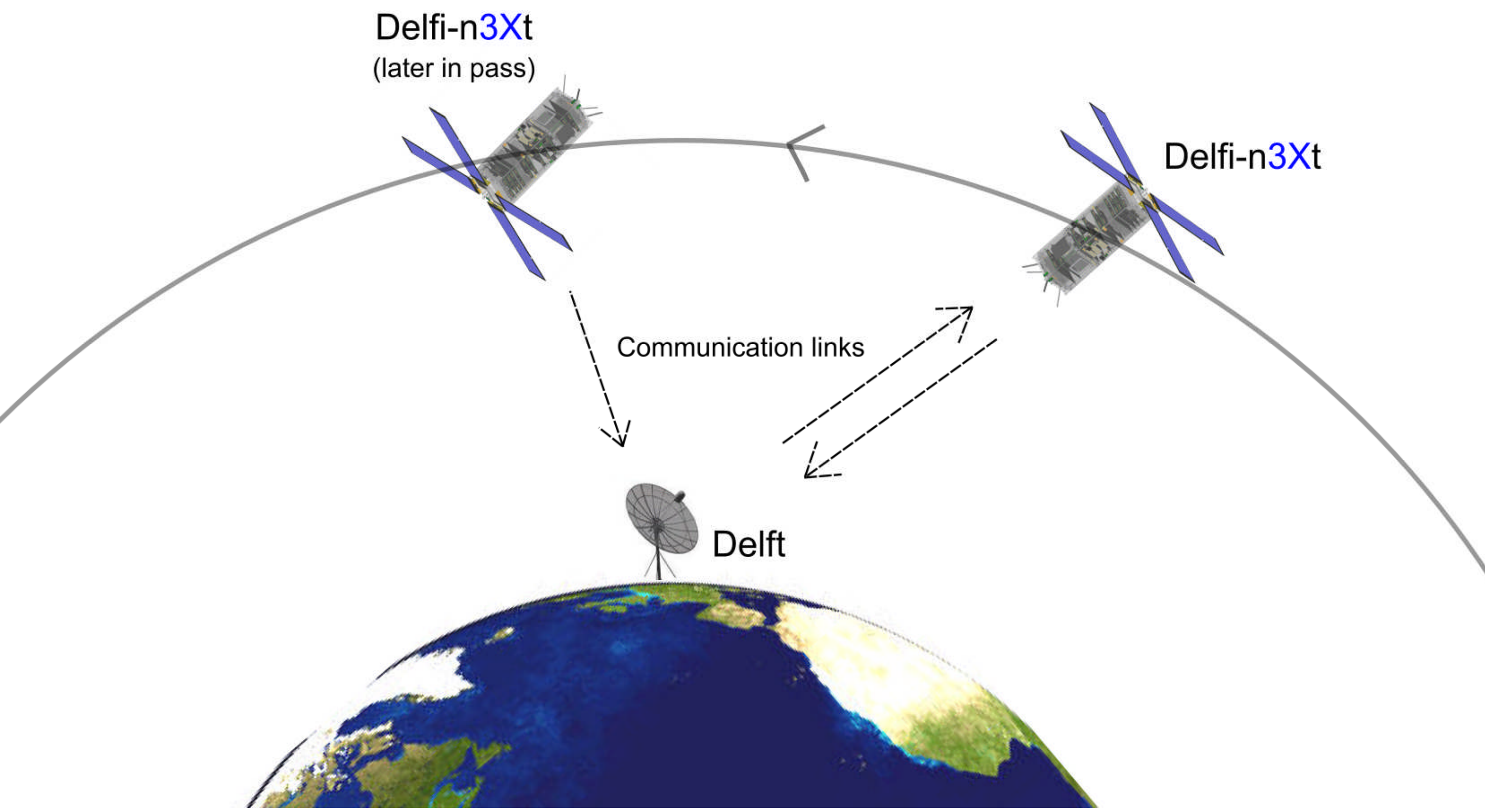
Delfi-n3Xt

Thesis paper

End-to-end analysis and design of the satellite communication links

System design of the communication subsystem of the Delfi-n3Xt nanosatellite



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Challenge the future

Preface

This thesis document represents all work performed during the second and final year of the Master degree of Aerospace Engineering. I have started this work in December of 2008, currently one-and-a-half year ago. In the middle of my thesis duration I have been away for exactly six months, to do an internship (unrelated to the thesis work). That means, all-in-all, I have taken some 13 months. Also, I have made common practice of having average days of over 10 hours, in order to fulfill my own desires to deliver high-quality work.

Thesis work performed within the Delfi-n3Xt programme adds to a general thesis assignment a couple of extra dimensions. First of all, there is a certain product on which work is performed. This product is an actual satellite that is to be launched in the very near future. Second of all, there are many students working on this one final product at any given time; team-work. Finally, there is a definite link to the Dutch space industry. Payloads within the satellite are provided by external parties.

In fact, there is one more aspect. A thesis assignment within the Delfi-n3Xt project is one of freedom and diversity. One is not assigned a main research question and a means of research; one is assigned a function. In my case, this function was to *be the systems engineer of the communication subsystem*. The purpose: to ensure that the final satellite will be able to communicate with us and us with it.

This input to my work has been twofold. A top-level design of the radios and other systems required on the Delfi-n3Xt satellite had been made by my predecessor. This specified the global identity of the new satellite, but no detailed design had taken place. At the same, there was and still is a working satellite in the sky right now, Delfi-C^3, communicating with us around noon on a daily basis. And whereas the Delfi-C^3 mission was a success, it is one realized by in some cases using design shortcuts and in most cases not having solid documentation to argue for the choices made. And, of course not everything worked as it should have.

The thesis work has proven to be a dynamic and insightful adventure. Using the documentation and equipment already in place, I have had to dig into many subjects related to communication theory to discover why choices have been made and what alternatives there are. And of course, what options there are for Delfi-n3Xt and its systems and operations. I have made sure to document these options and these choices, and to establish design options for Delfi-n3Xt. It is my hope that using this knowledge, not only the end-to-end communication system of Delfi-n3Xt can be finalized, fabricated and operated without problems, but that also future students will be able to acquaint themselves with the matter without requiring extensive research.

The body of this document is somewhat unconventional, due to the notion of 'technical notes'. Project documentation is kept and these documents consequently contain all knowledge. For the thesis however, a separate document is used to introduce the thesis and summarize the work done. This is the document currently in your hands; the thesis paper. All project documentation is added to the thesis paper in the form of appendices.

I would like to thank both Jasper Bouwmeester and Chris Verhoeven, as well as all other staff members on the Delfi project for being open to questions and discussions on Delfi-related or non-Delfi-related topics. Also I would like to thank all Delfi-C^3 experts and non-Delfi staff members who have allowed me to meet with them to discuss certain expertise. Finally I would of course like to thank my project members and friends for their continuing presence, enthusiasm, support and distraction.

Arthur Tindemans

A Tindemans

Delft, 21st of June 2010

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1 Introduction

The title of this thesis paper is the: *End-to-end analysis and design of the satellite communication links*. The subtitle is: System design of the communication subsystem of Delfi-n3Xt. Both Delfi-n3Xt and its communication subsystem (COMMS) are introduced in sections 1.1 and 1.2 respectively, to properly introduce the context of the thesis subject. The notion of communication links and the communication links of the Delfi-n3Xt mission are consequently introduced at the end of section 1.2. Section 1.3 finally explains the content and direction of the thesis work performed and presents the document structure.

1.1 Delfi-n3Xt

In order to facilitate low-cost in-space technology demonstrations and applications, the CubeSat standard has been developed. Small satellites of 10x10x10 cm can be launched jointly or together with a large satellite with unoccupied space in the launcher. Furthermore, three single-unit CubeSats can combined to form one triple-unit CubeSat. These CubeSats generally have a mass of only several kilos, and therefore fall in the class of *nanosatellites*. The Delft University of Technology has, or more specifically the faculties of Aerospace Engineering (AE) and Electrical Engineering, Electronics and Computer Science (EEMCS) have adopted this standard within the *Delfi programme*. The first satellite, Delfi-C^3, has been successfully launched in 2008 in is still operational.

Delfi-n3Xt is the successor of Delfi-C^3. Its mission statement:

> *"Delfi-n3Xt shall be a* ***reliable*** *triple-unit CubeSat of the TU Delft which implements* ***substantial advances in 1 subsystem*** *with respect to Delfi-C^3 and allows technology demonstration of* ***2 payloads*** *from external partners from* ***2012*** *onwards"*

The clauses in **bold** can be further specified:

- <u>Reliability</u> of Delfi-n3Xt shall be improved with respect to that of Delfi-C^3 by offering improvements within the satellite bus
- <u>Substantial advances</u> shall be aimed to take place in the Attitude Determination and Control Subsystem (ADCS); Delfi-C^3 applied no active attitude control which is aimed for with Delfi-n3Xt
- <u>Two payloads</u> are:
    - o $T^3\mu PS$: a cold-gas micro propulsion experiment provided by TNO
    - o ITRX: a power-efficient, reconfigurable transceiver provided by ISIS
- <u>2012</u> is the anticipated launch date

Next to the goals stipulated above, a number of other experiments and technology demonstrations are anticipated, depending on their status. One of these technology demonstrations is the inclusion of a high speed communication downlink, as further explained below.

1.2 COMMS

In order to facilitate the payloads, a satellite requires what is generally referred to as the *satellite bus*. Within this bus a division is usually made between *subsystems*. In case of Delfi-n3Xt, these subsystems are defined to be the STS, MechS, TCS, EPS, CDHS, ADCS and COMMS. The latter of course is of prime importance to this thesis, and stands for the **Communication Subsystem**.

The main functionalities of the COMMS are to get:

- Payload data from the satellite to the customer
- Housekeeping data from the satellite to the satellite operator
- Telecommands from the satellite operator to the satellite.

As can be seen, payload data *and* housekeeping data are to be transmitted to ground by the satellite. No distinction shall be made between the two for the communication system and the two are grouped together as *telemetry data* (TM). The satellite should in turn be able to receive *telecommands* (TCs) from the ground segment. In result, this means there should be both a *downlink* and an *uplink*. Together, these can be said to form the primary communication link. As a technology demonstration, the secondary high speed communication downlink has been designed. These two links together form the communication links of Delfi-n3Xt.

1.3 Document structure

This document aims to provide an executive summary of the work that has been performed by the author during the thesis duration and the conclusions that have followed. The detailed work is documented within project documentation. The resulting documents are added to this thesis paper as appendices. This thesis paper will use the *communication links* of Delfi-n3Xt as a red line, by discussing their design and status.

As illustrated by the above, a satellite communication system is not limited to the architecture required on the satellite, but also that required on-ground. Furthermore, the *link itself* should be specified, along with the rules that both transmitter and receiver should adhere to.

In contrary to more conventional theses, there has been no main research question. The goal of the work performed has been to contribute to the design and realization of the Delfi-n3Xt. More specifically: the design and realization of the communication subsystem of Delfi-n3Xt and its communication links.

The two communication links introduced above are discussed in two respective chapters:

- The primary communication link — Chapter 2
- The high speed communication downlink — Chapter 3

Consequently, two chapters aim to illustrate the work performed that has lead to the conclusions presented in chapters 2 and 3. Chapter 4 discusses the application of systems engineering tools. Chapter 5 provides a sketch of the definition of the Delfi-n3Xt COMMS before the author arrived, and presents a (short) description of *all* work that has been performed during the thesis duration. Finally, chapter 6 concludes with a statement of the general value of the work performed to Delfi-n3Xt, and what the consequent steps within the project are. Also, a short list of lessons learned and recommendations is given.

The appendices contain four different documents. These documents are labeled *technical notes*. These are:

1. COMMS – Top-level Design of Communication System [SLR 0014]
2. COMMS – STX System Design [SLR 0387]
3. COMMS – Antenna System Design [SLR 0036]
4. Delfi-n3Xt Link Budget [SLR 0106]

Most content of these documents is unique, that is to say written by the author. In case content results from earlier versions of the project document, the respective sections are indicated on the title page of the appendix.

Whenever relevant, this thesis paper refers to chapters and sections of these documents, as indicated by the *SLR code* mentioned above, to facilitate looking up more detailed information. Within the documents itself, general SLR codes are used to refer to all references. Whenever an SLR code refers to one of the documents written by the author, the respective SLR code is underlined.

2 Primary communication link

The primary communication link is the essential one. Its functions are exactly those as introduced in the previous chapter; it should allow for TM to be transmitted by the satellite, and TCs to be received by the satellite. In fact, the link can be interpreted to consist of two different links; uplink and downlink. Nevertheless, equipment for the two links is shared both on the satellite and on-ground, so that the two are logically grouped and discussed simultaneously.

The important components of the primary communication link are illustrated Figure 2.1 below.

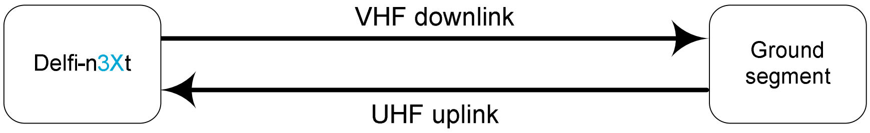


Figure 2.1: Primary communication link

The consequent four sections explain the aspects of this link. The characteristics of the link, common to both ends are introduced first. The primary radio system of Delfi-n3Xt and the ground segment are discussed consequently. Finally, operational aspects are discussed.

2.1 Link characteristics

2.1.1 Transmission bands and frequencies

Frequency transmission of the primary communication link will take place in the:
- VHF radio amateur band (144-148 MHz) for the downlink
- UHF radio amateur band (420-450 MHz) for the uplink

The use of these frequencies requires a return service to the radio amateur community. A primary and secondary return service will be available in case of Delfi-n3Xt, being:
- A linear transponder service
- A software data reception client, DUDe

These return services are in fact equal to those of Delfi-C^3. The linear transponder operates in both the VHF and UHF bands. The VHF transponder bandwidth can be used for the transmission of OLFAR signals as well (OLFAR is introduced in section 2.2 below). Graham Shirville is part of the committee of AMSAT (Amateur Satellite community) that coordinates the frequency coordination within the radio amateur bands. He has been heavily involved in the arrangement of the frequencies for Delfi-C^3. It has in discussion with him been decided that *Delfi-n3Xt can reuse the transmission frequencies of Delfi-C^3*.

The following Figure 2.2 demonstrates the resulting transmission frequencies anticipated for Delfi-n3Xt. The uplink transmission frequencies are masked for reasons of secrecy.

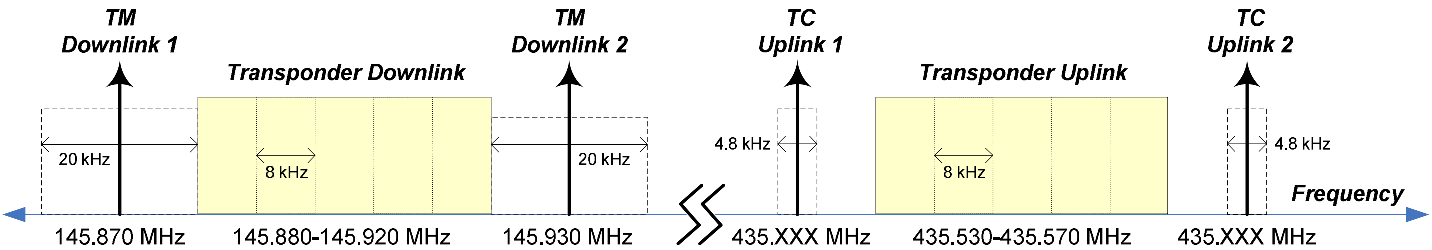


Figure 2.2: Transmission frequencies in the UHF and VHF band [SLR 0014, Figure 8.1]

In comparison to the Delfi-C^3 frequency allocation, the transmission bandwidths have been increased:

- **TM downlink**: increased from 2.4 kHz to 20 kHz. The increased bandwidth will allow for data rates of up to 9.6 kBaud.
- **TC uplink**: increased from 2.4 to 4.8 kHz. The increased bandwidth will allow for data rates of up to 2.4 kBaud.

The data rates are indicated in *Baud*. Baud stands for symbols/s. Usually, Baud equals b/s. Nevertheless, in some cases, as in the uplink of Delfi-C^3 (and Delfi-n3Xt), *two* symbols are required for the transmission of *one* bit due to application of a *line code*. This effectively reduces the bit rate. The above changes to the original Delfi-C^3 frequency allocation have been discussed with AMSAT and should pose no problem. The next step is for *official frequency filing to commence*, to confirm the (re)use of the above frequencies. It is fully acceptable if technical constraints in the end will not result in the maximum data rates to be met.

There is one remark to be made on the topic of the allotted bandwidth. Officially, a certain power reduction is to be achieved at the sides of these bandwidths. As is shown by detailed spectrum measurements performed [SLR 0014, section 7.4 and 8.4], Delfi-C^3 did not achieve the required power reduction, for both the up- and downlink. It has however been agreed upon that within the radio amateur bands these reductions are to be taken more as an indication, as long as proper effort is made to reduce the spectrum. This is important for the selection of the modulation scheme.

2.1.2 Link budget

The link budget for Delfi-n3Xt has been worked out in much detail in [SLR 0106]. It takes into account all gains, losses and noise additions along the communication link, as well as final required bit energy to noise spectral density ratio E_b/N_0. The link budget thereby takes into account the anticipated *modulation approach* as introduced in the next section. Conversely, the selection of a modulation scheme takes into account the margin in the link budget. This allows concluding on both the achievable data rate, as well as the available link margin. The margins predicted by the link budget for minimal data rates are shown in Table 2.1 below, for the minimum and maximum orbit altitudes of Delfi-n3Xt.

Table 2.1: Link margins of the primary link for the Delfi-n3Xt range of orbit altitudes [SLR 0014, Table 8.2]

Link direction	***Orbit altitude***	***Data rate***	***Received E_b/N_0***	***Required E_b/N_0 incl. min. margin***	***Link margin***
Down (VHF)	600 km	1200 Baud	26.45 dB	13.30 dB	*13.15 dB*
	850 km	1200 Baud	24.23 dB	13.30 dB	*11.02 dB*
Up (UHF)	600 km	1200 Baud	42.27 dB	16.30 dB	*25.97 dB*
	850 km	1200 Baud	40.14 dB	16.30 dB	*23.84 dB*

If the data rate is doubled, the link margin is halved. A maximum data rate of 9.6 kBaud then comes at the cost of 9 dB link margin. In conclusion, *the link budget closes with sufficient margin for both the up- and downlink for all achievable data rates*.

Delfi-n3Xt is no longer assumed to perform sun-pointing. As a result, available power has dropped significantly. At the same time, Delfi-n3Xt operations will likely consume more power than Delfi-C^3 and the inclusion of a battery yields to prefer continuous operations, even during eclipses. As a result, it can be decided to *lower the power output of the primary transmitter of Delfi-n3Xt*, the PTRX. Even in case the maximum data rate is used, a minimum of 2 dB additional margin would be available.

In order to verify the accuracy of the link budget of the downlink, the signal strength of Delfi-C^3 was measured during a pass [SLR 0106, chapter 3]. The received signal strength (-91.40 dBm) is perfectly in line with the prediction of the link budget (-87.41 to -94.33 dBm). The latter range is specified by an unknown satellite attitude and on-ground pointing error, both related to the antenna gain.

2.1.3 Modulation and data rate

Digital modulation is a process that impresses a digital symbol onto a signal suitable for transmission. In case of a satellite link, this signal is an electromagnetic wave. The frequency of the final wave to be transmitted is to be equal to the transmission frequency allocated to it. Different digital modulation schemes are introduced in [SLR 0014, section 7.2]. There are three parameters important for the selection of a modulation scheme. These are:

- Bandwidth efficiency
- Power efficiency
- Complexity

It has been shown above that the link budget closes for all data rates that are allowed within the radio amateur bands; the links are thus *not* power limited. The RAP of Delfi-C^3 however only performed at 1200 Baud, up- and downlink. The primary transceiver of Delfi-n3Xt is based on the RAP. As such, the primary communication link is **complexity limited**. But, as explained above, the frequency spectrum should be used respectfully and an effort should be made to reduce the required bandwidth, *especially on the downlink*. The uplink spectrum only affects a local region of the Earth and within very limited timeslots. Also, higher data rates than 9.6 kBaud are not allowed due to the limited available bandwidth.

A detailed analysis of existing modulation schemes has been performed. For Delfi-C^3, modulation schemes of minimum complexity were chosen. Also, as these modulation schemes have been successfully implemented, relative complexity is even less. As such, it makes sense for Delfi-n3Xt to reuse the modulation schemes applied in Delfi-C^3 [SLR 0014, sections 8.3-8.4]:

- *Downlink*: BPSK or binary phase shift keying
- *Uplink*: FSK or (binary) frequency shift keying

Related to the choice of modulation scheme are **line coding** and **pulse shaping**. Line codes specify the manner in which consequent bits are represented; normally consequent 'high's' and 'low's' represent 1's and 0's respectively. An adapted line code can allow lowering demodulation complexity at the cost of power or bandwidth efficiency, amongst others. Line codes are introduced in [SLR 0014, section 7.3]. In order to reduce the demodulation complexity, the following line codes are applied, with their indicated 'price':

- *Downlink*: NRZ-S (-0.8 dB of link margin)
- *Uplink*: NRZ-S and Manchester (-3.8 dB of link margin and a doubled symbol rate in Baud)

Pulse shaping serves to lower the produced bandwidth spectrum, but requires a certain additional modulation complexity [SLR 0014, section 7.4]. It involves shaping a (normally rectangular) bit shape to a smoother one. The consequently modulated signal will then have a lower frequency spectrum. For the VHF downlink, it is decided to apply pulse shaping:

- *Downlink*: (time-domain) raised cosine

For both the uplink and downlink the possibility of increasing data rates has consequently been assessed, by analyzing the hardware implications [SLR 0014, sections 8.3-8.4]. In result, the following targets are stated:

- *Downlink*: 2400-9600 b/s
- *Uplink*: 600-1200 b/s

High downlink data rates (9.6 kb/s) are likely achievable but should be proven by means of final integration tests. The weakest link is the *microcontroller*. At the same time, the data budget should be updated to determine the operational need or use for high data rates. For the uplink, a higher data rate requires a replacement of the *demodulator chip*. For data rates higher than 1.2 kb/s (2.4 kBaud) the processor would likely become a bottleneck, so that 1.2 kb/s is set as a maximum. TCs can in any case be transmitted using both 0.6 and 1.2 kb/s, so that an increase of data rate would serve mainly as proof of performance.

In both cases, there are no reasons to have a data rate varying during the mission *as the link budget closes for all elevation angles*. This is further explained in [SLR 0014, subsection 8.3.5].

2.1.4 Data protocol

A data protocol adds digital data to the content that is to be transmitted, in order to [SLR 0014, section 7.6]:
- Facilitate demodulation
- Ensure the proper reception of the digital content

In order to achieve this, a data protocol introduces a certain *overhead*. This overhead is expressed as a number of bits and is dependent of frame size. In case of Delfi-C^3, the AX.25 protocol was used. This protocol requires an overhead of:
- *Data fields*: 160 bits per frame
- *Bitstuffing*: An additional 1 bit per 6 bits (the flag is not allowed to occur within the frame)

In case of Delfi-C^3, an additional counter was added to each frame. [SLR 0014, section 7.6] introduces the concept of **throughput efficiency** of a data protocol, taking into account the frame length (relative overhead) as well as the anticipated amount of transmission errors (rejected frames). In result the throughput efficiency for the Delfi-C^3 downlink was only **66%**. If packet sizes would have been increased to 4800 bits with the same protocol, this efficiency could have been **76%**.

For Delfi-n3Xt, a customized replacement is proposed; the DelfiX protocol. It is customized for bi-directional (satellite) communication between two transceivers. It requires only an overhead of:
- *Data fields*: 80 bits per frame
- *Bitstuffing*: An additional 1 bit per 14 bits

The DelfiX protocol attains, for data frame size sizes of 1200 bits, throughput efficiency of **82%**. If frame sizes are increased to 4800 bits, the throughput efficiency becomes **86%**. A frame size of 2400 bits yields the same efficiency.

The data fields of the DelfiX protocol are shown in Figure 2.3 below.

Flag *2 bytes*		**Sender Address Field** *3 bytes*						**Frame Type** *1 byte*	**Information Field** *n bits*	**FCS** *2 bytes*	**Flag** *2 bytes*	
01111111	11111110	1	7-bit ASCII character	1	7-bit ASCII character	1	7-bit ASCII character	8-bit number		16-bit CRC value	01111111	11111110

Figure 2.3: DelfiX frame, optimized version [SLR 0014, Figure 8.9]

In summary:
- **Flags**: The length of the flags has been doubled with respect to the AX.25 protocol. This requires *less bitstuffing* over the entire frame content; a maximum of 1 per 14 bits as opposed to 1 per 6 bits.
- **Sender Address field**: The address field has been reduced from 14 bytes to 3. Satellite communication requires only a sender to be known (either satellite or GS) and requires fewer characters to do so. For Delfi-n3Xt and the Delft GS respectively DNX and DGS are suggested.
- **Frame Type**: This type field has been added for clarity; Delfi-C^3 communication required a separate indicator as part of the frame to indicate the content type of the frame (e.g. TC, TM1, TM2, etc.).
- **FCS**: The frame-check sequence is equal to that of AX.25. Three unique errors can be detected in message sizes up to over 32,000 bits, and 99% of all larger amounts of errors.

Reducing the amount of bitstuffing has effects on demodulation facility as less bit transitions are guaranteed. Tests using the digital demodulation core of Rascal/DUDe however have shown that flag sizes of at least 25 bits should be no problem [SLR 0014, section 8.5].

The above data protocol should be applied to two scenarios; the *TM downlink* and *TC uplink*. Of course other applications are possible, a *TM request frame* and *software upload frame* are discussed in [SLR 0014, section 8.6]. The two main scenarios require different additional fields and (data) operations and are discussed in more detail below.

For the **TC uplink**, the TC can be specified [SLR 0014, section 8.6], see Figure 2.4:

Encryption

Flag *2 bytes*	Address Field *3 bytes*	Frame Type *1 byte*	TC Code *1 byte*	Telecommand *8 bytes*	FCS 2 *2 byte*	FCS *2 bytes*	Flag *2 bytes*
01111111 11111110	3 ASCII Characters	8-bit number	8-bit number	64-bit value	16-bit CRC value	16-bit CRC value	01111111 11111110

Figure 2.4: DelfiX TC frame construction [SLR 0014, Figure 8.14]

From left to right a number of aspects can be seen:

- **TC code**: See below.
- **Telecommand**: The actual telecommand is assumed to be 64 bits long. For Delfi-C^3 the same value was used. Not all bytes were used for Delfi-C^3 but more TCs might be defined for Delfi-n3Xt. Also, size is not much of an issue on the uplink.
- **Encryption**: An ARC4 cipher with a certain offset is applied over the actual TC.
- **FCS 2**: Delfi-C^3 applied multiple different CRCs for TCs, also depending on the type of TC. For Delfi-n3Xt, a clearer approach is suggested. This is to have only *one* additional CRC to ensure proper transmission. The selected CRC is optimized for message lengths up until 151 bits, and allows detecting 5 unique errors. Both FCS's in a TC frame each allow for 99% of larger numbers of errors to be detected as well, guaranteeing near-faultless error detection.

Additional data aspects:

- **Preamble**: It is suggested to add 20 flags before the transmission of multiple TCs
- **Repeat**: It is suggested to repeat every TC at least 2 or 4 times, if not (much) more, to ensure proper reception

For Delfi-C^3, the (default) telecommand procedure requires 1) the transmission of a TC, 2) a confirmation on the downlink of the exact received TC and 3) the transmission of a TC confirmation. This ensures that people do not try to command the satellite by copying the transmission signal (without knowledge of the protocol) *and* gives extra reception verification.

For Delfi-n3Xt, it is suggested that functionalities are separated because verification of the message is sufficiently performed by the CRCs. The approach is therefore suggested to be changed. Whenever a TC is sent up, the **TC code** that is sent up should match the TC code which was previously sent or is being send down to the ground via the TM downlink. After a successfully received TC, the satellite then changes this TC code. This similarly ensures people cannot command the satellite without knowledge of the protocol. At the same time, commanding the satellite *only requires one* properly received TC.

For Delfi-C^3, blind commanding was also possible, requiring only a single transmission. For Delfi-n3Xt, in case commanding without a required TM downlink is desirable, a predefined and semi-constant *golden TC code* can be used, which can be changed during operations. This gives additional security over Delfi-C^3 operations as the blind command signal can change over time (preventing copying of the transmission signal).

The TC code can also be used as the *offset* for the encryption applied, as mentioned above when introducing the applied encryption.

For the **TM downlink**, the content of the packages should still be fully specified by the CDHS systems engineer. Some fields are however suggested in [SLR 0014, subsection 8.6.2], such as a:

- Unique frame ID (5 bytes)
- TC code (1 byte)
- Last processed TC *type* and *address* (2 bytes)

The frame size of the downlink can be determined taking into account data rate, throughput efficiency and speed of internal data generation. From a transmission point-of-view, throughput efficiency is most important assuming the satellite is rotating only slowly. Taking into account the optimum frame size and the knowledge that Delfi-C^3 successfully applied 1-second frames, it is suggested that:

- Frame sizes of **2400** bits are used for a data rate of **2.4 kb/s**
- Frame sizes of **4800** bits are used for data rates of **4.8 kb/s or higher**

Of course, the CDHS does not have to supply a frame every 0.5 second, even if the frame rate is 0.5 frame/s; it can simply supply (the content of) two frames *once* every second. For Delfi-C^3, a *flag frame* was also added every 4 other frames due to the limited throughput of the command and data handling system. This reduced overall data efficiency by another 20%. For Delfi-n3Xt, this frame is not deemed necessary. Additional flags are only transmitted in between frames when non-maximum bitstuffing allows for them to be added.

2.2 Primary radio system on Delfi-n3Xt

On the satellite, two components are essential:

- **PTRX**: The Primary Transceiver (PTRX) is exactly what the name implies. It adheres to all functionalities established above. It interfaces digitally with the on-board computer (OBC), and in an analogue fashion with the *antenna system*.
- **Antenna System**: The antenna system consists of those components required on the satellite other than the PTRX, in order to establish a communication link.

As a back-up radio the ITRX payload is used. If necessary, this payload could also be replaced by a second PTRX. As the design of the ITRX is not the responsibility of the TU Delft, it will not be further discussed below. The ITRX shall be prepared to use the above data protocol, as has been discussed with ISIS.

OLFAR might be flown as a technology demonstration. It is an integrated circuit designed to realize the reception of low-frequency signals over the antennas, in order to transmit these over the VHF band. The PTRX should take care of final preparation for transmission, but as [SLR 0014, chapter 11] argues: if OLFAR is to be flown, the bulk of the required circuitry should be integrated within the antenna system.

2.2.1 PTRX

The PTRX is based on the design of the RAP. Nevertheless, a number of things have been changed:

- **Only digital interfaces to the OBC**: The RAP had an analogue line between the transceiver and the OBC. In other words, the OBC was required to extract the actual (digital) TC. The PTRX takes care of all actions itself and thereby form a more *modular radio*.
- **Separated TX and RX sections**: By using two different processors, the TX and RX sections are completely separated functionally and electrically. Only the transponder forms a direct link between the two. Using two separate *standard system bus* interfaces, this allows switching off one without the other.
- **Increased transmission data rate of possibly 9.6 kb/s**: Analysis of the design has shown that the only component affected by a change of data rate is the processor. As a more powerful MSP430 processor is used for the PTRX compared to the PIC of the RAP, higher data rates are certainly possible. It is assumed that a data rate of 9.6 kb/s can even be achieved.
- **Improved electrical design**: A number of filters, crystals and other components have been changed. This has improved (temperature) stability, performance and areal requirements.

Possibly, the reception data rate can also be increased as explained, depending on the final receiver design.

The following Figure 2.5 clearly shows the separated sections and their interfaces.

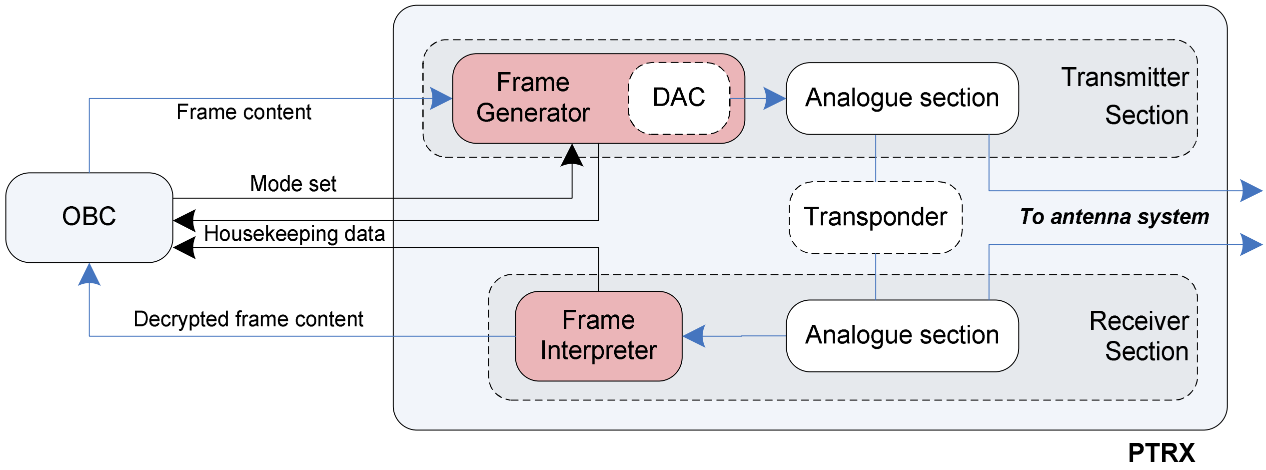


Figure 2.5: Integration of the PTRX processors [adapted from SLR 0014, Figure 9.11]

In the middle of the figure the transponder can be seen, with its analogue link between the RX and TX sections. The OLFAR interface is not shown, but would provide an analogue signal to the transmitter section.

The *Frame Generator* is able to switch between modes, to activate either one of:

- TM mode
- Transponder mode
- OLFAR mode
- TX OFF mode

Neither one of these modes switches off the receiver; the latter should always be on.

A much more detailed diagram has also been made and can be found in [SLR 0014, Figure 9.8]. It shows:

- **All main functional components** (amplifiers, mixers, oscillators, processors, etc.) including the components required for the generation of housekeeping data.
- **All required interfaces**. This includes those required for the transmission and reception of data, for the generation of housekeeping data and for switching between transmission modes.

2.2.2 Antenna System

All analogue RF interfaces of the PTRX (and ITRX) connect to the Antenna System. The antenna system logically receives and transmits these signals from and to the wireless medium. The antenna system supplements the PTRX with:

- Antennas
- Phasing circuit
- Coax cables and connectors

The **antenna** configuration (per frequency band) is assumed to be identical as for Delfi-C^3; applying four antennas in a *canted turnstile* configuration. The resulting configuration is illustrated by Figure 2.6 below. This configuration approaches an *omnidirectional gain pattern*. The antennas are deployed using Modular Antenna Boxes, or MABs. The MABs are part of the heritage of Delfi-C^3. These MABs in turn are integrated on the DAB, the Deployment and Antenna Board, which is a PCB. The MABs and the DAB can also clearly be seen in Figure 2.6. This DAB is placed at the bottom of the satellite, logically with the openings of the MABs facing outwards.

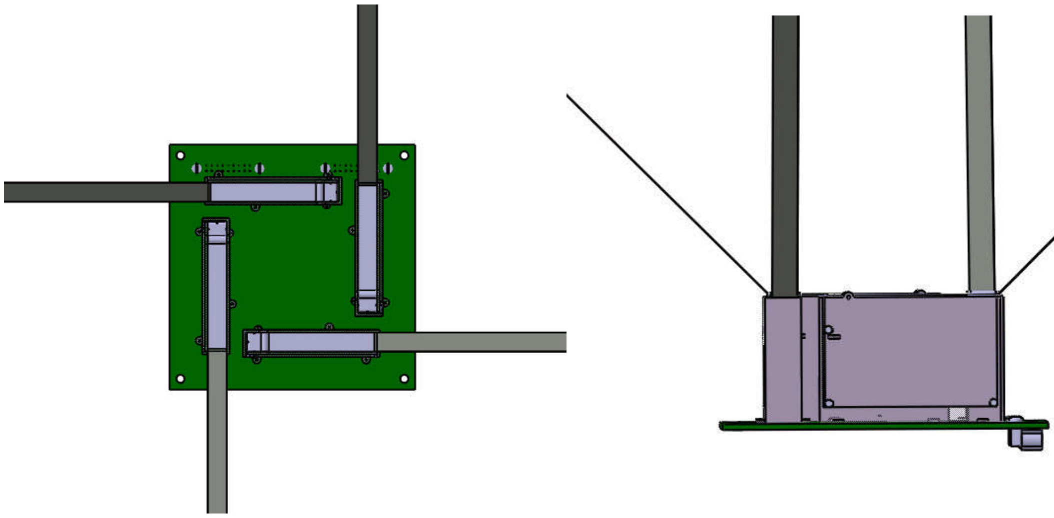


Figure 2.6: DAB including MABs and antennas; top-view (left), side-view (right) [SLR 0036, Figure 3.5]

The UHF/VHF antenna system of Delfi-n3Xt will be similar in concept to that of Delfi-C^3, with one important change; no separate UHF and VHF antennas will be used but **combined antennas**. There are two good reasons for this [SLR 0036, chapter 3]:

- *Only four antennas and deployment mechanisms are required*; this allows a volume saving of about 10% of the satellite given the height of the MABs of about 3 cm, next to a reduction of mass and complexity.
- *All components can be located on one side of the satellite*; this additionally reduces the amount of components required and complexity involved.

The combination of the antenna is possible because the wavelength of the UHF transmission frequencies (~69 cm) is more or less one third of the VHF wavelength (~206 cm). Maximum power performance can be obtained for antennas of either a quarter-wavelength or three-quarter-wavelength. The length of the VHF antennas would satisfy both constraints for VHF and UHF frequencies respectively. Two effects of combining the antennas should be prepared for:

1. The VHF transmission signal should not interfere with the UHF signals to be received
2. The length of the UHF antennas becomes slightly non-optimal

The gravity of the first effect has been measured using the Delfi-C^3 spare model. The power amplifier of the RAP/PTRX in fact causes harmonics of the transmission frequency to have significant signal strength. According to the link budget, a worst case reduction of -116 dB is required due to the differences in signal strength between up- and downlink. The third harmonic, which might interfere with the UHF frequencies, has measured signal strength of *-58.20 dB* with respect to the main carrier. However, the third harmonic of the VHF transmission frequency and the UHF transmission frequency are in fact spaced apart by *~2 MHz*. A signal strength reduction of *another -40 dB* of the third harmonic has been measured within 17.5 kHz; lower values could not be measured due to the noise floor of the test set-up. In conclusion however, it seems highly likely the worst case required reduction is met within 2 MHz.

The second effect results in impedance mismatch loss if not corrected for. Given the fact that the VHF antenna length is only slightly non-optimal for the UHF frequencies (2 MHz / 436 MHz ≈ 2%) circuitry can be introduced to compensate for this. In fact, this is one of the roles of the new Delfi-n3Xt *phasing circuit*.

The **phasing circuit** is required to:

- Introduce proper phase shift between the antennas for the resulting gain pattern to be omnidirectional and *circularly polarized*
- Isolate the UHF frequencies from the VHF frequencies
- Correct for the non-optimal length of the UHF antennas when combined with the VHF antennas

The first two aspects were in fact also required of the Delfi-C^3 phasing circuit. However, as the antenna systems were separated the interference of the VHF transmission was less severe. The updated Delfi-n3Xt phasing circuit has been tested and built by Wolter van der Kant. Some losses are incurred in the phasing circuit, but these are comparable to those on Delfi-C^3. The phasing circuit is also integrated on the DAB.

Coax cables and connectors logically connect the PTRX (and ITRX) and the antenna system. Rather large vertical SMA connectors were used for Delfi-C^3, as well as smaller MMCX connectors to connect the MABs to the antenna board. Several suggestions for Delfi-n3Xt have been made in [SLR 0036, chapter 5]. The cable used can be the same as that used in Delfi-C^3 (RG178BU, Ø ≈ 2 mm), but a slightly thicker cable (RG-316, Ø = 2.5 mm) can instead be used to halve the signal attenuation losses; most important for the S-band.

Finally, **OLFAR** can be placed on the DAB. This would require only a small space and the DAB already has a standard system bus interface required for OLFAR operations. This would however require the DAB processor to be used other than during deployment and to be programmed to operate the OLFAR chip. If OLFAR is placed on the DAB, an additional coax cable would be required to connect OLFAR and PTRX.

2.3 Ground station

On-ground, transmission and reception is primarily taken care of by one system: the Delft ground station (GS). The Delft GS will be the only station that is configured to command the satellite. Radio amateurs worldwide however are motivated to receive and forward TM data transmitted by the satellite, as satellite transmission is continuous. In order to facilitate this, a software telemetry client named DUDe will be spread, similar in function to Rascal in the Delfi-C^3 mission.

The Delft GS has been analyzed in detail, in order to:
- Properly identify the effects of changes to the transmission techniques
- Specify all values in the link budget

The main components of the Delft GS can be seen in Figure 2.7 below. In fact, more components are necessary to operate the antennas and to relay the signals between the roof and the actual transceiver hard- and software. *All* components are introduced and discussed in more detail in [SLR 0014, chapter 15].

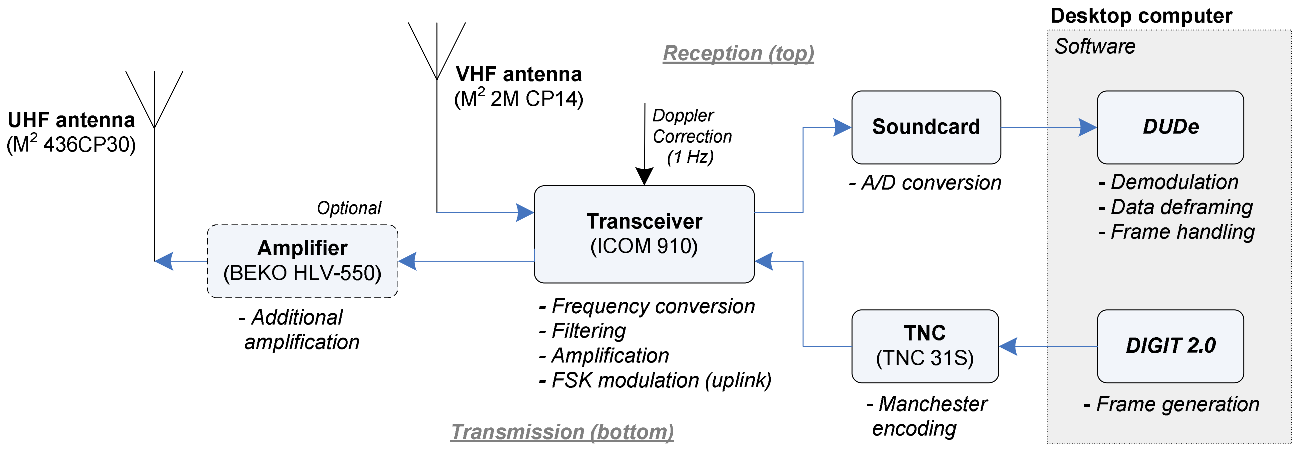


Figure 2.7: The Delft GS main equipment [SLR 0014, Figure 15.2]

The following updates to the GS are required for the Delfi-n3Xt mission:

Software:
- *DIGIT 2.0*: The software that has been written for Delfi-C^3 telecommanding should be updated.
- *DUDe*: Some parameters should be changed in the software to allow for higher data rates.

- *TNC 31S*: The software should be updated to allow for the use of the DelfiX protocol.

