for TM, spurious sequences will be generated on the I and Q channels due to filtering and hard-limiting, which are codewords of the same Gold code (see [i.7]). These codewords will be omitted from the code library. This requirement is mandatory for TM but not for TC.

2) All codewords should be balanced. This means that they contain exactly one more 1 than 0's. This implies that an un-modulated spread signal is nearly DC free and therefore has only a very small spectral line at zero frequency. However, an un-modulated spread signal is a periodical signal with period equal to the codeword length. The spectrum is therefore a line spectrum with lines separated by the inverse of the codeword length. The magnitude of some of these lines may be significantly larger than the DC line of an un-balanced codeword. Therefore, the balance-criterion may be considered as not so important.

The standard code library is based on exactly the same rules [i.7] and [i.8].

If none of these criteria needs to be met, then the initial values of Register A and C can be chosen arbitrary but different from each other. If the all zeros sequence is also excluded, $2^{n-1} - 1$ different codewords can be assigned.

The algorithm for searching suitable register initialization is a bit heuristic. As described in [i.7], for each register initialization two spurs, the so-called I-channel spur and the Q-channel spur, are generated which are codewords of the same Gold code. Therefore these spurs correspond to specific initializations, which then are blocked. According to [i.8]: Let a_k be the maximal length sequences produced by register A or register C of 6, respectively. k is the relative phase shift from the initial condition 00…01 ($k > 0$ implies a phase delay of k chips). Further, let b be the maximal length sequence produced by register B. Then all Gold codewords on I or Q can be denoted as $g^k = a_k \oplus b$. Let g^k and g^l represent the I and Q channel codeword, respectively. Then the spurs on the I and Q channel can be written as:

$$\text{I channel spur} = g^{k-1+\phi_a(l-k+\phi_a(1))}$$
$$\text{Q channel spur} = g^{l+1+\phi_a(k-l-1+\phi_a(1))}$$

where $\phi_a(j)$ is the "shift-and-add" function of sequence a, the maximal length sequence produced by Register A (or C) of Figure 5. If the argument of ϕ_a is zero then the spur is equal to the sequence produced by the all zeroes initial state of Register A (or C) and is therefore not a member of the Gold code. This is called the "zero spur" criterion. This (only applied to the Q channel spur) was used to select 255 balanced codeword pairs to be used for the MTM3 mode. This set is called the set of assigned codeword pairs [i.8].

NOTE: The modulo 2 sum of two maximal length sequences a_l and a_k from the same code with different delays k and l is again a maximal length sequence from the same code with delay $k + \phi_a(l-k)$. If $l = k$, then the all zeroes sequence results (because the modulo 2 sum of two identical sequences is the all zeroes sequence).

Further, a set of spare codeword pairs has been reported in [i.8]. These pairs produce spurs which do not meet the "zero spur" criterion, but which are unbalanced and therefore cannot interfere with any assigned balanced codeword. Finally, all pairs were accepted which produce balanced spurs that have been already identified as spurs of previous selected pairs.

With the above approach, 255 assigned codewords and 213 spare codewords, i.e. a total of 468 codewords, have been reported in [i.8]. As stated above, this procedure is somehow heuristic, which means that there may be search strategies which deliver a larger number of codewords. E.g. the "zero spur" criterion could in addition be applied to the I channel.

Further, codeword pairs which produce unbalanced spurs could be preferred. These refinements were tested but only 3 more codeword pairs could be for the same Gold code.

For a given shift register length n several Gold codes can be found. The number of codewords meeting the spur criterion depends on which code is chosen. Therefore the feedback taps should be chosen in order to maximize this number. Table A.1 shows the minimum and maximum number of codewords (over all Gold codes of the given shift register length n) which fulfill only the spur but not the balance criterion. Further, the polynomials of the code with maximum number of codewords are listed.

Table A.1: Minimum and maximum number of codewords fulfilling the spur criterion

n	min	max	Polynomial A	Polynomial B
8	81	89	615	453
9	185	184	1 175	1 151
10	322	386	3 441	2 773
11	644	750	5 373	4 423
12	1 262	1 518	10 175	13 611

Table A.2 shows the minimum and maximum number of codewords which fulfil both the spur and the balance criterion.