Hardware:

- *ICOM 910*: The transceiver should be replaced to allow for higher downlink data rates. A suitable replacement can be delivered by ISIS.
- *TNC 31S*: This component should be replaced if higher data rates are applied on the uplink. A suitable replacement can be delivered by ISIS.
- *Antenna tracking system*: Currently, the antenna sometimes has to rotate by 360° during mid-pass as the rotor begins halfway through its rotation range. This should be changed by updating the driving software or by changing the orientation of the (mechanical) rotator connection.

A further suggestion is to:

- ***Add error correcting capabilities to DUDe***: The CRC that is applied on the TM downlink can actually be used to correct for single errors, using a look-up table approach. It is rather computation intensive, but other than during a pass the GS computer performs no serious functions.

2.4 Operations

In order to fulfil all main tasks on the primary communication link, a number of COMMS modes are necessary. These COMMS modes are in fact combinations of all operational modes of the PTRX and ITRX. These modes are shown in Table 2.2.

Table 2.2: COMMS modes and respective PTRX and ITRX modes [adapted from SLR 0014, Figure 14.1]

COMMS mode	Effects
PTRX mode (default)	• **TCs** received by PTRX and ITRX and **TM** transmitted by PTRX
ITRX mode (secondary)	• **TCs** received by PTRX and ITRX and **TM** transmitted by ITRX
Transponder mode	• **TCs** received by PTRX and ITRX and Transponder activated
ITRX test mode	• **TCs** received by PTRX and ITRX and ITRX tests activated
OLFAR mode (experimental)	• **TCs** received by PTRX and ITRX and OLFAR activated
All TXs OFF	• **TCs** received by PTRX and ITRX

In order to asses the anticipated performance of the VHF downlink, the TM frames received from Delfi-C^3 and stored in the telemetry database have been analyzed. The entire analysis is described in [SLR 0014, section 14.2]. The main conclusions are:

- Almost **2%** of the maximum theoretically generated amount of data has been received over two years time of Delfi-C^3 operations.
- Radio amateurs have increased the received amount of unique data by **124%**, or when including non-unique (duplicate) data, even by **293%**.
- In the first month alone, radio amateurs have received more data than has been collected by the Delft GS *in over two years time*.
- The Delft GS is able to monitor passes at very low elevation angles, in some cases up to 1°.
- The Delft GS itself has been able to receive between **15% and 35%** of all data maximally transmitted within its range.

For Delfi-n3Xt, a higher performance is anticipated due to:

- A **more stable CDHS** that does not 'hang' until an eclipse has passed or a TC has been received
- A **battery** that possibly allows transmission to continue during eclipse

At the same time, radio amateurs should definitely be motivated to participate again.

3 High speed communication link

The high speed communication link is one that is aimed at *technology demonstration*. The link is not based on any heritage. The inclusion of a high-speed transmitter on Delfi-n3Xt should be pursued in order to:

- Have a proven S-band transmitter for future missions
- Increase the novelty and appearance of Delfi-n3Xt with respect to Delfi-C^3

The link is designed to achieve a data rate, or more specifically *bit rate*, as high as possible. At the same time, a working link can allow for the total Delfi-n3Xt mission data yield to be increased largely; this is however a secondary function. The high speed communication *down*link is a *uni-directional* one. The link is shown in Figure 3.1 below.

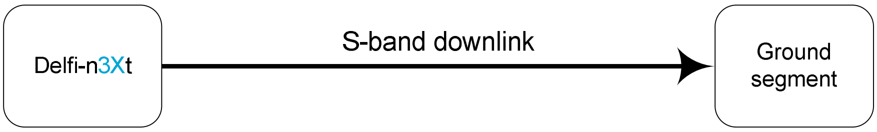


Figure 3.1: High speed communication link

This chapter follows the same structure as the previous one, discussing the link, the high speed radio system on Delfi-n3Xt, the GS and link operations (including *resulting data rate*) in four respective sections.

3.1 Link characteristics

3.1.1 Transmission band and frequency

The high-speed communication downlink will take place within the:

- S-band radio amateur band (2400-2450 MHz)

The radio amateur portion of the S-band overlaps with the larger ISM band (2.4-2.5 GHz). Within this band the bandwidth is *shared*; no restrictions in overall bandwidth exist as long as power values are sufficiently low. Radio amateur equipment is not allowed to disturb other equipment, which is of no concern for low-power satellite transmitters. Along these lines, available bandwidth is not limited, if properly used.

A discussion with AMSAT has lead to the conclusion that:

- Frequency allocation should be as close to the bottom of the band, while taking into account a maximum Doppler shift. This minimizes disturbances from other equipment to the satellite signal.
- The maximum data rate determines the required bandwidth and exact transmission frequency.

The following Figure 3.2 shows the preliminary transmission frequency and bandwidth as determined in [SLR 0014, section 12.1]. The bandwidth is based on a spectrum measurement of the transmission bandwidth of the modulator introduced below.

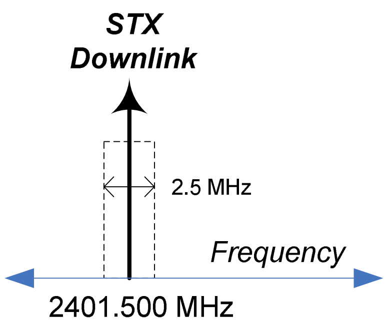


Figure 3.2: STX downlink frequency (bandwidth) [SLR 0014, Figure 12.2]

Furthermore, it has been agreed upon that even if the frequencies are filed for, it is no problem if it turns out a working transmitter will not be flown due to limited resources or technical constraints.

3.1.2 Modulation scheme

As illustrated by the above, the S-band communication link is not bandwidth limited. As bit rates as high as possible are strived for, the system is both:

- **Power limited and complexity limited**

In terms of power efficiency, a number of simple modulation schemes perform equally: MSK, BPSK and QPSK. The latter two however perform much worse in terms of bandwidth efficiency, if no pulse shaping is applied. Pulse shaping in turn requires a linear amplifier, as a non-constant envelope is introduced. The possible (future) use of high-efficiency non-linear amplifiers therefore leads to prefer MSK. Modulation schemes with better power efficiency are for MFSK schemes with $M \geq 8$; these are however very bandwidth inefficient and require increased complexity. Simple COTS MSK modulators are available, so that:

- The preferred modulation scheme for the S-band communication link is **MSK**.

No additional pulse shaping or line coding is required.

3.1.3 Forward Error Correction

Forward error correction (FEC) allows decreasing power efficiency at the cost of bandwidth efficiency; it involves adding extra symbols to the bit stream in order for the combined total to be demodulated with a lower required signal-to-noise ratio. When applying FEC, data rate goes up (stated in Baud) as opposed to bit rate (stated in b/s).

When it comes to introducing error correction, there are two other advantages on the S-band link:

- There is no need for real-time or little-delay communication
- The satellite involves only an encoder

These arguments are relevant as FEC coded signals logically take time to prepare and decode, and as most complexity is incurred in the decoder. In result, it makes sense to try and apply FEC to the S-band link. Three options are suggested in [SLR 0014, section 12.3], with the achievable coding gains indicated:

1. **Preset CC2500 FEC coding scheme** **2-3 dB**
2. **BCH block code** **3-4dB**
3. **The AO-40 FEC coding scheme** **7 dB**

The top option is the most simple to implement, with the bottom one being the most complex. In all cases however software is available in some form or another, and the coding schemes have been well-defined. As such, the preferred option logically is the AO-40 FEC coding scheme. The CC2500 FEC coding scheme forms an absolute minimum and can be directly applied using selected COTS hardware.

3.1.4 Data protocol

When FEC is applied, application of a data protocol becomes more complicated. This is due to the fact that the two functions of a data protocol, as introduced before, should be separated. These were to:

- Facilitate demodulation
- Ensure the proper reception of the digital content

The reason for the required separation is that facilitation of demodulation is required *before* decoding and content verification will take place *after* decoding. Also, an additional required functionality of the data protocol is to:

- Facilitate decoding: The data protocol should ensure that the decoder is in fact synchronized with the received encoded signal.

The raw data that is to be send over the S-band is assumed to be the *exact same* as that to be send over the primary communication link. As such, raw data consists of the *information field* and *frame type* fields from the DelfiX protocol. Figure 3.3 below shows the general data protocol functionalities that should be performed on this received data, irrespective of the applied FEC coding scheme. On-ground, these functionalities are performed in reverse.

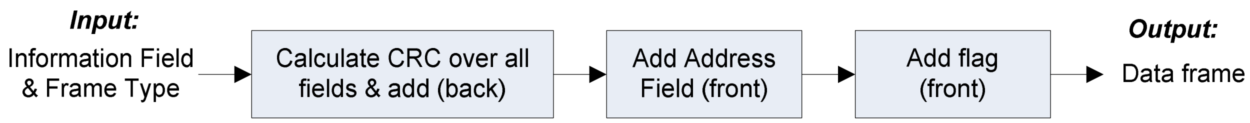


Figure 3.3: Primary data protocol operations [SLR 0014, Figure 12.3]

These fields are in fact identical in name and content to those of the DelfiX protocol. In case the AO-40 FEC coding scheme is applied, the following Figure 3.4 demonstrates the functionalities additionally required.

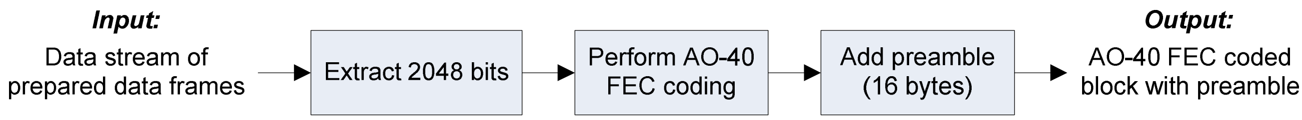


Figure 3.4: Secondary data protocol operations [SLR 0014, Figure 12.4]

The preamble as added in the scheme above is in the form of a stream of alternating 0's and 1's. The exact operations of the AO-40 FEC coding scheme are introduced in detail in [SLR 0014, section 7.8]. Similarly, data protocol operations for the other FEC coding schemes are presented in [SLR 0014, section 12.4].

3.2 High speed radio system on Delfi-n3Xt

On Delfi-n3Xt again two configuration items can be distinguished:

- **STX**: STX is short for S-band transmitter. It receives digital data from the OBC and delivers analogue data to the *antenna system*.
- **Antenna System**: For the S-band, a completely separated antenna system is required.

3.2.1 STX

The transmitter that is to provide communication over the S-band is the STX. Significant effort of the author has been put into designing the STX. The ambition of the STX is to provide a bit rate of *at least* 9.6 kb/s.

Four design goals have been established for the STX:

- **No satellite pointing shall be required**
- **The STX shall have power consumption no higher than the PTRX (~1.75 W currently)**
- **Commercial-off-the-shelf (COTS) components shall be used whenever possible**
- **All components in the design should be as modular as possible**

[SLR 0387] uses the minimum data rate and these design goals to establish required functionalities, derive from these the required component types, and finally establish for every component type a suitable component.

The following Figure 3.5 shows the resulting suggested component diagram.

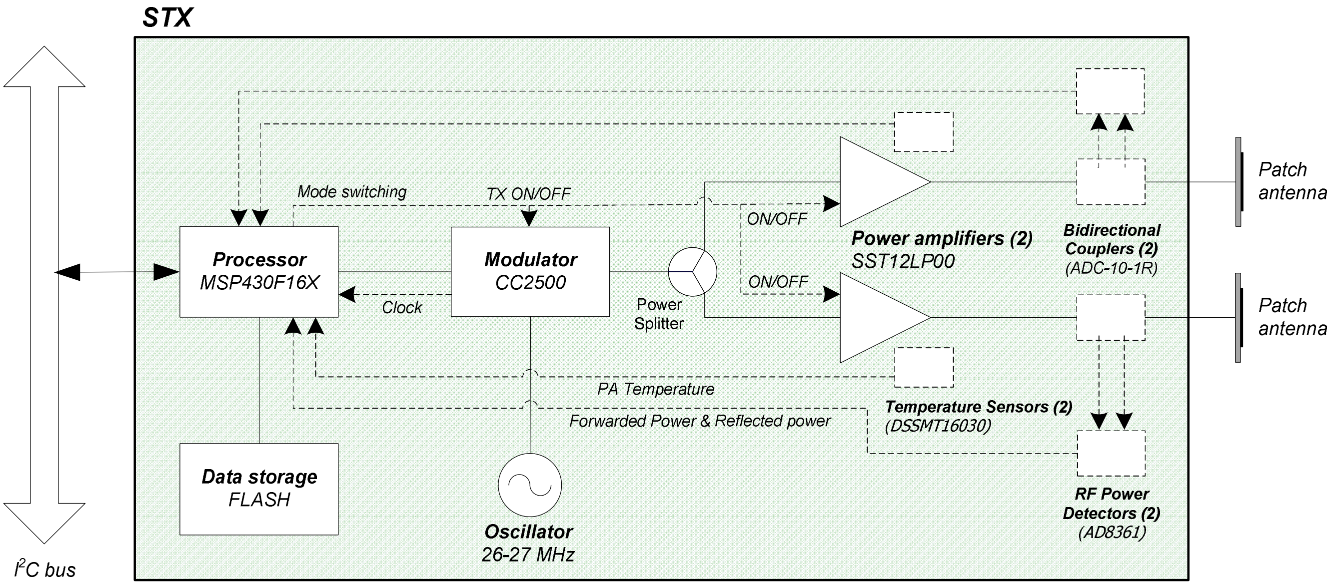


Figure 3.5: STX hardware components and their interfaces [SLR 0387, Figure 13.1]

The following components can be identified:

- **Processor**: The processor anticipated is the default Delfi-n3Xt processor. It will therefore require minimum integration complexity and enjoys the radiation proven performance of the latter.
- **Data storage**: External data storage is required as integrated processor memory is not sufficient. FLASH memory allows for large storage sizes and COTS components exist in the form of SD cards for example. At the same time, FLASH memory is a form of non-volatile storage, so that data is not lost in an eclipse without continued operations.
- **Modulator**: The selected COTS modulator chip delivers MSK up until data rates of 500 kBaud. It is designed for optimal cooperation with the MSP430 processor.
- **Oscillator**: *One* 26-27 MHz crystal is integrated to provide the required clock signals to both the modulator and processor, at the same time providing the frequency required to up-convert the transmission signal to RF.
- **Power amplifiers (2)**: Two high-efficiency amplifiers (58%) can *each* provide up to 25 dBm or 315 mW of power. This allows making optimal use of low-power COTS amplifiers as well as two patch antennas, as explained in the next section.
- **Housekeeping sensors (2)**: Three components are used to provide three values per signal path: a temperature value of the PA and values of power reflected by and forwarded to the antenna system. The components in the top half of the diagram are logically equal to those at the bottom half.

In the above figure it can be seen that the processor is able to:

- Switch the transmission of the CC2500 on and off
- Switch one *or* two PAs on and off using integrated on/off switches

This allows the processor to switch between operational modes *and* submodes as introduced below. At the same time, the processor receives sensor data from the housekeeping sensors.

The final STX is expected to consume ~100 mW during storage operations, ~700 mW with one PA activated and ~1450 mW with two PAs activated [SLR 0387, section 14.3]. All components, most notably processor, data storage, modulator and power amplifier perform one main functionality and have simple interfaces. When the design constraints change in the current or a future mission, these components can easily be changed. All components are COTS components. The independence of satellite pointing is further illustrated below. In conclusion, the above configuration *satisfies all four design goals as stated*.

The STX is most logically integrated on a PCB similar to the PTRX. However, given the low number of required components, it might be possible to opt for a more exotic approach. One can think of integration on the backside of the ground plane of the S-band patch antennas. Of course, these possibilities will depend on the final electrical design and its physical extent. For Delfi-n3Xt, they are currently not assumed as volume is no problem and regular PCB integration simplifies mechanical integration.

3.2.2 Antenna system

Several types of antenna exist and can be used to transmit S-band signals. Parabolic and helical antennas would however require sizeable volume on-board of the satellite. Wire antennas can be smaller, but would require to be deployed. As a result, patch antennas are the logical choice. Also, all (three) successfully operated CubeSats with S-band transmission capability have applied *patch antennas*. A number of larger microsatellites similarly did so.

Two available patch antennas are found and presented in [SLR 0036, chapter 4]. Both patch antennas function in the proper frequency range and deliver right-handed circularly polarized radiation. The measured gain patterns of the patch antennas for or angles off-boresight of $-90° \leq \theta \leq 90°$ are plotted in Figure 3.6 below. The patterns are approximately rotationally symmetric. For a definition of the axes and angles, see [SLR 0036, chapter 4].

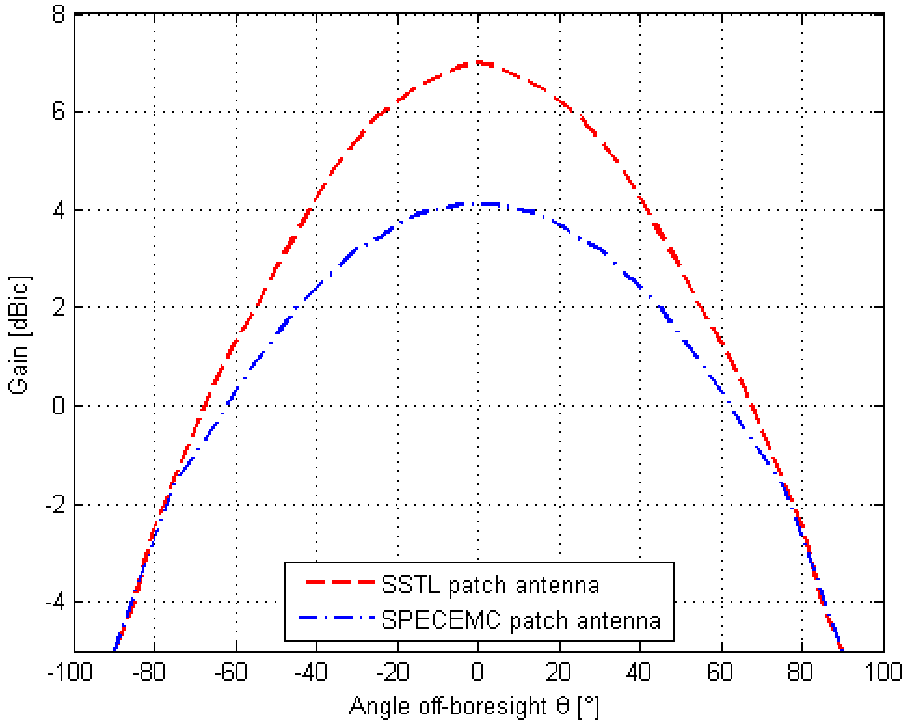


Figure 3.6: Gain values in dBic for different angles off-boresight of the SSTL and SPECEMC patch antennas [SLR 0036, Figure 4.2]

It can be seen that the two patch antennas found, labeled the SSTL patch antenna and SPECEMC antenna have similar gain patterns, although that of the SSTL patch antenna is more directed at no cost of gain at lower angles off-boresight. The antenna is however a lot larger and heavier than the other. The gain performances of the two patch antennas have been used to draw conclusions on possible antenna configurations. As mentioned by one the design goals of the STX, no satellite pointing shall be required for the operation of the latter. As such, the most logical antenna gain configuration is *an omnidirectional one*.

Patch antennas only radiate significantly to one side, perpendicular to the antenna. As such, an omnidirectional pattern gain be achieved using either two or six patch antennas, applied to respectively two opposite sides or all six sides of the satellite. Logically, the application of six patch antennas would become very difficult on a satellite as Delfi-n3Xt due to reasons of area, volume, mass and complexity. Therefore *two patch antennas should be applied on two opposite sides*. In fact, this is the configuration applied by one of the three successful CubeSats mentioned above, CanX-2. The other two, GeneSat and Pharmasat used just one patch antenna for sporadic communication possibilities.

The application of two patch antennas can ideally be combined with the use of two PAs on the STX. In this manner:

- The power output of the STX is effectively doubled, while satisfying power constraints of Delfi-n3Xt
- The two different amplified signals do not have to be *exactly* in-phase as the signals will feed two antennas on opposite sides of the satellite; no negative interference will occur.

Using this implementation, three options become possible, *switchable by telecommand*:

- **Omnidirectional pattern, two PAs activated**
  The minimum gain value is -8 dB, but the use of two PAs effectively adds 3 dB.
- **Near-omnidirectional pattern, two PAs activated**
  With the available patch antennas, a minimum gain value of -5 dB can be assumed for ~80% of the time assuming a (slowly) rotating satellite (other minimum gain values can of course also be chosen). Again, the use of two PAs effectively adds 3 dB.
- **Hemispherical pattern, one PA activated**
  One PA and thus one respective patch antenna can be used at a time. This *lowers overall power consumption*, with a minimum gain of -5 dB within just one hemisphere.

In result, a very flexible solution is proposed, allowing maximum amplification using high-efficiency COTS amplifiers and switching between antennas for pointed or semi-pointed operations (switching off the patch antenna that does not happen to look at the ground station). At the same time, the implementation of two patch antennas and PAs adds failure-resistance.

On which sides the patch antennas should be attached remains to be determined. In fact, the largest insecurity is the final solar panel configuration. Currently work is performed on determining the Delfi-n3Xt solar panel configuration by Johannes Bürkle. Solar panels will incur negative reflections if placed in front of the antennas. Nevertheless, two general options for the patch antenna integration:

- **One patch antenna at the TOP, one patch antenna at the BOP** (top panel and bottom panel respectively; these are the short sides of Delfi-n3Xt)
- **Two patch antennas on opposite long sides**

On all suggested sides some area is available, as well as some volume 'behind' the side. The SSTL patch antenna as introduced above is likely too large as it requires 80x80x20mm. The SPECEMC patch antenna is significantly smaller as it requires some 45x45x7.5 mm, where the size and thickness of the ground plane can still be adjusted.

A test model of the SPECEMC patch antenna has been requested and has been sent from the Germany office to the Netherlands. Whenever it arrives, performance testing can be performed. Also, it should be determined whether the design is space-worthy.

Finally, the S-band antenna system requires **cables and connectors** as does the UHF/VHF antenna system. The same cable and connector suggestions apply that have been made in the previous chapter for the latter antenna system.

3.3 Ground station

In order to receive the S-band signal, a dedicated ground station set-up is necessary. Currently, no S-band equipment has been installed at the Delft GS. A minimum equipment set-up can be seen in Figure 3.7 below. Suggestions have been taken from the components used in the UHF/VHF set-up.

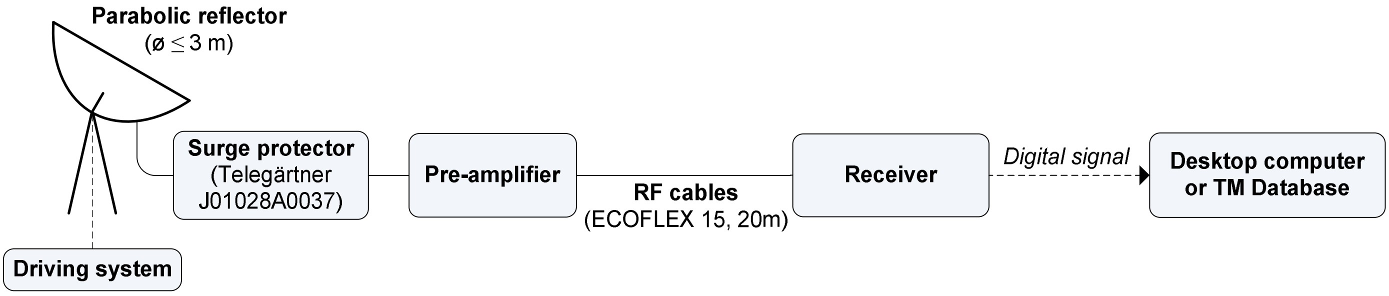


Figure 3.7: The proposed minimum S-band ground station set-up [SLR 0014, Figure 15.4]

An introduction to all components indicated above can be found in [SLR 0014, chapter 15].

3.4 Operations

As already suggested above, the STX will have two main modes. These are:

- **TX OFF mode**
- **TM mode** (1 *or* 2 PAs active *and* an adjustable data rate)

A complete OFF mode is automatically provided by the standard system bus connection as the 'PCB' can be switched off. The default mode is the TX OFF mode if the final power budget allows for it. In this mode, no transmission shall take place but data sent to the PTRX or ITRX (as well) is continuously stored in the data storage of the STX. When however Delfi-n3Xt is above Delft, the TM mode can be activated. It is most desirable to do this by means of a large number of *submodes*. The option to activate either one or two PAs has already been commented on above. The preference to adjust the data rate is further explained below.

Activation of the STX should be done by means of telecommand. If possible, the STX should be *scheduled by telecommand*. This would allow for the STX to be activated automatically and in advance, making optimal use of a possible pass. The STX will, due to its experimental status, not likely be integrated in nominal operations. Optimal use of the data storage could be made if specific data could furthermore be requested by means of telecommand.

The most significant loss incurred in the link budget is the path loss. This path loss however changes drastically during a pass; between an elevation angle of 10 and 90° the path loss changes reduces over 10 dB. This 10 dB equals an achievable data rate of a factor 10(!).

The following Figure 3.8 shows the normalized data volume that can be downloaded for a pass, given its *off-zenith angle* (for its definition, see [SLR 0014, section 7.9]), in case:

- The data rate is constant and based on a closed link budget for a minimum elevation
- The data rate is optimally changed between passes to maximize downloadable data volume during that pass, but kept constant during the pass
- The data rate is continuously changed during a pass (theoretically maximum downloadable data volume)

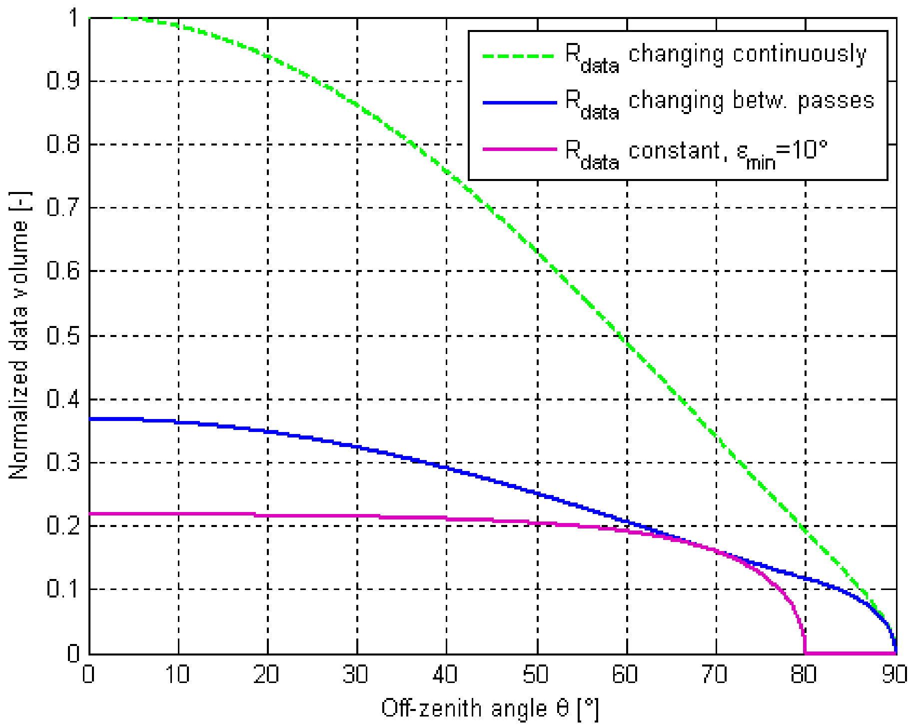


Figure 3.8: Maximum downloadable data volume versus off-zenith angle, plotted with two other scenarios [SLR 0014, Figure 7.54]

It can be seen that very large differences in achievable performance exist. It should however be said that the conventional approach of basing the data rate on a low minimum elevation angle is still the best approach to maximize downloadable data content if, and only if:

- The *data rate is kept constant* throughout the mission *and* all passes are utilized and monitored

The difficulty in *changing the data rate* lies with:

- Having an amount of data available that allows for a varying data rate transmission
- Having a transmitter and receiver that are able to change the transmission data rate
- Having intelligence to determine the data rate

For the STX however:

- Data storage basically forms an infinite pool of data
- The CC2500 allows for a very large number of data rate presets to be defined, and variable on-ground data rate reception is likely neither a problem
- Telecommands should be used to switch on the STX anyhow

This last aspect is where the notion of *submodes* comes in. Whenever the STX is activated, it makes sense to command it to do so with a certain data rate. The intelligence will then be located on-ground. In this manner, data rates can easily be changed in between passes. In other words, the blue (middle) line in Figure 3.8 can be approached.

[SLR 0014, section 7.9] presents the theory required to establish the optimum data rate (by means of the minimum elevation angle), given the off-zenith angle θ of a pass.

If possible, the transmission data rate of the STX can even be changed *during* a pass. This can allow increasing the downloadable data volume by adjusting the data rate:

- A single time, to correct for the momentarily attitude of the satellite (related to antenna directivity)
- Continuously, to compensate for the decrease in path loss

Of course, changing the data rate continuously or following discrete steps during a pass requires more complicated automated or manual procedures. Nevertheless, feasibility can be tested with the STX.

It remains to determine what data rates can actually be achieved on the S-band communication downlink. The link budget [SLR 0106] again presents in much detail the anticipated losses, gains and noise values, and presents achievable data rates for a *pessimistic* and *optimistic* case. The main differences between the two cases are the:

- Ground antenna directivity: either 22 or 32 dB depending on final antenna size
- Coding gain: either 3 or 7 dB based on the above FEC coding schemes

The optimistic case *does* however take into account a minimum antenna gain value based on a closed link budget for *all* satellite attitudes. As explained above, 3 dB gain can be achieved by accepting the link budget to close for only ~80% of the time. The following Table 3.1 presents the achievable data rates for the *optimistic case*, depending on the final orbit altitude of Delfi-n3Xt and the elevation angle at which communication is commenced.

Table 3.1: Optimistic STX implementation [adapted from SLR 0014, Figure 12.4]

Orbit altitude	*h* = 600 km		*h* = 850 km	
Minimum elevation	10°	40°	10°	40°
Bit rate	93 kb/s	576 kb/s	57 kb/s	298 kb/s
Symbol rate	233 kBaud	1441 kBaud	143 kBaud	744 kBaud

In the above table it can be seen that very sizable data rates can be achieved, of course assuming the optimal FEC coding scheme as well as a 3 m parabolic reflector as ground antenna. It is important to note that higher data rates than 500 kBaud are not supported by the modulator (CC2500). Similarly, the processor (MSP430) will have a finite maximum output data rate.

Given the fact that sizeable bit rates might be reached, it is tempting to try and see if the STX can in fact provide *all data generated in the Delfi-n3Xt mission* to be downloaded, given correct operations and using a single ground station. The following Table 3-2 presents the data rates required to download all data for passes with different characteristics; these characteristics in turn specify the indicated pass time. Multiple PTRX data rates are indicated, where for each data rate it is assumed that the entire data rate is used for the transmission of unique data. A 1.2 kb/s PTRX data rate will yield about 1050 b/s of raw data, when removing the application of the DelfiX protocol. In fact, the transmission by the STX also requires a certain data protocol overhead but this is ignored for the moment. The maximum achievable data rates as presented in Table 3.1 are repeated in Table 3-2 in *italics*. The required data rates that are less than the maximum achievable data rates are indicated in **bold**.

Table 3-2 illustrates that in fact it might be possible to download all mission data, but only if the unique data generation within Delfi-n3Xt will end up being between 1.2 and 2.4 kb/s.

Table 3-2: STX required bit rates for a complete data dump, using a constant data rate [SLR 0014, Figure 14.5]

	Required constant data rate			Maximum achievable data rate
PTRX data rate	1.2 kb/s	2.4 kb/s	9.6 kb/s	-
Daily data volume	90,029 kb	186,278 kb	763,776 kb	-
All passes, ε_{min}=10°,h=850 km → *48 min*	**31 kb/s**	62 kb/s	249 kb/s	*57 kb/s*
All passes, ε_{min}=10°,h=600 km → *31 min*	**49 kb/s**	98 kb/s	391 kb/s	*93 kb/s*
All passes, ε_{min}=40°,h=850 km → *7 min*	223 kb/s	445 kb/s	1.8 Mb/s	*298 kb/s*
All passes, ε_{min}=40°,h=600 km → *4 min*	401 kb/s	802 kb/s	3.2 Mb/s	*576 kb/s*

The above table however does not take into account a data rate that changes *during* a pass, or even *between passes*. Therefore, if indeed the STX is continuously operated, and the intelligence is implemented to actually optimize the data rate per pass, Figure 3.8 above shows that the total downloaded data volume can be increased to anywhere between two times and five times. In that case, it might be possible *to downlink all data generated within the Delfi-n3Xt mission*.

Finally, if the final ADCS will allow for some sort of satellite pointing, the effective gain of the communication link can be increased by anywhere between 8 and 10.5 dB, as concluded by looking at Figure 3.6 above and assuming a 30° pointing accuracy.

In conclusion, the S-band communication link is:

- Primarily, one to perform end-to-end technology demonstration. The ability to communicate on the S-band with a minimum data rate can be proven.
- Secondly, an experimental link that can be used to generate varying data rates and analyze the performance and difficulties involved.
- Thirdly, a link that can provide a significant increase in downloaded Delfi-n3Xt mission data.

4 Systems engineering tasks

A discussion on the work performed as a systems engineer of the Delfi-n3Xt team would not be proper without an assessment of typical systems engineering tasks performed. Systems engineering deals with designing and managing complex engineering projects. A general systems engineering approach in the authors perception, and that as applied on different levels within the COMMS is that shown in Figure 4.1 below.

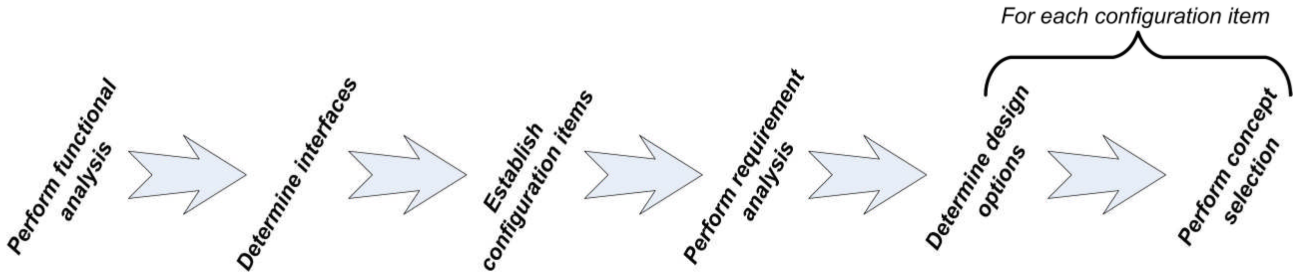


Figure 4.1: Systems engineering design flow

The first four steps illustrated above can be applied to any product, be it the ultimate satellite or simply a (sub)system of it. A configuration item (CI) tree with respective requirements will result. If a single CI is split into several other ones, the collective of these CIs should perform the function(s) of the higher level CI. If a sufficiently low level is reached, concrete design options can be determined, and concepts in the form of *hardware* can be selected. The final step then in all cases will be electrical and mechanical assembly, integration, testing and verification; <u>this is the current stage of most COMMS components</u>.

In the Delfi-n3Xt project a large number of configuration items (CIs) has been defined following a detailed functional analysis. In fact, this has lead to the notion of having *subsystems*. Each subsystem is part of the satellite bus and performs a dedicated function. On the level of the COMMS the same analysis has been performed to determine all components of the COMMS. These are specified in [SLR 0014, chapter 5]. The components are as introduced above the PTRX, STX, antenna system and OLFAR. The ITRX can be seen as a COMMS component although it is in fact a payload; no detailed design is performed on it.

The exact same sequence as shown Figure 4.1 has been performed by the author for every single one of the COMMS components, except for the ITRX. This has allowed the *low-level design* to take place. In other words, this has allowed the step to be made *from top-level design to detailed design*.

In case of some components, detailed design solutions already existed; the PTRX is a good example. As it turned out that the PTRX would not be the simple update of the RAP that it theoretically could have been, it has been important to analyze its design using a top-down approach as well. In the end, this has allowed to *determine the interfaces to OLFAR* and to decide to *separate the TX and RX sections*.

For the STX, no functional or hardware design existed yet. As such, the entire flow of systems engineering tasks has been performed. The STX only had one established main function:

- Provide a data rate of over 9.6 kb/s over the S-band.

In the previous chapter, a component diagram of the STX (Figure 3.5) has been presented. [SLR 0387] presents the entire analysis as performed, following a systems engineering approach. It concludes with this component diagram. As an illustration, the systems engineering diagrams leading up to this component diagram are shown in respectively Figure 4.2, Figure 4.3 and Figure 4.4 below.

Per figure:

- **Figure 4.2** shows the subfunctions required to perform the final main function
- **Figure 4.3** shows the hardware components required to fulfil these functions with proper respect for interfaces
- **Figure 4.4** shows the resulting CIs, excluding antenna system but including housekeeping sensors

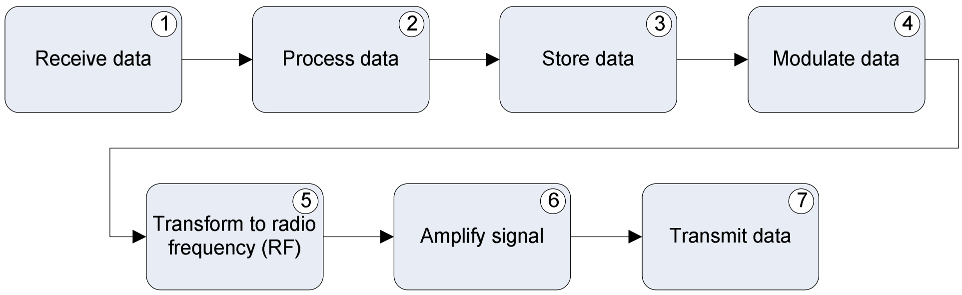


Figure 4.2: STX functional flow diagram [SLR 0387, Figure 5.1]

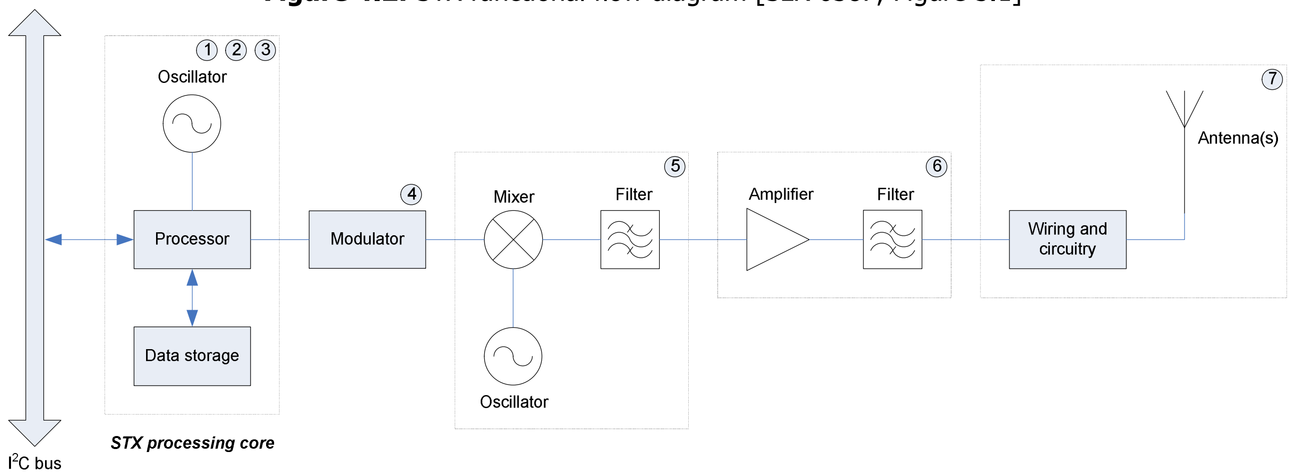


Figure 4.3: STX hardware diagram [SLR 0387, Figure 5.2]

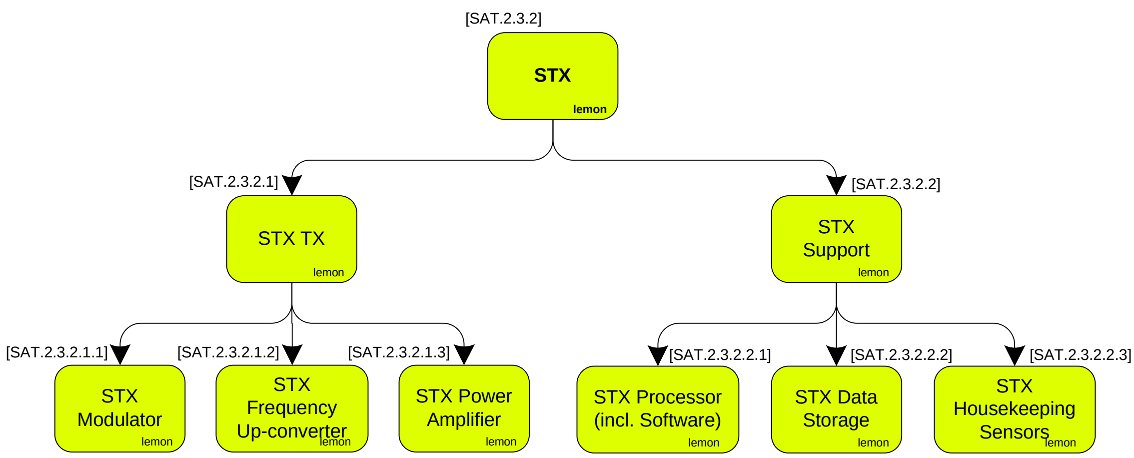


Figure 4.4: STX Configuration Item tree [adapted from SLR 0387, Figure 5.3]

Consequently, requirements have been established for every CI, and design options are explained in [SLR 0387]. The different components have been selected taking into account the interfaces. The component diagram of the STX (Figure 3.5) is the result.

5 Description of work performed during thesis

The work performed by the author has in some sense been a continuation of the work performed by the previous COMMS systems engineer, or even of the COMMS work performed for Delfi-C^3. To properly illustrate the added value of the author, the two sections of this chapter establish the status of the Delfi-n3Xt COMMS before the author's arrival, and the exact work performed during the thesis.

5.1 Status of COMMS before arrival

The previous COMMS systems engineer, Martijn de Milliano, has established the global architecture and objectives of the Delfi-n3Xt COMMS. The main conclusions and decisions made are summarized below. Possible limitations which are logically improved upon during the author's work are indicated.

Top-level

- **Main COMMS components**: The higher level components of the COMMS had been established; PTRX, ITRX, STX, Antenna System. OLFAR was still envisioned to be completely integrated on the PTRX, and was thus part of the latter.
- **Frequency bands**: The frequency bands that were to be used had been established, and the reuse of the linear transponder had been confirmed. Discussion with the radio amateur community about frequency allocation had however not taken place yet.
- **Requirements**: A (too) large list of requirements had been made.
- **CI tree**: A CI breakdown had been made (without proper functional analysis performed)

PTRX

- **RAP-based**: The PTRX had been decided to be based on the RAP. This meant that its functions were taken for granted without further analysis, applying to modulation schemes including line codes and pulse shaping approaches as well as the data protocol (AX.25).
- **Demodulation on PTRX**: It had been decided to move the demodulation capability to the PTRX, moving it away from the OBC as on Delfi-C^3. At the same however it was decided to let it perform back-up TC interpretation functionality without proper trade-off.
- **Increase of data rates suggested**: The suggestion had been made to investigate the increase of data rates

STX

- **Purpose**: The main function of the STX had been established (deliver a data rate above 9.6 kb/s). No functional design had taken place.
- **Design suggestions**: Some design suggestions had been made:
  - Include data storage
  - Include a high-efficiency switching amplifier
  - Include a high-speed and low-speed mode
- **Pointing**: Pointed operations were assumed and some pointing scenarios were roughly sketched.

OLFAR

- **Experiment definition**: The OLFAR experiment had been established, i.e. the scientific value. No functional design had taken place.

Antenna system

- **UHF/VHF antenna configuration**: Possible configurations had been assessed in much detail. Unfortunately these configurations were based on the *old solar panel configuration* using four panels pointed towards the sun at all times. Also, the option of integrating the antennas had not been assessed.
- **STX antenna configuration**: A theoretical design of a patch antenna has been made and these patches were used to establish a (high-directivity) antenna configuration. Unfortunately, this was again based on the old solar panel configuration and assuming three-dimensional pointing.

Link budget

- **Preliminary link budget**: An eye-catching template had been established and a preliminary link budget was made. However:
    - Variables in the link budget were not explained
    - Values were not argued for in many cases
    - Many values were to be specified in more detail, but most significantly: the downlink PA output was taken too high (-2 dB), the phasing circuit losses were forgotten (-2 dB on both the up- and downlink) and no noise calculations were performed
    - The template had stability issues ('corrupted file' errors)
    - The S-band link budget was highly tentative due to its design status
    - The presentation was rather messy and the calculation of its values non-transparent

Also, in the author's opinion, the general COMMS document structures were rather unclear. Furthermore, the ground station and its functionalities or performance had not been analyzed; In other words, the work had been focused on the *top-level* design of (only) the *COMMS of Delfi-n3Xt*.

5.2 Work performed

The author has expanded this work in two directions:

- **Downwards**: The functional components of the COMMS have been analyzed, and in turn the components of these components. In other words, *detailed design* was started.
- **Sideways**: The scope was expanded to include the complete communication links and the ground segment.

The added scope further explains the title of this thesis paper. Detailed analysis of the COMMS of Delfi-n3Xt goes hand-in-hand with detailed analysis of the on-ground systems and operations; thus involving the *entire communication link(s)*.

The following consists of a near-exhaustive list of all work that has been performed. Given the lack of a main thesis research question, many different subjects have been addressed during the thesis period. These are all tasks on which days if not several weeks have been spent. All topics are documented in different concrete sections of the project documentation, as indicated. Topics that already have been assessed in more detail in the previous chapters are not explained. This list can be used as a look-up table for certain topics.

Top-level work

- **Requirements definition**: After the redefinition of the mission, all requirements have been redefined. In fact many requirements were unstructured as they were derived by different people in different phases. The top-level requirements *as well as* all COMMS requirements were restructured and redefined.
- **CI definition**: All COMMS CIs have been defined for all low-level components following a proper functional analysis.

Researching communication techniques [SLR 0014]

- **Modulation schemes** [section 7.2 and appendix A]: Detailed research has taken place into the workings of different schemes and their power, bandwidth and (de)modulation complexities.
- **Line codes** [section 7.3 and appendix B]: Characteristics of different line codes such as NRZ-I and Manchester have been researched.
- **Pulse shaping** [section 7.4]: Commonly applied pulse shaping filters as well as that applied in Delfi-C^3 have been analyzed in detail.
- **Data protocol** [section 7.6]: The *requirements* of a data protocol are established, and the AX.25 and FX.25 protocols are introduced. Also, the concept of *throughput efficiency* is introduced.
- **CRCs** [section 7.7]: Research has been performed to establish applicable CRCs, involving exhaustive analyses performed in several papers discussing detection performances of CRCs up to 16 bit.
- **FEC codes** [section 7.8]: The purpose of FEC coding is explained, and different FEC codes, FEC coding techniques and FEC coding scheme performances are given.

- **Optimizing downloadable data volume** [section 7.9]: Continuing on a suggestion made by Martijn de Milliano, approaches to maximize downloadable data volume are discussed. Optimum elevation angles are given for different orbit altitudes *as well as* off-zenith angles.

Analyzing Delfi-C^3 heritage

- **Spectrum measurements** [SLR 0014, section 7.4 and 8.4]: The spectral performance of the downlink signal, when transmitted by the satellite and when received on-ground, as well as the transmitted uplink frequency spectrum have been measured and analyzed.
- **Power measurement** [SLR 0106, chapter 3]: Using the received downlink signal strength of Delfi-C^3, the feasibility of the (Delfi-n3Xt) link budget has been assessed.
- **Ground station set-up** [SLR 0014, section 15.1]: All components present on the ground station and relevant to the UHF/VHF communication link have been analyzed and mapped.
- **Telemetry data base analysis** [SLR 0014, section 14.2]: The actual data yield performance of Delfi-C^3 has been analyzed.
- **RAP functionalities** [SLR 0014, chapter 8 and 9]: The components taking care of different functionalities have been analyzed in detail to assess possible improvements, most notably related to the choice of modulation scheme.
- **Telecommand procedures** [SLR 0014, section 8.6]**:** The procedures and the specific data exchange involved when sending TCs has been determined and documented as this was not previously done.

Design of the UHF/VHF communication link [SLR 0014]

- **Transmission frequencies and bands** [section 8.1]
- **Modulation scheme** [section 8.3 and 8.4]
- **Customized (DelfiX) data protocol** [section 8.5]
- **Additional data fields and operations** [section 8.6]: The telecommand procedures and the general data content of a TC have been established, including aspects as preambles and TC repetition. For the downlink, suggestions have been made as to what fields are required.

Design of the PTRX [SLR 0014]

- **Functional analysis** [section 9.1-9.4]: Following a functional analysis, the interfaces and hardware components of the PTRX have been mapped in detail.
- **Housekeeping values** [section 9.5]: Given the actual systems on the PTRX, the different housekeeping variables and the required components have been mapped, along with several suggestions to increase the amount of produced housekeeping data.
- **Processor functionalities** [section 9.6]: The different functionalities of the processors have been established to facilitate software writing.
- **Linear transponder** [section 9.8]: The linear transponder as well as its transponder beacon have been functionally specified and documented, based on the final (undocumented) operations of the RAP.
- **Analysis of new features** [section 9.7 and 9.9]: Options such as TC interpretation, a variable voltage bus connection and software upload capability are discussed.

Design of the S-band communication link [SLR 0014]

- **Transmission frequencies and bands** [section 12.1]
- **Modulation scheme** [section 12.2]
- **FEC coding scheme options** [section 12.3]
- **Data protocol** [section 12.4]
- **Data rate (approach)** [section 12.6]
- **Operations** [section 14.3]: The link activation by means of TC and the operational modes are explained.
- **Data yield assessment** [section 14.4]: Taking into account achievable data rates, the possibility to download all generated Delfi-n3Xt TM data is assessed.

Design of the STX [SLR 0387]

- **Functional analysis** [chapter 5]: The CIs of the STX have been established using a functional and interface analysis, after which requirements have been stipulated for each CI.
- **Component selection** [chapters 6-11]: For each CI concrete component selection has been performed for as far as possible to yield a final component diagram.
- **Housekeeping values** [chapter 8]: All required housekeeping data has been established as well as the consequently required components.
- **Integration options** [chapter 13]
- **Power consumption estimation** [section 14.3]
- **Experimental features** [chapters 11-12 and appendix A]: Different features such as the inclusion of a high-efficiency (switching) amplifier, a low data rate (beacon) mode, the connection to a variable voltage bus and the application of an FPGA have been assessed.

Design of the Antenna system [SLR 0036]

- **Combination of UHF and VHF antennas** [section 3.2]
- **UHF and VHF antenna configuration** [chapter 3]
- **S-band antenna system** [chapter 4]
    - **Constrained pointing scenarios** [deleted]: Assuming the old solar panel configuration as well as sun-pointed operations, a (preliminary) analysis has been done of the optimal patch antenna gain pattern and S-band antenna configuration, assuming not all axes are available for S-band pointing.
- **Wiring and connectors** [chapter 5]: Suggestions have been made to improve upon the wires and connectors applied in Delfi-C^3.

Link budget [SLR 0106]

- **Introduction of variables *and* explanation of values** [chapter 2]: All different variables/values used in the link budget are carefully explained and documented.
- **Detailed noise calculations** [section 2.4]: Instead of giving overall noise assumptions the total noise has been calculated for all communication links using the characteristics of different receiver components.
- **Template and presentation**: The stability issues of the template and the overall transparency of the link budget has been improved upon.

Other

- **OLFAR** [SLR 0014, chapter 11]: Using the component design of the OLFAR chip its integration on the satellite has been reassessed and determined to be on the DAB, if at all.
- **ITRX** [SLR 0014, chapter 12]: Changes to the UHF/VHF communication link of relevance to the ITRX have been discussed (with ISIS) and can be incorporated.
- **Ground station upgrades** [SLR 0014, section 15.1]: Different (required or optional) upgrades are presented, including a possible set-up required to receive S-band signals.
- **Doppler shift (rate) calculation** [SLR 0014, section 3.4.3]: A more detailed analytical calculation of the size of Doppler shift and Doppler shift rates has been performed.
- **Frequency allocation**: Discussions with Graham Shirville have allowed the allocation of the preliminary frequencies and bandwidths. Also, bandwidth regulations were researched in more detail.

- **Delfi-n3Xt Promotion**: Finally, this amounts to a 'softer' type of work performed. On different accounts stands were manned, presentations were given, articles were written and other project-related promotion tasks were performed.

6 Conclusions and recommendations

As a result of the performed work, the final status of the COMMS as well as the required next steps can be established. Some concrete achievements of the thesis work shall first be stated in section 6.1. Afterwards the recommended next steps to be taken are assessed in section 6.2. Finally, some concrete lessons learned or words of caution are expressed by the author in section 6.3.

6.1 Achievements of thesis work

Three different achievements can be stated of relevance to Delfi-n3Xt:

All higher level Delfi-C^3 design choices relevant to COMMS have now been documented, at least indirectly. All original choices made relevant to current Delfi-n3Xt components have been taken into account for the design of these components, in which case the decision processes have been properly documented. Also, systems such as the GS have been properly mapped in terms of all their components.

A reference work for communication techniques has been established. All design choices for communication techniques applied on Delfi-n3Xt have been introduced with required theory, not being limited to only explaining the design choices at hand. A large chapter of some 100 pages [SLR 0014, chapter 7 incl. appendix A and B] has been devoted to introducing all theoretical and practical knowledge of communication techniques required to make the relevant design choices afterwards. This same material can be used for the definition and design of any new satellite communication link.

The 'paper phase' has been finalized as far as the Delfi-n3Xt COMMS is concerned. For all components such as PTRX, STX, OLFAR and the antenna system it holds that the functional design choices have been made and documented, and that the next step is (electrical) integration and testing. The antenna system does however still require knowledge of the new solar panel configuration as assessed below. A lot of interfaces do exist with the CDHS, which should still be defined from a top-level by the responsible CDHS engineer. Examples of topics are the gathering of housekeeping data, the definition of telecommands and the content of telemetry frames. For the COMMS, a dedicated top-level systems engineer should no longer be required.

6.2 Follow-up work

On a top-level, only little work remains to be performed. Three tasks:

- **Frequency licensing**: The frequencies should be and can be filed for a soon as possible, using the established information. The author can still be involved in this process given required knowledge and correspondence.

- **Antenna System**: As soon as the new solar panel configuration has been defined, the antenna configurations for both the UHF/VHF and S-band antenna systems should be determined and/or confirmed given the suggestions made in the respective documentation.

- **OLFAR**: The future of OLFAR on Delfi-n3Xt should be established given the anticipated different opinions on the manner within the management. OLFAR:
    - *Can provide valuable technology demonstration for a future OLFAR mission.*
    - *Requires an allocation of resources to test and possibly adapt the design*; an on-chip solution design has been delivered by Edwin Wiek but it yet to be produced.
    - *Has definite effects on data architecture, power consumption and structural implementation* given its required implementation on the DAB. This necessitates a separate (cable) connection to the PTRX as well as a processor to operate it; the DAB processors are otherwise inactive after deployment.

The general focus of the COMMS can and should now be shifted to lower-level integration and testing. This holds for all (in-house) components of the COMMS:

Primary communication link

- **PTRX**: The components design is finished and different hardware has been tested. A complete transmitter, receiver or PTRX has however not yet been delivered.
- **UHF/VHF antenna system**: The system can be built and tested using a mock-up of Delfi-n3Xt, awaiting the determination of the final solar panel configuration.
- **Ground station**: The software of the GS should be updated for Delfi-n3Xt operations, pending the further specification of the CDHS architecture. UHF/VHF hardware should be replaced if a higher data rate on the downlink is supported; ISIS can likely support these activities.
- **Link budget**: the UHF/VHF links are specified with sufficient precision.

High speed communication link

- **STX**: The component design has been specified for as far as possible. Integration and testing should be performed to confirm and improve its design.
- **S-band antenna system**: The preferred antenna configuration has been established. The final solar panel configuration can confirm the antenna configuration after which integration tests and performance measurements can be performed, using available patch antennas.
- **Ground station**: The ground station should be expanded to support the S-band communication link. Again ISIS can likely support the actual integration.
- **Link budget**: The S-band link mainly depends on the configuration of the ground station, the feasibility of the FEC coding scheme in the STX and the final S-band satellite antenna system performance. Also, when more is known on the ground station set-up this information should be used to perfect the link budget.

6.3 Lessons learned

As a systems engineer of a subsystem of Delfi-n3Xt, a firm understanding is gained of the subsystem at hand, the COMMS in the author's case. But at the same time, all other subsystems of the satellite are heavily involved in the daily work. As such, a Delfi-n3Xt systems engineer expands his or her knowledge of the entire satellite bus design. Even more interestingly, COMMS knowledge as required to design a satellite relates directly to general (tele)communication processes on-ground such as internet and telephone. And finally, general systems engineering as applied on a complex project such as designing a satellite, is useful for any complex (professional) project, with many people, (sub)systems and interfaces.

A couple of concrete lessons learned of various natures:

Bandwidth restrictions within the radio amateur bands are relatively unimportant: Considerable time has been spent by the author on first of all determining the regulations, second of all determining the allocated bandwidth for Delfi-C^3, third of all discovering that these regulations are not met for both the down- and uplink and finally that they are in fact not so important. The ITU specifies a minimum reduction of power at the edge of the spectrum, but radio amateurs quite simply seem to accept that these measures are not reached within their bands. Although it is good to have the matter sorted out now including actual measures of performance for Delfi-C^3 and anticipated measures for Delfi-n3Xt, it seems unlikely that the relative un-importance of these rules was not known by people and thus *not documented*.

Delfi-n3Xt is delayed by the lag in Delfi-C³ documentation: The effect of documentation that has not been made for Delfi-C³ should not be underestimated. In case of the COMMS, decisions and features turned up during the detailed design process that were previously unknown due to a lack of documentation (the final design of the RAP, the implementation in hard- and software of the transponder beacon, TC uplink procedures incl. encryption and CRCs, etc.). The time-profit that was made at the end of Delfi-C³ by not documenting, has added to the time-schedule of Delfi-n3Xt.

The initial over-ambitiousness of Delfi-n3Xt has caused a large time-delay: The over-ambitiousness has consisted of the choice to rely on 3-axis attitude control, *as well as* having a large number of additional novelties. The removal of the guarantee of sun-pointing has caused a large delay *on top of* the delay caused by its initial inclusion. In many cases the design change (which was indeed necessary) has forced design to be redone; currently the redesign of the solar panel configuration takes place, and will have structural and operational impacts. Within the COMMS, this has caused detailed analyses performed on both the UHF/VHF and S-band antenna systems to be outdated. At the same time the many possible novelties have caused 'possible constraints' everywhere. The variable voltage bus is a good example.

The Delfi programme lacks a good approach for electrical integration and general AIVT: These types of 'time-consuming' and possibly 'simple' activities (from an academic point-of-view, yet requiring definite expertise) are not properly supported within the Delfi programme. (Aerospace and Electrical Engineering) students can not readily be used for this due to the lack of academic depth and the required expertise. This has held for the electrical integration and testing of the PTRX and this will again hold for final soldering and PCB design that will have to be performed, as well as AIVT activities near the end. The attempted solution of using HBO students has only marginally worked, given the fact that already five HBO students have performed work on the PTRX without delivering a working radio; they have only a very limited duration of stay of which considerable time goes into familiarizing oneself with the matter.

The value of the Delfi-experience increases when more systems engineers are active, and conversely decreases when less systems engineers are available. Needless to say, the Delfi programme suffers from the lack of (Aerospace Engineering) students. But more importantly so, the dynamicity of the programme decreases if there are less interfaces and if many estimations have to be made for other subsystems. Unfortunately, it turns out to be difficult to have a constant influx of students.

The faculty of EEMCS is generally disconnected from the AE faculty, where the actual development of a satellite takes place. Although arguably many top-level decisions can be relatively unimportant for low-level electrical designs, the boundaries of information exchange are too high. This causes electrical decisions to not be properly considered from a higher-level point-of-view (moving OLFAR to the DAB or not), and at the same time does not allow the Aerospace Students to casually tap into the knowledge of electrical systems essential for final design choices to be made. It might be hard to solve this problem, but an optimal solution would of course be a (perhaps only periodically) shared location.

The SLR quickly grows very large due to the continuous addition of documents and files of various natures. Although this is not a problem from a reference point-of-view, this decreases the possible functionality of the SLR as a source of knowledge for beginners. The current SLR contains 164 documents related to COMMS; logically these documents (including books that are not instantly available) can never all be scanned by a beginner. In other words, more emphasis will be placed on the available (top-level) technical notes and their introduction the beginner, as well as *consistent use of references whenever knowledge results from a reference*. In much old *and* new documentation facts or concepts are stated without proper reference. In case of the COMMS top-level document, the author had made an effort to refer to all COMMS related project documentation, and at the same time introduce all new concepts and facts with proper references throughout the documentation.

Symbol list

1. Units

Many units are used with the prefixes μ, m, k, M to indicate an extra factor of receptively 10^{-6}, 10^{-3}, 10^{3}, 10^{6}.

Unit:	***Abbreviation:***	***Description:***
Ampere	A	unit of electric charge
Baud	Baud	symbol per second
Bit	b	1 or 0
Byte	B	8 bits
Decibel	dB	10 log *X*, where *X* is any number
Degree	°	unit of angle
Hertz	Hz	times per second
Joule	J	unit of energy
Kelvin	K	unit of temperature
Meter	m	unit of distance
Ohm	Ω	unit of electrical impedance
Second	s	unit of time
Steradian	sr	unit of solid angle
Volts	V	unit of electromotive force
Watt	W	unit of power

Unit:	***Description:***
dBc	power ratio in decibels of the signal power relative to the carrier signal
dBd	forward gain of an antenna relative to a half-wave dipole antenna
dBi	forward gain of an antenna relative to the hypothetical isotropic antenna
dBic	forward gain of an antenna relative to a circularly polarized isotropic antenna
dBm	power ratio in decibels of the measured power relative to 1 mW
dBw	power ratio in decibels of the measured power relative to 1 W
ppm	parts per million

2. Constants

Constant:	***Value :***	***Unit:***	***Description:***
μ	$3.986004418 \cdot 10^{14}$	m^3s^{-2}	gravitational parameter of Earth
c	$2.99792458 \cdot 10^{8}$	$m\ s^{-1}$	speed of light
k	$1.3806504 \cdot 10^{23}$	$J\ K^{-1}$	Boltzmann constant

3. Variables

Variable:	***Unit:***	***Description:***
α	°	angle
α	-	value between 0 and 1
α		signal transmission efficiency between antenna and LNA (1-total loss)
β	°	angle
β	-	value between 0 and 1
β	-	signal transmission efficiency between LNA and second amplifier (1-total loss)
ε	°	elevation angle
$\varepsilon_{\min}$	°	minimum elevation angle
ε_{opt}	°	optimum elevation angle
$\lambda_{\min}$	°	minimum Earth central angle
$\lambda_{\max}$	°	maximum Earth central angle
λ	m	signal wavelength
$\eta_{throughput}$	-	throughput efficiency
θ	°	off-zenith angle
θ	°	phase angle
θ	°	angle off-boresight
ϕ	°	azimuth angle
ψ_s	-	spectral density
Ω		solid angle
ω	Hz	frequency
A	-	amplitude
B	-	bandwidth
$B_{99\%}$	Hz	bandwidth in which 99% of the signal energy is contained
B_T	Hz	transmission bandwidth
BER	-	bit error ratio
BR	-	bitstuffing ratio
BT	Hz s	bandwidth time product
D_{pass}	b	downloadable data volume during a pass
DL	b	average data loss
E_b / N_0	dB	bit energy to noise spectral density ratio
E_X	-	*x*-component of the electric field vector
E_Y	-	*y*-component of the electric field vector

f_0	Hz	centre frequency
$f_{carrier}$	Hz	carrier frequency
f_{IF}	Hz	frequency at IF
$f_{oscillator}$	Hz	frequency of the oscillator
FS	b	frame size
G_{LNA}	-	Gain of the LNA
h	m	orbit altitude
h	-	frequency separation (in FSK schemes)
I	-	amplitude of a waveform
k	b	message length
L_p	-	path loss
M	-	positive integer
m	b	CRC length
N	Hz	Nyquist bandwidth
n	b	total message length
n	b	block length
n	b	positive integer
M	-	positive integer
OV	b	overhead
P	s	orbital period
P_b	-	bit error ratio
R	m	radius of the Earth
R	-	code rate
R	Baud	data rate
R_b	b/s	bit rate
R_{data}	Baud	data rate
$R_{\max}$	Baud	maximum data rate
S	m	slant range
T_0	K	system temperature and reference temperature (physical temperature)
$T_{2ndStage}$	K	noise temperature of the second amplifier (not a physical temperature)
T_a	K	antenna or sky temperature (not a physical temperature)
$T_{coverage}$	s	coverage time
T_{LNA}	K	LNA noise temperature (not a physical temperature)

T_{max}	s	maximum duration of pass
T_{pass}	s	duration of pass
T_{s}	K	The system noise temperature
T	s	duration
t	b	integer number of correctable bits
t	s	time
V	-	amplitude
v_c	m/s	circular velocity
v_r	m/s	radial velocity with respect to the ground station
x	-	any value
y	-	any value

Acronyms and abbreviations

3D	3-Dimensional
A/D	Analog to digital
ac	alternating current
ADCS	Attitude Determination and Control Subsystem
AE	Aerospace Engineering
AES	Advanced Encryption Standard
AGC	Automatic Gain Control
AHP	Analytical Hierarchy Process
AM	Amplitude Modulation
AMI	Alternate Mark Inversion
AMSAT	Amateur Satellite
ARQ	Automatic Repeat Request
ARS	Antenna Rotor System
ASK	Amplitude Shift Keying
AWGN	additive white Gaussian noise
BCH	Bose-Chaudhuri-Hocquenghem
BER	Bit Error Rate
BFSK	Binary FSK
BM	Beacon Mode
BOP	Bottom Panel
BPSK	Binary PSK
CD	Compact-Disc
CDHS	Command and Data Handling Subsystem
CDMA	Code Division Multiplexing
CF	Carrier Frequency
CI	Configuration Item
CL	Connectionless
CO	Connection-Oriented
COMMS	Communication Subsystem
COTS	Commercial-Off-The-Shelf
CPM	Constant Phase Modulation
CRC	Cyclic Redundancy Check
CRC-16	16 bit Cyclic Redundancy Check
CW	Continuous Wave
D/A	Digital to Analog
DAB	Deployment and Antenna Board
DAC	Digital-to-Analog
dc	direct current
DES	Data Encryption Standard
DRAM	Dynamic Random Access Memory
DSB-SC	Double Sideband, Suppressed Carrier
DUDe	Delfi Universal Data extractor
EEMCS	Electrical Engineering, Mathematics and Computer Science
EEPROM	Electrically Erasable Programmable Read-Only Memory
EER	Envelop Elimination and Restoration
EIRP	Effective Isotropic Radiated Power
EM	Electromagnetic
EPROM	Erasable Programmable Read-Only Memory
EPS	Electrical Power Subsystem
FCS	Frame-Check Sequence
FDMA	Frequency Division Multiplexing
FEC	Forward Error Correction
FIR	Finite impulse response
FM	Frequency Modulation
FPGA	Field Programmable Gate Array
FSK	Frequency Shift Keying
GENSO	Global Educational Network for Satellite Operations
GFSK	Gaussian FSK
GMSK	Gaussian MSK
GS	Ground Station
GSE	Ground Support Equipment
GSM	Global System for Mobile Communications
GSN	Ground Station Network
HD	Hamming Distance
HDM	High Data rate Mode
HK	Housekeeping
I(-frame)	Information(-frame)
I^2C	Inter-Integrated Circuit
IARU	International Amateur Radio Union
IC	Integrated Circuit
IF	Intermediate Frequency
IIR	Infinite Impulse Response
IMD	Intermodulation Distortion
ISI	Intersymbol Interference
ISIS	Innovative Solutions In Space
ISM	industrial, scientific and medical
ITRX	ISIS Transceiver
ITU	International Telecommunication Union
KISS	Keep It Simple and Stupid

L-EPS	Local EPS
LHCP	Left-Handed Circularly Polarized
LNA	Low-Noise Amplifier
LO	Local Oscillator
LOS	Launch and Orbit Segment
LTDN	Local Time of the Descending Node
MAB	Modular Antenna Box
MASK	M-ary ASK
MechS	Mechanisms Subsystem
MFSK	M-ary FSK
MMCX	Micro-Miniature Coaxial
MOST	Microvariability and Oscillation of Stars
MPSK	M-ary PSK
M-QAM	M-ary QAM
MSK	Minimum Shift Keying
NASA	National Aeronautics and Space Administration
NRZ	Non-Return-to-Zero
NRZ-I	Non-Return-to-Zero-Inverted
OBC	On-Board Computer
ODFM	Orthogonal Frequency Division Multiplexing
OLFAR	Orbiting Low Frequency Antennas for Radio astronomy
OOK	On-Off Keying
OQPSK	Offset QPSK
OSI	Open Systems Interconnection
OVTA	Optimized VTA
PA	Power Amplifier
PAE	Power Added Efficiency
PCB	Printed Circuit Board
PID	Protocol Identifier
PM	Phase Modulation
PROM	Programmable Read-Only Memory
PSD	Power Spectral Density
PSK	Phase Shift Keying
PT	Pseudoternary
PTC	positive thermal conductor
PTRX	Primary Transceiver
QAM	Quadrature AM
QPSK	Quadrature PSK
RAP	Radio Amateur Platform
RAP	Radio Amateur Platform
RC-BPSK	Raised Cosine BPSK
RF	Radio Frequency
RHCP	Right-Handed Circular Polarization
RS	Reed-Solomon
RSSI	Received Signal Strength Indicator
RTC	Real Time Clock
RX	Receiver
RZ	Return-to-Zero
S(-frame)	Supervisory(-frame)
SCL	Serial Data Line
SDL	Serial Clock Line
SEU	Single Event Upset
SFL	Space Flight Laboratory
SLR	Standard List of References
SMAD	Satellite Mission Analysis and Design
SMSK	Serial MSK
SPLASH	Space FLASH
SQL	Structured Query Language
SRAM	Static Random Access Memory
SSB	Standard System Bus
SSB-SC	Single Sideband, Suppressed Carrier
SSID	secondary station identifiers
SSTL	Surrey Satellite Technology Ltd.
STK	Satellite Tool Kit
STS	Structural Subsystem
STX	S-band Transmitter
TC	Telecommand
TCM	Trellis Coded Modulation
TCS	Thermal Control Subsystem
TLA	Table Lookup Algorithm
TLE	Two-Line Element
TM	Telemetry
TNC	Terminal Node Controller
TOP	Top Panel
TRASH	The Radio Amateur Satellite Handbook
TRX	Transceiver
TRXVU	VHF/UHF Transceiver
TX	Transmitter
TXS	S-band Transmitter of ISIS
U(-frame)	Unnumbered(-frame)
UHF	Ultra High Frequency
UI(-frame)	Unnumbered Information(-frame)
USB	Universal Serial Bus
UTIAS	University of Toronto Institute for Aerospace Studies
VHF	Very High Frequency
VTA	Virtual Table Algorithm
WLAN	Wireless Local Area Network
WP	Work Package

References

Living Project Documentation written by the author

[SLR 0014] COMMS – Top Level Design of Communications Subsystem, by Tindemans A.M.
V 3.1, 13-6-2010

[SLR 0036] COMMS Antenna System Design, by Tindemans A.M.
V 3.0, 18-6-2010

[SLR 0106] Delfi-n3Xt Link Budget, by Tindemans A.M.
V 2.1, 13-6-2010

[SLR 0387] COMMS - STX System Design, by Tindemans A.M.
V 3.0, 13-6-2010

Living Project Documentation

[SLR 0017] Delfi-n3Xt Power budget, by Teuling, R.
V 2.4, 17-9-2009

[SLR 0018] Delfi-n3Xt Mass budget, by Perez-Lebbink, L.
V 3.1, 3-7-2009

[SLR 0019] EPS – Top Level Design of Electrical Power Subsystem, by Teuling, R.
V 1.5, 16-3-2009

[SLR 0029] CDHS – Top Level Data Bus Architecture, by Jong, S. de
V 1.6, 1-6-2009

[SLR 0104] CDHS – High level Software Design, by Jong, S. de
V 1.0, 30-12-2008

[SLR 0105] CDHS – OnBoard Computer System Design, by Jong, S. de
V 2.0, 29-12-2008

[SLR 0167] Delfi-n3Xt Requirements Management Tool, by Genbrugge, M.
V 2.7, 15-12-2009

[SLR 0169] STS – Top Level Design of Structural Subsystem, by Go, S.Y.
V 2.2, 1-7-2009

[SLR 0170] OLFAR Experiment Definition, by Wiek, E.
V 1.3, 15-6-2009

[SLR 0182] Interface Control Document ITRX, by Perez-Lebbink, L.
V 1.1, 28-11-2008

[SLR 0263] Delfi-n3Xt – Mission Definition, Systems Overview and General Design Approach by Bouwmeester J.
V 0.6, 6-11-2008

[SLR 0281] COMMS - PTRX Design, by Li, S.
V 1.1, 17-6-2009

[SLR 0282] Delfi-n3Xt Data Budget, by Jong, S. de
V 1.3, 15-12-2008

[SLR 0285] Delfi-n3Xt Operational Modes, by Teuling, R.
V 1.1, 14-6-2009

[SLR 0300] CDHS – OnBoard Computer Microcontroller Selection, by Noroozi, A.
V 0.4, 28-10-2008

[SLR 0303] Delfi-n3Xt Volume Budget, by Go, S.Y.
V 1.4, 10-5-2009

[SLR 0317] Systems Engineering Team Members Manual, by Genbrugge, M.
V 1.1, 10-7-2009

[SLR 0325] Delfi-n3Xt Configuration Tree, by Tindemans, A.M.
V 2.2, 25-3-2010

[SLR 0591] COMMS - PTRX - Transmitter Overview, by Bock, R. de
V 1.0, 20-4-2009

[SLR 0592] COMMS - PTRX - BPSK Modulator, by Bock, R. de
V 1.0, 21-4-2009

[SLR 0659] PTRX - Electrical Circuit Diagrams, by Li, S.
V 0.5, 6-8-2008

[SLR 0715] COMMS - Advanced Phasing Network Design, by Kant, W. van der
V 0.1, 8-4-2010

[SLR 0722] PTRX - Frame Generator & Frame Interpreter Design, by Hartman, C.
V 1.0, 19-3-2010

Standing Project Documentation

[SLR 0244] Thesis on RAP Transmitter, by Stoopman
V 1.0, 1-3-2007

[SLR 0245] RAP Integration and Testing, Bart, R.J.
V 1.0, 1-1-2007

[SLR 0248] Thesis on RAP, by Ubbels, W.J.
V 1.0, 1-10-2005

[SLR 0492] Roel Reijerse, The Delfi-C3 Distributed Ground Station Network, by Reijers, R.H.
V 1.0, 1-9-2005

[SLR 0643] Het communicatiesysteem van Delfi-n3Xt in de S−band, by Eijk, R. van
V 1.0, 5-6-2009

Project Software

[SLR 0654] PTRX - Telecommand Processor Software, by Hartman, C.
V 0.9, 7-1-2010

[SLR 0655] GSN - RASCAL Telemetry Client Software, by Aalbers, G.T.
V 1.1, 1-6-2009

Tools

[SLR 0219] AMSAT-IARU Link Budget Tool, by King J.A.
V 2.4.1, 14-10-2008

External literature

[SLR 0166] Digital Communications, by Proakis, J. G.
5th Edition, McGraw-Hill, 2008

[SLR 0261] Antenna Theory, by Balanis, C.A.
3rd Edition, 2005

[SLR 0293] AX.25 Specification, by McGuire, J.
V 2.2, 1-7-1998

[SLR 0294] FX.25 Specification, by McGuire, J.
V draft, 1-9-2006

[SLR 0310] Information for Amateur Radio Satellites
V 15.5, 6-10-2006

[SLR 0311] Swisscube CMDS Phase B Design, by Pierre-André, T.
V 1.0, 24-2-2007

[SLR 0316] Space Mission Analysis and Design, by Wertz, J.R.
3rd Edition, Space Technology Library

[SLR 0384] FPGAs: Instant Access, by Maxfield, C.
1st Edition, Elsevier Ltd., 2008

[SLR 0389] Space Systems Engineering (AE4-S12) Lecture Notes, by Hamann, R.J.
V 4.0, TU Delft, 1-12-2006

[SLR 0396] Relative Measurement and its Generalization in Decision Making: (…), by Saaty, T.L.
Real Academia de Ciencas, 2008

[SLR 0443] Digital and Analog Communications, by Couch, L.W.
7th Edition, Pearson – Prentice Hall, 2007

[SLR 0447] Digital Modulation in Communication Systems, an Introduction,
14-3-2001

[SLR 0552] Digital Modulation Techniques, by Xiong, F.
2nd Edition, Artec House Inc., 2006

[SLR 0573] Turbo Coding, Turbo Equalisation and Space-Time Coding, by Hanzo, L.
1st Edition, Wiley, 2002

[SLR 0575] Error Correcting Codes Lecture Notes, by Weber, J.H.
V 1.0, TU Delft, 1-2-2006

[SLR 0683] Error Detection, Correction and Related Topics, by Pfeiffer, J.J.
29-4-2009

[SLR 0684] Optimum Cyclic Redundancy-Check Codes with 16-Bit Redundancy, by Castagnoli, G.
1-1-2009, IEEE Transactions on Communications Vol 38. No. 01

[SLR 0685] Undetected error probability performance of cyclic redundancy-check codes of 16-bit redundancy by Baicheva, T.
5-10-2000, IEE Proceedings-Communications

[SLR 0686] 32-Bit Cyclic Redundancy Codes for Internet Applications, by Koopman, P.
1-1-2002, Proceedings of The International Conference on Dependable Systems and Networks,

[SLR 0687] Cyclic Redundancy Code (CRC) Polynomial Selection For Embedded Networks, by Koopman,P.
1-1-2004, Proceedings of The International Conference on Dependable Systems and Networks

[SLR 0688] Efficient High Hamming Distance CRCs for Embedded Networks, by Ray, J.
25-6-2006, Proceedings of The International Conference on Dependable Systems and Networks

[SLR 0689] Optimization of Cyclic Redundancy -Check Codes with 24 and 32 Parity Bits, by Castagnoli, G.
1-6-1993, IEEE Transactions on Communications Vol. 41 No. 06

[SLR 0694] Signals & Systems, by Oppenheim, A.V.
2nd Edition, Pearson – Prentice Hall, 2007

[SLR 0695] Practical GMSK Data Transmission, by Kostedt, F.
1-1-1998

[SLR 0718] Error Correction Coding, by Moon, T.K.
1st Edition, Wiley, 2005

[SLR 0730] Transmission Systems Engineering (ET4036), by Arnbak, J.C.
TU Delft, 1-10-2006

[SLR 0731] Proposed Coded AO-40 Telemetry Format, by Karn, P.
V 1.2, 7-1-2002

[SLR 0732] Proposal for a FEC-Coded AO-40 Telemetry Link, by Karn, P.

[SLR 0733] Oscar-40 FEC Telemetry, by Miller, J.
1-8-2003

[SLR 0738] Performance and Reliability Evaluation of the S-Band Vienna Satellite Ground Station, by Keim, W.
28-8-2008, Proceedings of the 5th IASTED International Conference

[SLR 0739] A New Approach for Enhanced Communication to LEO Satellites, by Gill, E.K.A.
Proceedings of the SpaceOPS Conference, Issue 9, 1-9-2009

[SLR 0745] GeneSat-1 Post Flight Report

[SLR 0746] Canadian Advance Nanospace Experiment 2: Scientific and Technological Innovation on Thee-kilogram Satellite, by Sarda, K.
19-4-2006

Websites

http://www.ka9q.net/ao40/
Last accessed on 21-06-2010

Data Sheets and manuals

[SLR 0110] AD8361 Datasheet, 1-1-2004

[SLR 0138] MAX2640-MAX2641 Datasheet, 1-4-2007

[SLR 0139] MAX2630-MAX2633 Datasheet, 1-2-2007

[SLR 0199] MC3362-D Datasheet, 2-1-1996

[SLR 0213] Saw Filter Specification, 10-12-2004

[SLR 0437] S-band Class E Power Amplifier Preliminary Specification, by Harbers, J.W.L. V 1.0, 1-4-2008

[SLR 0454] MSP430 Microcontroller Datasheet,2-1-2008

[SLR 0556] CC2500 2.4 GHz RF Transceiver Note, V 1.3, 9-5-2007

[SLR 0559] eZ430-RF2500 S-band Development Tool User's Guide, 1-2-2009

[SLR 0576] AM42-0055 2 Watt S-band Power Amplifier, 2.2-2.4 GHz, V 1.0

[SLR 0578] CGB 240 2-Stage Bluetooth InGaP HBT Power Amplifier, V 1.5, 16-12-2003

[SLR 0579] SST12LP00 2.5-2.5 GHz Power Amplifier, V 3.0, 1-2-2009

[SLR 0581] MSP430x15x, MSP430x16x Microcontroller Notes, 1-8-2006

[SLR 0692] SMT16030 Digital Temperature Sensor, 1-7-2005

[SLR 0719] Directional Couplers 2 - 2600 MHz

[SLR 0720] Aircell 7 & Ecoflex 15 Coaxial Cable

[SLR 0721] RF1-LIM RF power transmission protector

[SLR 0723] RG178BU coaxial cable, 1-7-2005

[SLR 0724] RMS-1 0.5 to 500 MHz Frequency Mixer

[SLR 0725] SP-2000 & SP-7000 Pre-amplifiers, 21-10-1994

[SLR 0726] RF-components for Mobile Radio Base Stations, 1-1-2006

[SLR 0734] 2MCP14 Circular Polarized Yagi, 30-8-1999

[SLR 0736] 436CP30 Circular Polarized Yagi, 21-8-2001

[SLR 0737] IC-910H VHF/UHF Transceiver, V 4.0, 1-1-2004

[SLR 0741] RF2713, 29-1-2001

[SLR 0742] S-band Patch Antenna, V 5.3, 1-7-2009

[SLR 0743] PA-28-2450 Patch Antenna Gain Test Data, 18-3-2004

[SLR 0744] Patch Antenna Elements & Assemblies, 1-1-2006

[SLR 0747] RG316/U 50 Ohm Coaxial Cable, 24-2-2005

Appendix A

COMMS – Top-level Design of Communication System

SLR 0014

Within this document, the following sections are based on texts from the earlier versions of the project documentation. These earlier versions have not been written by the author. Per section a short line explains the extent of the adoption. In all cases have the texts been updated and edited for correctness and relevance.

- **Chapter 3 Available frequencies** (~5 pages)
  The content is largely old, with the exception of the section on Doppler shift. This section has been expanded with calculations and better values.
- **Chapter 5: COMMS space segment design** (~5 pages)
  The content of this chapter is based on the old content, however with newly written texts and updated figures.
- **Chapter 9: PTRX** (~1/2 page)
  The section on ranging (9.8.3) is adopted.
- **Chapter 10: ITRX** (~2 pages)
  The introduction of the ITRX is based on the old ITRX chapter.
- **Chapter 15: Ground Segment** (~2 pages)
  The sections on the worldwide radio amateur network, Eindhoven ground station and GENSO are results of the old documentation (15.2-4).

COMMS – Top-Level Design of Communication System

Description:	Top-Level Design of the Delfi-n3Xt Communication System

Subsystem(s) involved:

ADCS	CDHS	COMMS	EPS	MechS	STS	TCS	ITRX	MPS	$T^3\mu PS$	SDM	Splash	GSE	GSN	Launch
		x											x	

Revision Record and Authorization

Issue	Date	Author / Editor	Reviewer checked	PM approved	Affected Chapter (s)	Description of change
0.1	31/03/2008	Martijn de Milliano	SB	JB	All	First issue
1.0	10/09/2008	Martijn de Milliano	MG	JB		Outdated changes
2.0	27/11/2008	Martijn de Milliano	RT			Outdated changes
2.1	29/01/2009	Martijn de Milliano	AT	JB		Outdated changes
3.0	31/02/2010	Arthur Tindemans			***All:***	***Complete update***:
					1	Completely rewritten
					2	New chapter
					3	Old chapter 2: -Expanded Doppler section -Improved references -Added available band explanation
					4	New chapter
					5	Completely rewritten (old chapter 4), except for "frequency selection"
					6	New chapter
					7	New chapter
					8	New chapter
					9	Completely rewritten (old chapter 6)
					10	Completely rewritten (old chapter 7)
					11	New chapter
					12	New chapter
					13	Completely rewritten (old chapter 8)
					14	New chapter
					15	New chapter
					16	Completely rewritten (old chapter 6), except for "GENSO"
					17	Completely rewritten (old chapter 11)

3.1	13/06/2010	Arthur Tindemans			All 13-17 1 7 & A, B 7.7 7.8 8.1 8.3 & 8.4 8.5 9.7 9.9 12 13 14.3/.4 15 16	Random technical and visual updates Old chapter 13 deleted Removed notion of 'experiments' Moved introduction of modulation schemes to appendix A, introduction of line codes to appendix B Added theory on FEC codes Rearranged and expanded section after comments from E. Gill Established transmission bandwidths Restructured and clarified Notion of 'default' and 'optimized' DelfiX removed Section on TC interpretation added Section on PTRX power reduction added Content written Content written Content written Upgrades extended and added S-band support Next steps written

Table of Contents

1 Introduction

The Delfi-n3Xt satellite consists of a number of payloads and a number of subsystems, in order to make the satellite work and facilitate the payloads. Per usual satellite convention, Delfi-n3Xt has been subdivided in a number of subsystems, being STS, Mechanisms, ADCS, TCS, CDHS, EPS and COMMS. This document describes the top level design of the latter, being the Delfi-n3Xt communication system (COMMS). The communication system forms the interface between the ground station(s) on the surface of the Earth and the satellite.

The context of the COMMS on Delfi-n3Xt can be seen in Figure 1.1 below. It shows the Space Segment, being Delfi-n3Xt, and the Ground Station Network, being part of the larger Ground Segment:

- **Delfi-n3Xt**: Within Delfi-n3Xt the COMMS is shown, with active interfaces to the electrical power subsystem (EPS), and the command and data handling subsystem (CDHS), providing power and data communication respectively. These interfaces connect to a number of *radios* within the COMMS. These radios in turn connect to the Antenna System for data transmission and reception.
- **Ground Station Network**: The Ground Station Network primarily consists of a Ground Station, located in Delft, which is in charge of commanding the satellite and receiving data from the satellite. Also, a number of Radio Amateurs worldwide are anticipated to receive data from Delfi-n3Xt as well.