Table A.2: Minimum and maximum number of codewords fulfilling the spur and the balance criterion

n	min	max	Polynomial A	Polynomial B
8	53	57	561	551
9	114	119	1 275	1 225
10	325	337	2 363	3 575
11	463	477	7 113	4 745
12	881	944	13 505	14 357

From Table A.2 it can be seen that for the MTM3 mode the number of suitable codeword pairs can be increased to 477, if another Gold code is selected.

The code library is just a table of suitable feedback tap connections for all register length from $n = 9$ to $n = 24$. The results are reported in Table A.3. All values are in octal. Number of codewords means number of balanced and spurious free codewords. For n up to 12 the optimum code, i.e. the code with maximum number of balanced spurious free codeword pairs, has been determined. For $n > 12$, the code with the minimum number of feedback tap connections for register A (and C) has been selected. From Table A.2 it can be seen, that the difference between the minimum and maximum number of codewords is not very large. Further, the computational effort to find the best code increases dramatically with n. Therefore the search for the best codes was only performed for shift register lengths up to 12.

Table A.3: Gold code feedback taps for $m = 9, \ldots, 24$

n	Register A (and C)	Register B	Number of Codewords
9	1 275	1 225	119
10	2 363	3 575	337
11	7 113	4 745	477
12	13 505	14 357	944
13	33 001	30 733	1 884
14	60 421	74 573	5 376
15	140 001	135 131	7 601
16	320 021	226 135	15 146
17	440 001	605 177	30 223
18	1 004 001	1 643 255	85 542
19	3 440 001	2 470 377	120 788
20	4 400 001	4 563 321	239 579
21	12 000 001	16 776 011	483 840
22	30 000 001	32 454 353	1 384 502
23	63 000 001	40 104 011	1 931 403
24	160 400 001	104 401 661	3 881 062

Then, for all n, the list of initial register settings should be provided. As an example Table A.4 reports these settings for $n = 9$.

Table A.4: PN code generator initial register states, $n = 9$

Codeword	Register A	Register C	Codeword	Register A	Register C	Codeword	Register A	Register C
1	001	401	41	575	303	81	353	646
2	473	246	42	577	300	82	236	165
3	700	440	43	665	157	83	721	517
4	003	402	44	536	361	84	002	635
5	352	637	45	606	505	85	051	226
6	514	752	46	060	050	86	167	340
7	144	526	47	743	422	87	565	310
8	021	031	48	432	227	88	462	010
9	214	312	49	741	421	89	106	301
10	431	625	50	164	116	90	147	175
11	372	207	51	624	136	91	176	205
12	254	772	52	042	463	92	750	742
13	062	053	53	235	323	93	257	343
14	720	070	54	313	656	94	537	114
15	412	617	55	274	342	95	420	022
16	127	174	56	146	525	96	760	441
17	761	011	57	647	564	97	302	206
18	040	460	58	534	362	98	155	666
19	370	204	59	125	177	99	330	137
20	023	032	60	256	771	100	574	566
21	604	506	61	350	634	101	762	322
22	410	614	62	217	710	102	247	470
23	215	713	63	237	320	103	071	267
24	333	266	64	763	012	104	546	255
25	430	224	65	471	245	105	461	616
26	276	341	66	105	547	106	731	030
27	626	135	67	145	043	107	524	677
28	555	733	68	124	063	108	773	033
29	645	567	69	576	452	109	657	275
30	451	675	70	701	277	110	403	615
31	516	751	71	041	740	111	674	607
32	702	443	72	433	411	112	013	472
33	667	154	73	061	020	113	134	450
34	722	073	74	664	554	114	636	052
35	107	544	75	413	371	115	515	545
36	331	265	76	556	332	116	156	504
37	166	115	77	644	234	117	423	442
38	311	655	78	216	605	118	654	627
39	453	676	79	373	126	119	535	400
40	557	730	80	104	703			

Annex B: Validation of Doppler and Doppler Rate requirements

B.1 Summary of Previous Work

In [i.3], annex A, an analysis of the maximum Doppler shift and Doppler rate has been made, under the following conditions:

- The considered orbits are a GTO (200 x 35 788 km), a drift (1° per day or 3° per day) and GSO (< 3m/s radial velocity).
- The Ground Station is located in the Equatorial Plane and the orbit inclination is 0°.
- The position of the satellite in its orbit (for the GTO case) was restricted to a True Anomaly higher than 40°.
- The effect of the Earth rotation was not included in the simulation.