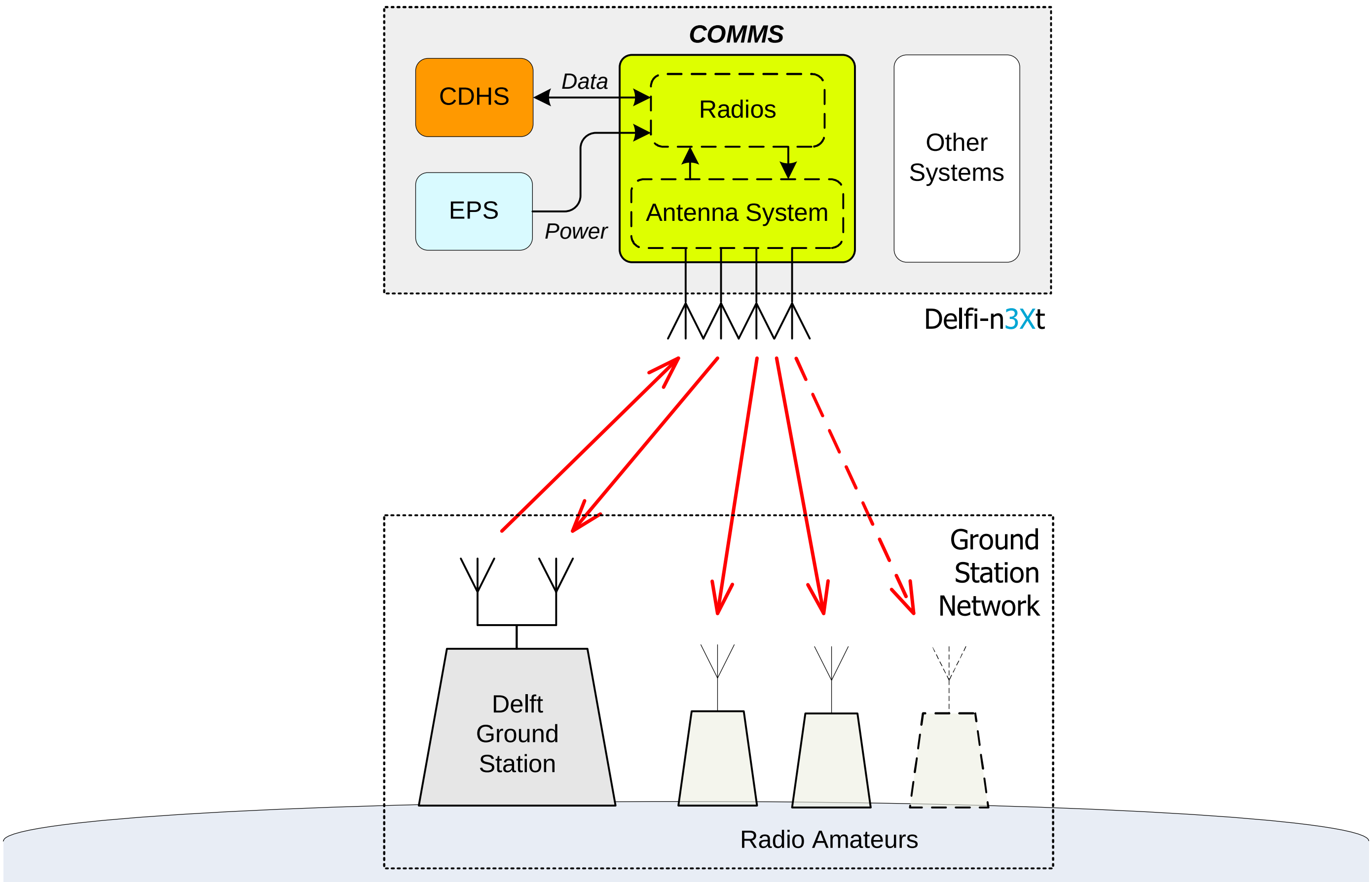


Figure 1.1: COMMS context

The primary tasks of a satellite communication system are to get:

- Payload data from the satellite to the customer
- Housekeeping data from the satellite to the satellite operator
- Telecommands from the satellite operator to the satellite.

These functional streams can be divided in a *downlink* and *uplink* transmission, in which the first two types of data can be jointly called *telemetry* and the latter involves only *telecommands*. In order to transmit and receive data a *transmitter* and *receiver* are required, generally combined in a single *transceiver* unit. At the same time, in order to physically receive and transmit the signal, an *antenna system* is required.

The overview of the Delfi-n3Xt COMMS is presented in section 1.1 below, after which the document structure is presented in section 1.2.

1.1 Overview of the communication system

The entire COMMS from an organizational point of view can effectively be subdivided in a number of components. This document or subdocuments introduced in this document serve to specify the design of these components. The components are:

1) **Transceiver(s)**: The transceivers take care of the ground communication. The transceivers have to be specified in terms of transmission frequency, modulation scheme, transmission/reception power, data rate and data protocol. The number of transceivers required has to be determined as redundancy is required to ensure reliability of the Delfi-n3Xt mission.
2) **Antenna system**: The antenna system transmits or receives transceiver data. The antenna system has to be specified in terms of physical configuration, number of antennas and antenna gain pattern, amongst others.
3) **Ground Station(s)**: At least one ground station is required to receive transmitted telemetry, and to transmit telecommands. The ground station is not a part of the COMMS, but logically the communicational aspects of the satellite should be properly tuned to those on-ground.
4) **Delfi-n3Xt technology demonstration**: In the Delfi-n3Xt COMMS two non-essential systems are included, both with the primary purpose of technology demonstration for future applications.These are:
   - ***The STX***: A high speed transmitter that would allow, now or in the future combined with satellite pointing, to transmit a large amount of data during a pass.
   - ***OLFAR***: A science proof-of-technology that would allow low frequencies that exist in the universe to be intercepted and measured for studying.
5) **Link budget**: The link budget takes into account transceiver characteristics on the satellite and on-ground, as well as properties of the channel used for transmission, in order to conclude on the required transmission power levels and thus data rate. This budget should be taken into account in all other components. The link budget is however not discussed or introduced in this document, but in [SLR 0106]. Its results are however presented and used.

1.2 Document structure

In order to properly constrain the COMMS its higher-levels will be defined in chapter 2, along with all high-level requirements. Consequently, the choice of frequency licence type is made and commented on in chapter 3, as this affects the architecture and constraints of the COMMS. Chapter 4 then presents the resulting requirements on a COMMS level.

Chapter 5 moves into the design and determination of the COMMS Space Segment. In fact, COMMS is defined to be a space segment subsystem, but the link with the ground segment is also discussed in a later chapters. After all COMMS space segment parts have been introduced, its requirements are presented in chapter 6.

Chapter 7 gives a significant introduction to radio transmission techniques. This chapter is rather large, spanning almost 100 pages (including the appendices). It presents all theory required to determine the characteristics of the transmission links between Delfi-n3Xt and the ground station(s). The information is the result of literature research, performance calculations and performance measurements from Delfi-C^3. A reader with sizeable knowledge of transmission techniques can skip through it and focus on the general conclusions.

As the primary communication link, the only link being non-experimental is the UHF/VHF link; chapter 8 determines the communication link's characteristics. With this done, the PTRX and ITRX can be introduced in chapter 9 and 10 respectively. Chapter 11 describes the OLFAR experiment, due to its link with the PTRX. Chapter 12 then moves on to the secondary (S-band) link and its specification.

As all COMMS communication links and required components have been established, chapter 13 discusses the integration of the COMMS components on Delfi-n3Xt. Chapter 14 discusses COMMS operations and performance.

Chapter 15 moves into the topic of the ground segment, and discusses the part of ground segment required for proper communication.

Finally, chapter 16 presents the next steps that are important with respect to the COMMS.

Appendix A and Appendix B are in fact extensions of the matter discussed in chapter 7, in these appendices several modulation techniques and line codes are introduced in more detail.

2 The top-level and its requirements

The Delfi-n3Xt mission is specified by a large number of Configuration Items, CIs. There is one top-level CI, the Delfi-n3Xt mission, and a large number of CIs on lower levels. In fact, as CIs contain other CIs, a Configuration item Tree is formed [SLR 0325]. The CIs are based on hardware; a CI generally represents a physical system. Consequently, persons are, whenever possible, allocated to a CI on any level, the lower the CI, the more specified and tangible the work-package that can be derived from the CI. The same holds for documentation, documents are whenever possible also linked to the CIs; this document logically describes all things relevant to the COMMS CI. Finally, per CI requirements are determined. This chapter introduces the higher levels down to the COMMS, as well as the requirements of these; requirements are listed in [SLR 0167].

Figure 2.1 demonstrates the higher level CIs of the Delfi-n3Xt, which are relevant to the context of COMMS. The figure is relatively self-explanatory. The requirements belonging to the first three levels are introduced in sections 2.1 through 2.3 below. These requirements are *input requirements*, as they have been defined on a higher level. COMMS requirements are derived from these, as will be done in chapter 4.

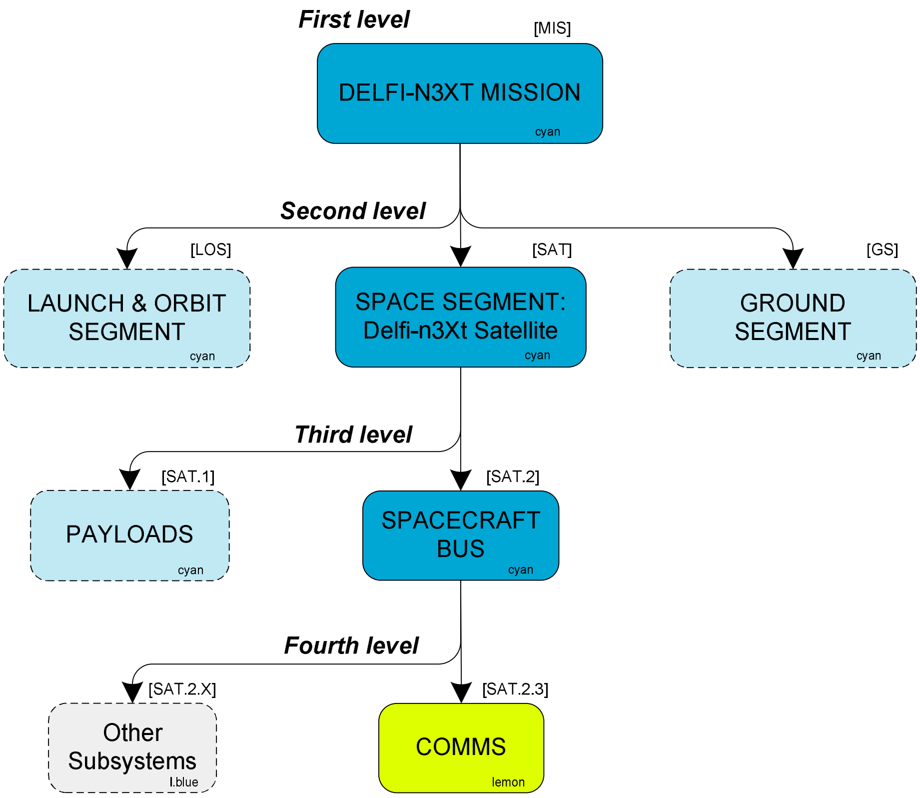


Figure 2.1: Top-level CI tree relevant to COMMS

2.1 First level: mission

The Delfi-n3Xt mission has a low number of top-level requirements, which are therefore for clarity reasons included in this document. Table 2-1 shows these requirements; they are derived in [SLR 0263]. Categories of requirements are indicated, as well as a requirement number. For more information on categories and numbering, see [SLR 0317].

Table 2-1: Delfi-n3Xt mission requirements

Category	Req. #	Requirement
FUNCTIONAL	MIS-F.01	The mission shall facilitate payloads from external partners
FUNCTIONAL	MIS-F.02	The mission shall facilitate bus advancements w.r.t. Delfi-C3
FUNCTIONAL	MIS-F.03	The mission shall facilitate additional experiments under strict conditions, as specified in [SLR 0263]
GENERAL	MIS-G.02	The mission shall facilitate educational objectives

2.2 Second level: satellite

The requirements of the second level can be seen in Table 2-2 below. These are important mainly because they influence all satellite bus systems, thus also the COMMS. Some comments are made below the table.

Table 2-2: Delfi-n3Xt satellite requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT-F.01	The satellite shall be able to communicate with the Ground Segment (GS)
FUNCTIONAL	SAT-F.02	The satellite shall be able to be put in orbit by the Launch and Orbit Segment (LOS)
FUNCTIONAL	SAT-F.03	The satellite shall be able to switch off any transmitter on the satellite
CONSTRAINT	SAT-C.01	All satellite systems shall comply with the mass budget, as given in [SLR 0018]
CONSTRAINT	SAT-C.02	All satellite systems shall comply with the volume budget, as given in [SLR 0303]
CONSTRAINT	SAT-C.03	All satellite systems shall comply with the power budget, as given in [SLR 0017]
CONSTRAINT	SAT-C.04	All satellite systems shall comply with the data budget, as given in [SLR 0282]
CONSTRAINT	SAT-C.05	All satellite systems shall comply with power and data bus interfaces, as specified in [SLR 0263]
CONSTRAINT	SAT-C.06	All satellite systems shall be able to withstand the launch environment
CONSTRAINT	SAT-C.07	All satellite systems shall be able to withstand the space environment

Some remarks:

- **Requirement SAT-F.01**: This functionality is completely covered by the COMMS.
- **Requirement SAT-F.03**: This requirement follows from ITU regulations; any transmitter using a transmission frequency should be able to be switched off in case of an emergency or harmful interference caused by a defect for example [SLR 0310]. It is however not a COMMS requirement; it can be taken care of using power supply switching.
- **All constraint requirements**: These requirements are applicable to all satellite systems, no matter on what level.

2.3 Third level: satellite bus

Finally, the third level or satellite bus requirements can be seen in Table 2-3 below. Again two requirements are important as they influence all satellite bus systems.

Table 2-3: Delfi-n3Xt satellite bus requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2-F.01	The satellite bus shall facilitate the payloads
FUNCTIONAL	SAT.2-F.02	All satellite bus systems shall generate housekeeping data when of interest to satellite operation
CONSTRAINT	SAT.2-C.01	All satellite bus systems shall adhere to reliability standards, as specified in [SLR 0263]

Some remarks:

- **Requirement SAT.2-F.02**: Any satellite bus system with an interface to the CDHS shall provide the latter with any housekeeping data that is of interest to satellite operation.
- **Requirement SAT.2-C.03**: A number of reliability standards are stated in [SLR 0263], which should be adhered to by all satellite bus systems. Most important is the demand of single-point-of-failure free systems.

3 Available frequencies

Before discussing the communication system and its requirements, it is important to present the topic of transmission frequency licences. The frequency bands and the consequent costs, be it money or return services, as well as the bandwidth restrictions pose requirements for the communication system architecture.

Section 3.1 introduces the licenses available, after which section 3.2 decides on the license type to be used for Delfi-n3Xt. Section 3.3 introduces the different frequency bands available to radio amateurs and finally section 3.4 presents advantages and disadvantages of these bands in order to decide on the bands used later in chapter 5 in determining the space segment of Delfi-n3Xt.

3.1 Types of frequency licenses

There are basically two types of licences available for a programme such as the Delfi programme with unspecified payloads. These two options are *commercial licenses* and *radio amateur licences*.

A **commercial license** is one option for a frequency license, but this is relatively expensive compared to the entire costs for a nanosatellite and the total budget available for building a university nanosatellite. No exact data could be obtained, but the range is reported to be € 10,000 – € 100,000 (2008) for one frequency (bandwidth) for a reasonable data rate for just one year. The license for a commercial frequency is less flexible in terms of time: if the launch of the satellite is delayed, the frequency license still needs to be paid for. Also, a delayed launch could cause the licensed time window to be missed.

A **radio amateur licence** forms the second option. With such an agreement, it is allowed to use certain frequencies within the radio amateur bands. As a return favour, however, the satellite needs to provide a service to the radio amateur community. This service is generally a store-and-forward system for messages, or a linear transponder, as in the case of Delfi-C^3. However, other options are also possible. This return service adds some complexity to the satellite.

Two restrictions apply to the use of amateur bands. The data transmitted over the amateur frequencies may not be used for any commercial purpose, and when using an amateur license data encryption is not allowed except for uplink telecommands. All data on what bit rates, encodings, modulation techniques and data protocols must be made available to the public, so that anyone is able to decode the information in the signal. These two restrictions pose some issues when payloads for industry are flown, as is the case for Delfi-n3Xt, but it is assumed that this can be solved. For Delfi-C^3 it was no problem.

3.2 Selection of a frequency license type

In order to make a decision on which bands to use, initially the following criteria were selected:

- **Cost**: the total cost for being able to use one frequency, including the additional costs for providing a return service in the case of using amateur bands.
- **Coordination time**: the time required from the start of the frequency coordination process to the point where the frequencies are actually available for use and the satellite is allowed to be launched.
- **Use for commercial purposes**: the ability to use the frequencies to transmit data that are used for commercial purposes.
- **Return favour requirement**: the requirement for having to provide a service to radio amateurs in return for using a frequency in one of their frequency bands.
- **Ground segment**: consequences the choice of frequency license has for the ground segment, i.e. coverage, heritage, and support.
- **Encryption**: whether or not it is allowed to use encryption to protect data.
- **Flexibility**: whether there is a penalty in some form when the launch is delayed.

Using the Analytical Hierarchy Process (AHP) [SLR 0389, SLR 0396] weights were assigned to these criteria. The results are shown in Table 3.1. The overall consistency ratio according to the definition of the AHP is 9%, which is less than 10% so it is acceptable [SLR 0389, SLR 0396]. It is concluded that the criteria *Coordination time*, *Commercial usage* and *Encryption* are not relevant, and are ignored. After removing these criteria, the consistency ratio is improved to 7%.

Table 3.1: Trade-off criteria

Criterion	Normalized weight (%)
Cost	37
Coordination time	6
Commercial usage	4
Return favour requirement	14
Ground segment support	28
Encryption	2
Flexibility in schedule	10
Total	100

The criteria *Cost*, *Return favour requirement*, *Ground segment support*, and *Flexibility* are used in the trade-off. The results are presented in Table 3.2.

Table 3.2: Trade-off table of frequency license types

	Cost	Return favour requirement	Ground segment	Flexibility in schedule	Total score
Weight (%)	**42**	**16**	**31**	**11**	**100**
Commercial	1 red Expensive	4 green No	1 red Expensive, no global coverage, no heritage	2 ora. Stricter license	1.59
Amateur	3 lime Return favour only	2 orange Yes	3 lime Support from GENSO and radio amateurs	3 lime Amat. flexible	2.84

red	Strongly disadvantageous
orange	Slightly disadvantageous
lime	Somewhat advantageous
green	Strongly advantageous

The above table clearly shows that for Delfi-n3Xt the most favourable option is to use frequencies in the radio amateur band.

The main reasons are the fact that a commercial license is expensive, and that when using the amateur bands the ground segment has global coverage and there is heritage from previous missions. The main disadvantage of using an amateur license is that the satellite needs to provide a service for radio amateurs in return, but a linear transponder has already been developed and flown on Delfi-C^3.

In case payload providers would not agree with the openness of the data, a backup option would be to use one commercial frequency over which payload data is transmitted. However, in that case it would more readily be a reason not to fly the respective payload; unless they would be willing to pay for the commercial license (partly). For Delfi-n3Xt, there are no objections to using a radio amateur licence.

As such, *it is decided to pursue a license for frequencies in the amateur band for Delfi-n3Xt.*

3.3 Available frequency bands

The radio amateur frequency bands that are available for use are listed in Table 3.3. Only those bands are shown that are of interest to Delfi-n3Xt. After the frequency range is selected, detailed frequency coordination can take place. This is done at first via the Amateur Satellite community (AMSAT), which divides and manages frequencies within the radio amateur bands. Official frequency conventions are then to be made via the International Amateur Radio Union (IARU), which serves as an intermediary between the satellite developer and the International Telecommunication Union (ITU). The ITU is a part of the United Nations, in which all countries are represented. The exact frequencies are determined by investigating possible interference within the satellite as well as with other users of the spectrum, which can either be other satellite operations or operations on the surface of the Earth.

Table 3.3: Selection of the frequency bands in the amateur satellite service [SLR 0310]

Band	Frequency range [MHz]	Remarks on the band
VHF band for satellites	145.80—146.00	Primary
UHF band for satellites	435.00—438.00	Non-interference basis
L-band	1260.0—1270.0	Earth-to-space only, non-interference basis
S-band	2400.0—2450.0	Secondary as it is part of ISM band

The VHF band is the generally used band for satellite communications; standard equipment is readily available, but it is also crowded. In the UHF band an extra rule is in place that stipulates that no interference should be caused to a number of (terrestrial) services operating. This has proven not to be a problem with Delfi-C^3. The L-band is uni-directional, and has the same non-interference rules. The S-band finally is part of the ISM band, being one designated for industrial, scientific and medical applications. This requires that harmful interference from these applications must be accepted, and is therefore a secondary band. Nevertheless, the ISM band is from 2400 to 2500 MHz, with a centre frequency set at 2450 MHz. Therefore, interference is limited at frequencies near 2400 MHz.

3.4 Decision factors

There are a number of factors that play a role in frequency selection. These include, but are not limited to: path loss, pointing accuracy, Doppler shift, available bandwidth and distribution of equipment among radio amateurs. These aspects are shortly commented on in subsections 3.4.1 through 3.4.5 below.

3.4.1 Path loss

Path loss of a radio signal occurs when RF energy is emitted into free space. As the distance between transmitting and receiving antenna increases, the power flux density decreases. This effect is stronger for increasing wavelengths. The path loss factor L_p is given by [SLR 0316]:

$$L_p = \left(\frac{\lambda}{4\pi S}\right)^2 = \left(\frac{c}{4\pi f S}\right)^2, \qquad (2\text{-}1)$$

Where λ is the wavelength, S the distance between transmitting and receiving antenna, f the frequency and c the speed of light. The path loss increases with increasing frequency and increasing distance. This would favour lower frequencies.

3.4.2 Antenna gain and pointing accuracy

At higher frequencies, it is easier to create antennas with high gain. An increase in gain, however, also increases the required accuracy of antenna pointing (the antenna pattern lobes of transmitting and receiving antennas must overlap, which becomes more difficult with narrower lobes) [SLR 0261].

3.4.3 Doppler shift

Due to the fact that the satellite is moving with respect to the ground station, a Doppler shift will occur. Due to the Doppler shift, the frequency at which the downlink signal is received and the frequency at which the uplink signal must be transmitted must be corrected for Doppler shift during a transmission. The Doppler shift is given by the following equation:

$$f = \frac{-v_r \cdot f_0}{c} \qquad (2\text{-}2)$$

Where f_0 is the centre frequency of the signal, v_r is the speed of the satellite relative to the ground station in line-of-sight direction and c the speed of light.

Assuming a perfectly spherical, non-rotating Earth, as well as a perfect circular orbit, the radial velocity with respect to the ground station can calculated relatively easily. This scenario is illustrated by Figure 3.1 below.

A big blue arrow that is indicated by v_c signifies the circular velocity of a satellite body. It is decomposed in two components, one signifying the tangential velocity component with respect to the ground station, v_t and the blue component signifying the required radial component with respect to the ground station, v_r.

Furthermore the radius of the Earth *R* and the satellite orbit altitude *h* are indicated, as well as angles *α* and *β,* with *α* signifying the *satellite elevation*.

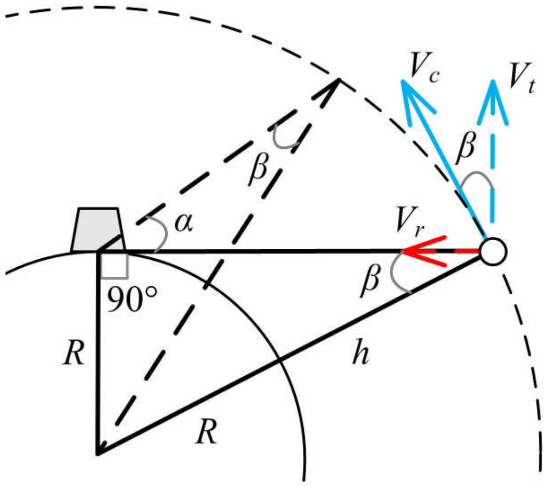


Figure 3.1: The radial velocity component of a satellite with respect to the ground station

The circular velocity can be calculated using the following relation [SLR 316]:

$$v_c = \sqrt{\frac{\mu}{R+h}} \tag{2-3}$$

Where μ is the gravitational parameter of the Earth. The tangential component can then be calculated using:

$$v_r = v_c \cdot \sin\beta \tag{2-4}$$

In order to calculate angle β then the sine rule can be used, expressing β in terms of the satellite elevation α .

$$\frac{\sin(90+\alpha)}{R+h} = \frac{\beta}{R} \tag{2-5}$$

Substituting relations 2-3, 2-4 and 2-5 in 2-2, yields the following equation for the Doppler shift:

$$Doppler\ shift = \frac{-v_c \cdot f_0}{c} \cdot \sin\left(\frac{R \cdot \sin(90+\alpha)}{R+h}\right) \tag{2-6}$$

Also derived can be the expression for rotational speed, being α / t , given the circumference of the orbit and its circular velocity:

$$\frac{\alpha}{t} = \frac{V_c}{2\pi(R+h)} \cdot 360 \tag{2-7}$$

Then, differentiating equation 2-6 and multiplying with relation 2-7 gives the Doppler shift rate:

$$Doppler\ shift\ rate = \frac{-v_c \cdot f_0}{c} \cdot \cos\left(\frac{R \cdot \sin(90+\alpha)}{R+h}\right) \cdot \frac{R \cdot \cos(90+\alpha)}{R+h} \tag{2-8}$$

The two relations 2-6 and 2-8 can be plotted to yield the following results, see Figure 3.2 (top and bottom). An elevation angle of 90° per definition means that the satellite is right above the ground station. It can be seen that total *Doppler shift* is maximum at 0 and 180° angles, with no Doppler shift occurring during an overhead pass. *Doppler shift rate* however is the highest when the satellite is passing overhead. Knowledge of the Doppler shift is important as receiver and/or transmitter should have a certain tuning range for frequency adjustment. Knowledge of Doppler shift rate on the other hand is important as it gives information on how frequent the frequency should be re-tuned.

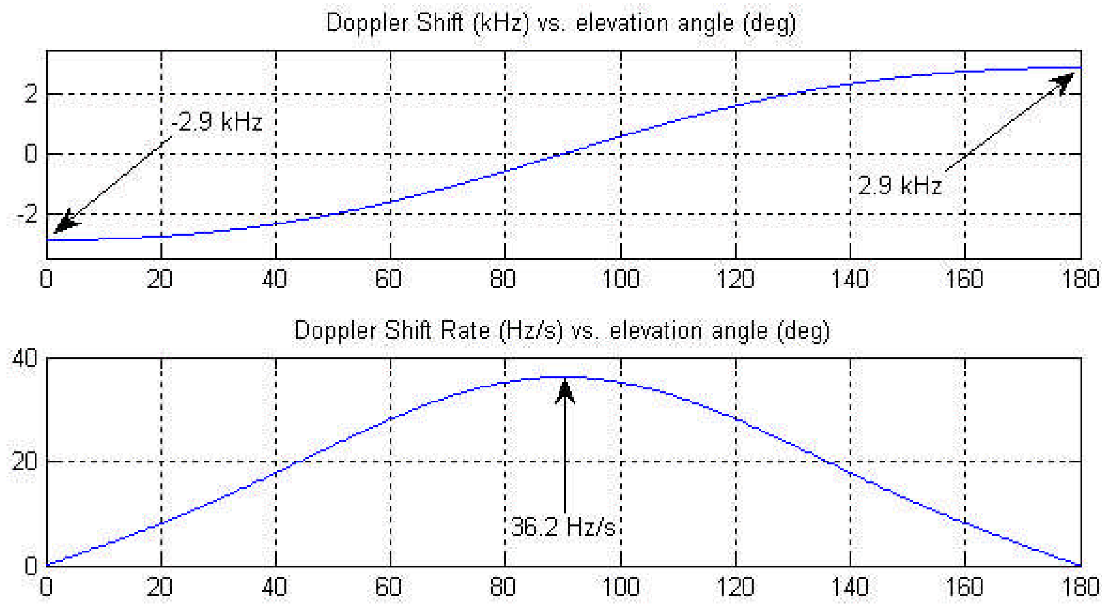


Figure 3.2: Doppler shift and Doppler shift rate vs. elevation angle during a pass; *h* = 600 and f_0 =146 MHz

Figure 3.2 shows the Doppler shift and Doppler shift rate for a transmission frequency of 146 MHz, and an orbit altitude of 600 km. Two maximum values have been indicated, being 2.9 kHz and 36.3 Hz/s. The orbit altitude of 600 km gives a worst case, as an orbit altitude of 850 km only yields 2.8 kHz and 32.3 Hz/s respectively. Also, the frequency used in the above plot is 145 MHz; Doppler shift and Doppler shift rate scale linearly with the frequency.

However, a number of assumptions have been made and it turns out actual Doppler shift and Doppler shift rate values are higher. Neglected aspects that have a significant influence:

- The Earth is actually rotating
- The Earth is actually slightly elliptical
- The orbit will be slightly elliptical, gradually increasing with time during the mission

As a result, Satellite Tool Kit or *STK* was used to give Doppler shift data or actually *radial velocity values* with respect to the ground station. These radial velocity values can be recalculated to Doppler shift values using relation 2-2 above. This was done for the Delfi-n3Xt range of orbits, having an orbit altitude between 600 and 850 km and being sun synchronous. Also, the radial velocities for Delfi-C^3 have been determined, having an orbit that is sun synchronous at 630 km altitude. The resulting values are presented in Table 3.4 below; Doppler shift in reality turns out to be some 15% higher than according to the theory, due to the assumptions made to obtain theoretical values.

Table 3.4: Maximum ***Doppler shift*** values for the Delfi-n3Xt range of orbits and Delfi-C^3

	***h* = 600 km**	***h* = 630 km (Delfi-C^3)**	***h* = 850 km**
f_0 = 146.00 MHz	3.41 kHz	3.39 kHz	3.22 kHz
f_0 = 436.00 MHz	10.20 kHz	10.11 kHz	9.62 kHz
f_0 = 2400.0 MHz	56.12 kHz	55.65 kHz	52.97 kHz

STK was also used to calculated radial accelerations with respect to the ground station, in order to give Doppler shift rate values again by multiplying with frequency and dividing by the light speed. The resulting values are shown in Table 3.4 below; again values are some 15% higher than the theoretical ones.

Table 3.5: Maximum ***Doppler shift rate*** values for the Delfi-n3Xt range of orbits and Delfi-C^3

	***h* = 600 km**	***h* = 630 km (Delfi-C^3)**	***h* = 850 km**
f_0 = 146.00 MHz	41.75 Hz/s	40.24 Hz/s	26.77 Hz/s
f_0 = 436.00 MHz	124.67 Hz/s	120.16 Hz/s	79.93 Hz/s
f_0 = 2400.0 MHz	686.25 Hz/s	661.42 Hz/s	439.98 Hz/s

3.4.4 Available bandwidth

The size of the different bandwidths of the revenant frequency bands has been presented in Table 3.3. It can be seen that the total bandwidth available in the VHF range is only 200 kHz. The actual VHF band is 2 MHz large but only a small portion is reserved for satellite communication. Many amateur satellites transmit on this frequency, so only a narrow band is available. It has been generally agreed upon within the radio amateur community that no bit rates higher than 9.6 kb/s shall be used in this band. This leads to an indicated bandwidth of some 20-25 kHz.

For UHF frequencies the radio amateur satellite bandwidth is larger, with 3 out of 10 MHz available. Still, maximum data rate are generally set at 9.6 kb/s, again leading to a general 20-25 kHz allocated bandwidth.

In reality, the frequency allocation is less precise. The allocated bandwidth is of little real importance, and overlapping frequency licenses are appointed, to cope with the limited bandwidth. For satellites, there is little change that overlap actually takes place. Finally, Doppler shift also shifts the actual transmission frequencies. With respect to the latter, it is most important that frequencies to do not deviate to *outside* of the radio amateur bands.

It can be concluded that for higher data rates a higher frequency band is necessary in order to comply with regulations restricting the frequency bandwidth and data rate. The L-band has some 10 MHz available, and the S-band even has 50 MHz available. Both are currently used by little satellites, and thus would allow larger bandwidth allocations.

3.4.5 Equipment of radio amateurs

The satellite will be part of the radio amateur service, and therefore it must provide a service for the radio amateur community. VHF bands are primarily supported by radio amateurs. UHF bands as well, but less commonly so. S-band support is growing over the last years. In conclusion, most participation and support can be expected at the lower frequencies.

4 COMMS requirements

In chapter 2 the top-level requirements were presented, belonging to the first three levels introduced in the same chapter. In this chapter the requirements for the fourth level, the COMMS itself, shall be derived and presented. In some cases these requirements follow from the top-level requirements, and another number of requirements follow from the choice to use radio amateur frequencies. Also, a number of requirements necessary on the COMMS level related to performance and security can and should be stated. Finally, a number of organizational requirements are stipulated.

The first type of requirement is labelled *technical requirement* and is presented in section 4.1. The second type, logically called *organizational requirement* is presented in section 4.2.

A number of top-level requirement stated in chapter 2 in fact also apply to all COMMS components, but these will not be restated. An example is the requirements concerning budgets and reliability.

4.1 Technical requirements

The technical requirements are presented in Table 4-1 and Table 4-3 below. As before, categories are given followed by a requirement number and the actual requirement. If applicable, the parent(s) is/are given. Underneath each requirement a rationale is given.

Table 4-1: COMMS technical requirements (part one)

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3-F.01	**The COMMS shall be able to receive telecommands transmitted by the Ground Station Network** ***Parent:*** *SAT-F.0.1* ***Rationale:*** *As COMMS is in charge of communicating with the GS, this requirement follows from the parent.*
FUNCTIONAL	SAT.2.3-F.02	**The COMMS shall transmit telemetry (payload and housekeeping data) from the satellite to the Ground Station Network** ***Parent:*** *SAT-F.0.1* ***Rationale:*** *As COMMS is in charge of communicating with the GS, this requirement follows from the parent.*
FUNCTIONAL	SAT.2.3-F.03	**The COMMS shall provide a return service to the radio amateur community** ***Rationale:*** *In the previous chapter the choice for radio amateur frequencies was explained. The logical consequence of this is that a return service should be provided.*
FUNCTIONAL	SAT.2.3-F.04	**The COMMS may be able to receive other data such as software updates transmitted by the Ground Station Network** ***Parents:*** *SAT-F.0.1 & MIS-F.0.3* ***Rationale:*** This requirement is a further specification of the parent requirement of being able to communicate with the GS. However, it also follows from the requirement to facilitate additional experiments, such as the ability to receive software upload data.
FUNCTIONAL	SAT.2.3-F.05	**The COMMS shall facilitate the STX** ***Parent:*** *MIS-F.0.3* ***Rationale:*** *The STX should be flown as a technology demonstration*

Table 4-2: COMMS technical requirements (part two)

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3-F.06	**The COMMS may facilitate the OLFAR experiment** ***Parent:*** *MIS-F.0.3* ***Rationale:*** *OLFAR is an experiment that may be flown.*
PERFORMANCE	SAT.2.3-P.01	**All transmitters shall provide a maximum permitted power level for spurious emissions in dBc according to -43 - 10 log (P), where P is the transmitted power** ***Rationale:*** *ITU regulations give boundaries for the transmission bandwidth out-of-band for any officially allocated transmission frequency. However, given the use of radio amateur bands and limited transmission power, the actual requirement is less strict, and more 'best-effort'.*
PERFORMANCE	SAT.2.3-P.02	**The bit error rate of all digital data transmission links shall be designed to be at most 10^-5** ***Rationale:*** *Bit errors follow from probabilities related to channel noise. A BER value of 10^{-5} is a common value in spaceflight and for non-critical information [SLR 0316].*
PERFORMANCE	SAT.2.3-P.03	**In the link budget a minimum link margin of 3 dB is required for all transmission links** ***Parent:*** *MIS-F.0.3* ***Rationale:*** *This value is again a common value in spaceflight, accounting for inaccuracies in the system model [SLR 0316].*
CONSTRAINT	SAT.2.3-C.01	**The frequencies for communicating with the satellite shall lie within the frequency bands allocated to the amateur satellite service** ***Parent:*** *SAT-F.0.1* ***Rationale:*** *As was decided in the last chapter, the radio amateur frequency bands shall be used for data communication.*
CONSTRAINT	SAT.2.3-C.02	**Any data transmitted to and from the satellite other than telecommands shall not be encrypted in any way** ***Parent:*** *SAT-F.0.1* ***Rationale:*** *This requirement follows from radio amateur regulations and is a consequence of the choice for the radio amateur frequencies.*
CONSTRAINT	SAT.2.3-C.03	**All receivers that can receive telecommands shall be turned on by default** ***Rationale:*** *This requirement is one that underlines the essential role of the receivers; in case of any malfunctions the ability to command the satellite can save the mission. Therefore special attention should be paid to the availability of the receivers.*
CONSTRAINT	SAT.2.3-C.04	**The COMMS shall facilitate a redundant set of transceivers** ***Parent:*** *SAT.2-C.01* ***Rationale:*** *Following from the parent requirement of reliability, a single-point-of-failure free communication system results in a desire for two transceivers; which makes more sense then doubling every component on one radio.*
CONSTRAINT	SAT.2.3-C.05	**All UHF and VHF antenna connections and transmission lines on the satellite will have an impedance of 50 ohm** ***Rationale:*** *This value is the commonly used value with COTS components. It has also been used in the Delfi-C^3 architecture.*
GENERAL	SAT.2.3-G.01	**It shall be prevented that parties other than designated satellite operators shall be able to command the satellite** ***Rationale:*** *This requirement follows from mission security. It can be implemented by using unpredictable or unknown (encryption) protocols, which consequently should be kept secret. This requirement of secrecy is further addressed in the next section on organizational requirements.*

4.2 Organizational requirements

A small amount of organizational requirements follow from the COMMS; these are listed and argued for in Table 4-3 below. No strict numbering system is determined (yet) so they are logically numbered ORG.1-3.

Table 4-3: STX TX requirements

Category	Req. #	Requirement
ORGANIZATIONAL	ORG.1	**A downlink specification for each transmission link shall be made publicly available, including transmission frequency, data rate, modulation scheme, data protocol and content** ***Rationale:*** *This requirement follows again from the choice of using the radio amateur transmission frequencies, as the latter allows no transmission of nonpublic information.*
ORGANIZATIONAL	ORG.2	**Uplink transmission frequencies, intermediate frequencies and oscillator frequencies in the receiver, telecommand codes and encryption keys and offset shall be kept secret and be available only in restricted documentation** ***Rationale:*** *This requirement follows from the desire of having no unauthorized people operating the satellite.*
ORGANIZATIONAL	ORG3	**Documentation on daily operations of the ground station as well as the ground station itself shall have restricted access** ***Rationale:*** *This requirement follows from the desire of having no unauthorized people operating the satellite.*

5 COMMS: space segment design

As stated in the introduction of this document, COMMS hardware includes a number of transceivers, experiments, the antenna system and the ground segment. As the latter is called the *ground segment*, the first three components logically form the *space segment*.

Section 5.1 introduces all CIs that make up the COMMS. Sections 5.2 and 5.3 consequently argue for the exact frequency bands that are to be used as well as the radio amateur return service that is to be provided. Using this knowledge, a resulting overview of the COMMS can be given in section 5.4, and a further specification of the CIs and their documentation is presented in section 5.5. Finally, sections 5.6 and 5.7 describe failure cases and Delfi-C^3 heritage amongst the COMMS components.

5.1 Description of the communication platform

It has been presented in the previous chapter that there are a number of payloads. One of these payloads is the ITRX; being a transceiver built by ISIS. It is built to function as a primary transceiver for CubeSats.

A primary radio, if possible being non-experimental and thus proven is required for Delfi-n3Xt. The ITRX however is a new design and is little proven. Also, it is a payload and as such little control on the actual status is exerted. This is unacceptable for a mission critical system and as such the *primary transceiver* or PTRX is decided to serve as the primary radio. In order to make the radio non-experimental in as far as possible, it has been decided that the PTRX is based on the RAP of Delfi-C^3 (requirement SYS.REQ.G.008). In reality, more changes have been necessary but nevertheless large parts are based on proven concepts with Delfi-C^3 flight heritage. The PTRX is used to transmit payload data and housekeeping data, together being: telemetry (TM), and to receive telecommands (TCs).

The implementation of the ITRX payload however allows the latter to be used as a redundant radio. Whereas the ITRX is not yet a finished design, the PTRX also requires more changes then anticipated. As such, the use of two different radios spreads risks but also possibly allows the possibly more power-efficient ITRX to be used during nominal operations. If in the end the ITRX turns out to be not ready, of course a second PTRX can be flown as a redundant radio.

As a result, there will be two main radios on board of Delfi-n3Xt:

- **PTRX**: The primary transceiver for transmitting telemetry and receiving telecommands;
- **ITRX**: The secondary transceiver for redundancy, also a payload.

Each of the radios is integrated on PCBs. As the signals produced or to be processed by the transceivers are to be translated to and from wireless signals by antennas, also required is the:

- **Antenna system**: Depending on the transmission frequencies, a set of antennas is required for the radios on Delfi-n3Xt, as well as all connections from the PCBs to the antennas. This antenna system would also include any required STX antenna parts, explained below.

Within the COMMS, two *technology demonstrations* are further more specified. These systems have no mission-critical purpose, and no resources should be devoted to them if these resources can otherwise be used for mission-critical systems. The demonstrations related to the COMMS:

- **STX**: An optional high-speed transmitter of telemetry. STX stands for S-band transmitter. The selection of frequency band will be explained below in section 5.2. A transmitter in a different frequency band requires extra antenna(s) and connections, thus enlarging the content of the antenna system.
- **OLFAR**: Using the same antennas as the PTRX, signal of frequencies between 1 and 30 MHz are to be received in order to investigate space-borne radiofrequency waves.

The following figure shows all configuration items or *CIs* that represent parts of the total communication system. This figure introduces one extra level to the CI tree presented in Figure 2.1 above.

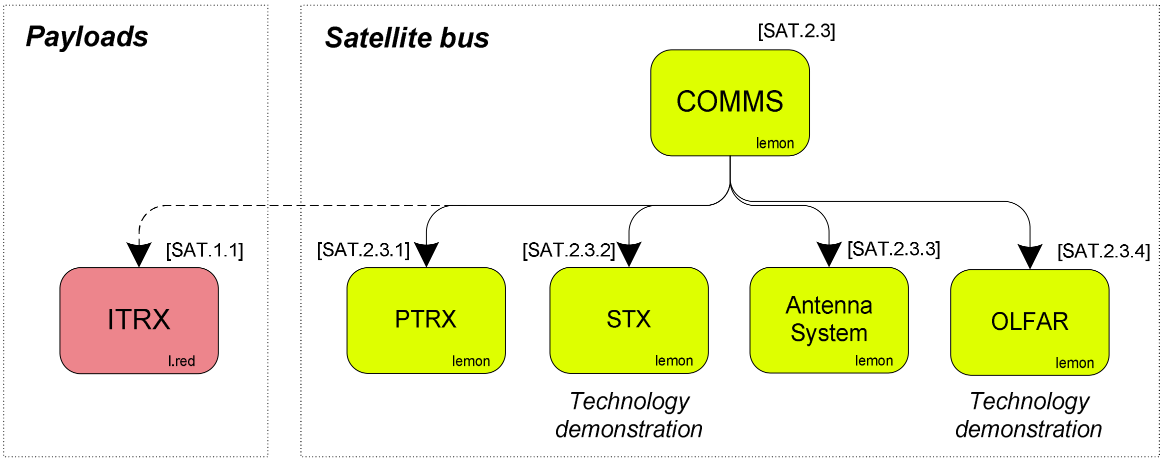


Figure 5.1: The COMMS CIs

5.2 Frequency band selection

In chapter 3 the choice was made to use radio amateur bands. However, it remains to confirm the exact bands, as multiple radio amateur frequency bands exist. Subsections 5.2.1, 5.2.2 and 5.2.3 present the choice of band for respectively PTRX and ITRX, STX, and OLFAR. The exact transmission frequencies are presented in chapters 8 and 12 on the UHF/VHF and S-band transmission links.

5.2.1 PTRX and ITRX

The PTRX and the ITRX are based on the RAP of Delfi-C^3. It is a requirement that for the PTRX the RAP is not changed to a large extent, similarly the ITRX will be designed to use the popular UHF/VHF bands. A change of frequency would incur design changes, thereby increasing complexity. Also, as will be commented on in chapter 8, reusing the same frequency bands simplifies the frequency filing procedure as Delfi-C^3 frequencies can be reused. Thus, both the PTRX and the ITRX will use the UHF band for uplink and the VHF band for downlink. More precisely, they will use the satellite transmission portions of the radio amateur band, in turn being part of the UHF and VHF bands. In terms of frequencies:

- PTRX/ITRX downlink: 145.80-146.00 MHz
- PTRX/ITRX uplink: 435-438 MHz

5.2.2 STX

The high-speed downlink transmitter is not allowed to operate in the UHF band, since this can cause interference with both the receiver of the PTRX as well as the receiver in the ITRX. This leaves the VHF band and the S-band (2—4 GHz) as possible frequency bands. The L-band (1—2 GHz) can only be used in the Earth-to-space direction in the amateur satellite service [SLR 0310].

Since for VHF only a very limited bandwidth is available, this is also not a suitable frequency band for a high-speed link. Frequencies higher than S-band are not considered because the higher the frequency, the more difficult the design will be. Due to the increased path loss more power or a higher antenna gain (with the resulting pointing requirements) is needed. Operation in S-band is considered to be ambitious already, and therefore this will be the maximum frequency that can be used for the STX on Delfi-n3Xt. At the same time, different CubeSats, such as GeneSat and CanX-2, and the S-band transmitter HISPICO have demonstrated large data rates to be achievable within the s-band, using small transmitters. Considering all these arguments, *it is concluded that the frequency band of the experimental high-speed transmitter (STX) is the S-band.* In terms of frequencies:

- STX downlink: 2400-2450 MHz

5.2.3 OLFAR

The OLFAR experiment (see [SLR 0170]) receives frequencies in the band 10—30 MHz, inherent to the experiment. Transmission of the signals will be done in the VHF band (144—146 MHz), using *the same transmission architecture as the PTRX*. Since the satellite only receives at 0.3—30 MHz, no frequency coordination needs to be done (this only applies for transmitting).

5.3 Return service to radio amateurs

Delfi-n3Xt will operate in the amateur satellite service. A requirement is that the satellite provides a service for the radio amateur community in return for use of radio amateur frequency bands, as explained in chapter 3. Examples of return favours are:

- ***Carrying a linear transponder***: A transponder receives signals in some frequency band, amplifies these signals and retransmits the signal in another frequency band. A *linear* transponder works in such a way that the shape of the signal is not altered. A linear transponder can therefore be used for all kinds of signals: amplitude modulated signals, frequency modulated signals as well as phase modulated signals. The use of an Automatic Gain Control amplifier however might remove the support for amplitude modulated signals.
- ***Providing a store-and-forward service***: This type of service allows users to up- and download message to and from the satellite.
- ***Testing of new radio equipment to advance radio science***: New equipment that can, if successful, be of use to the radio amateur community, can be tested.
- ***Providing a beacon***: This is the simplest return service. Basically any signal that is transmitted in the radio amateur bands forms a beacon. The content of the transmission can be altered to satisfy radio amateurs.

[SLR 0248] describes the trade-off performed for Delfi-C^3. It mentions that discussions with the radio amateur community have brought forward a number one preference of a linear transponder as the return favour. It:

- Accommodates a large number of signal types due to its linearity
- Allows multiple radio contacts at the same time
- Can easily be integrated in the transceiver architecture
- Has proven itself on a number of amateur satellites, the latter of course being Delfi-C^3.

As the linear transponder functionality consequently has been implemented successfully on Delfi-C^3, its reimplementation on Delfi-n3Xt is rather simple. Of course a linear transponder has already been flown (on Delfi-C^3), so its application is less novel and its inclusion could be perceived to be less valuable. Nevertheless, the Delfi-C^3 transponder is no longer active, and there seem to be no reasons to think the feature might be less interesting to radio amateurs. Early discussions with the radio amateur community have not presented any doubts with respect to the re-inclusion of a linear transponder. As such:

- ***Delfi-n3Xt will provide a linear transponder service as a return service to radio amateurs***

It will be included on the PTRX, which is an update to the original RAP, which featured the linear transponder as well. In fact, a beacon will also be provided to announce the activation of the linear transponder. The transponder has no direct functionality for the mission; its implementation is only as return-of-service for the use of the radio amateur frequencies and the forwarding of our telemetry by radio amateurs during normal operations. The linear transponder is further discussed in chapter 9 on the PTRX.

Furthermore, in the Delfi-C^3 mission the RASCAL client was conceived. This program allows demodulating sampled telemetry data, interpreting and visualizing the data and forwarding it to the TM (telemetry) Database. This program was welcomed with open arms by the radio amateurs and at the same time provided us with invaluable data. As such:

- ***A second return service will be the Delfi-n3Xt data reception client, named DUDe***

In addition, a number of radio amateurs have expressed their interest in having an S-band beacon on a satellite. Whenever the STX will be on, telemetry will likely be transmitted, thereby posing as a beacon and a return service. This can be seen as a third return service, but will not be explicitly stated to be one as it is in fact simply our TM transmission. Also, no link between the transponder and the STX for a more complicated S-band transponder system shall for now be assumed, given the additional complexity required. This might be an option for the future.

For now, the aforementioned return services are deemed sufficient. Indeed, a conversation with Graham Shirville, who is in charge of the frequency conversation within the AMSAT community, confirms that the return services stated would be sufficient.

5.4 Overview of the communication platform elements

All main functionalities and frequency bands of the radios have been specified. One other functionality exists and has not been mentioned above, which is the payload functionality of the ITRX; a number of tests can be performed by the latter involving transmission and reception. Combining all radio functionalities and distinguishing between transmission and reception yields the following Figure 5.2. It also shows the transmission frequencies. Finally, OLFAR and the STX are indicated as being a *technology demonstration*.

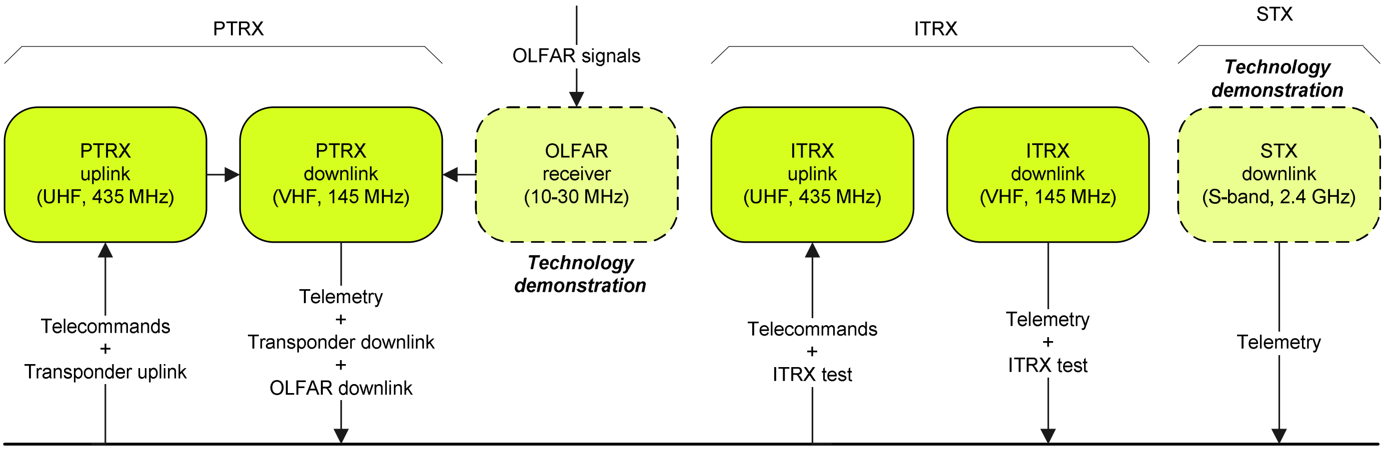


Figure 5.2: Delfi-n3Xt radios and their inputs and outputs

Figure 5.3 below consequently shows not only the radio segments, but also the antenna system, as well as the OBC, or on-board computer of Delfi-n3Xt. The latter is the communication point for all digital data. In this figure it can be clearly seen that if the STX is to be flown, it requires a separate antenna system.

Only transmission related interfaces are shown; other interfaces such as those related to housekeeping data are commented on in later chapters. It can be seen that both PTRX and ITRX can interface with the OBC to exchange TM and TCs. Both the PTRX and ITRX also exchange signals with the Antenna System, which can be seen to consist of Phasing Circuit, S-band Antenna Connection and Antennas. It should be noted that OLFAR is most likely largely integrated on the UHF/VHF Phasing Circuit; it is however *not* part of the Antenna System, it is defined as a separate CI. This is of course assuming the OLFAR will be flown.

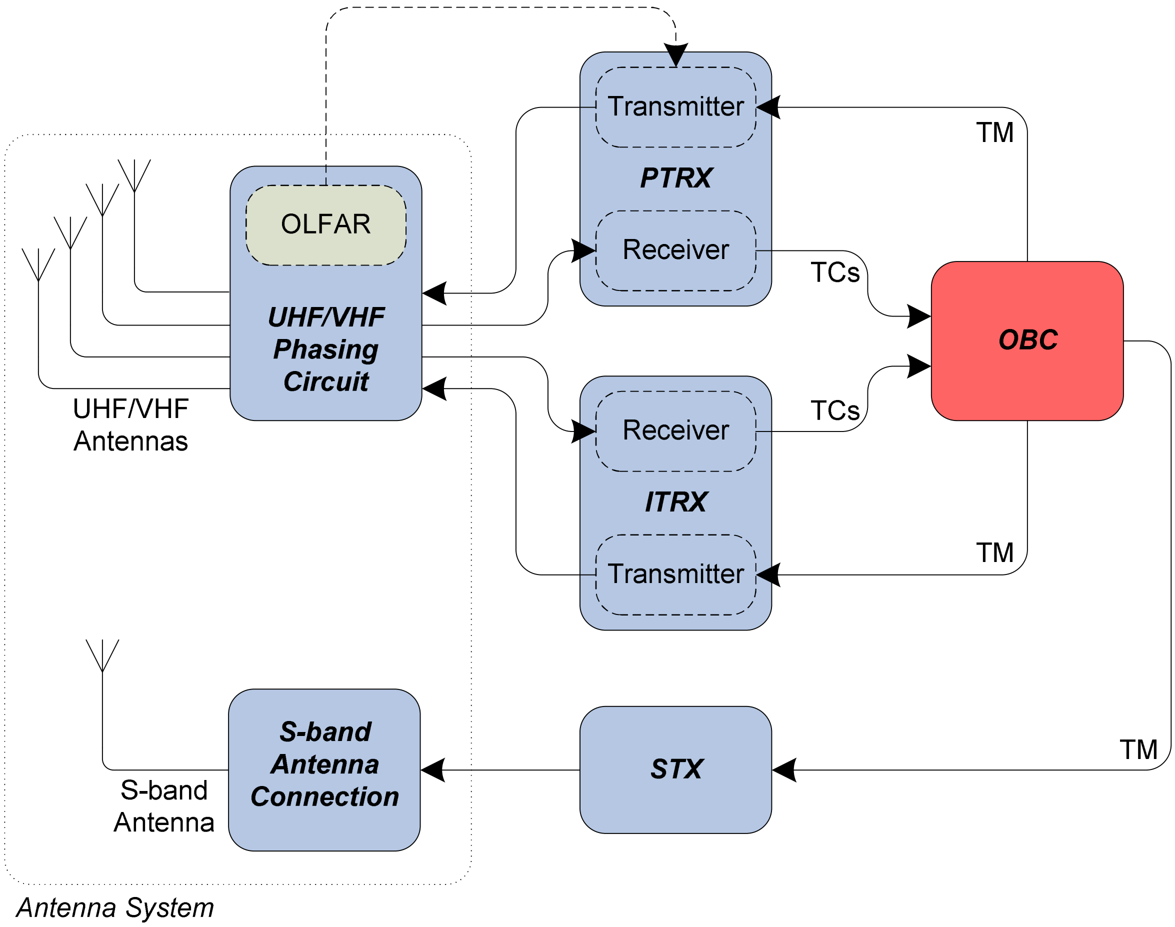


Figure 5.3: Delfi-n3Xt radios and their inputs and outputs

5.5 Specification of the communication platform elements

This document aims to provide a first specification of all systems, at least in terms of subfunctionalities, interfaces and integration on Delfi-n3Xt. This section will specify what exactly is discussed in later chapters, and *what is discussed in other (project) documentation*.

In the sections 5.5.1 through 5.5.4 below the 5 lower level CIs are further commented on, starting with the transceivers, followed by the experimental transmitter, in turn followed by OLFAR experiment and finishing off with the required antenna system.

5.5.1 PTRX

This document describes the top-level design of the PTRX, including the specification of its (sub)functionalities and its CIs. In order to do so, the entire UHF/VHF link is first specified with regard to power, bandwidth and complexity effects, taking into account the ground station effects as well. In other words, important to the PTRX are:

- Chapter 8: The UHF/VHF communication link
- Chapter 9: PTRX

A large number of other documents describe the PTRX or parts of it, on a lower level and generally related to electrical aspects. An extensive list of current PTRX-centred documentation:

- [SLR 0281]: PTRX Design
- [SLR 0591}: PTRX Transmitter
- [SLR 0592]: PTRX Modulator
- [SLR 0722]: PTRX Frame Interpreter and Frame Generator Design
- [SLR 0659]: PTRX Electrical Circuit Diagrams

Also, PTRX Processors software is integrated in the SLR:

- [SLR 0654]: Frame Interpreter software
- [SLR 0655]: Frame Generator software

5.5.2 ITRX (payload)

This document introduces the ITRX and its functionalities, modes and interfaces. However, as it is a payload, it is treated more or less as a black box. Therefore the main design is not part of Delfi programme. In this document the following chapters are of interest:

- Chapter 8: The UHF/VHF communication link
- Chapter 10: ITRX

There is one and only on other document that describes ITRX aspects, and that is a general interface document which is used for all payloads:

- [SLR 0182]: ITRX Interface Control Document

5.5.3 STX (experiment)

The STX is only a transmitter as opposed to a transceiver. It has no Delfi heritage and has to be designed from scratch. Hopefully, the design will have advanced to a point where it can be flown at launch. A functional and operational specification, as well as a possible design of its lower-level components has been made.

This document determines aspects with respect to the S-band communication link, involving not just the STX but also the ground segment, and similarly the operations of the STX will be discussed. This is done in:

- Chapter 12: S-band communication link
- Chapter 14: Operations and performance

The functional specification of the STX, its configuration items, its interfaces and its operational modes are discussed in a separate top-level document. The same document presents the *design philosophy* of the STX given the project status, as well as a possible hardware design and its status:

- [SLR 0387]: Top-level design of the STX

5.5.4 OLFAR (experiment)

This document introduces the functionalities and architecture of the OLFAR experiment, as well as its impacts on the other systems. In order to so it is mainly discussed in two chapters:

- Chapter 9: PTRX
- Chapter 11: OLFAR

Furthermore, there is one main document discussing the entire OLFAR experiment:

- [SLR 0170]: The OLFAR experiment

Finally, the relation of the OLFAR system to the Antenna System is discussed in:

- [SLR 0036]: Antenna System Design

5.5.5 Antenna system

The antenna system is not discussed in this document, but in a separate one. This document is:

- [SLR 0036]: Antenna System Design

The failure cases of the antenna system are however quickly assessed below.

5.6 Failure cases

Without a functioning communication system it is not possible to retrieve the data produced by the payloads or to control the satellite. Therefore, it is important to consider the impact of failures of components of the communication system, given the vision of complete redundancy of all mission critical satellite elements.

Failure of PTRX receiver – If the PTRX receiver fails, the ITRX receiver should still function, therefore telecommands can still be received. At the same time, the PTRX is designed to have its receiver and transmitter sections separately connected to the power and data bus, therefore a failure of the receiver would not automatically cause failure of the transmitter section. The transponder functionality however requires both, and is lost in case of failure of any of the two. The latter however is not mission-critical and is thus acceptable.

Failure of PTRX transmitter – A failure of the PTRX transmitter leaves the ITRX with the transmission responsibility. Possible the STX can be used as well. The PTRX receiver is not automatically lost, but the linear transponder is.

Failure of ITRX – Telemetry is, as per default, transmitted by the PTRX or possibly the STX. Receiving telecommands is done, also per default, by the PTRX. The ITRX experiments are lost. The ITRX is not designed to have separately connected receiver and transmitter sections.

Failure of STX – If the STX will be flown, and it fails, PTRX or ITRX only transmit telemetry data. The STX in any case is not mission-critical.

Alos, with respect to the antenna system the following can be stated. The details of the antenna system are explained in [SLR 0036].

Failure of one UHF/VHF antenna – A failure of one UHF/VHF antenna will not cause loss of the link, as four different antennas are in place. Two opposite antennas are actually only required. A failure of one antenna (for example because the deployment fails) will likely only cause an assymetrical gain pattern and consequently slightly deteriorated performance (depending on the nature of the failure).

Failure of one S-band antenna – A failure of one out of two S-band antennas or its connection will still allow for communication in more or less 50% of the time. For the purpose of technology demonstration, this would still be sufficient.

It is concluded that the presented communication subsystem concept fulfils the mission goals, and for the critical bus systems there is at least one backup. Failure cases are detected by either faulty or not received housekeeping data, or by the Local EPS systems on-board of all bus-connected systems.

5.7 Delfi-C3 heritage

The following components of Delfi-n3Xt have Delfi-C^3 heritage:

- The PTRX (the general transmission and reception architecture are based on the RAP, and the linear transponder is equal in design)
- The VHF antennas and the deployment MABs (but will now also be used for UHF signal reception)

New components of the Delfi-n3Xt communication system are:

- The STX
- The STX Antenna System
- The phasing circuit required for the integration of VHF and UHF antennas
- The OLFAR experiment and its integration

6 Lower level requirements

Given the CIs presented in the last chapter, requirements can be stated for each of them. In chapters 2 and 4 the top-level requirements have been presented. A number of requirements on higher levels are equally requirements on lower levels. Examples on the highest level are all those related to budgets, as well space and launch environment requirements. On the COMMS however as well, performance requirements related to bandwidth occupancy and bit error rates for example are equally applicable to COMMS radios. As such Table 6-1 below gives all requirements that are generally relevant for the lower level systems, without repeating the exact requirements.

Table 6-1: Generally applicable higher level requirements

Level	**Requirement(s)**
Satellite	SAT-C.01, SAT-C.02, SAT-C.03, SAT-C.04, SAT-C.05, SAT-C.06, SAT-C.07
Satellite Bus	SAT.2-F.02, SAT.2-C.01
COMMS	SAT.2.3-P.01, SAT.2.3-P.02, SAT.2.3-C.02, SAT.2.3-C.03, SAT.2.3-C.05

Sections 6.1 through 6.5 discuss respectively the PTRX, ITRX, STX, Antenna System and OLFAR requirements.

6.1 PTRX

The PTRX requirements are presented and argued for in Table 6-2 and Table 6-3 below.

Table 6-2: PTRX requirements (part one)

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.1-F.01	**The PTRX shall be able to receive telecommands** ***Parent:*** *SAT.2.3-F.01* ***Rationale:*** *With the PTRX being the primary transceiver, this requirement follows logically from the parent.*
FUNCTIONAL	SAT.2.3.1-F.02	**The PTRX shall be able to transmit telemetry (payload and housekeeping data)** ***Parent:*** *SAT.2.3-F.02* ***Rationale:*** *With the PTRX being the primary transceiver, this requirement follows logically from the parent.*
FUNCTIONAL	SAT.2.3.1-F.03	**The PTRX shall have linear transponder functionality** ***Parent:*** *SAT.2.3-C.02* ***Rationale:*** *In the previous chapter the choice for a linear transponder was explained, along with the logic of integration it on the PTRX.*
FUNCTIONAL	SAT.2.3.1-F.04	**The PTRX may facilitate the OLFAR experiment** ***Parent:*** *SAT.2.3-F.06* ***Rationale:*** If the OLFAR experiment is indeed flown, the PTRX should (partly) facilitate its integration
FUNCTIONAL	SAT.2.3.1-F.05	**The PTRX may be able to receive other data such as software updates transmitted by the Ground Station Network** ***Parent:*** *SAT.2.3-F.04* ***Rationale:*** *If software data frames are uploaded with characteristics differing from TC packets, these should be properly received and forwarded to the OBC.*
INTERFACE	SAT.2.3.1-I.01	**The PTRX shall have an interface with the Antenna System to forward and receive an analogue data signal** ***Rationale:*** *This interface is logically required.*
PERFORMANCE	SAT.2.3.1-P.01	**The PTRX shall allow for transmission of data with a data rate of at least 1200 bits/s** ***Rationale:*** *The RAP on which the PTRX is based attained a downlink data rate of 1200 b/s; as such this is the minimum that is to be achieved for the PTRX.*
PERFORMANCE	SAT.2.3.1-P.02	**The PTRX shall allow for reception of data with a data rate of at least 600 bits/s** ***Rationale:*** *The RAP on which the PTRX is based attained an uplink data rate of 1200 b/s; as such this is the minimum that is to be achieved for the PTRX.*

Table 6-3: PTRX requirements (part two)

Category	Req. #	Requirement
CONSTRAINT	SAT.2.3.1-C.01	**The centre frequency of the VHF signal transmitted by the PTRX transmitter shall lie within the frequency band 145.80-146.00 MHz** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *As the parent requirement specifies, the radio amateur bandwidths shall be used. This requirement is expanded here by including exact frequencies.*
CONSTRAINT	SAT.2.3.1-C.02	**The centre frequency of the UHF signal received by the PTRX receiver shall lie within the frequency band 435-438 MHz** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *As the parent requirement specifies, the radio amateur bandwidths shall be used. This requirement is expanded here by including exact frequencies.*
CONSTRAINT	SAT.2.3-C.02	**Received telecommands shall be encrypted** ***Parent:*** *SAT.2.3-G.01* ***Rationale:*** *Given the requirement of restricting the commandability of the satellite, encryption shall be applied as was done with Delfi-C^3. In fact, this choice is formally made in chapter 8.*
CONSTRAINT	SAT.2.3-C.03	**Changes of the PTRX with respect to the RAP design shall be minimized** ***Rationale:*** *The RAP in practice has been successful although with a number of issues. As the PTRX is a mission-critical system, it makes sense to try and limit the changes to the proven design to essential or valuable ones.*

6.2 ITRX

The ITRX requirements are largely the same as those of the PTRX, but with the experimental features and transponder functionality removed. Also, as the Delfi-team is not involved with the actual design, no lower-level design requirements are specified. The resulting reduced list of requirements is presented in Table 6-4 below.

Table 6-4: ITRX requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.1.1-F.01	**The ITRX shall be able to receive telecommands** ***Parent:*** *SAT.2.3-F.01* ***Rationale:*** *See PTRX requirements description*
FUNCTIONAL	SAT.1.1-F.02	**The ITRX shall be able to transmit telemetry (payload and housekeeping data)** ***Parent:*** *SAT.2.3-F.02* ***Rationale:*** *See PTRX requirements description*
INTERFACE	SAT.1.1-I.01	**The ITRX shall have an interface with the Antenna System to forward and receive an analogue data signal** ***Rationale:*** *See PTRX requirements description*
PERFORMANCE	SAT.1.1-P.01	**The ITRX shall allow for transmission of data with a data rate of at least 1200 bits/s** ***Rationale:*** *See PTRX requirements description*
PERFORMANCE	SAT.1.1-P.02	**The ITRX shall allow for reception of data with a data rate of at least 600 bits/s** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *See PTRX requirements description*
CONSTRAINT	SAT.2.3.1-C.01	**The centre frequency of the VHF signal transmitted by the PTRX transmitter shall lie within the frequency band 145.80-146.00 MHz** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *See PTRX requirements description*
CONSTRAINT	SAT.2.3.1-C.02	**The centre frequency of the UHF signal received by the PTRX receiver shall lie within the frequency band 435-438 MHz** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *See PTRX requirements description*
CONSTRAINT	SAT.2.3-C.02	**Received telecommands shall be encrypted** ***Parent:*** *SAT.2.3-G.01* ***Rationale:*** *See PTRX requirements description*

6.3 STX

The STX requirements are presented and argued for in Table 6-5 below.

Table 6-5: STX requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.1-F.01	**The STX shall be able to transmit telemetry (payload and housekeeping data)** ***Parent:*** *SAT.2.3-F.02* ***Rationale:*** *With the PTRX being a transmitter, this requirement follows logically from the parent.*
FUNCTIONAL	SAT.2.3.1-F.02	**The STX shall have a buffer to store data that must be transmitted** ***Parent:*** *SAT.2.3-F.02* ***Rationale:*** *As STX transmission data rate is supposed to be higher than data is received from the OBC, data storage is required.*
FUNCTIONAL	SAT.2.3.1-F.03	**The STX shall have an interface with the Antenna System to forward an analogue data signal** ***Rationale:*** *This interface is logically required.*
FUNCTIONAL	SAT.2.3.1-F.04	**The STX shall allow for transmission of data with a data rate of at least 9600 bits/s** ***Parent***: *SAT.2.3-F.06* ***Rationale:*** *This performance goal is based on the fact that CubeSat radios on UHF/VHF frequencies generally have data rates not higher than 9.6 kb/s. Similarly, the required assumed bandwidth (up to 20 kHz) is generally the maximum allocated within radio amateur transmission bands. Therefore, a significant improvement by a (high-speed) transmitter in a different frequency bandwidth is made and only made by improving on this limit.*
FUNCTIONAL	SAT.2.3.1-F.05	**The STX Data Storage must have a size that represents at least 24 hours of all unique payload data and housekeeping data** ***Parent:*** *SAT.2.3-F.04* ***Rationale:*** *This requirement is based on the fact that (at least) one good day-light pass over Delft is assumable per day. Multiple passes per day can of course be utilized, and larger data storage could be used if available at little cost.*
INTERFACE	SAT.2.3.1-I.01	**The STX is assumed to transmit the exact same raw data as the PTRX, with raw data indicating data excluding transmission protocol specific data** ***Rationale:*** *This requirement reduces the impact of the STX on the satellite, as the data can simple be sent to both PTRX/ITRX and STX.*
CONSTRAINT	SAT.2.3.1-C.01	**The STX signal shall be within the S-band frequency band between 2400 and 2450 MHz** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *As the parent requirement specifies, the radio amateur bandwidths shall be used. This requirement is expanded here by including exact frequencies.*

6.4 Antenna System

The Antenna System requirements are presented and argued for in Table 6-6 below.

Table 6-6: Antenna System requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.3-F.01	**The Antenna System shall be able to receive the uplink signal over the UHF frequencies** ***Parent:*** *SAT.2.3-F.01 & SAT.2.3-C.01* ***Rationale:*** *This requirement follows logically from the parents.*
FUNCTIONAL	SAT.2.3.3-F.02	**The Antenna System shall be able to radiate the downlink signal over the VHF frequencies** ***Parent:*** *SAT.2.3-F.02 & SAT.2.3-C.01* ***Rationale:*** *This requirement follows logically from the parents.*
FUNCTIONAL	SAT.2.3.3-F.03	**The Antenna System may be able to radiate the downlink signal over the S-band frequencies** ***Parent:*** *SAT.2.3-C.01* ***Rationale:*** *This requirement follows logically from the parents.*
FUNCTIONAL	SAT.2.3.3-I.01	**The Antenna System shall have an interface with the PTRX to forward and receive an analogue data signal** ***Rationale:*** *This interface is logically required.*
FUNCTIONAL	SAT.2.3.3-I.02	**The Antenna System shall have an interface with the ITRX to forward and receive an analogue data signal** ***Rationale:*** *This interface is logically required.*
INTERFACE	SAT.2.3.3-I.03	**The Antenna System may have an interface with the STX to forward an analogue data signal** ***Rationale:*** *This interface is logically required.*
CONSTRAINT	SAT.2.3.3-C.01	**The polarization of the Antenna System shall be circular** ***Rationale:*** *For a satellite it is desired to have circular polarization, because with linear polarization the orientation of the polarization at the ground station is unknown due to Faraday rotation in the atmosphere.*
CONSTRAINT	SAT.2.3.3-C.02	**A single antenna deployment failure should not cause loss of the link** ***Parent:*** *SAT.2-C.01* ***Rationale:*** *This follows from the parent requirement, stating a single-point-of-failure free design.*

6.5 OLFAR

The OLFAR requirements are presented and argued for in Table 6-7 below.

Table 6-7: OLFAR requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.4-F.01	**OLFAR shall receive signals in the frequency range 0.3 – 30 MHz and prepare these for retransmission**
		Rationale: *This is the functionality as defined for the OLFAR experiment.*
FUNCTIONAL	SAT.2.3.4-I.01	**OLFAR shall have an interface with the Antenna System to receive an analogue data signal**
		Parent: *SAT.2.3-F.02 & SAT.2.3-C.01* ***Rationale:*** *This interface is logically required.*
FUNCTIONAL	SAT.2.3.4-I.02	**OLFAR shall have an interface with the PTRX to forward an analogue data signal**
		Parent: *SAT.2.3-C.01* ***Rationale:*** *This interface is logically required.*
FUNCTIONAL	SAT.2.3.4-C.01	**OLFAR is classified as an experiment, as specified in [SLR 0263]**
		Parent: MIS-F.03 ***Rationale:*** *OLFAR is classified as one of the experiments indicated by the parent requirement.*

7 Transmission techniques

In layman's terms, a radio often refers to a receiver, a transmitter, or both: a transceiver. In all these cases its functionality can be described quite simply, for example a transmitter takes a data input (bits, a voice signal, sounds, but in any case at some point in the process portrayed by an electronic signal), and converts this to a transmittable signal, and finally transmits it. A receiver does the reverse, and logically a transceiver can do both. Due to the increased use of computers in all radios, at some point or another, the data usually consists of bits. As such, the functionality of the same transmitter can be reduced to "*preparing a digital signal for transmission, and transmitting it*".

In terms of hardware, a logical distinction can usually be made between antenna, and the rest. The radio is made up of electronics, integrated circuitry, whereas the antenna is usually connected to these electronics via a cable, and has a distinct shape and location on the 'radio'. The same holds for Delfi-n3Xt. As such, three radios can be said to be integrated on it, together with the antenna systems to take care of actual physical transmission. Therefore, the functionality of our transmitter has reduced to "*preparing a digital signal for transmission*". This functionality is exemplified by Figure 7.1 below.

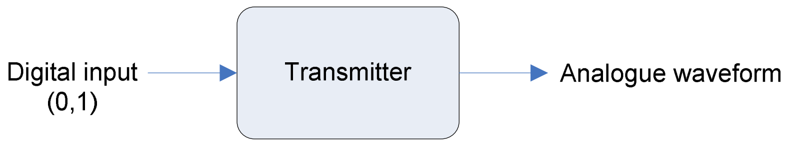


Figure 7.1: Input and output of a simple transmitter

This exact functionality in fact takes place in not just radios, but also in (wired/wireless) internet connections, or even USB connections. As such, a standardized model has been described to subdivide transmission techniques and functionalities, called the OSI-model (Open Systems Interconnection). This model applies layers to conceptually similar functions, interfacing solely with the layer above and under it. As such, communication functionalities are often described in layers (for a more detailed explanation, see [SLR 0574]).

Whereas the OSI-model has 7 layers in total, only 2 are required to accurately describe Delfi-n3Xt communication with the ground station. Other layers, such as those required for data routing are addressed within the software on-ground. The two required layers are the ***physical layer*** and the ***data link layer***.

The **physical layer** deals with *how to portray bits*. It works on a bit-per-bit level. It describes the interaction of a single device with a medium.

The **data link layer** deals with *how to portray data*. It works on the level of chunks of bits. It describes the interaction of two or more devices with a shared medium; it adds functionality required for the reception of the data by a second device. It is dealt with in software.

In fact, all layers other than the physical layer deal with data in software. A third layer, the network layer, would encompass functionalities required to communicate between multiple devices. As for Delfi-n3Xt there is only 'itself' and 'the ground station' no routing protocols are required. As mentioned above, possible forwarding of received data by radio amateurs is handled by DUDe, the program used to receive and forward data.

Different layer one functionalities relevant to Delfi-n3Xt radios are shown in Figure 7.2 below.

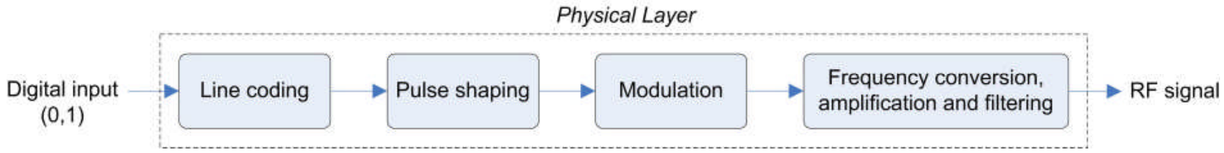


Figure 7.2: Physical layer (layer one) functionalities

The following processes are demonstrated:

- ***Line coding***: Line codes are a way of portraying bits, by two or three power levels for example. In fact, a bit sequence in a computer is already portrayed by a specific line code, which might be changed for beneficial modulation or demodulation properties, such as reduced bandwidth, improved carrier or symbol synchronization, polarity independence and even simple error detection. Examples of line codes are Manchester and NRZ-I.
- ***Pulse shaping***: Line codes include abrupt changes at edges, which gives rise to large occupied bandwidths. As such pulse shaping, also more vaguely called *filtering*, can be applied before modulation to smoothen the signal changes. Effects depend on the modulation scheme selected. Examples of pulse-shaping filters are the raised cosine and Gaussian filters.
- ***Modulation***: After line coding and filtering the signal is modulated, the information signal is mapped onto a waveform; a carrier signal of a certain frequency. This frequency does not have to be the ultimate transmission frequency, generally it is not. The modulator then modulates the bit stream onto an intermediate frequency (IF) carrier, which has at least the frequency of the bit stream. This IF frequency is chosen in relation to the modulation scheme and electronics. Examples of (digital) modulation are BPSK, FSK, and MSK.
- ***Frequency conversion, amplification and filtering***: Through a process often realized with multiple steps, the IF frequency is up-converted to transmission frequency or radio frequency (RF). This is called the *superheterodyne principle*. After frequency transformation resulting frequency by-products should be filtered out and the signal should be amplified to sufficient strength for transmission. Afterwards again unwanted amplification by-products should be filtered out. Multiple amplification or filtering steps can take place. In this stage, if proper filters are used, the information content of the signal is not changed; only the signal frequency is changed and strength is increased.

As layer two functionalities take place in software, they effectively change the bit sequence. As such in the transmission process these take place on the bits, before line coding. As such Figure 7.2 above can be extended to Figure 7.3.

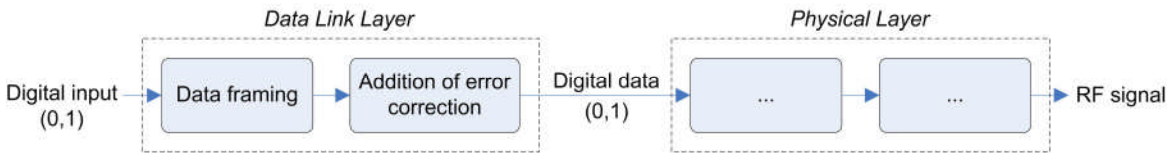


Figure 7.3: Data Link layer (layer two) functionalities in relation to those of the physical layer

Applicable layer two functionalities in case of Delfi-n3Xt are data framing, the addition of forward error correction (FEC) and possibly the application of data security measures such as encryption and error detection. Data framing is usually specified by means of the *data protocol*.

This chapter aims to describe techniques involved with preparing data for transmission, which leads to be able to select techniques for the transmission links on the VHF, UHF and S-bands in the following chapters. In fact, these sections are structured to serve as theoretical foundation for a selection for any similar satellite radio. The information presented focuses on the important conclusions to serve as an overview of these techniques, without loss of the general perspective.

Frequency conversion, bandpass filtering and amplifying move into the domain of electronic design. The intermediate frequencies, filter specifics and amplifiers need to be designed and chosen with proper bottom-level knowledge. These aspects are consequently not discussed in detail in this top-level document, but other documents serve this purpose, such [SLR 0281] in case of the PTRX. The superheterodyne principle will be explained however, to properly explain the architecture of the transmitter and receiver, and to give a complete review of the transmitter or receiver physical layer. The selection of other physical layer functionalities can be done without thorough electrical knowledge.

Looking at Figure 7.3 this chapter deals with the layers described, but more or less in reverse. This is for reasons of clarity, as certain functionalities depend on and support others. Firstly, in section 7.1, the transmission medium and its effects are introduced.

The heart of the physical layer is modulation, which therefore is explained next in section 7.2. Line coding and pulse shaping supplement modulation schemes and are therefore discussed in sections 7.3 and 7.4 respectively. The superheterodyne principle, explaining the use of frequency converters and filters is introduced in section 7.5.

Afterwards data link layer functionalities are discussed, commenced by an introduction to data frames (data protocol) in section 7.6 and followed by a discussion on data security measures, dealing with ensuring data integrity, in section 7.7. Error correction, not just involved with ensuring data integrity but also allowing reducing (modulation) power requirements is discussed in section 7.8.

Finally, theory on maximizing downloadable data volume by changing or setting the data rate accordingly is presented in section 7.9.

7.1 The transmission medium

Before different transmission techniques are introduced, a general medium and its *information-changing* effects are first introduced. Transmission techniques should take into account these effects, and how to overcome them. *Amplifying* is applied to overcome the effects of signal *attenuation*, which is not shown as it does not change the information content of the signal. Figure 7.4 below shows a model of data transfer, starting with modulation and ending with demodulation. When the (wireless) channel is passed certain effects are imposed on the signal as is explained below.

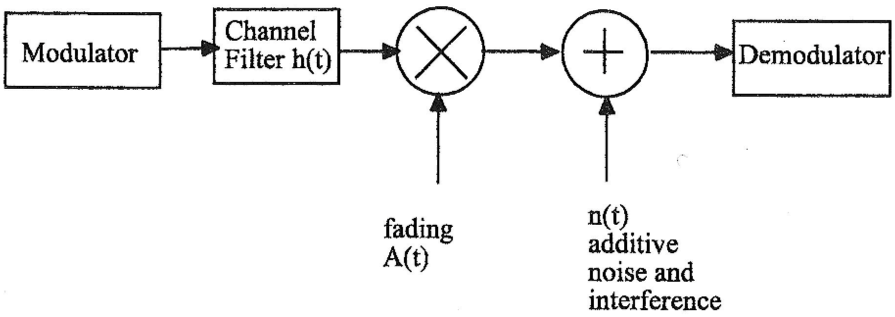


Figure 7.4: Digital communication system model for modulation and demodulation [SLR 0552]

The first block after modulation shows the *Channel Filter*, which represents the transmitter filter, the actual channel filter and the receiver filter. Due to hardware filters implemented in both the transmitter and receiver the actual bandwidth that is produced by the modulator is limited, in order to meet bandwidth restrictions and to cut off parasitic frequencies. The actual channel can also have certain bandwidth limiting effects, for example the atmosphere filtering out certain EM waves. Proper carrier frequency selection and sufficiently narrow signal bandwidth takes care of this effect. Finally the receiver includes a filter to limit the received bandwidth, in order to cut off noise and interference outside of the used bandwidth.

The second block after modulation indicates *fading effects*, which refer to multiple versions of the transmitted signal being received at slightly different times (multipath effects). This happens due to reflections in rural areas for example. This effect is minimized and often not important in space communication.

Finally the third block after modulation indicates the *addition of noise and interference* within the used bandwidth. Interference can follow from interfering signals within the frequency bandwidth. Noise is the usually the most important channel characteristic. The channel adds noise to the signal, on all frequencies. The noise channel is usually modeled by an additive white Gaussian noise (AWGN) channel, implying signals experience no amplitude loss or phase distortion, but white Gaussian noise is added to the signal. The amplitude frequency response is flat in this model, which means that equal noise is added at all frequencies. This assumption is also used in results shown in this as well as the nest section. *Space can almost exactly be modeled by an AWGN channel*.

The demodulator must receive a signal of sufficient quality, taking into account all mentioned effects, as well as decreasing signal strength as distance (including circuitry) is traversed. The quantity used to describe this quality is the ratio $\boldsymbol{E_b/N_0}$, which is the *bit energy to noise spectral density ratio*. When talking about digital signal transmissions, the quantity is inversely related to the *bit error rate* (***BER***); per modulation scheme a minimum E_b/N_0 is required to achieve a certain maximum *BER*. The *BER* indicates the chance that a transmitted bit is received incorrectly. A usual *BER* accepted for data transmission is 10^{-5}, which means that 1 bit per 100,000 bits is corrupted. This *BER* value is also used for Delfi-n3Xt system design.

7.2 Digital modulation

In order to transmit data over a wireless channel, modulation is applied. In case of Delfi-n3Xt all data is digitized, therefore requiring digital modulation techniques. Usual techniques are BPSK, BFSK, MSK, OOK or QAM for example. This section gives an introduction to digital modulation, in order to yield enough background information to argue for modulation techniques to be applied on Delfi-n3Xt. An important source of provided information is [SLR 0552].

Digital modulation is a process that impresses a digital symbol onto a signal suitable for transmission. In wireless transmission, the applied modulation technique is called bandpass modulation, or carrier modulation. A sequence of digital symbols is used to alter the parameters of a high-frequency sinusoidal signal, also called carrier. A sinusoidal signal has three parameters, being amplitude, frequency and phase. Therefore the three basic carrier modulation methods are:

- **Amplitude Modulation (AM)**
- **Frequency Modulation (FM)**
- **Phase Modulation (PM)**

When talking about binary carrier modulation these modulation methods are more specifically called:

- Amplitude shift keying (ASK)
- Frequency shift keying (FSK)
- Phase shift keying (PSK)

Using these three types of modulation methods a variety of schemes can be derived. The inverse of modulation, the process of extracting information from the received signal, is logically called demodulation.

The three basic types of modulation are illustrated in Figure 7.5 below. (a) and (b) are line codes, or simply methods of portraying bits in digital systems. OOK is a version of ASK, where one of the two used amplitudes is equal to zero. BPSK shifts the phase of the carrier signal at phase boundaries, and FSK shifts the frequency to portray a different bit. Finally double side band, suppressed carrier, or DSB-SC (with pulse shaping of the baseband digital system) is shown, which is a type of AM. Since BPSK suffers from energy percentage efficiency as will be explained below, shaping the bits beforehand can drastically improve this efficiency. This modification of BPSK is called DSB-SC (with shaping), or Raised Cosine BPSK. Pulse shaping is introduced in section 7.4.

One can also notice in Figure 7.5, a 1 b/s signal (1 Hz bit stream) does not require a 1 Hz carrier. In fact, the carrier should *at least* be the bit frequency, but can also be much more. In practice, the modulator modulates onto a certain 'temporary' carrier defined by modulation advantages and electronics; the intermediate frequency or IF, as introduced above.