The drift and the GSO operations induce negligible Doppler shifts and Doppler Rates compared to the GTO phase.

Under these conditions, the technical report concluded to a maximum Doppler shift of 22 ppm and a maximum Doppler rate of 1,7 ppm/s of the RF frequency.

B.2 Additional Analysis

Due to the fact that [i.1] applies to many different operating scenarios, including Low Earth Orbit constellations, an analysis of different orbits has been conducted, using Ephemeris Propagator software. This software takes into account the full dynamics, including the effect of the Earth rotation.

For inclined orbits, the simulation has been performed for 16 values of the Argument of the Ascending Node (every 22,5°).

Moreover, for elliptical orbits, the simulation has been performed for 16 values of the Argument of Perigee (every 22,5°) for each Argument of the Ascending Node, thus totalling 256 different relative positions of the ground station to the Argument of Perigee and Argument of the Ascending Node.

For elliptical orbits, the limitation on the True Anomaly higher than 40° has been applied to the analysis. This eliminates the part of the orbit closest to the perigee, in which a ground station has a very short visibility.

Table B.1 provides the results of this analysis, in which "Visibility" means Elevation > 0° and True Anomaly between 40° and 320°:

Table B.1: Doppler and Doppler rate vs. selected orbits

Orbit name	GTO	SSTO	Molniya	GlobalStar	OneWeb	Iridium
Perigee	200 km	470 km	1 000 km	1 414 km	1 200 km	781 km
Apogee	35 750 km	65 000 km	40 000 km	1 414 km	1 200 km	781 km
Inclination	0°	50°	63,4°	52°	90°	86,4°
Ground Station Latitude	0°	0°	0°	0°	0°	0°
Ground Station Long.	0°	0°	0°	0°	0°	0°
Maximum Doppler shift	32,1 ppm	31,1 ppm	27,9 ppm	18,6 ppm	20,4 ppm	22,1 ppm
Percentage of Visibility with Doppler < 22 ppm	99,1 %	99,7 %	99,8 %	100 %	100 %	97,5 %
Maximum duration of Doppler > 22 ppm in Visibility	20 min	22 min	17 min			2 min
Maximum Doppler Rate	0,9 ppm/s	1,0 ppm/s	0,3 ppm/s	0,09 ppm/s	0,13 ppm/s	0,24 ppm/s
Percentage of Visibility with Doppler Rate < 1,7 ppm/s	100 %	100 %	100 %	100 %	100 %	100 %

As an example, Figure B.1 shows all Visibilities calculated for the SSTO orbit, which is the most critical in terms of maximum Doppler Rate and contiguous duration of a Doppler shift higher than 22 ppm. This plot cumulates all Visibilities in the 256 tested conditions.