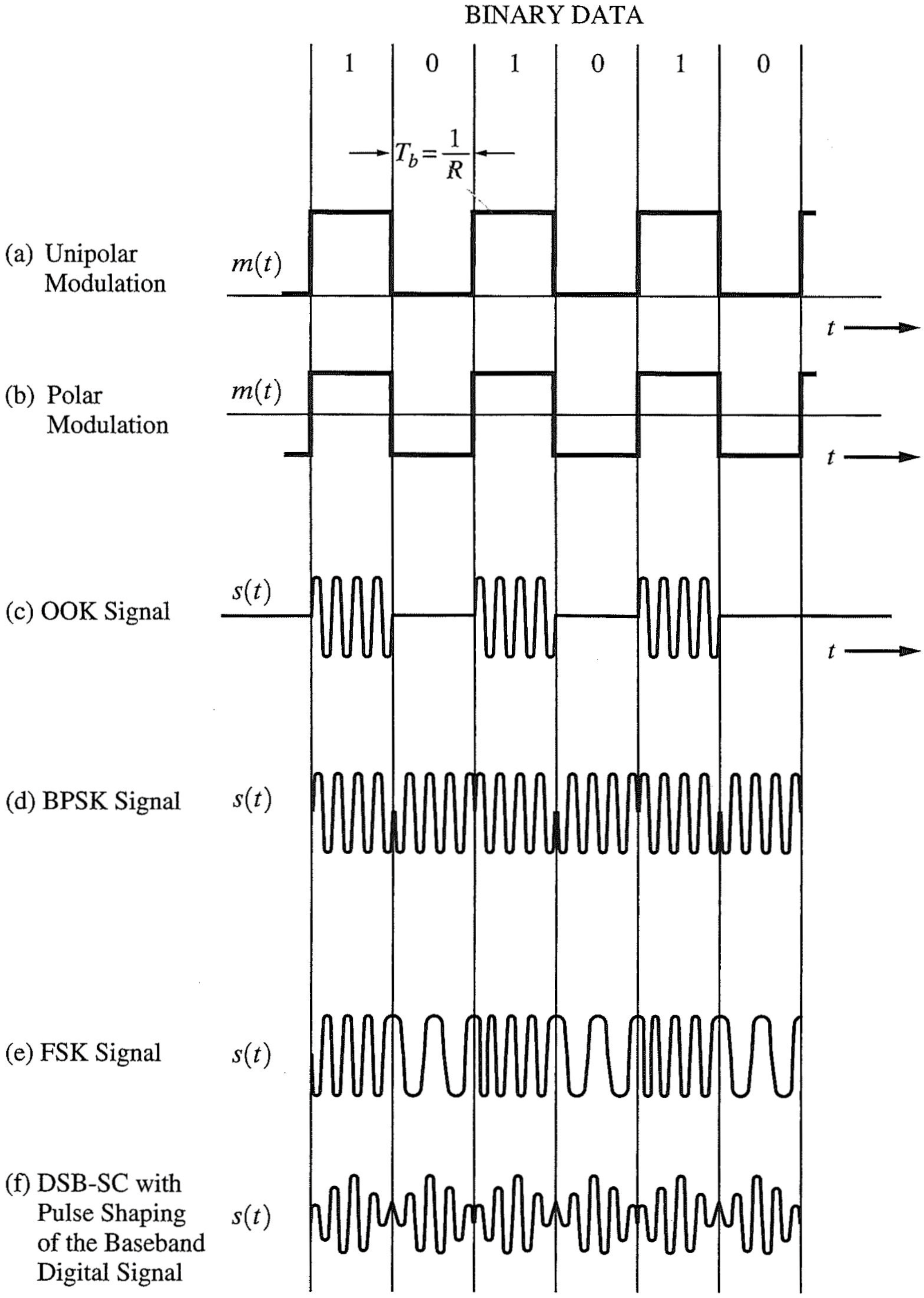


Figure 7.5: Bandpass digitally modulated signals [SLR 0443]

Modulation selection criteria

Three criteria are relevant for the selection of a modulation method. These criteria are:

- **Power efficiency,**
- **Bandwidth efficiency,**
- **System complexity.**

The power efficiency and the resulting power requirements are elaborated on above; a certain E_b/N_0 is required per modulation scheme to obtain a certain *BER*. In other words, some schemes require a relatively low signal-to-noise ratio, being power-efficient. Power efficiency is also referred to as error performance, or *BER* performance.

The bandwidth efficiency is defined as the number of bits per second that can be transmitted in one Hertz of system bandwidth. The necessity for this criterion can be visualized by remembering that a finite signal in time space will yield an infinite signal in frequency space, as thought by basic courses on signals and Fourier transforms [SLR 0443]. The power distribution in frequency space will usually also be taken into account in determining bandwidth efficiency, for example using the *Nyquist efficiency*. This efficiency takes into account minimum Nyquist filtering, also called *raised cosine filtering*, as well as a noise-free modulator, demodulator and transmission channel, and gives the required bandwidth to have symbol demodulation without intersymbol interference (ISI). This efficiency can be reached (theoretically) using pulse shaping, as introduced in section 7.3.

A more obvious and clear type of bandwidth (efficiency) indicator is the *null-to-null bandwidth*. It indicates the bandwidth of the main lobe, the lobe between the two first points of zero spectral energy; this is clearly explained by looking at a figure such as Figure A-4 below. As the latter efficiency gives no definite indication of the energy contained within this lobe, and not all frequency spectra contain nulls, the *energy percentage bandwidth* is the most unambiguous. It indicates the bandwidth within which a certain percentage of all signal energy is focused. These definitions will become clearer when reading this subchapter. By using near-optimal (pulse shaping) filters, the optimum bandwidth efficiency can be approached, which for some modulation schemes is better than for others.

Finally system complexity refers to the amount of circuitry involved and the technical difficulty of the system. The system complexity is logically related to development time and cost. Some modulation schemes require (significantly) more circuit complexity than others.

Three aspects are related to system complexity. Demodulators can require the following; listed in increasing order of impact on system complexity:

- ***Symbol synchronization***
  For the symbols (representing one or more bits) to be extracted, the demodulator should know the symbol start and duration, in other words symbol phase and frequency.
- ***Carrier frequency synchronization***
  This implies that the frequency of the reference signal in the demodulator should be exactly to that used to generate the received signal.
- ***Carrier phase synchronization***
  This implies that not only the frequency of the reference signal should be equal to reference signal of the modulator, but the phase as well.

Symbol synchronization is required to extract data; in other words in all demodulators. Modulation schemes do not automatically yield enough information for symbol synchronization to be guaranteed, but it can be achieved by guaranteeing sufficient transitions in the received signal, for the demodulator to 'lock' onto the signal. This can be taken care off by applying a specific line code (such as Manchester coding) or by involving higher level functionality; a specific sequence of symbols can allow the demodulator to get a lock on the signal through the use of a synchronizer circuit. The AX.25 protocol provides exactly this through the use of flags (see section 7.6.2). Also, the protocol assures sufficient symbol transitions during communications through the use of bit stuffing.

Carrier frequency synchronization requires knowing the exact carrier frequency. Reference signals can be extracted from the received carrier, or a separate oscillator can be used. Nevertheless, if Doppler shift takes place in the carrier frequency, carrier frequency tracking is required and complexity is increased; its extent depending on the modulation scheme. To be clear, for proper communication of course the transmission frequency should be known; electronics will be designed for this frequency and proper filters are in place. However, certain demodulators can get away with having just a sufficiently-wide-band filter tuned to the transmission frequency with margins for the Doppler shift (requiring no reference frequency). In these cases, carrier frequency synchronization is said not to be necessary.

Carrier phase synchronization simply requires not only knowing the exact carrier frequency, but also its phase. The reference signal should be exactly in phase with the oscillator used in the modulator. If indeed both carrier frequency and phase synchronization are realized, one speaks of *coherent demodulation*. Its opposite, *noncoherent demodulation* involves only carrier frequency synchronization or no form of carrier synchronization. If a modulation scheme is to be demodulated noncoherently, one speaks of a *noncoherent modulation scheme*.

A final criterion, related to both complexity and power efficiency is having a *constant envelope modulation scheme*. This means that the envelope of the transmitted signal, in other words the spectral power at the targeted frequency, does not change. If it does not, more efficient non-linear power amplifiers can be used such as class C and D amplifiers, without dedicated feedback systems. This is because no information is carried by the amplitude for those modulation schemes. ASK schemes per definition have non-constant envelopes. PSK schemes do not, but as they usually require (amplitude varying) pulse shaping to improve bandwidth efficiency, they are not suitable either. As such, only FSK schemes generally provide this feature.

The following Figure 7.6 gives a good overview of the power and bandwidth efficiency of some of the most important digital modulation techniques.

The horizontal axis indicates the required signal E_b/N_0 for successful demodulation for a *BER*, or P_B of 10^{-5}. The vertical axis indicates the perfect theoretical bandwidth efficiency, or Nyquist efficiency, given by a ratio of bit rate (*R*) over bandwidth (*W*). It should thus be realized, that even without changing modulation scheme, a doubled data rate will lead to a doubled required bandwidth.

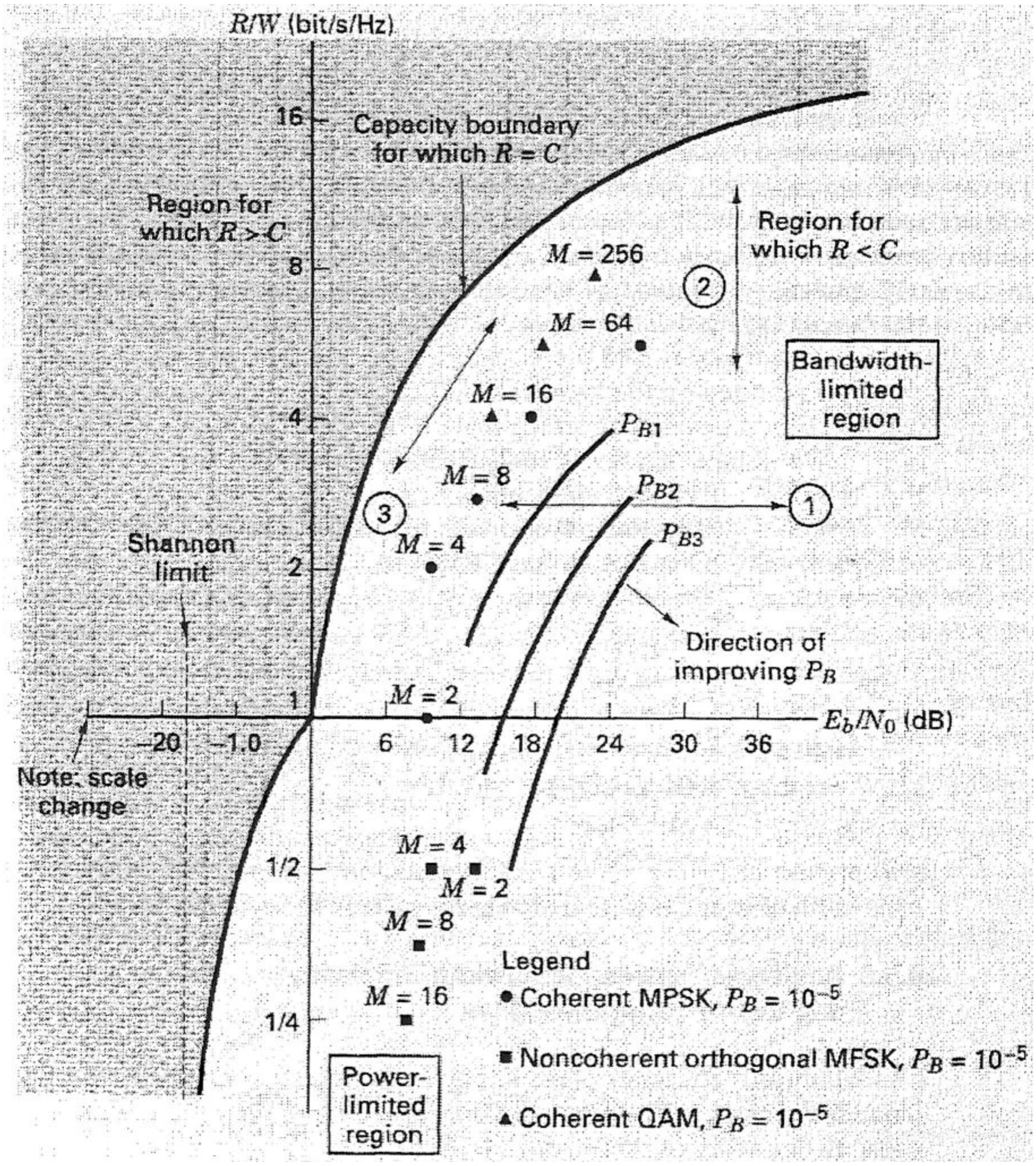


Figure 7.6: Bandwidth and power efficiencies of modulation techniques, $P_b = 10^{-5}$ [SLR 0552]

A lot of information is presented in Figure 7.6 but attention will be placed on only a certain few characteristics. A complete description can be found in [SLR 0552].

It can be seen that MFSK techniques tend to become more power efficient as the amount of applied frequencies *M* increases, whereas MPSK techniques tend to become more bandwidth efficient with increasing amounts of phase values *M*. At a certain point (M-)QAM techniques outperform MPSK techniques, from *M*=16 onwards. MSK is not shown in the figure, but actually it has the same theoretical efficiency as BPSK. In terms of energy percentage bandwidth efficiency however it outperforms BPSK. Noncoherent MFSK as opposed to its coherent version is indicated, even though it theoretical performance is the same as its counterpart.

All modulation schemes shown in the figure above plus some more advanced onces are reviewed in more detail in Appendix A. Given the size of the resulting text the content has been moved to an appendix. Also, exact knowledge of these schemes is not required to follow the reasoning of selecting modulation schemes for Delfi-n3Xt in later chapters.

The appendix presents the exact characteristics of the different modulation schemes and presents in more detail the power performance, bandwidth performance and required complexity. Also, different options of complexity which in turn improve on either power or bandwidth efficiency are introduced. Modulator and demodulator schematics are presented. The level of detail for each of the modulation schemes is dependent on the assumed relevance to Delfi-n3Xt.

For the following, it is useful to introduce the term *Baud*. In a modulated signal one symbol can signify multiple bits; in this case the rate in Baud is only half the bit rate. Indeed, Baud is another word for *symbols per second*. When Manchester coding or FEC coding is applied, as introduced in respectively the next section and section 7.8, more symbols are required per bit. In those cases, the rate in Baud is *increased* with respect to the bit rate.

7.3 Line codes

As introduced in the last section, digital modulation is a matter of mapping digital data onto a signal suitable for transmission. Before that, it can be decided how the data is represented in the first place. Whereas usually '1' or '0' bits within digital logic (computers) are represented by respectively positive signals or signals with zero magnitude (unipolar NRZ-L code as introduced below), there are different conventions possible. In fact, different codes have different properties which can be desirable, also when combined with a variety of modulation methods. As these codes deal with direct transmission without frequency transform, it is labeled *baseband modulation*. Other names are baseband formats, baseband waveforms, or *line codes*.

This section introduces different line codes and their characteristics. This in turn allows proper consideration of line codes for Delfi-n3Xt transmission systems. Unless indicated otherwise all data has been extracted from [SLR 0552].

Figure 7.7 below demonstrates a quite complete list of interesting and basic binary line codes. Four basic classes are represented; the nonreturn-to-zero class (NRZ), the return-to-zero class (RZ), pseudoternary (PT) codes and the biphase codes. More advanced codes are substitution codes and block codes. Aliases can be seen as well in the figure below, such as the *NRZ-I* and *Manchester codes* applied on Delfi-C^3.

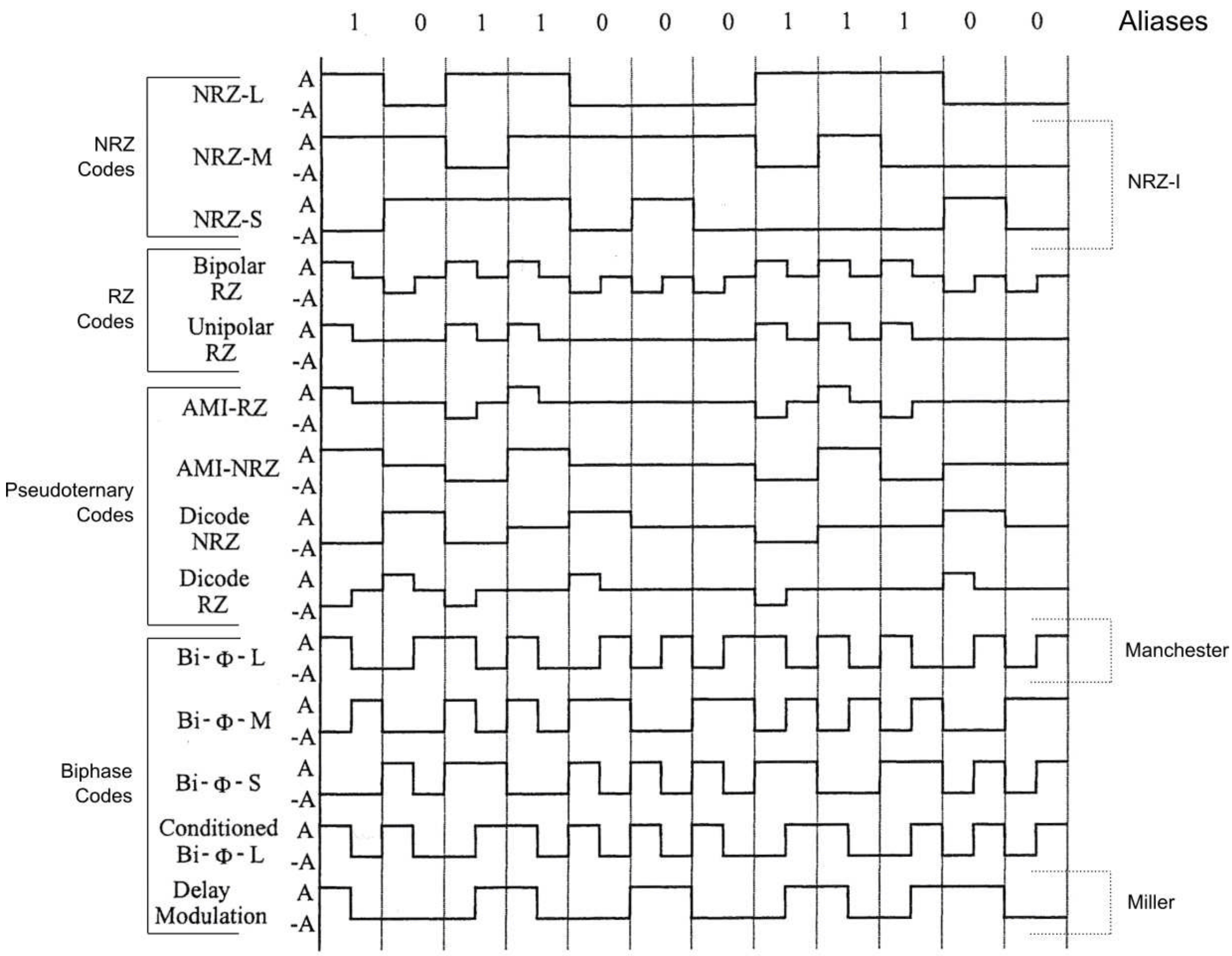


Figure 7.7: Basic line codes [adapted from SLR 0552]

Line code properties

In order to select a proper line code, a number of characteristics are important. The improvement of some properties of course directly degrades performances in other respects. The most important properties are:

- **Low error probability**
  As for digital modulation schemes, the E_b/N_0 required to achieve a certain *BER* can be established.
- **Adequate timing information**
  Bit or symbol timing is usually recovered from the received data sequence. This requires an adequate transition density in the codes sequence, provided by the line code. A high transition density facilitates timing recovery. Again, this is similar as for modulation schemes.
- **A channel-suitable spectrum**
  For example:
  - *No dc-component and small near-dc components in the Power Spectral Density (PSD)*: ac-coupled channels or systems using transformer coupling (which have very poor low frequency responses) require this characteristic. A long series of 0's or 1's, depending on its convention, can otherwise charge capacitors to give incorrect results. Also the knowledge of having zero dc bias can simplify the detection of a Doppler shift.
  - *Narrow bandwidth*: the bandwidth of a line code is logically preferred to be as narrow as possible, similar as for a modulation scheme. It can be reduced by filtering or multilevel transmission schemes, however at the cost of an increase in *BER*, due to an increase of ISI (inter-symbol interference) and a reduced signal-to-noise ratio.
- **Bit sequence independence**
  It is preferable that a line code is independent of the bit stream, in other words offers good similar performance even in extreme cases of continuously switching bits or constant 0's or 1's. This is often related to the ability to deliver timing information, as well as a resulting dc-component.
- **Differential coding**
  Differentially encoded schemes are independent of polarity inversion, as is further explained below. This however comes at the cost of a small increase in *BER*.
- **Error detection capability**
  Some schemes allow for error detection, by adding dependence on a previous bit occurrence in the way a bit is portrayed. Error correction however is not possible without introducing extra symbols.

More advanced line codes possess desirable qualities, such as sufficient transitions required for symbol synchronization, no or small dc component(s), the ability to detect errors and in case of differential codes independence of received signal polarity. These qualities come at the cost of reduced error performance and/or increased bandwidth.

It is important to realize, that generally (the effects of) line codes and modulation characteristics are *superimposed*. As can be seen, a large number of properties can be seen to be similar for both techniques. If a line code causes a lower *BER* performance to be achieved, the *BER* performance of a signal that is encoded with that line code and consequently modulated also has degraded *BER* performance. The same holds for bandwidth performance.

A concrete example is Manchester coding; it doubles the symbol rate to allow for better clock recovery. The effect of this is *that bandwidth performance is reduced*. For Delfi-C^3 this caused the uplink symbol rate to be 1200 Baud, whereas the bit rate was only 600 b/s.

Similar to modulation techniques, each line code itself yields a certain required E_b/N_0 for a required *BER* or P_b. For all classes of important codes the following Figure 7.8 demonstrates the error performances.

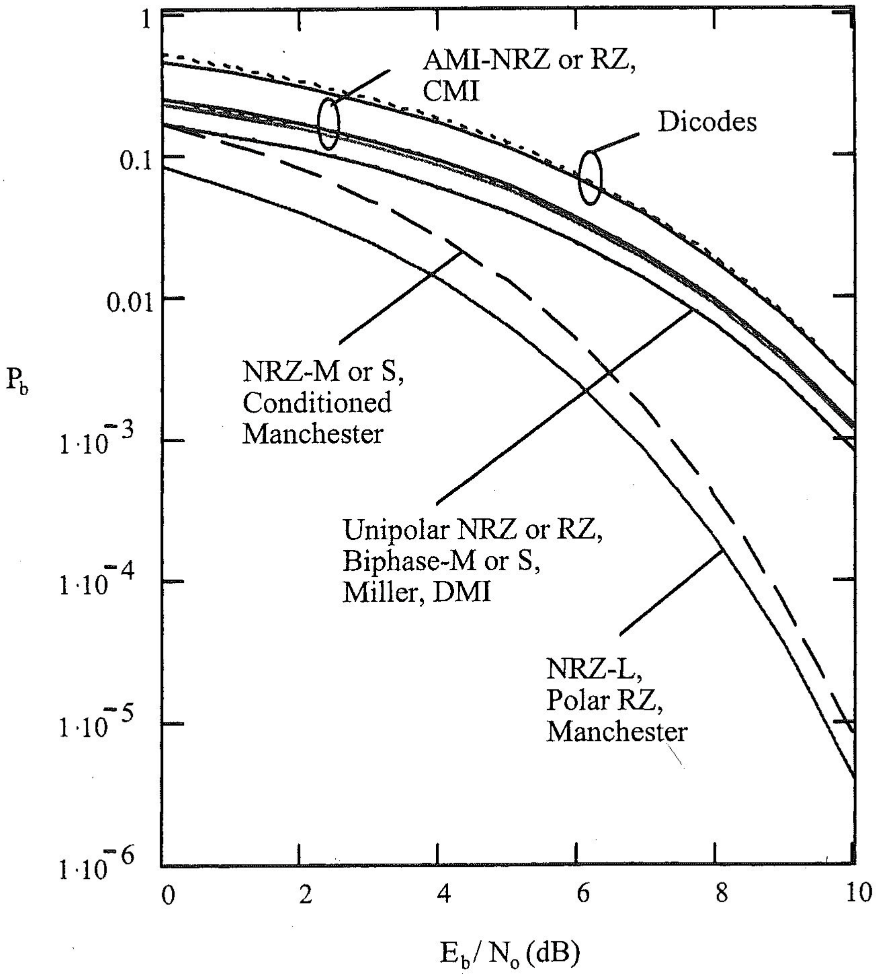


Figure 7.8: *BER* performances of line codes [SLR 0552]

Again, in the case of Delfi-n3Xt the values belonging to a P_b of 10^{-5} are relevant. Even though the figure is not centered on this region, a number of things can be concluded.

First of all, even without digital modulation on a signal ready for transmission a code already requires a 9.6 dB E_b/N_0! This value is the exact same as that for a number of standard and simple modulation schemes such as BPSK, QPSK, MSK and 4FSK. In fact, these modulation schemes implicitly take a power efficient code such as (unipolar or bipolar) NRZ-L and replace the level changes by phase or frequency shifts. However, this implies that these standard modulation schemes actually come *at no cost in terms of BER with respect to the line codes itself*. Some modulation schemes such as FSK (for values of *M* higher than 4), actually improve the *BER* performance. They spread the bit sequence to a number of different frequencies, thereby effectively decreasing the data rate on that frequency and with that the required signal-to-noise ratio for proper demodulation.

Second of all, it can be seen that there are 6 line codes in Figure 7.7 that have the best *BER* performance; indeed it is for that reason that they are generally the most useful for satellite missions. Moreover, while not properly shown in the graph, all other schemes yield an E_b/N_0 of at least 3 dB more than a simple NRZ-L coded bit stream, being 12.6 dB for Unipolar NRZ or RZ for example.

The four classes of basic line codes and some more advanced line codes are introduced in Appendix B. As the text is sizeable and its function in some cases is limited to giving proper background information to an interested reader the text has been moved to an appendix.

Attention is paid to the advantageous and disadvantageous properties of the different line codes. The level of detail is limited to an introduction. Line codes of most importance to Delfi-n3Xt are the (differential) NRZ codes and the Manchester code.

7.4 Pulse shaping

Any fast transition in a signal, whether in amplitude, phase or frequency, will require a large occupied bandwidth for correction transmission and reception. Techniques that help smoothen these transitions reduce the amount of required bandwidth. The idea of *pulse shaping* is to shape the bit shape, per definition one with abrupt transitions, in order to smoothen its edges.

All line codes as introduced in the previous section are based on a rectangular pulse shape. When a rectangular shape is mixed with a sinusoid, effectively, a *convolution* takes place. Remember that this basically takes place in any type of modulation; amplitude, phase or frequency shift keying. The only difference is that in amplitude modulation two consequent rectangular pulses take on different amplitudes, in phase modulation consequent signals have equal but opposite amplitudes, and in frequency modulation the modulated frequencies for consequent signals are different. The resulting frequency pattern is called the *Fourier transform*. [SLR 0694] describes basic signal theory, and shows the spectral characteristics of a rectangular shape; see Figure 7.9 below.

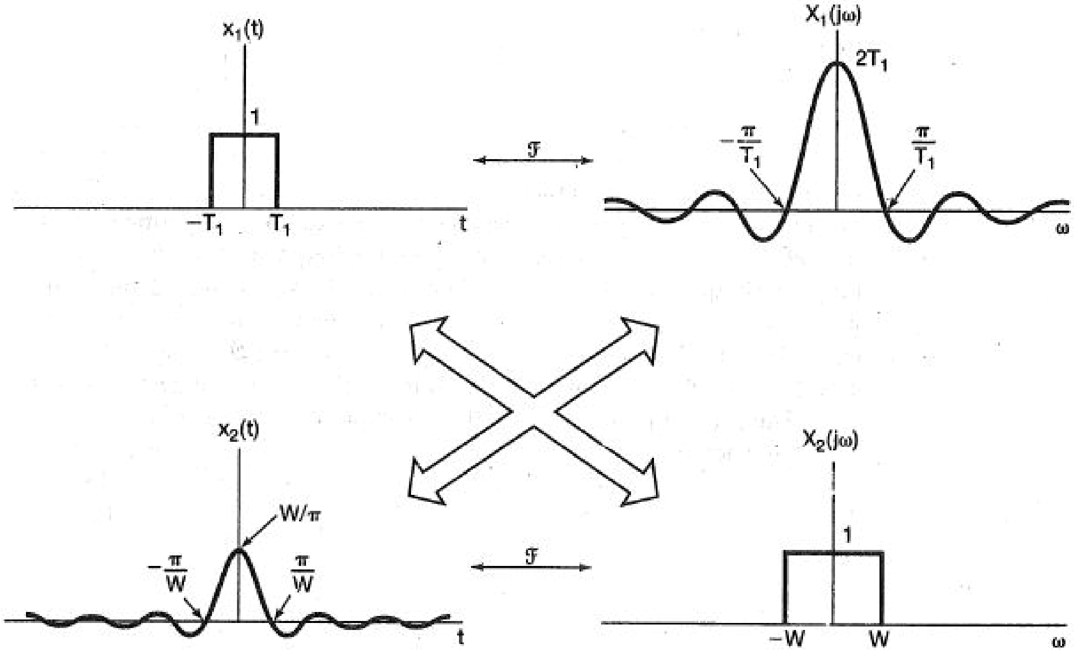


Figure 7.9: Fourier transform pairs involving a rectangular pulse [SLR 0694]

As Figure 7.9 (top) shows, a rectangular shape in the time domain, yields an infinite shape in the frequency domain; a $\sin(\pi/x)/\ \pi x$ or *sinc* shape. In fact, this frequency response is exactly the PSD of a BPSK signal, as shown in Figure A-16; a rectangular pulse of 1 second gives T_1 =0.5, whereas π equals 0.5f, and finally in case of BPSK the rectangular input has an apparent amplitude of '2', as a shift from -1 to +1 or vice versa has taken place. Replacing and entering these values as well as replacing ω on the axis by fT, and realizing that 'negative' frequencies are in fact frequencies with a flipped phase, gives exactly Figure A-16. Indeed, it is the frequency characteristics that are inherent to the rectangular pulse that determine the (bad) frequency characteristics of a BPSK signal. Therefore *improving these characteristics allows improving the bandwidth efficiency*.

The filter implicitly applied in case of a rectangular pulse shape is called a *Boxcar filter*. Figure 7.9 (bottom) shows its inverse, the *Sinc filter*. It can be seen that the sinc pulse very closely resembles the Fourier transform of the rectangular pulse, and vice versa. A difference in amplitude of 2π exists, which in the present discussion is unimportant. If an infinitively broad sinc pulse would be used to indicate a bit, an optimum limited bandwidth would result; one that equals *exactly the minimum Nyquist bandwidth*.

Of course, it is impractical or even impossible to apply an infinite pulse. As such, an optimum should be found. The characteristics offered by the sinc filter are approached by the *raised cosine filter*. In fact, it can be said that the sinc filter is a special type of the raised cosine filter. This filter is introduced below in subsection 7.4.1. A large number of other filters exist, such as Butterworth filters, Chebyshev filters or Bessel filters. These filters are however not generally used as pulse shaping filters in modulation schemes, as they are outperformed by the Gaussian filter. The latter is introduced in subsection 7.4.3 below.

7.4.1 Raised cosine filter

The raised cosine filter is designed to approach the optimum sinc filter, as this sinc filter is in fact impossible to implement. The sinc filter in fact has another important property, which is that of *no intersymbol interference*, or ISI. What this means is illustrated by Figure 7.10 below.

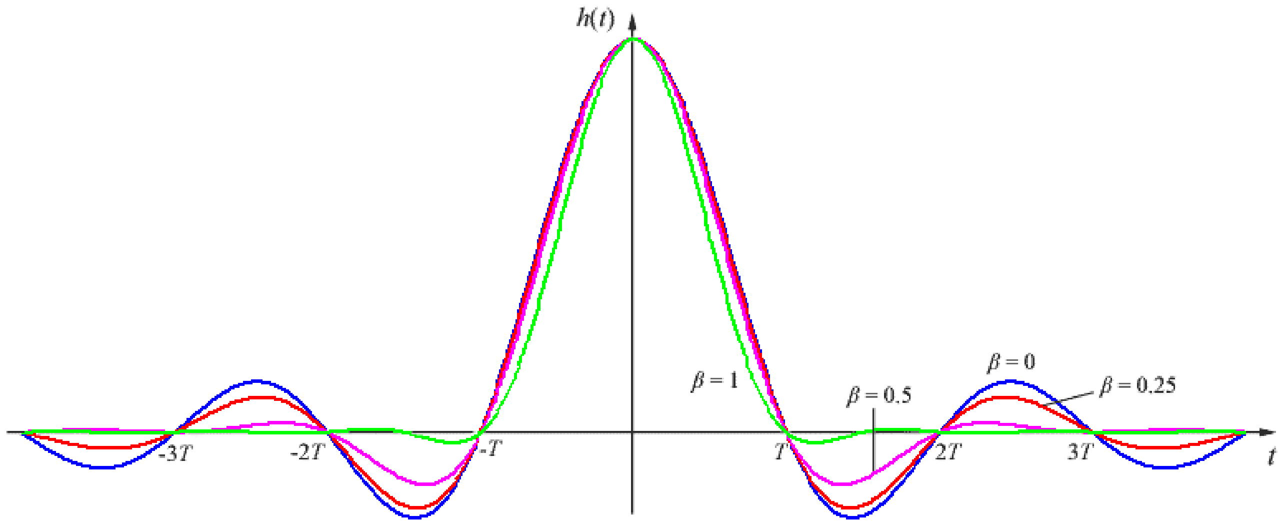


Figure 7.10: Time domain response of a raised cosine filtered pulse for different values of β

Figure 7.10 shows the time domain pulse strength of a raised cosine filtered bit. It can be seen that at every integer multiple of T except at $0*T$, the total power is zero. This means, that at the point where a next bit has its maximum power, on the *bit decision point*, the residual power of another bit is zero. A demodulator, basing its bit value on this point, will therefore not be influenced. This is called *zero ISI*.

In the same figure it can also be seen that multiple graphs are plotted, with different values of β (depending on the source, both α and β are used). A β of zero gives the optimum spectrum. Larger β's can more easily be implemented in reality, as their power for larger values of T quickly drops off and at some point can be ignored. Also, less total amplification power is then necessary (explained below) and the effects of a demodulator basing its bit decision on not the exact middle of a bit are less severe. But of course, the resulting bandwidth reduction is less. The resulting frequency spectrum can be seen in Figure 7.11 below.

The time domain shape, shown above, can be given by the following equation [SLR 0697]:

$$p(t) = \frac{\left(\operatorname{sinc}\frac{t}{T}\right)\left(\cos\frac{\beta\pi t}{T}\right)}{1-\left(\frac{2\beta t}{T}\right)^2}, \tag{6-1}$$

As a side note, if the above relation is used for calculations, care should be taken to avoid $\beta t/T=\pm 1/2$, as this will make the denominator go to zero. The limit of p(t) at $\beta t/T=\pm 1/2$ can be shown to be given by $(\pi/4)$ sinc (t/T); not infinity.

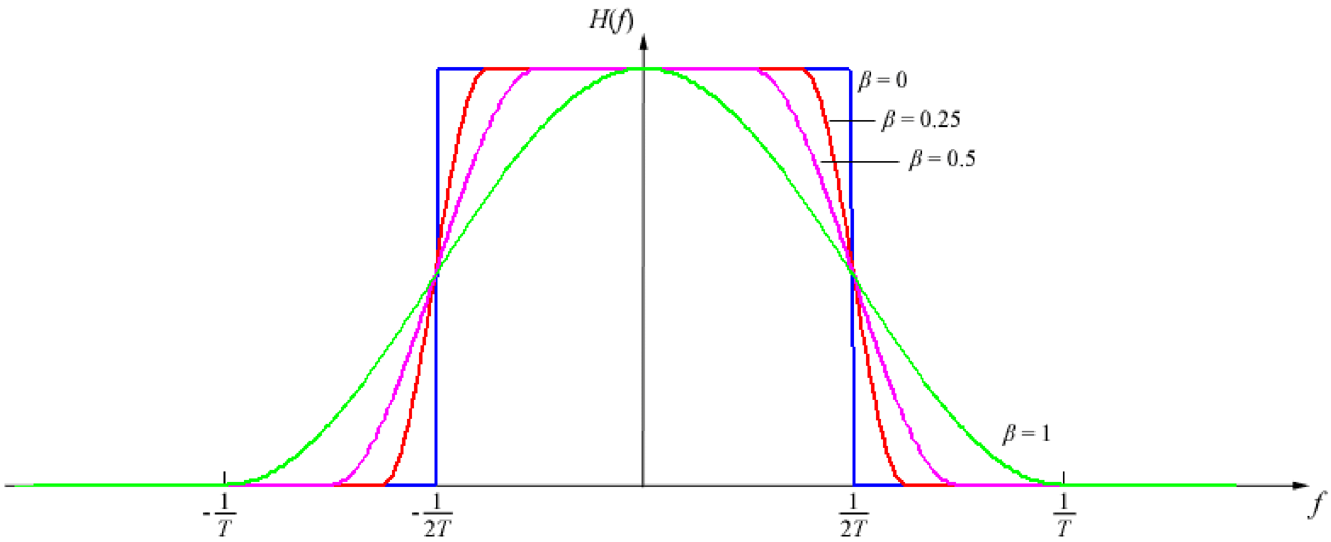


Figure 7.11: Frequency domain response of a raised cosine filtered pulse for different values of β

As Figure 7.11 illustrates, the bandwidth required to transmit a raised cosine filtered symbol stream of *X* Hz is between 1 and 2 times *X*. In fact, the relation between occupied bandwidth, symbol rate and β is:

$$\textit{Occupied bandwidth} = \textit{symbol rate} * (1 + \beta) \tag{6-2}$$

It can also be seen that the raised cosine frequency spectrum belonging to $\beta = 1$, actually is a raised cosine; indeed it is from the frequency spectrum response that the filter gets its name. The equations of frequency domain graph will not be presented but can be found in [SLR 0697] and [SLR 0696]. The spectrum for a rectangular pulse in fact is almost the same to the case where β is 1, being slightly larger in power content around $1/2T$. After $1/T$ however spectral regrowth occurs.

In Figure 7.10 it can be seen that, to avoid large bandwidth occupation, significant power outside of the actual bit duration must be transmitted; this is due to the *overshoot* necessary (the power cannot go to zero and stay zero). In fact, this overshoot causes more power to be necessary than the symbol itself requires. If this power would be filtered or limited, ISI will still be introduced. If the total power would be limited, the resulting signal-to-noise ratio would be reduced, whereas a minimum is required for correct demodulation. The effect of this can be sizeable; for example QPSK with a β of 0.2 requires some *5 dB more power* [SLR 0447].

In case of *constant envelope modulation schemes* (M-FSK and MSK), the situation is different; no extra power is required to implement the changes as the signal total strength is not dependant on the input, but the main transmission frequency is. As such, in that case, the signal-to-noise ratio at the bit decision points would remain identical independent of the pulse shaping scheme as long as no ISI is introduced. It should be said however, that constant envelope schemes are per definition less bandwidth efficient and therefore, even with a raised cosine filter, would require more bandwidth (as two or more transmission frequencies are used). Therefore, raised cosine filtering is not often applied for FSK schemes.

Filters can be implemented in an analogue and digital fashion. Digital filters offer serious advantages over analogue designs, as they can be integrated directly on silicon, thus in on-chip designs. Also, component drift due to temperature and aging is eliminated [SLR 0697]. Two types of digital filters exist; finite impulse response (FIR) and infinite impulse response (IIR). The fundamental difference between the two is the inclusion of feedback in the latter. Hence, the future of the output is dependent on the input history. Even long after the input signal is terminated, the IIR will continue producing a signal. As this is generally not desired, FIRs are used. These can then be used to approach the actual raised cosine spectrum, depending on the sampling rate of the FIR (as a multiple of the symbol rate) and the amount of symbol periods over which an input symbol is spread (see [SLR 0696]). The latter amount depends on the application, and a small delay is introduced at the start of signal transmission because of it.

To give a final example of the effects of raised cosine filtering, see Figure 7.5. The final demonstrated modulation scheme, DSB-SC or just BPSK, with applied raised cosine filtering is demonstrated. The effect is that at the transitions the signal amplitude is much reduced. As such, in can be understood that the bandwidth effects of the sudden change are reduced, as the power content of the sudden change is also reduced.

The downside of the raised cosine filter is that its implementation through the use of a FIR filter quickly becomes very demanding, as higher data rates are applied. The sampling rate required should be multiplied by the amount of bit periods over which the symbol is spread, and each additional bit period over which the signal is spread increases the required possible states exponentially. Therefore, less demanding Gaussian filters are usually applied for higher bit rates. Also, whereas realistic raised cosine filters will rarely decrease the bandwidth below 1.5 times the signal frequency, Gaussian filters can reduce this bandwidth to even less than 1 times the signal frequency.

Finally, Delfi-C^3 actually applied a simpler implementation of the raised cosine filters, which achieved lower bandwidth reductions than regular raised cosine filters, however coming at the cost of very little complexity. This scheme is introduced in subsection 7.4.2 below.

Root raised cosine

As a receiver requires the use of filters to limit the received spectrum, in some cases a root raised cosine filter is used. The idea is to integrate a root raised cosine filter shape at both the transmitter and receiver, so that the resulting total filter is the zero-ISI raised cosine filter. Its integration in the transmission link is shown in Figure 7.12 below.

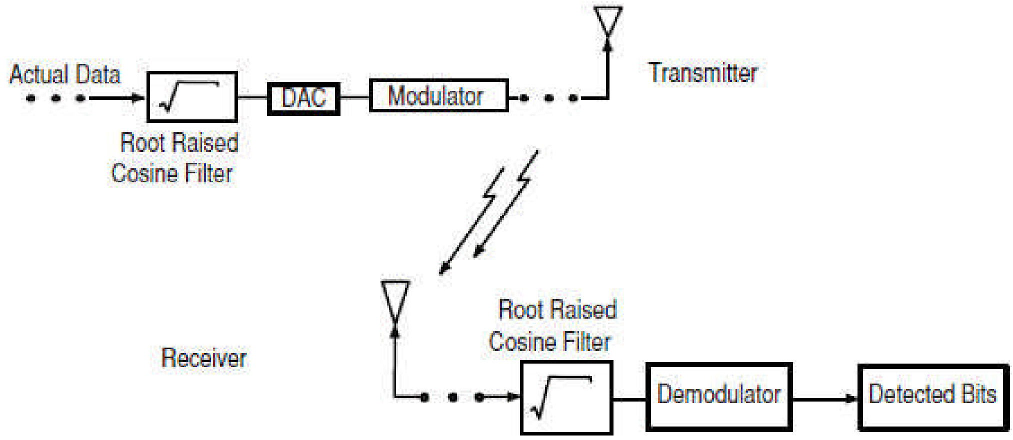


Figure 7.12: The implementation of a root raised cosine scheme [SLR 0447]

It can be seen that the data is first digitally filtered, after which a DAC (digital-to-analogue converter) is used to produce a useable signal to the modulator. After transmission the receiver filters the signal *in an analogue fashion*, after which demodulation yields the transmitted bits. The difficulty in this set-up is the inclusion of this analogue filter with the exact desired properties, and the matching of both the transmitter and receiver filter to actually create one full raised cosine. If multiple manufacturers are involved, with transmitters and receivers of different types and sizes, this is quickly complicated. In any case a trade-off should be made, as more bandwidth is required for transmission when only a root raised cosine is applied, but at the gain of more filtering of broadband noise and other transmissions in nearby channels. In case of space flight and pointed antennas however the latter aspect is luckily limited.

7.4.2 Raised cosine in Delfi-C^3

Although not represented by found literature, in case of Delfi-C^3 a variant of raised cosine filtering was actually applied. It likely has to do with more generally applied radio amateur approaches. In any case the approach has been to use the raised cosine shape, and not a sinc shape, in the *time domain* to represent a bit. This is illustrated by Figure 7.13 below. This means that no infinite pattern in the time domain is approached, thereby creating a clear infinite pattern in the frequency domain. In other words, less bandwidth efficiency is obtained.

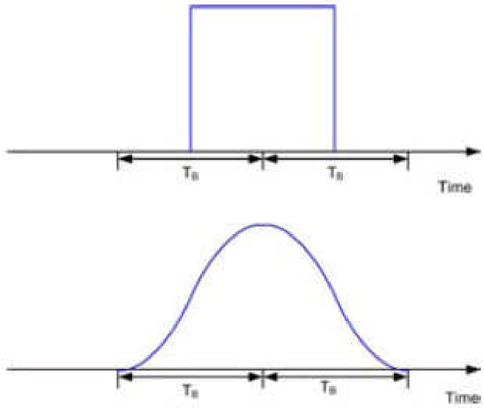


Figure 7.13: Time domain raised cosine filtering [SLR 0248]

The benefit of the approach however is that required power is optimal, as no more power is required for the transmission of one bit; no overshoot occurs. The integral area underneath both shapes is equal (in fact, a series of pulses even has a constant envelope when summed together, just as a series of rectangular pulses would have). The power spreading is maximized, as the pulse is spread up until the points of zero ISI. Also, complexity is limited as one bit is only spread out over two bit periods. Therefore, this method provides a low-complexity solution with significant reduction of the required bandwidth, at no power cost.

The actual bandwidth (reduction) effects can be calculated, using the relation 6-1 presented above, as in fact the frequency response of a time domain signal is equal to a time response of a frequency domain signal, adjusted for a value of $2*\pi$ as shown in Figure 7.9 above. Plotting the relation for βs of 0 and 1 gives the graphs in Figure 7.14 below, being in fact equal to those shown in Figure 7.10, but plotted on a logarithmic scale and for larger bandwidth. The frequency / bit rate value indicates the bandwidth required for the transmission of one bit per second and the spectral power in dBc indicates the spectral power with respect to the main carrier frequency.

As was already explained before, the frequency pattern created by a BPSK modulated signal is exactly equal to the Fourier transform of the modulated pulse.

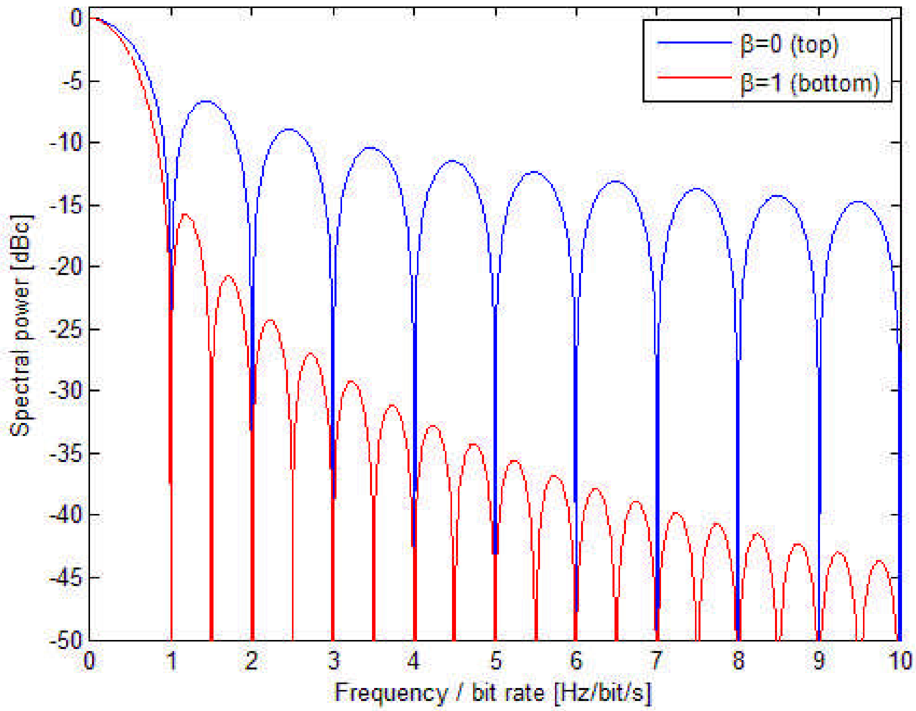


Figure 7.14: Spectral power vs. normalized bandwidth

It can be seen that, as already shown in the previous section, a rectangular pulse has very poor bandwidth characteristics. The time domain raised cosine pulse decreases the required bandwidth significantly as it introduces a clear decreasing trend (Figure 7.14).

In reality, not a single pulse will be transmitted, but a *pulse train*. This pulse train can in turn be assumed to be infinite, for a continuous data stream. As a pulse can be said to indicate a '1' bit, and 'no pulse' can be said to indicate a '0' bit, a pulse train of a certain frequency can be said to simulate a bit sequence of alternating 0's and 1's of double that frequency. In order to be able to analyze Delfi-C^3 bandwidth performance below, bitrates applied on Delfi-C^3 will be used: the case of a 1200 b/s signal. The pulse train will then have a 600 Hz frequency.

The resulting frequency spectrum of this rectangular pulse train is shown in Figure 7.15 below, generated using a signal generator and measured with a spectrum analyzer, both available at the faculty of EEMCS.

The frequency spectrum of an infinite raised cosine pulse train will not be shown, the latter would logically yield a sine function of constant frequency; this is further commented on below. The spectrum generated by a constant sine logically is a single impulse at the frequency of the sine.

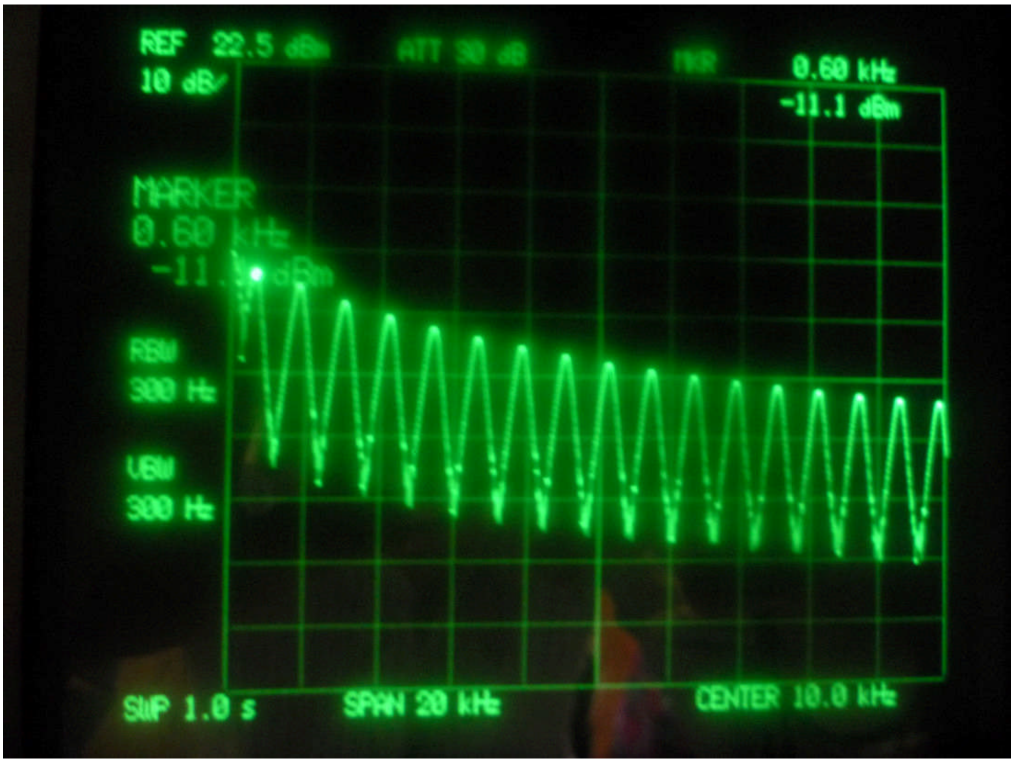


Figure 7.15: Measured frequency spectrum of a 600 Hz rectangular pulse train over a 10 kHz span

In the above figure a span of 20 kHz is portrayed, starting at 0 Hz. The pulse train main frequency is at 600 Hz, or 0.6 kHz. A large peak can be seen at 0 Hz, this one should be ignored as it represents the zero-dc component present in the spectrum analyzer; a by-product of the experimental set-up thus. The pure signal generated by the pulse train then has a peak at 0.6 kHz, and its signal strength actually steadily decreases towards 0 Hz. Signal strength is indicated on a logarithmic scale, with a 10 dB decrease per vertical square.

In the figure it can be seen that a pulse train similarly has bad spectrum characteristics. Compared to the spectrum of a single rectangular pulse, the pulse train frequency spectrum at first decreases less with frequency; only a small ~2 dB decrease from the first to the second peak compared to some -7 dB for the single pulse. However, the spectrum of the single pulse afterwards remains larger; within the 10 kHz span it seems to converge to a -15 dB value. The pulse train frequency spectrum reaches this same value at 10 kHz approximately despite its larger spectrum at low frequencies, but continues to decrease more significantly afterwards.

The behaviour that can be seen here can be explained by Fourier transforms: a finite pulse in the time domain will yield an infinite frequency spectrum, whereas an infinite signal in the time domain yields a frequency spectrum that is finite.

As explained, Delfi-C^3 applied a time-domain raised cosine filter. The time-varying characteristics of the pulse train then depend on the bit sequence. In order to explain the resulting pattern slightly better the implementation of the time-domain raised cosine filter is better explained below.

In Figure 7.10 above the pulse shape for a 0 *or* 1 bit value is shown, depending on convention a 0 bit can be portrayed by the exact same shape but mirrored in the x-axis. The total amplitude values along the time axis are then summed together. The time domain raised cosine pulse shape, showed in Figure 7.13, follows the same convention. The resulting pattern, for a random bit sequence, can be seen in Figure 7.16 below. It can be realized that using positive and negative raised cosine shapes, or a positive raised cosine shape for 1 bit, and no shape for a 0 bit, yield *the exact same pattern* when summed together, due to the symmetrical properties of the cosine shape. The only difference is the introduced dc-level in the pattern, which can be removed by simply subtracting a certain voltage. The actual pulse shaper also takes into account two consecutive bits, as two consequent ones should yield a line of constant amplitude, and actually shapes either half a cosine or a constant line. *The resulting signal sent to the modulator is then a constant level signal for a stream of only 0's or only 1's, or a perfect sine for a stream of alternating 0's and 1's.* For any other pattern, it will be a combination of the two extremes.

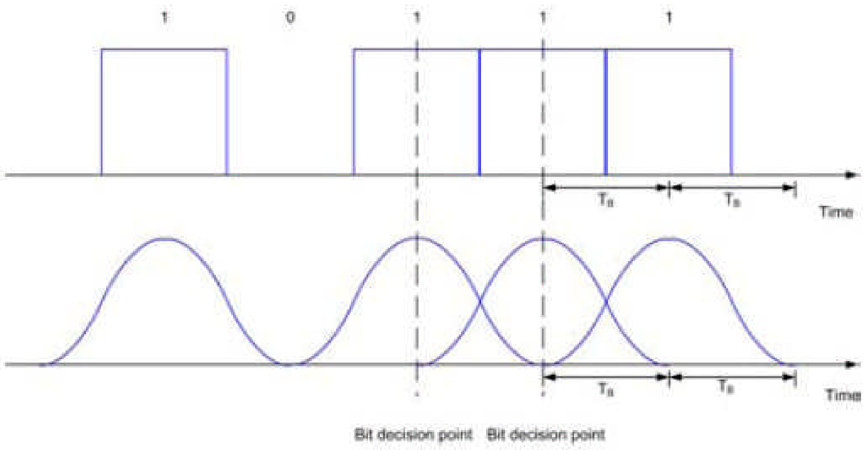


Figure 7.16: Delfi-C^3 implementation of a raised cosine filter applied on a bit sequence [SLR 0248]

In conclusion, the two possible extreme signals are a constant sine, or a constant line. In the frequency domain this would be indicated by an impulse at either the sine frequency, or at zero Hz. The actual signal will be no single impulse, but also will never have a frequency spectrum as bad as that of the rectangular pulse train.

In order to obtain the actual transmitted frequency spectrum, one should consider the *modulation process*. As explained earlier this chapter, BPSK modulation in fact is the mixing of a bipolar bit stream. The same holds if pulse shaping is applied. What is important to realize, is that (ideal) mixing creates two frequencies; one being the sum of the two mixed frequencies, and one being the difference. This is explained in the next section on the *superheterodyne principle*.

When a zero Hz or 600 Hz information signal are then modulated onto a carrier, the resulting signal is a pure carrier signal or a double-peaked sum of the two. In reality, the information signal is similar in behavior to that shown in Figure 7.15 above. The resulting frequency spectrum for Delfi-C^3 is presented below.

As Delfi-C3 has been built and is operational, there are two sources to determine the actual transmitted spectrum. These are Delfi-C3 itself (during a pass over the ground station), and its flight spare which is still in the clean room of the AE building. A spectrum analyzer has been used to determine the bandwidth characteristics of both in Science Mode, therefore generating a representative stream of 'random' mission data. The resulting measurements are presented below.

For illustration, three pictures are shown which demonstrate the frequency spectrum on different scales of the transmission signal as measured on the spare model. The y-axis is kept constant and shows a signal reduction of –10 dB per vertical square. The x-axis spans respectively 800, 60 and 20 kHz in Figure 7.17, Figure 7.18 and Figure 7.19 respectively.

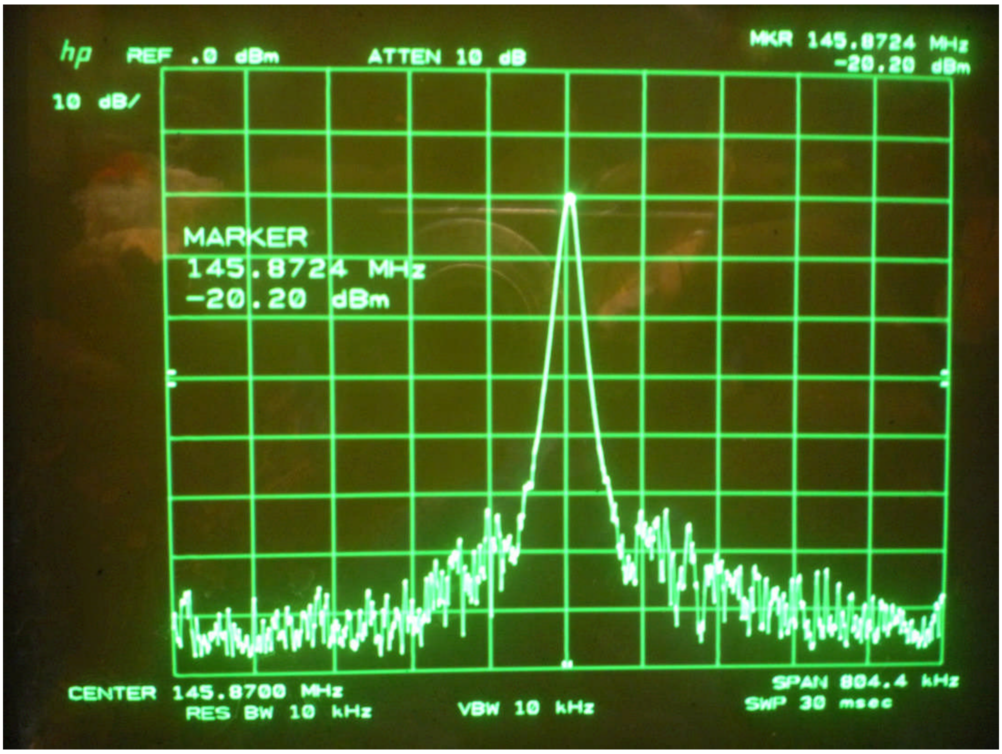


Figure 7.17: Measured frequency spectrum of the Delfi-C3 spare model transmission over a 800 kHz span

In Figure 7.17 above, it can be seen that the spectrum as 'zoomed out' basically is one simple peak, which has a span of some 160 kHz, and power levels at the outer edges of some 60 dB less than in the center. If one zooms in more, as in Figure 7.18 below, three peaks are shown. The two-side peaks are likely the result of non-linear effects in the modulation and/or up-conversion process. They have a value of some 40 dB less than the main peak. The final figure, Figure 7.19, then shows a pattern similar to that caused by the rectangular pulse train, but more steep. This then, is *the effect of the added time-domain raised cosine filter*.

As compared to the spectrum of the rectangular pulse train, the signal is mirrored in the carrier signal; two main peaks result at the carrier frequency (CF) plus 600 Hz, and CF minus 600 Hz. The carrier itself is *suppressed*. In fact, here it can be seen that BPSK modulation is in fact a form of *Double Sideband, Suppressed Carrier* (DSB-SC) modulation.

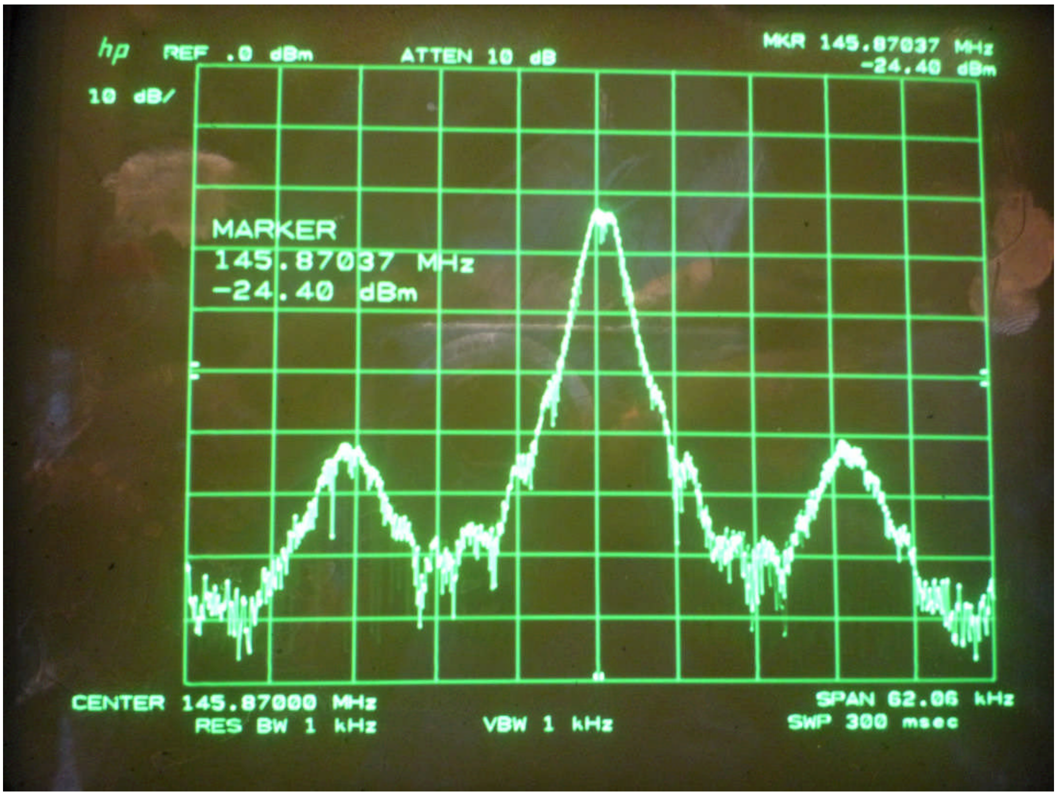


Figure 7.18: Measured frequency spectrum of the Delfi-C^3 spare model transmission over a 60 kHz span

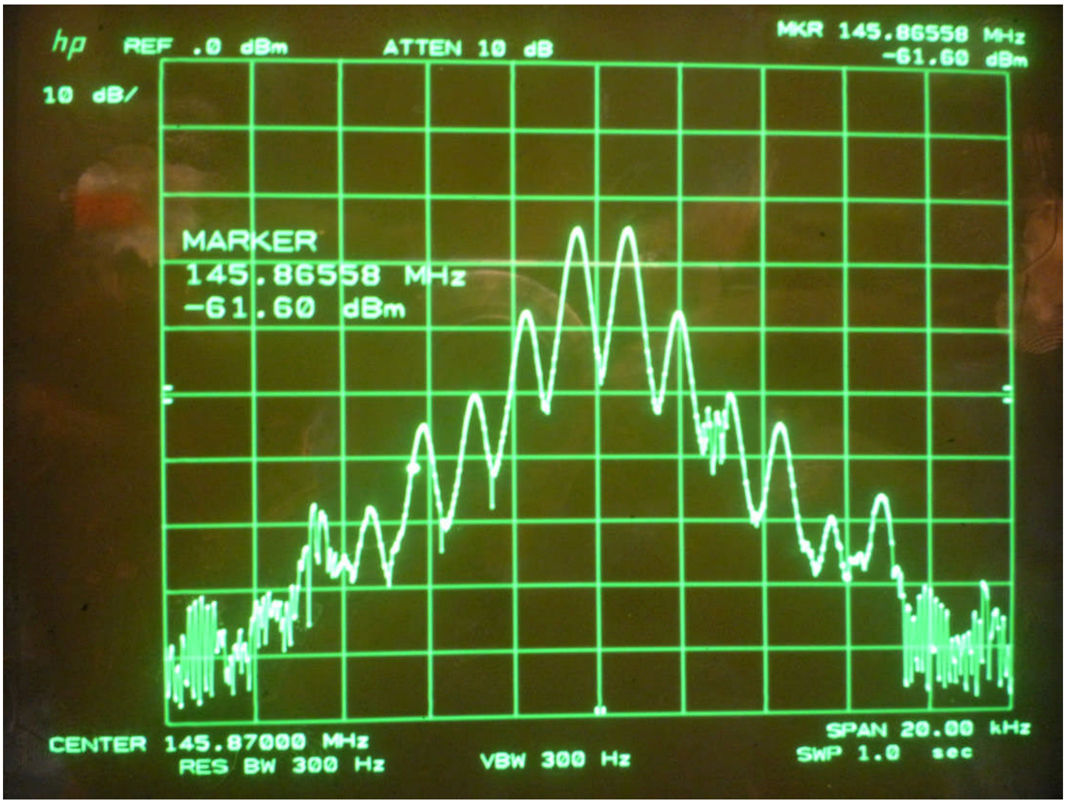


Figure 7.19: Measured frequency spectrum of the Delfi-C^3 spare model transmission over a 20 kHz span

What is important to take into account when looking at these figures is requirement SAT.2.3-P.01. It follows from ITU regulations and states that the spectral component at the edge of the allotted bandwidth should have a certain reduction of strength with respect to the strongest frequency component. As is calculated in chapter 8, the 200 mW power output of both the RAP and PTRX yields this reduction to be -36 dB. A reduction with respect to the strongest components, usually the carrier, is indicated by the unit dBc.

In Figure 7.19 above a marker has been place at the point where -36 dBc has been achieved. It might be badly visible in the figure, but the resulting required bandwidth is some **9 kHz**. *However, only **2.4 kHz** was actually allotted to Delfi-C^3*.

This is an important conclusion as it shows that Delfi-C^3 actually did not adhere to the spectrum requirements imposed to it by the radio amateur community and ITU. Whereas verification was apparently done before launch, it seems this was only done for the signal strength of the harmonics of the transmission frequency (introduced by the PA) [SLR 0244], but not for the actual modulation signal. As no narrow-band filters are in place after this modulator, the resulting spectrum is not sufficiently reduced.

Figure 7.20 however shows a slightly more satisfying result. It illustrates the spectrum received on-ground, during a near-zenith pass. The measurement was made when Delfi-C^3 was more or less above Delft, resulting in maximum received power and minimal background noise as the sky is faced.

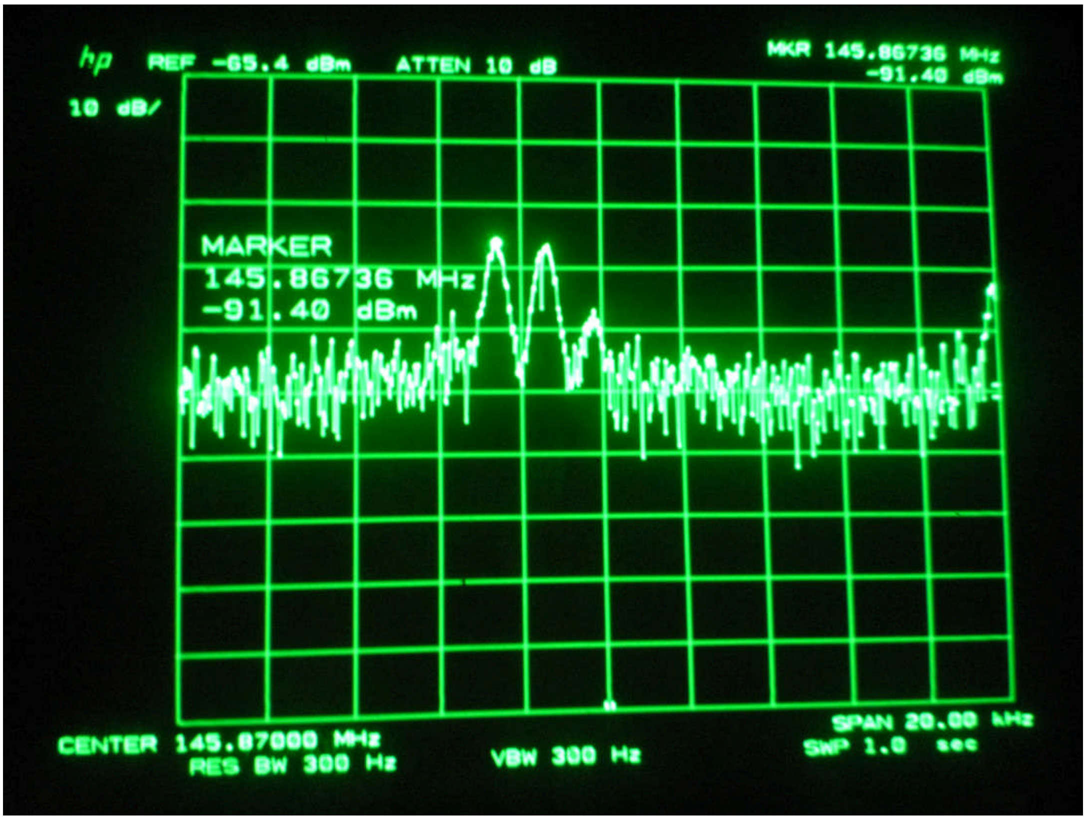


Figure 7.20: Measured frequency spectrum of the Delfi-C^3 transmission during a near-overhead pass

The received signal is in fact surrounded by background noise. The effect of this background noise is that the spectrum at lower power levels, thus that at the edges of the transmitted signal, is *immersed* in the noise. Then, Earth receivers with sensitivity similar to that of the Delft GS will never have disturbances more significant than the noise already present. Radio amateur ground stations will in general perform worse or similar to the Delft GS, given its equipment and excellent altitude. The only exception would be for remote areas with little radio activity, such as in deserts. The signal power levels of only 4 peaks can be said to be above the noise floor, resulting in a bandwidth of some **5.0 kHz**. While still not compliant to the regulations, or below the value of 2.4 kHz, it is significantly less than 9 kHz.

The latter measure, using the bandwidth *above noise levels*, is not stated by regulations but is a more satisfying one for reasons of common sense. In fact, what seems to be a downside of the regulations that state the required power reductions at the sides of the spectrum, is that they do this *with respect to the transmitted center frequency power*. Whereas many (space) transmitters have transmission powers of many watts, that of the Delfi-C^3 RAP and similarly the Delfi-n3Xt PTRX is *only 0.2 W*.

The use of a *reference bandwidth* in order to determine the signal strength at the edge of allotted bandwidth, as further explained in chapter 8, reduces the 'official' value as an average power value is taken within a 4 kHz span. Still, this is not enough to be compliant.

However, there is a big comment that can be made. In fact, as confirmed by a conversation with multiple radio amateurs, such as also Graham Shirville who heads frequency coordination, bandwidth is indeed a fairly ambiguous term and should not be weighed too heavily. Especially on the radio amateur bandwidths, frequency spectra of different projects per definition can overlap, which is a necessary evil of the limited resources. Also, as commented on above, the power levels of many radio amateur satellites are moderate. A general rule of thumb for the supposed occupied bandwidth then becomes some 2-2.5 times the data rate, irrespective of whether this requirement is actually formally met. And indeed, no complaints have ever been received on the topic of Delfi-C^3.

For Delfi-n3Xt, it becomes important to know what the spectrum will do when the data rate goes up. As data rate and occupied spectrum are directly correlated, a doubled data rate would logically yield a doubled required spectrum. Of course, the absolute violation of the regulations becomes larger then, assuming a doubled allotted bandwidth for doubled data rates. A discussion on the effects and possibilities for Delfi-n3Xt is included in chapter 8.

7.4.3 Gaussian filter

Gaussian filters apply a different approach than that of raised cosine filters. All energy is chosen to be concentrated within a period slightly longer than the bit period, using a Gaussian shape. A Gaussian filter thereby does not have zero ISI, but it is easier to implement. Also, the Gaussian filter does not require significant amounts of power to support the overshoots required in raised cosine filtering, which in turn simplifies symbol synchronization. Similarly, as symbols can only interact with generally 2 symbols left or right to it, amplifiers can be easier matched to the input. The occurring ISI however does require a slightly larger signal-to-noise ratio. The Gaussian filter creates a Gaussian shape both in time and frequency domain.

Gaussian filters are most notably applied in FSK schemes, which use the constant envelope of FSK to be able to use highly efficient class C amplifiers. GSM (the telephone system) and Bluetooth are good examples of the Gaussian filter implementation, the latter applying GFSK and the former increasing power efficiency through the use of GMSK. The shape of the Gaussian filter is specified by the BT value, which is the bandwidth time product. It is determined as [SLR 0695]:

$$BT = \frac{f_{-3dB}}{BitRate}, \tag{6-3}$$

It gives the bandwidth within the signal strength has reduced with 3 dB, or 50%, related to the bit rate. GSM uses a *BT* of 0.3, whereas Bluetooth uses one of 0.5. Figure 7.21 below demonstrates the resulting pulse shapes.

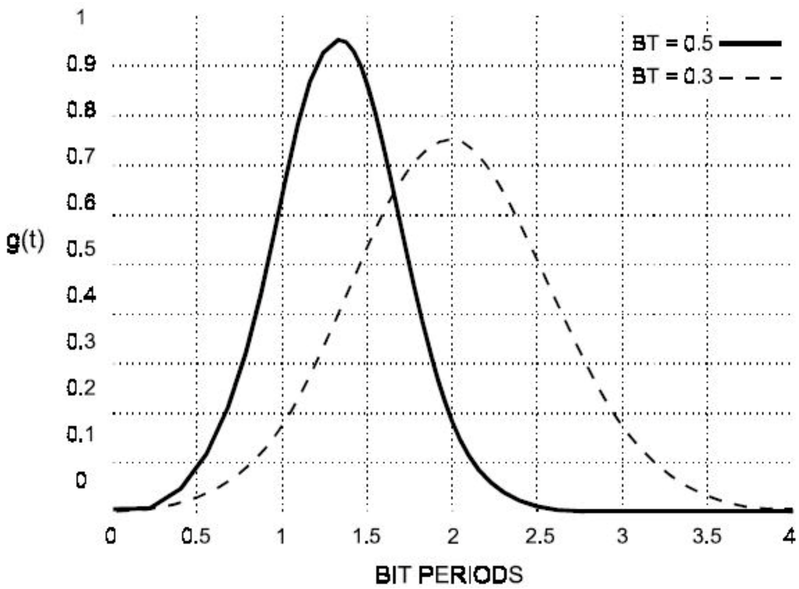


Figure 7.21: Gaussian filter shapes with BT of 0.3 and 0.5 [SLR 0695]

It can be seen that a smaller BT spreads the symbol over more time, a value of 0.5 causes the pulse to be spread over two bit periods approximately, and with a value of 0.3 it is spread over three. This spreading then gives rise to ISI. However, of course this gives better bandwidth efficiency. The bandwidth reduction for MSK schemes is shown in Figure 7.22 below.

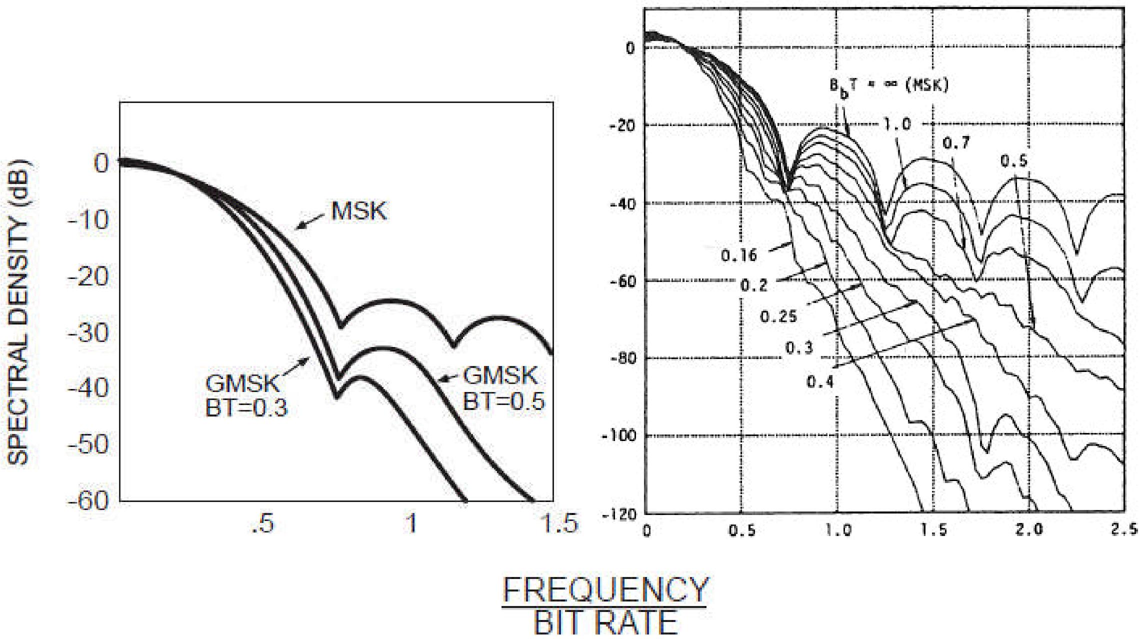


Figure 7.22: Gaussian filter frequency responses for different values of BT [SLR 0695 & SLR 0552]

Figure 7.22 (left) shows clearly the reduced spectrum for the two most well-known *BT* values of Gaussian filters, applied to MSK. A second lobe is allowed to appear, just as with regular MSK, but further lobes are suppressed. In Figure 7.22 (right) it can be seen that for smaller BT values this second lobe even disappears altogether. The result of the introduced ISI finally is a reduced error performance. [SLR 0552] gives some results of this phenomenon, as can be seen in Figure 7.23 below.

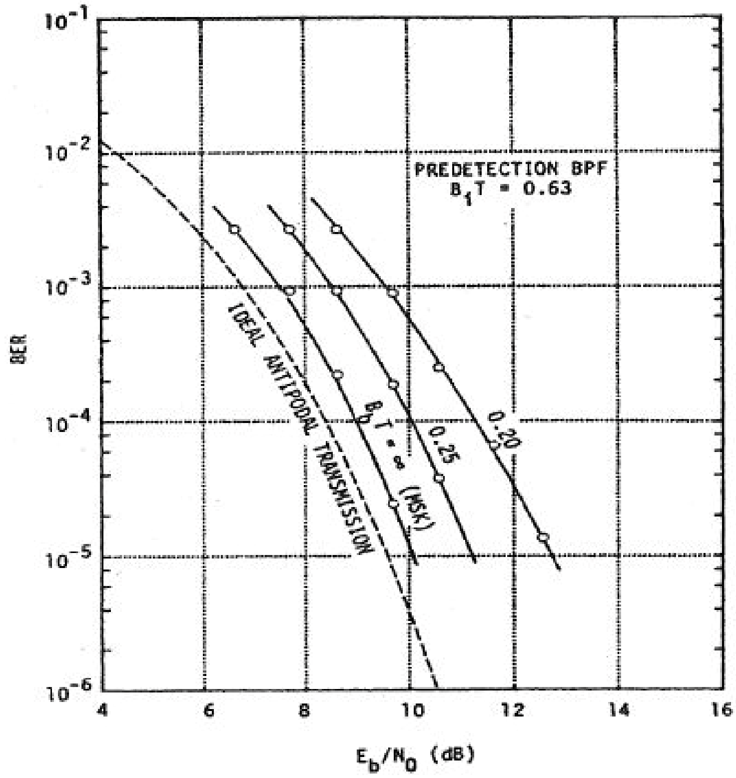


Figure 7.23: GMSK *BER* performances for different values of BT [SLR 0552]

First of all, it can be seen that regular MSK (with a BT value of infinity) yields some required 10.1 dB E_b/N_0 for a *BER* of 10^{-5}, even though theoretically it should only be 9.6 dB. This can be specific to the scenarios of the test, including modulator and demodulator. Therefore, the resulting *difference* between different BT values is most relevant. It shows that the BT value of 0.25 has only an error performance reduction of about 1 dB, compared to regular MSK. A BT value of 0.2 gives more significant differences, of some 2.5 – 3 dB more than MSK. However, as these *BT* values start to affect not 2 but 4 symbols next to that being transmitted, the 0.3 value is the lowest practically used. Depending on the allowed bandwidth, GMSK signals 'fit' within a bandwidth of some 1 times the bit rate, given the sizeable power reductions at the edges of between 30 and 40 dBc.

7.5 The superheterodyne principle

Present-day radios commonly apply what is referred to as the heterodyne principle. The idea of superheterodyne designs is to perform essential functionalities as modulating, but also consequent (pre-)amplifying and filtering at a far lower frequency than that that will be used for transmission; the *intermediate frequency* or IF. The effect is that the signal path thereby has to be sensitive to only a small range of frequencies. This principle is best explained by referring to the transmitter pictured in Figure 7.24 below, indicating several hardware blocks.

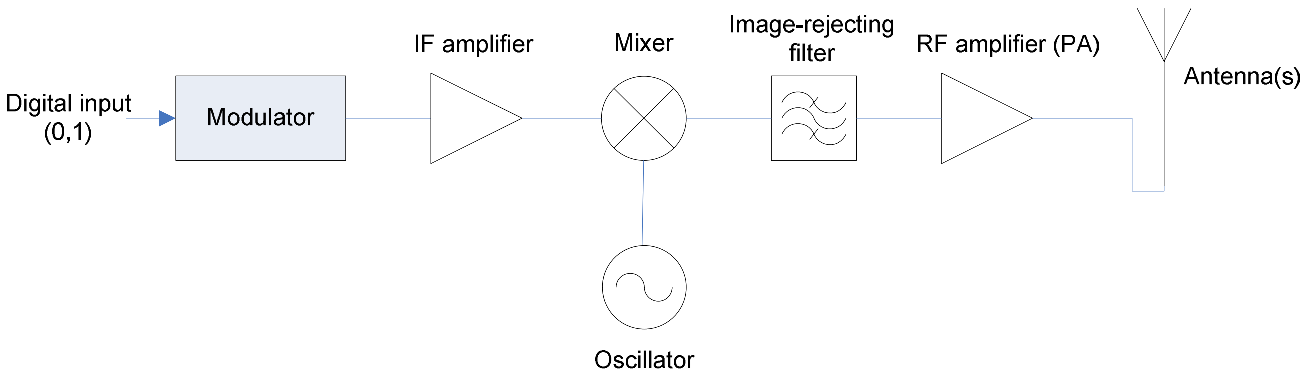


Figure 7.24: A superheterodyne transmitter

The complete modulation scheme, also including line coding and pulse shaping, is performed at *baseband*. Baseband indicates that frequencies close to zero are used, such as the bit transmission rates. During modulation, these digital signals are modulated onto a carrier, being an intermediate carrier, or IF. IF amplifiers can be inserted to amplify this low-frequency signal, using low-complexity linearized IF amplifiers.

An IF signal is consequently converted to a higher frequency, either being a secondary IF frequency if multiple conversion steps are used, or the ultimate *radio frequency*, RF. The generation of new frequencies by mixing or multiplying two oscillating waveforms is called *heterodyning*. The created frequencies are similarly called heterodynes. This process is further illustrated by Figure 7.25 below.

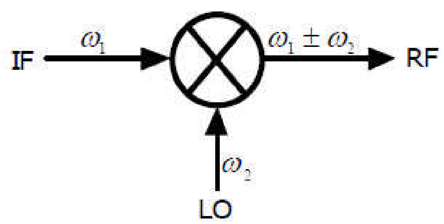


Figure 7.25: The process of heterodyning [SLR 0244]

An IF, having frequency ω_1 is mixed with a frequency ω_2 originating from a LO, or *local oscillator*. The resulting frequencies are created according to the relation:

$$\left(V_1 \cos \omega_1 t\right)\left(V_2 \cos \omega_2 t\right) = \frac{V_1 V_2}{2}\left[\cos\left(\omega_1 + \omega_2\right)t + \cos\left(\omega_1 - \omega_2\right)t\right], \qquad (6\text{-}4)$$

Where V_1 and V_2 are amplitude values and t is time logically.

Therefore, ideally, two frequencies result, being $\omega_1+\omega_2$ and *ω1-ω2*. In reality however mixing is performed by a non-linear process as mixers are not ideal. As such, the original frequencies remain present in the output, as well as *harmonics* of them and the sum of the frequencies; being integer multiples of ω_1, ω_2, and ω_1 *and* ω_2. The unwanted frequency resulting in the process described by relation 6-5, usually the lower one in the transmitter and the higher one in a receiver, as well as all other unwanted mixing products should be filtered out. This is done by what is usually called an *image-rejecting filter*, again see Figure 7.24.

As in this manner a one *sideband* can be said to be cut off, and as the carrier, being the LO frequency is suppressed, the resulting transmitted signal is of the form *Single Sideband, Suppressed Carrier* (SSB-SC).

After one or more frequency conversion steps using available electronics and after filtering the signal, the signal can be amplified to contain sufficient power for transmission. Therefore a dedicated RF amplifier or *power amplifier* (PA) is used. It is only this amplifier which adds significant amounts of power to the signal, and therefore efficiency becomes very important only for this amplifier. The signal should then be passed on via radio frequency circuitry and cables to the antenna, where a wireless signal is created.

It can be seen that by applying the heterodyne principle, modulation frequency and ultimate RF are functionally separated. This has advantages for the application of COTS components, and modularity in general. The same holds for intermediate components, as several IF frequencies as 455 kHz, 10.7 MHz and 70 MHz are highly standardized. The PTRX in fact applies exactly these IF frequencies, and thereby as well a larger number of mixers, filters and amplifiers than are suggested by Figure 7.24. The electronic design of the PTRX as said before is not discussed in this document, but in documents as [SLR 0281]. The consequent hardware block diagram including all IFs is presented in chapter 9 on the PTRX.

The collective of (de)modulation and superheterodyne design of mixers, filters and amplifiers establishes the entire physical layer involved in a transmitter or receiver. The specification of the digital input/output is discussed next in the following sections on data link layer functionalities.

7.6 Data protocol

Whereas modulation and line coding deal with bit-per-bit transmission, or physical layer functionalities, wireless communication requires more control. If a communication channel is only used for one-directional communication, thus in the absence of varying noise, and if transmitter and receiver are perfectly synchronized, there would be no direct need for further control. Since this is rarely the case, data frames are being made us of in all practical digital communication types, adding *layer two functionalities*. The functions performed on the data to form the data frames are specified by the *data protocol*.

Data frames are introduced in section 7.6.1. Afterwards the popular radio transmission data protocol AX.25 is introduced in subsection 7.6.2. A suggested improvement, allowing for error correction, FX.25, is introduced in subsection 7.6.3. Finally in subsection 7.6.4 some theory is presented with respect to data throughput efficiency of data protocols, yielding conclusions on optimum frame length.

7.6.1 An introduction to data frames

There are two general types of communication; connection-oriented (CO) and connectionless (CL). In the first type, small packets of information sent between transmitter and receiver negotiate a channel with certain characteristics, after which all raw data can be sent in one stream. In connectionless communication however raw data is just sent; usually in multiple packets with enough extra data to address the packets properly and to confirm whether the correct data was sent. CO communication thereby allows negotiating a certain quality of service conforming to the type of information sent. CO communication requires specific data packages in the negotiation of the channel, whereas CL communication uses data frames to package all information. The use of these data frames is further elaborated on below.