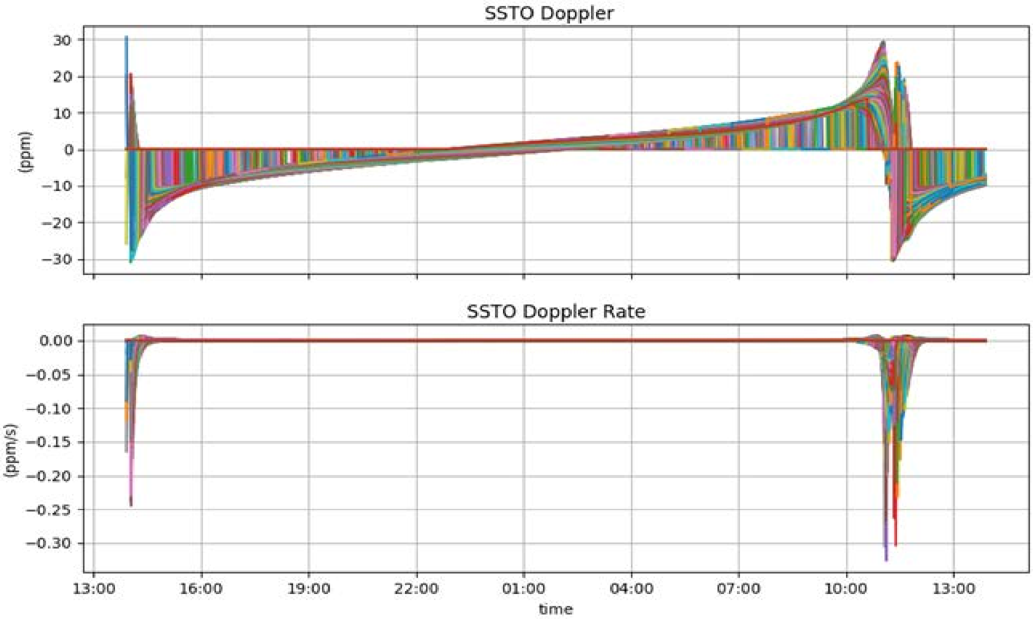

Figure B.1: Visibilities for SSTO orbit

Figure B.2 shows the longest Visibility (18,7 hours) for the SSTO orbit. The visibility begins when the satellite is moving up towards the apogee, and ends when the satellite is moving back towards the perigee. The Doppler shift during this visibility is between -8,0 and +17,2 ppm, much below the 22 ppm limit.

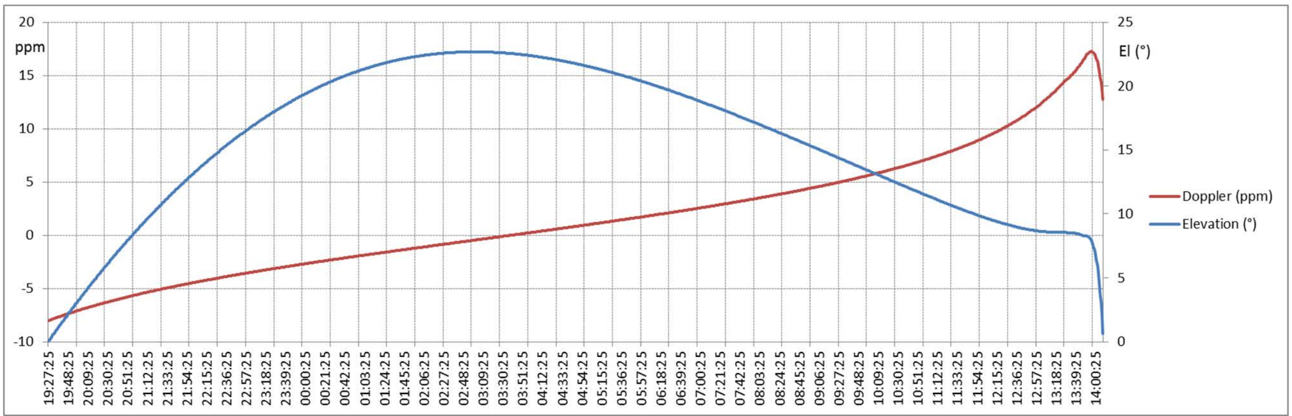

Figure B.2: Longest visibility for SSTO orbit

Figure B.3 shows the visibility in the SSTO orbit in which the maximum contiguous time with a Doppler above 22 ppm is observed. This visibility starts when the True Anomaly just becomes higher than 40° and the satellite remains visible for only 22 minutes. The operational advantage of such a short visibility is limited, in comparison to the long operational time offered by other visibilities of the same orbit.

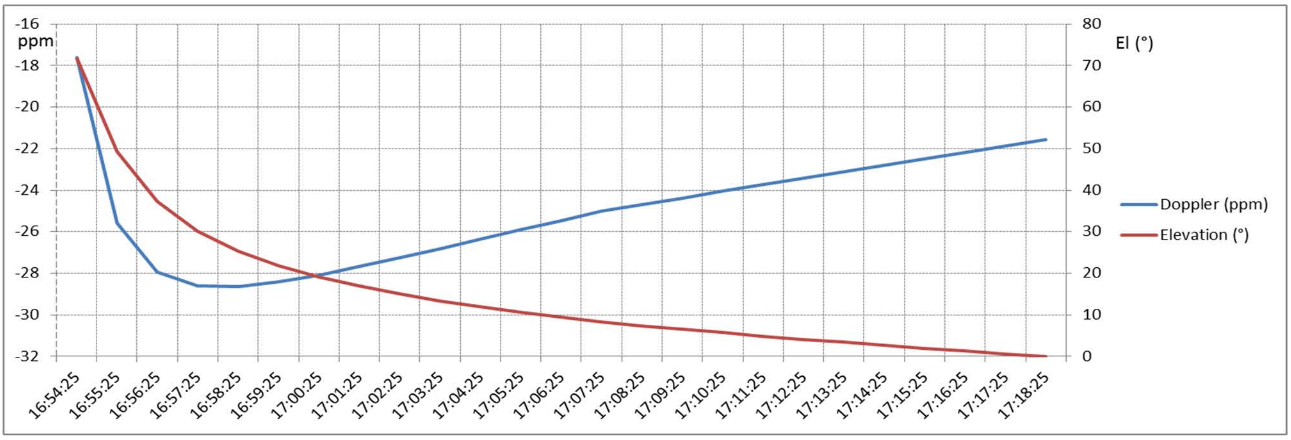

Figure B.3: SSTO orbit with maximum time interval with Doppler above 22 ppm

Figure B.4 shows a visibility in the Iridium orbit in which the 22 ppm Doppler is exceeded. It can be observed that this occurs at very low elevations (below 5 degrees), and therefore when it is not possible to command the spacecraft.

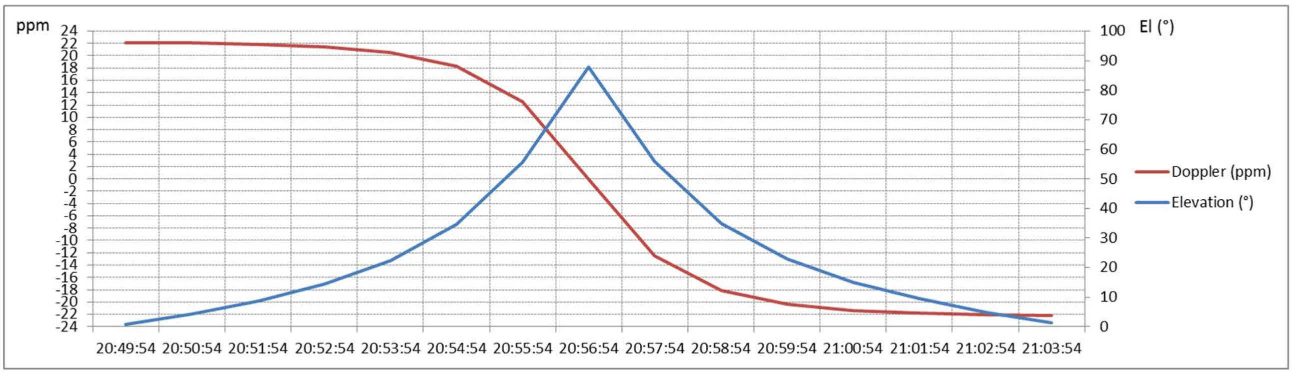

Figure B.4: Iridium orbit with 22 ppm Doppler exceeded

B.3 Conclusion

The analysis performed on the various orbits confirms the validity of the 22 ppm maximum Doppler shift, provided that the True Anomaly is between 40° and 320°. This 22 ppm condition is met over more than 99,5 % of the visibility time for all tested orbits, with the exception of the lowest orbit (Iridium) in which the Doppler shift exceeds 22 ppm during 2,5 % of the visibility time, but without exceeding 22,1 ppm and at elevations lower than 5°.

The 1,7 ppm/s maximum Doppler Rate is also confirmed by this analysis.

Annex C: Cryptographic Pseudo-random Codes

The popularity of Gold codes comes from their property that the cross-correlation of each pair of codewords is well defined and can only assume 3 different values. It can be shown that the average cross-correlation is very close to the square root of the codeword length, $\sqrt{L}$. The same average is obtained if two sequences are randomly picked from the set of all possible sequences of the same length L. However, these sequences do not have a well-defined cross-correlation; instead it is considered as a random number. It is easy to show that this random number is nearly Gaussian distributed.

If more than two sequences are involved, e.g. in a CDMA signal ensemble, then cross-correlation of one Gold codeword with the sum of all others is not so simple to define. In an ensemble of N sequences the number of different cross-correlations is 3^{N-1}. Therefore, for large N, the cross-correlation should also be considered as a random variable. Due to the central limit theorem, this variable is also nearly Gaussian distributed. This can easily be verified by simulation. Figure C.1 shows the cross-correlation probability distribution for 1 to 10 simultaneous interferers (with equal power) for Gold code and random code interferers.

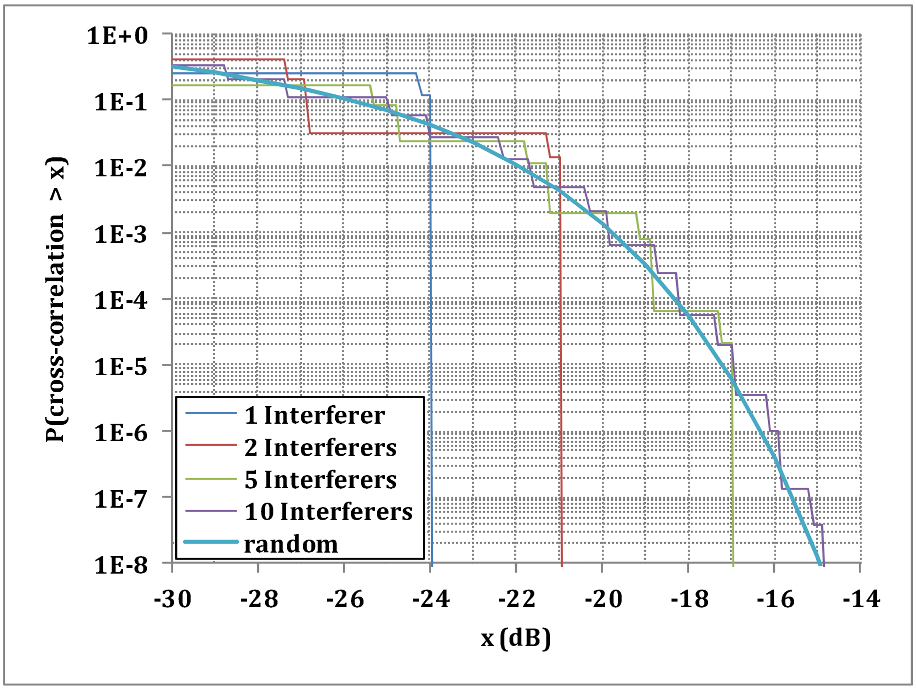

Figure C.1: Cross-Correlation Probability Distribution for simultaneous Gold code interferers; normalized total interfering power

It can be seen, that the cross-correlation level of Gold codes is in principle upper bounded. The bound however increases with the number of interferers. In a situation with only one interferer there is a high probability that the cross-correlation of a random code interferer is much larger than the worst case Gold code interferer. But for 10 or more interferers, there is practically no difference between Gold codes and random codes. From this it can be concluded, that the use of Gold codes offers no benefit over pure random codes, if CDMA systems with a large number (> 10) of collocated subscribers are considered.

The advantage of random codes is that their number can be considered as unlimited. Consider for example a CDMA system with in total 10 000 subscribers. With shift register length 10, at most 1 025 different single channel Gold codewords can be constructed, so that it is not possible to assign a unique codeword to each subscriber. Since at any time only a small fraction of these subscribers are collocated, it is possible to assign the same codeword to subscribers which are not collocated. Then however, codes are dynamically assigned if the system is dynamic. If random codes are used, this is not necessary. A fixed code can be assigned to each subscriber.

Pseudo-random codes with characteristics similar in practice to random codes can be generated by using a block cipher, for instance according to the Advanced Encryption Standard (AES) [i.18], in counter mode ([i.19], section 6.5). The schematic of such a block cipher is shown in Figure C.2. An input string consisting of 128 bits is encrypted into an output string of the same length using a 128 bit key. The output of a good cipher can be considered as a white random sequence independent of the input string. If the input is connected to a periodic counter, then periodic output sequences are produced. For example, the periodic counter $0,1,\ldots,7,0,1,\ldots$ produces an output signal with period 1 024.

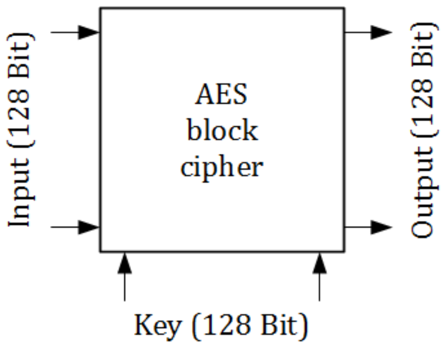

Figure C.2: 128-bit block cipher

History

Document history		
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