In space to earth communication or vice versa, only CL communication is seriously feasible. The direct reason is the delay introduced in their communication by the long distance in between. At the very least this type of communication will require a certain preamble to allow the raw data to be sent. As the receiver will not likely know precisely when or at what precise frequency the data will be sent (due to a Doppler shift, but potentially also variations in the transmitter caused by temperature for example); preamble can provide these details. The reception of a signal could trigger a receiver to 'pay attention', but also the preamble can provide details for carrier and symbol synchronization, depending on the requirements of a modulation scheme. These topics along with modulation have been further discussed in section 7.2, but in summary carrier synchronization implies tuning to the exact centre frequency (or the main centre frequency in case of frequency shift keying, or FSK), whereas symbol frequency allows the receiver to lock on the 'clock' of the signal, each tick signifying one symbol.

Whereas raw data with a certain preamble could consequently provide a working communication link, packaging is usually applied. Next to synch

Next to synchronization, more control is required as the transmission channel is not disturbance free for the duration of the entire transmission. Control of the data can be provided by wrapping pieces of data in concrete frames. Each of these frames will only require a short limited time to be sent, and as such not all information has to be received in one stream, and by the same receiver. Furthermore the accuracy of the received data can be confirmed or increased by adding redundant error detection or error correction data to selected blocks of information; usually with a block equalling the data content of one data frame (excluding possible flags; these are explained below).

Now the question is: "what comprises a data package?"

In fact, for a satellite mission, a data package can be quite simple. The essential parts:

- ***Start/end-indicators***
  A start-of-frame indicator, and, if the frame length is variable and not specified within the frame, an end-of-frame indicator are necessary to properly read out the frame configuration. An end-frame indicator can thus be replaced by a length field or a length previously agreed upon.
- ***An address field***
  Assuming the frame construction is not unique (for the mission at hand) and/or there are multiple sources/destinations that apply the same protocol, a sender/destination address is necessary to verify that the source/destination of the data is correct and that thus the frame is accepted or discarded. One or two addresses specifying the sender and/or receiver can then be integrated in a designated address field.
- ***Information content***
  The actual message, consisting of any predefined or variable number of bits.
- ***Bits to allow for demodulation***
  Depending on the type of modulation, carrier synchronization (see section 7.2) might be required. At the same time, depending on the bit stream and/or line code, additional symbol synchronization (bit synchronization) might be required. Both can be achieved by implementing bit patterns yielding preferred characteristics; an 111111 pattern with BPSK modulation yields a pure carrier signal useful for carrier synchronization, and a 010101 with FSK modulation yields a large number of transitions usable for symbol synchronization.
- ***Error detection***
  As there is no way to know a package is corrupted given the above fields, some form of error detection is simply required as space communication is prone to errors due to small transmission powers and long distances traversed. Depending on the frame length, a simple code of several bytes calculated over the frame can usually guarantee no small amounts of errors in the frame. As an error detection code might expect a certain number of bits, bits can be added to yield a certain multiple of 2, 4 or 8 bits for example.

The above fields basically comprise a minimum. Other things that can be included are listed below. As these are not automatically required for successful communication, they can also be seen as part of the information frame.

- ***Error correction***
  Instead of just error detection, (many) more bits can be added to not only allow for error detection, but also correction. This is the topic of the next section, 7.8.
- ***Extra addressing information***
  More addresses, required for routing for example might be required.
- ***Higher level control information***
  In the same vein as the previous point, additional addressing might be required to specify the higher level protocols; meaning for example software / application protocols.
- ***Frame type identifier***
  The frame protocol might include different frame types (corresponding to different functionalities, such as downlinking information, uplinking telecommands, confirming telecommands); in this case the type can be specified by a field.
- ***Information type identifier***
  Payload data, housekeeping data, and possible different configurations / combinations of these can be identified by a field.
- ***Frame number***
  To keep track of which frames are successfully received, a unique number can be added to each one.

- ***Real Time Clock (RTC) tag***
  To give more details about the information content, a real time clock tag can be added; which can be a relative or absolute time tag (a number of seconds since a certain date and time, or a date and time directly for example).
- ***Encryption / security measures***
  The content of a data frame, any number of fields, can be encrypted. Similarly, a (frame-content-dependent) key or encrypted hash (bit sequence calculated over the frame content) can be added to verify the sender and data integrity.

7.6.2 The AX.25 protocol

The AX.25 protocol [SLR 0293] is a set of rules that defines data framing for simple communication channels that do not require extensive routing, signalling and error correction. It does however allow for both CO and CL types of communication. It is widely used by radio amateurs as well as CubeSat teams not too different from the Delfi team. Similarly it is applied in the Delfi-C^3 mission and is thus well-proven in its use. The AX.25 protocol offers the essential functionalities as introduced above and little else, and is as a result reasonably data efficient. Adding the given advantage of its widespread application within the radio amateur community, whom Delfi-n3Xt is supposed to provide a favour for, for using the radio amateur frequencies, the AX.25 protocol provides a good option. It is further introduced below.

The AX.25 protocol offers 3 types of frames: information, supervisory and unnumbered (S, I and U) frames. I frames carry the actual information in normal flow conditions. S frames in turn, as [SLR 0293] states, "…provide supervisory link control such as acknowledging or requesting retransmission of I frames, and link-layer window control". The latter (window) indicates an approach in which multiple I frames are sent after which the first should still be acknowledged. As such, basically I frames automatically trigger an acknowledgement or a repeat request, in normal operations. U frames, can then be used to establish and terminating connections, in other words, in a CO approach. These functionalities are *not required or desirable* in a CL or connectionless approach. As such, UI frames can be used.

UI frames are classified as U frames, but mimic the functionality of an I frame, but without including a frame number and thereby requesting a confirmation. A UI frame in the AX.25 protocol has the construction as shown in Figure 7.26:

Flag	Address	Control	PID	Info	FCS	Flag
01111110	112/224 Bits	8/16 Bits	8 Bits	N*8 Bits	16 Bits	01111110

Figure 7.26: AX.25 UI frame [SLR 0293]

As can be seen the AX.25 incorporates exactly what is required for a simple one directional traffic stream, and not much more:

- ***The flags*** indicate the start and end of a frame, whereas the first flag also allows for carrier synchronization depending on the modulation scheme. For this purpose the first flag of the first frame is usually repeated a number of times. It should be noted that AX.25 is designed to be used with NRZ-I encoding or more precisely NRZ-S encoding (see section 7.3); in other words 6 consecutive '1' bits cause no signal transitions, thus a pure carrier signal.
- ***The address field*** of 112 bits gives two addresses, exactly that of the sender and receiver. More functionality can be added by extending it to 224 bits. The address field merits some further explanation as is given below.
- ***The control field*** stipulates the type of AX.25 frame used, and at the same time offers some more functionality as including the frame counter. In case of a UI frame it has an 8 bit length, and is set to 00000011 or 03 in hexadecimal (base 16 as opposed to base 2 with bits); 03_{hex}.

- ***The PID (protocol identifier) field*** allows for higher layer protocols to be specified, to supplement specific applications and allow for routing for example. As this functionality is similarly not required, 11110000 specifies "no layer 3 protocol", thus $F0_{hex}$.
- ***The info field*** is the field where actual raw data is encapsulated, as indicated in Figure 7.26 the only constraint is that it should be an integer number of bytes.
- ***The frame-check sequence or FCS field*** finally provides the ability to check the contents of the transmitted frame. It is calculated over the entire frame with the exception of the flags (and the FCS of course). The applied 'generator polynomial' is called CCITT-16, and equals $x^{16}+x^{12}+x^5+1$, yielding $1+ 2^5+ 2^{12}+ 2^{16}$. This equals a bit sequence 1 00010000 00100001. Section 7.7 explains how this generator polynomial is applied. This FCS has the ability to detect all single, double and triple errors in messages shorter than 32752 bits, or 4094 bytes, as well a burst errors up to 15 bits of length.

The 112 bits address field is composed of a destination address block, followed by a source address block. The lay-out of the address field is illustrated in Figure 7.27 below. Per address 6 bytes are available. This allows to specify a callsign of maximally 6 128-bit ASCII characters (supplemented with spaces), being either uppercase alpha or numeric characters. The 8th bit in each address byte is set to 0, to indicate more address bytes to follow. However, as AX.25 reverses the order of all bits, this is shown as the 1st bit. Only in the last byte, A14, if no address extension (224 bit address field) is applied, this 1st bit is set to 0.

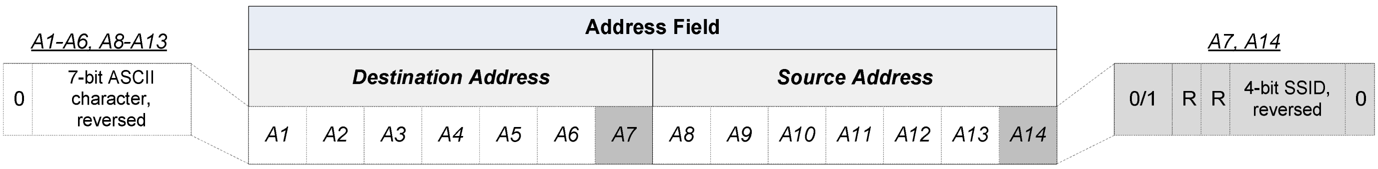


Figure 7.27: AX.25 Address Field

The bytes A7 and A14 as indicated allow for SSIDs, or secondary station identifiers to be added. Secondary stations are basically radios with the same callsign, but then an SSID of more than 0000. The primary radio indeed has a SSID of 0000. The second and third bit as indicated, officially being the 6th and 7th bit, are reserved for future use and are to be set to 1. The final bits of both fields A7 and A14 are per definition set to 0.

As can be seen from the description above, the start and end flag have efficiently been combined with a flag necessary for symbol synchronization. As the end flag bit sequence should not be allowed to occur within the frame, bit stuffing is applied. This makes sure all sequences of 5 1's are followed by a 0 bit within the frame body. This, at the same time, *yields a guarantee for sufficient bit transitions* during the frame to maintain symbol synchronization.

The frame length of an AX.25 frame is arbitrary, as *N* in the info field can take any (integer) number. A practical maximum is given by the FCS generator however. If *N* is sufficiently large, the overhead introduced (extra bits added) by using the AX.25 protocol is little. The fields added that would not be strictly necessary for simple communication are most notably the control field, the PID field, and the address field. The first one indicates the type of frame (UI), whereas the second basically says "no information". The address field, as argued before, can be said to be required to identify that the data actually comes from the proper source. If the source is trivial, it might not even be necessary. The same holds for the destination. In any case the address could be shorter, to indicate just the simple situation of frames coming from Delfi-n3Xt, and going to the ground station; which in the end indicates the same server.

In conclusion, this means that 16 bits are effectively useless but required due to the protocol, and 112 bits would be inefficiently used in case of a simple mission such as that of Delfi-n3Xt. Given a long enough frame this effect is minimized and AX.25 brings along the advantage of being a standard data protocol, at that one that is not very complicated. For shorter frames, or those of known length, it might not be desirable.

7.6.3 The FX.25 protocol

The AX.25 protocol does not provide error correction, only error detection. For this exact reason the FX.25 protocol was introduced [SLR 0294]. FX.25 is an extension to AX.25; FX.25 is designed not to replace the latter but to supplement it. A receiver capable of interpreting only AX.25 will interpret FX.25 data frames by AX.25 data frames surrounded by noise. For this exact reason the FX.25 protocol is also introduced here; it adds error correction to a well-known and well-integrated protocol.

The FX.25 data frame construction is shown in Figure 7.28 below. The entire AX.25 frame as shown in Figure 7.26, including flags, is integrated as indicated.

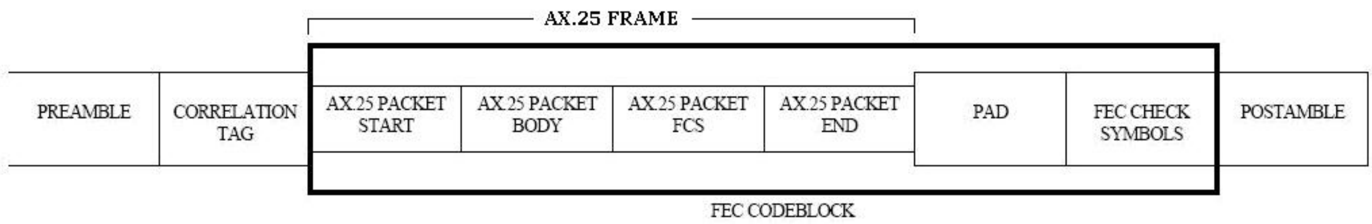


Figure 7.28: FX.25 frame [SLR 0294]

The four blocks in the middle represent the AX.25 data frame as illustrated in Figure 7.26. The idea of FX.25 is to pad the bits in the AX.25 frame with a number of bits, so that an FEC algorithm can calculate a number of FEC check symbols; these are consequently added to the padded AX.25 frame. The expected FEC algorithm is thus a block code; it takes a block of bits, and calculates a number of coding bits. This type of coding as well as its alternatives are introduced in section 7.8.

The resulting data block, of AX.25 frame plus FEC-related bits, is illustrated in Figure 7.28 under the name of the FEC codeblock. In order to specify the FEC algorithm used, a correlation tag is used. As the correlation tag itself should be interpretable even in a noisy environment, use of Gold Codes [SLR 0294] is advised, in a block of 8 bytes or 64 bits. As this use of FX.25 causes the original AX.25 flags to be hidden, there is a need for new preamble and postamble for frame start and end indication, and additional signal synchronization. As such it can be concluded that FX.25 introduces some practically unnecessary overhead, at the gain of interoperability with AX.25 interpreters. To be precise, the AX.25 flags, as well as the FCS are not strictly required anymore except for the recognition of the AX.25 frame. Nevertheless if the frame body is large enough these extra fields cause little unnecessary overhead.

7.6.4 (Delfi-C^3) data throughput efficiency and optimal frame length

The use of a data frame introduces a certain overhead. At such, it makes sense to reduce the relative overhead, by reducing the data protocol overhead itself, or by increasing the frame length. During transmission however, information is subject to a certain bit error rate or *BER*. Whenever a bit error occurs, a well-designed data frame will notice this, but due to overhead or complexity limitations will often not be able to correct this. Therefore, the entire frame is discarded. From a *BER* point of view then, the frame size should be minimized. The relation between these two parameters can be simply written out mathematically, yielding *data throughput efficiency*:

$$\frac{FS - OV_{total} - DL}{PS} = \eta_{throughput} \tag{6-5}$$

The effective frame size (*FS*) is reduced by two parameters, being the total overhead introduced by the data protocol (OV_{total}) and the average data loss (*DL*) due to transmission errors.

The total overhead can be divided in two separate losses, that incurred by data protocol fields (OV_{fields}) and that incurred by bitstuffing ($OV_{bitstuffing}$), if any. The first is just an invariant number, depending on the data protocol. The second can be written as a product of frame size and *bitstuffing ratio BR*, resulting in:

$$OV_{total} = OV_{fields} + OV_{bitstuffing} = OV_{fields} + BR \cdot FS \tag{6-6}$$

The data loss on the other hand is determined by the amount of data (not any form of overhead) lost on average. This can be written as equation 6-7, with *BER* being the *bit error rate* already introduced above:

$$DL = FS \cdot BER \cdot (FS - OV_{total}) \tag{6-7}$$

The product of *PS* and *BER* represents the amount of frames that will not arrive due to transmission errors, where the difference of *PS* and OV_{total} indicates the amount of data lost per transmission error.

Equation 6-6 can be substituted in equation 6-7, and equations 6-6 and 6-7 can then be substituted in equation 6-5 to yield equation 6-8:

$$\frac{FS - (OV_{fields} + BR \cdot FS) - FS \cdot BER \cdot (FS - (OV_{fields} + BR \cdot FS))}{FS} = \eta_{throughput} \tag{6-8}$$

To give some concrete results, equation 6-8 can be applied to the AX.25 protocol, *with Delfi-C^3 telemetry (TM) frames* as reference. TM frames profit the most from an optimized throughput, as data yield on-ground should be maximized. Telecommand (TC) frames on the other hand have no push to increase data volume throughput. Delfi-C^3 TM frames contained a maximum of 1200 bits, due to a 1200 b/s data rate and 1 frame per second. Bitstuffing is defined by the AX.25 protocol as 1 in 6 bits maximum, but frames were never created with more information that the worst-case would allow. Therefore, 200 bits out of 1200 were reserved for bitstuffing. Also, the AX.25 (UI) frame requires 160 bits of fields, next to the information field. Finally, the *BER* to be designed for is 10^{-5}. In summary:

$$FS = 1200,\ BR = \frac{1}{6},\ OV_{fields} = 160,\ BER = 10^{-5} \tag{6-9}$$

As was already explained above, the occurrence of overhead would favour an infinite frame length, whereas transmission errors would favour a bit-per-bit transmission. The optimum then is somewhere in between.

With the assumed *BER*, and the use of the AX.25 protocol, a throughput efficiency of **76%** can be obtained. In case of Delfi-C^3 however, this efficiency was only **69%**, due to the frame size of only 1200 bits. The optimum occurs at a frame size of 4400 bits, nearly four times as much. A frame size of 2400 bits however would have already raised the efficiency to more than 73%; some 40 bits per 1000 more.

In fact, Delfi-C^3 also applied a frame identifier (2 bits), and a larger frame counter (32 bits). As these fields are only useful once per frame, independent of its size, they effectively are overhead as well. Adding these bits to the overhead, thereby adjusting OV_{fields} to 194 bits, gives an efficiency of only **66%**, with a maximum achievable value of again nearly **76%** with frame sizes of 4800 bits. Again, 2400 bits per frame would have improved the efficiency to almost 74%, or 80 bits per 1000 more. In other words, *frame rate had likely better been set to 2 frames per second* (thereby integrating housekeeping and payload data). Given the current rotation rate of only a little over 0.2° per second, this would not have caused any problems related to polarization.

Both amounts of overhead discussed above are plotted for both *BER*s in Figure 7.29 below, the first case being "General AX.25, *BER*=10^-5" and the second being "Delfi AX.25, *BER*=10^-5". The two lines are therefore nearly superimposed middle lines.

Now what should also be mentioned, is that generally a link margin exists, which in case of Delfi-C^3 was sizeable. This additionally available transmission power effectively decreases the *BER*. As illustrated by the graphs in section 7.2, for BPSK modulation an extra signal-to-noise ratio of 3 dB already lowers the bit error rate to only 10^{-8}. The effect of such a *BER* is that overhead becomes the limiting factor; only an effective low-efficiency lower limit to frame size is set and *larger frames are favoured*.

It can clearly be seen in Figure 7.29 below that at small frame lengths, where overhead is relatively very sizeable, throughput efficiency is very low. From 1200 bits onwards it becomes acceptable, but 2400 is a better value. Afterwards, *BER* begins to play a role; a larger *BER* will push the graphs down again so that overall efficiency again goes down. A smaller *BER*, for example when the link margin is (momentarily) high, causes the throughput efficiency to keep rising slightly, in fact until a maximum optimum frame size is reached at **153,000 bits**, with a total efficiency of **83%** for the case of largest overhead. Effectively, the latter takes *BER* out of the equation.

What is also plotted is that at a *BER* of only 10^{-4}, only 60% efficiency can be obtained, and frame efficiency drops off quickly so frame sizes should be strictly limited. Not plotted, is that at a *BER* of 10^{-3} no positive average throughput can be obtained anymore; one in every 1000 bits will be in error.

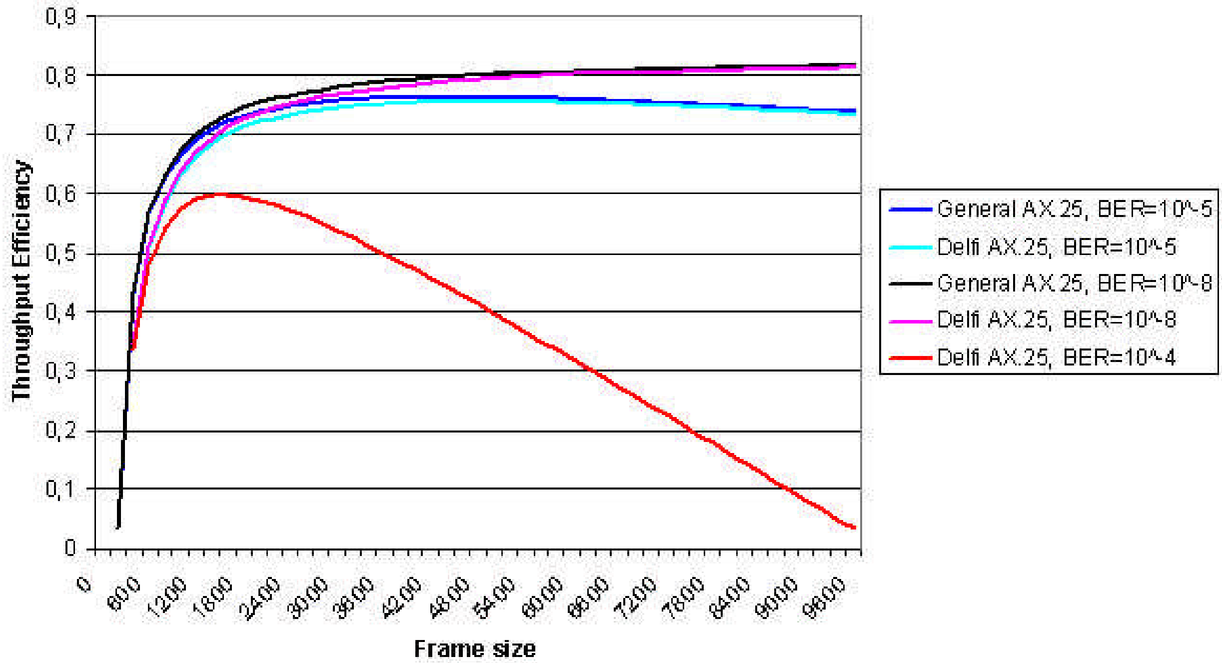


Figure 7.29: Throughput efficiencies vs. frame sizes for the AX.25 protocol

In general it can be said that:

- **Overhead should in any case be decreased as much as possible**, 160 bits and 1/6 of the total amount of message bits in AX.25 is a lot. The overhead is most notably caused by bit stuffing as it increases relatively to the frame size.
- **Data throughput can be optimized by increasing frame size.** In other words, by moving towards or over the maximum of this curve. The maximum can be approached relatively quickly, and surpassing the maximum will only decrease throughput slightly or increase it slightly, depending on the actual *BER* including margins.

In the next chapter the relation presented above is further used for the optimization of total data throughput for Delfi-n3Xt.

7.7 Data security

The AX.25 protocol introduced above applies a FCS, or frame-check sequence. This sequence is inserted to ensure the *data integrity* of the frame. In other words, this FCS is a form of data security. Data security is generally added as any real medium is prone to errors, be it in small amounts or large. These 'errors' can also be wilfully incurred by a third party, which tries to change the message conveyed.

This section introduces various applicable measures of data security, starting with the CRC, which is a general name for the type of security offered by the FCS, in subsection 7.7.1. Encryption is introduced in subsection 7.7.2. Afterwards, as the CRC is actually a form of a more general hashing algorithm, the (encrypted) hash function is the topic of subsection 7.7.3.

7.7.1 Cyclic Redundancy Check (CRC)

A CRC is a form or *error detection*. It is designed to warn for the occurrence of random channel errors, which, even in a very well-designed system can occur. In fact this system is designed to have a certain low bit error rate; in case of Delfi-n3Xt one of 10^{-5} bits. Another word for a CRC is a *polynomial code checksum*; by using a (generator) polynomial code it devises a checksum of fixed length. Depending on the length of the package to be transmitted, or *word length*, a specific number of bit errors can be detected. The number of unique errors is always limited, for example a well-chosen 16-bit CRC calculated over a 2048-bit word can never detect *more* than 3 random errors with 100% certainty, but depending on the location of them it can detect a (much) larger number. While this is prone to random success, the use of a specific polynomial in the calculation of the checksum, allows not only singular errors can be corrected, but also *burst errors of length less than the polynomial itself* [SLR 0683], and even more than **99%** of larger burst errors [SLR 0688]. As communication is usually the target of these errors exactly, this feature proves very useful. Furthermore CRC codes thank their popularity to their computational facility; both encoding and decoding can be done extremely fast in processors.

Calculation of the CRC

The word 'cyclic' in the name CRC comes from the fact that the message is cycled through while applying the same polynomial The latter has a length much smaller than the actual message. Computational facility comes from the fact that a simple shift register can be used, applying *modulo-2 arithmetic*. This latter means that when bits are added, regular addition rules apply, except for the case of two 1's being added; this results in a 0. The shift register is shown below in Figure 7.30 below.

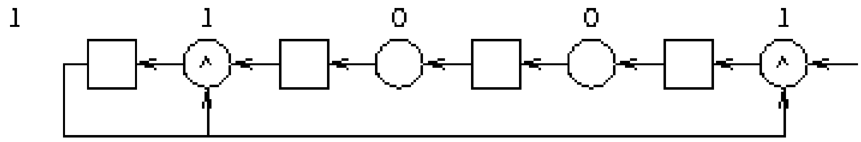


Figure 7.30: CRC shift register operation [SLR 0683]

The application of a 5-bit polynomial, 11001, is shown. In fact, the name polynomial comes from the possible binary polynomial representation of the sequence; in this case it would be x^4+x^3+1. A bit sequence, the message, is shifted through the register. Circles indicate operations, whereas squares indicate value store locations. Step after step, the incoming bits are at the same time passed on from square to square. At first, all squares contain zeros. The first square value is combined with the floating '1' value and is passed on to the circles with a value of 1 above them, or those with feedback loops connected to them; in other words whenever the first bit equals a zero, no changes are incurred to the bit stream. If however the first bit equals a 1, all circles with '1' values above them invert their input; a 0 is mixed with a 1 to become a 1, and a 1 is mixed with another 1 to become a 0. The result of this operation is a remainder of the length of the polynomial minus 1, which is the CRC.

The above process is rather time consuming, as it shifts and recalculates a value on a bit-per-bit basis. It has however been proven that simple PICs were able to live up to the task as well as processing the AX.25 frames on Delfi-C^3, applying the standard 16-bit AX.25 CRC. [SLR 0688] does however suggest another calculation approach (the Table Lookup Algorithm, or TLA) in which *n* bits can be processed at the same time, at the cost of a look-up table of size $k*2^n$, with *k* equalling the CRC size in bits; this yields a required 256 kB memory size for optimal 8-bit processing of the 16-bit CRC. An intermediate approach, the Virtual Table Algorithm (VTA), and a variant called Optimized VTA (OVTA) are also introduced in [SLR 0688] with speeds and memory sizes of intermediate values. The article finishes with typical execution times versus memory sizes.

On reception, the CRC can be recalculated over the message, to see if the received CRC is the same. Or, a second option is to perform the same CRC calculation over the message and the transmitted CRC; if the message and CRC are error-less, the result will be 0. If the result is not 0 however, this latter approach allows not only to detect errors, but to *correct* one error as well, if the error detection capability of the applied CRC is at least 2. The required processing application is sizeable however; the functionality is introduced further below.

Performance of CRCs

As introduced above, CRC capabilities depend on CRC length and word length. Any CRC will be able to detect single errors, as the minimal functionality is given by the *parity check*; if all bits are added together modulo-2, the resulting 1-bit 0 or 1 value will be able to detect all odd numbers of errors. Nevertheless larger CRC lengths are in general more useful, as they will allow for more errors to be detected and they add the burst error detection functionality. 4-bit CRCs are used in the USB protocol, but usual CRCs are integer multiples of 8, as (micro)processers generally calculate with 8 bit applications. Indeed also in Delfi-n3Xt this is the case.

Figure 7.31 below gives an impression of the maximum achievable amount of detected errors for a CRC of given size, for certain word lengths. It is the result of an exhaustive research into all possible CRCs for given lengths. *HD* stands for *Hamming Distance*; *HD*-1 yields the amount of detectable errors. The word maximum word length for the applicable scenario is given in the cells, with a hexadecimal representation of the CRC polynomial underneath it. Details of this representation can be found in [SLR 0687]. The pluses for *HD*=2 indicate that indeed every CRC can detect single errors in a word of indefinite length.

Max length at HD	CRC Size (bits)									
Polynomial	3	4	5	6	7	8	9	10	11	12
HD=2	2048+ 0x5	2048+ 0x9	2048+ 0x12	2048+ 0x21	2048+ 0x48	2048+ 0xA6	2048+ 0x167	2048+ 0x327	2048+ 0x64D	–
HD=3		11 0x9	26 0x12	57 0x21	120 0x48	247 0xA6	502 0x167	1013 0x327	2036 0x64D	2048 0xB75
HD=4			10 0x15	25 0x2C	56 0x5B	119 0x97	246 0x14B	501 0x319	1012 0x583	2035 0xC07
HD=5						9 0x9C	13 0x185	21 0x2B9	25 0x5D7	53 0x8F8
HD=6							8 0x13C	12 0x28E	22 0x532	27 0xB41
HD=7									12 0x571	none
HD=8										11 0xA4F

Figure 7.31: Optimal CRCs for a given CRC length, HD and word length [adapted from SLR 0687]

16-bit CRCs are those that have seen the most use over the years. As such they have logically seen a lot of research. [SLR 0684] aims to provide an exhaustive research into 16-bit CRCs. It concluded that there is no one 'optimal' 16-bit CRC, as with other lengths. A CRC focussed on a specific word length will underperform when compared to a similar-length CRC focussed on a different word length. The following Table 7-1 shows the maximum achievable performances for 16-bit CRCs, with given total lengths (including CRC).

Table 7-1: Maximum achievable performances for 16-bit CRCs [adapted from SLR 0684]

Total length n	***Hamming Distance***
17	17
18	12
19-21	10
22	9
23-31	8
32-35	7
36-151	6
152-257	5
258-32767	4
32768-65535	3
≥65536	2

It can be seen that *HD*=6, or **5 detectable errors**, can be achieved until total length of **151 bits**. The code that achieves this result however, with a generator of *236545* (in this polynomials codes are expressed as products of three octals; i.e. 23 stands for 8 bits), achieves only *HD*=2 above *n* = 151. Another code with generator *371237* achieves *HD*=6 until *n* = 130, *HD*=4 until *n* = 258, after which it drops to *HD*=2. *HD*=5 can be reached until *n* = 257, but also decreases to *HD*=2 for larger values of *n*.

The more power is used to transmit (higher link margin), the less errors are expected to occur. The uplink signal of Delfi-n3Xt has by far the most link margin as is further assessed in the next chapter. However, its reliable communication is more important as no telecommand should be wrongly interpreted. The above codes can therefore be very useful for (generally short) telecommands.

The downlink signal, with larger frame lengths, is more susceptible to errors. As such, 16-bit CRCs with performance optimized for larger word lengths are shown in Table 7-2.

Table 7-2: Maximum achievable performances for 16-bit CRCs with large *n* [adapted from SLR 0684]

HD	***Total length n per CRC polynomial***			
	200433	*321353*	*320227 (IBM-SDLC)*	*210041 (CCITT)*
10		17-18		
8		19-27	17-24	
6	17-115	28-109	25-83	
4	116-28658	110-32767	84-32766	17-32767
2	≥28659	≥32768	≥32767	≥32768

The fourth polynomial is the familiar $x^{16}+x^{12}+x^5+1$ polynomial applied in the AX.25 protocol, and thus also in Delfi-C^3. It can be seen that in fact other polynomials are strictly better. [SLR 0687] presents another polynomial by its generator value of 0xBAAD, which achieves *HD*=5 until *n* = 122, which turns out to be the maximum achievable for a code also providing *HD*=4 for message lengths of 1024; the article research is said to be exhaustive for all CRC codes up until this length.

Whereas [SLR 0687] mentions polynomial were once restricted by hardware complexity as polynomial 1's were more difficult to implement due to feedback loop, as illustrated by Figure 7.30, this is no longer the case as software is indifferent to the number of 1's. Nevertheless, the improvement of other polynomials can be said to be of pure theoretical value if the smaller message lengths are not used.

One thing to be added to the discussion above is the fact that CRCs are said to have *HD*=*N* up until the first word length where they lose this ability; afterwards a finite number of words exists where the *HD* value is no longer guaranteed. The CRCs *can* however still detect errors for larger messages. This number of words may vary per polynomial, and thereby also the chance that a larger number of bit errors is still detected. This is called *error performance of CRCs*.

[SLR 0685] gives a detailed analysis of these effects but only up until word lengths of 1024 bits. In this region the CCITT CRC is shown to outperform the IBM code. CRCs are generally applied to cope with random channel errors, which occur in the order of bits as opposed to bytes for example. A general rule of thumb [SLR 0686] says that if a *BER* of 10^{-2} is maintained on a link, the chance of two errors occurring is approximately 10^2 less. Therefore, as a *BER* of 10^{-5} is designed for in case of Delfi-n3Xt, which yields only one in 100,000 bits to be in error on average. The occurrence of two errors then is expected to take place a factor of 10^5 less. Therefore, the importance of this error performance to Delfi-n3Xt is limited.

24- and 32-bits are discussed in detail by [SLR 0689] which follows up the discussions on 16-bit CRC in [SLR 0684]. Three interesting 24-bit CRCs are introduced, which for convenience shall only be named by their 'birth names':

- ***CRC-24/6.1***: *HD*=6 for *n* = 84-2050, *HD*=4 for *n* = 2051-4098 and *HD*=2 for *n* ≥ 4099
- ***CRC-24/5.1***: *HD*=6 for *n* = 38-252, *HD*=5 for *n* = 253-4097 and *HD*=2 for *n* ≥ 4098
- ***CRC-24/6.1***: *HD*=6 for *n* = 62-846, *HD*=4 for *n* = 847-8,388,607 *and* HD=2 for larger *n*

For 32-bit CRCs total lengths of 306 bits can be reached with *HD*=8, at the same time yielding *HD*=6 until 32,768 bits and *HD*=4 until 65,534 bits. [SLR 0686] introduces yet another code class with *HD*=6 until only 16,360 bits, but extending the range of *HD*=4 to 114,663 bits. Little 32-bit CRC codes are presented that seem to be optimized for smaller word lengths, or these are simply of little practical interest as relatively larger overhead would also be incurred for these messages. The largest *HD* at intermediate total lengths was found to be *HD*=8 until *n* = 1024, but after *n* = 2046 quickly dropping of to *HD*=2, with an intermediate value of *HD*=4. It can be seen that a doubling of CRC length, with respect to 16-bit versions, *does not* directly double the amount of errors that can be corrected at intermediate lengths, but *it can* more than double the allowed word length.

Error correction

As introduced above, by calculating the FCS over a received message and its FCS a remainder results, being either 0 or not. In case of a 0, no errors can be assumed to have occurred. If the value is nonzero however, the remainder itself holds the information as to where the error has occurred. As the message consists of bits, knowledge of where exactly the error has occurred allows to simple 'flip' the bit value for a resulting correct message.

For large messages, the calculation of the error can be difficult. In fact, general targeted error correction approaches usually add relatively large overheads (as opposed to CRCs) so that errors can be extracted relatively easily. Nevertheless, for single error correction using CRCs some shortcuts can be used. The restrictions:

- The message length cannot be more that 2^{d-1}-1, where *d* is the degree of the generator [SLR 0691]. This is the largest factor in the generator; CCITT has an x^{16} term, so degree = 16. The maximum message length is then 32,768.
- Three bit errors can generate the same residual as a single bit error; therefore to make use of error correction the chance of a three bit error should be very low.
- The frame length should be known.

There are different manners of calculating the erroneous bit location, but only one class leads to an acceptably reduced calculation overhead for sizeable frames. This is by means of a table look-up algorithm. The idea is quite simple. For a given received message, due the properties of a CRC, there is only one optional received FCS per error location. In result however, a straight-forward table would have to store 2^{32} values of 32 bits for a 32 total length of message and CRC. For larger messages this is of course undoable. As such, smarter table look-up algorithms have been created; an optimized one is given in [SLR 0690].

The resulting table has a much smaller length, with a total amount of *n* rows, where *n* is the total length of message and FCS as above. The total amount of bits that need to be stored is then equal to:

$$2^{p} \cdot (m - p + 1 + (\log_2 k)) \tag{6-10}$$

Where *m* is the amount of CRC bits, *k* is the message length and *p* depends on the polynomial and message length; it has a value of at minimum $\log_2 n$, in other words the amount of bits needed to express the message length with. The trick with the algorithm is that an appropriate set of distinct rows should be obtained; if it is not found, *p* is increased by 1 and the search process is restarted. No results are given on typical iterations for sizeable message lengths.

A simple calculation gives for a downlink frame length of Delfi-C^3 (1000 bits total, excluding bit stuffing) a matrix with 1000 rows of course, and a total matrix size of at least 17,000 bits or 2.1 kB. Potentially this number should however be multiplied by 2 a number of times. In case of a 2400 and 4800 bit frame respectively with a 16 bit CRC the equation yields a minimum matrix size of 70,000 and 130,000 bits, or some 9 and 16 kB.

7.7.2 Encryption

The goal of encryption is to mask the information being transmitted. This subsection will present some terminology and gives a non-exhaustive introduction to some examples of cryptography.

In the process of masking original data, *plain text* is transformed into *cipher text*. A *cipher* controls the encryption and decryption, usually be means of a *key*. Whereas the cipher then is the algorithm, the key changes the variables of the encryption. The key therefore add security to the algorithm. There is a distinct difference between *symmetric* and *public* key schemes; a symmetric key system implies having a similar key at transmission and reception, whereas a public key system does not. Consequently, there is also a difference between *block* ciphers and *stream* ciphers; block ciphers operate on a block of predefined size such as 64 or 128 bits, whereas stream ciphers do not. Any content to be encrypted with block ciphers should therefore be a multiple of these block lengths; if it is not it can be stuffed with 'dummy' bits. The best-known symmetric key block ciphers are DES (Data Encryption Standard) and AES (Advanced Encryption Standard). RC4, also known as ARC4, is the most-widely used symmetric key stream cipher. Public key systems, such as Diffie-Hellman, RSA and DSA have the downside of involving a much larger computational load [SLR 0492].

Stream ciphers yield the largest flexibility, as the message size is not restricted. Using a 25-character key, a pseudorandom *key stream* is generated, which is used to transform the message. The key stream is combined with the message (the TC info field in the Delfi-C^3 case) using a bitwise exclusive-or operation. The resulting cipher text can only be decrypted by applying the same process, with the same key stream. RC4, next to being the most-widely used, is remarkable for its simplicity and speed in software. It has therefore successfully been applied in Delfi-C^3 for the TC uplink.

7.7.3 (Encrypted) hash function

Hash functions are a general class of functions that reduce a large amount of data to a small related amount of data. CRCs are in fact non-secure hash functions; multiple messages can transform to the same *hash*. Two identical hashes following from a different message create a *hash collision*. CRCs do not have the same purpose as a hash, as they are intended to protect against a small number of transmission errors. Hash functions have multiple uses, such as speeding up search processes by means of hash searching. Also, in case of data transmission, hash functions can provide much better protection against intentional altering of the message during transmission; if the hash algorithm is unknown that is. Compared to CRCs however they thereby incur a much larger overhead, depending on the amount of hash collisions allowed.

Hash sizes can be effectively reduced by applying encryption. A smaller, less effective hash function might produce many possible hash collisions, which can be misused by a potential attacker by changing the message while yielding the same hash. Encryption however can make use of the fact that the encryption algorithm is now known, thereby allowing a number of existing hash collision, without the attacker having sufficient knowledge to exploit them. Well-known encrypted has functions are those of the SHA family, or MD5 [SLR 0492]. Of course, the complete process of extracting a hash and encrypting it is rather computation intensive.

7.8 Forward error correction

The previous section has introduced means of error detection, through using the CRC or a more general hash function. Depending on the application, erroneous frames or content can then be simply discarded, or a discard can be followed by a new request for the corrupted date, requiring repeating transmitted data. This is the first of two methods that can be used to deal with frequently occurring errors in a digital communication system. These two methods are to make use of:

- An automatic repeat request (ARQ)
- Forward error correction (FEC)

The first type of error handling works by labelling blocks of data and adding redundant bits (parity bits) such as in the CRC approach. The calculation of these bits is based on the content of the data block, and if the same calculation on the received bits yields a different result, a request for a repetition of the erroneous data block is sent to the original transmitter. This is an example of error detection, followed by the actual ARQ.

In case of forward error correction or FEC, redundant bits can also be used to correct for errors. However, FEC is generally used for a different purpose than correcting sporadic errors. It is used to actually *increase the achievable bit rate, by lowering the required power needed to transmit a data bit*. This is explained with the use of two figures directly below.

It was introduced before that a certain E_b/N_0 is required to transmit a bit with a certain (low) probability of error, or *BER*. This E_b/N_0 depends on the modulation scheme, also including line coding and filtering. If the data rate is increased, but available power is unchanged, the E_b/N_0, or bit-energy-to-noise ratio goes down. Indeed, when error correction bits are added to a certain amount of data bits with fixed total power, the *BER* goes up. This is shown in Figure 7.32 below.

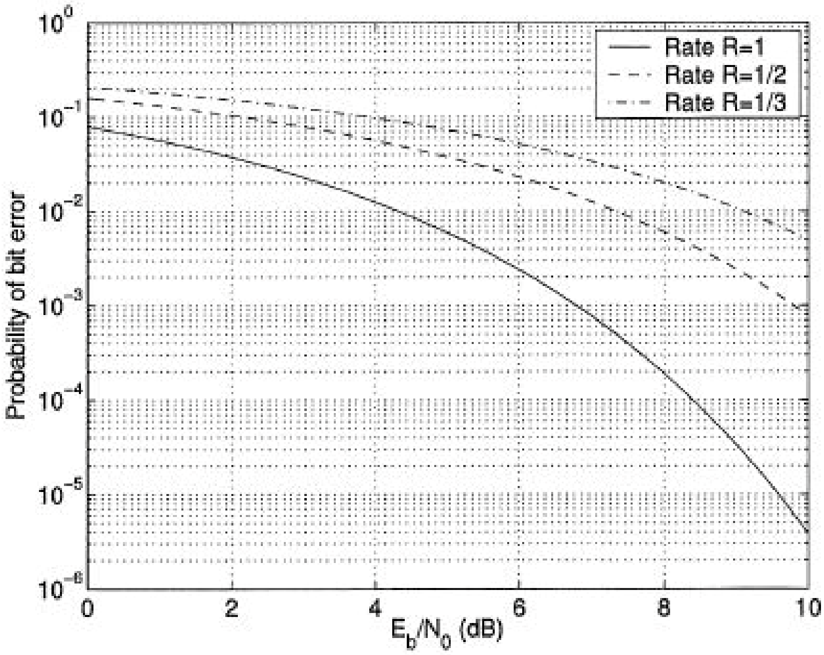


Figure 7.32: E_b/N_0 versus *BER* for different data rates [SLR 0718]

Figure 7.32 above shows the performance of a BPSK modulated signal. A *rate* (short for *code rate*) of R=1 signifies an uncoded signal, whereas an indicated rate of R=1/2 and R=1/3 signify 2 and 3 times as much transmitted bits respectively. Code rate is defined by *k/n*, where *k* is the amount of information bits and *n* is the total amount of bits transmitted. It can be said that the amount of *bits* is increased, but the amount of *data* remains the same; the same can be said of the *bit rate* and *data rate*. Logically, 3 dB of increased E_b/N_0 is required for a doubling and almost 5 dB for the tripling, for an unchanged *BER*. Clearly, for the added bits to be of any value, the coding bits should compensate for the lower transmission power per bit, *and* yield error correction capability. Fortunately, many types of codes exist that meet this criterion. An example of the resulting performance is shown in Figure 7.33 below.

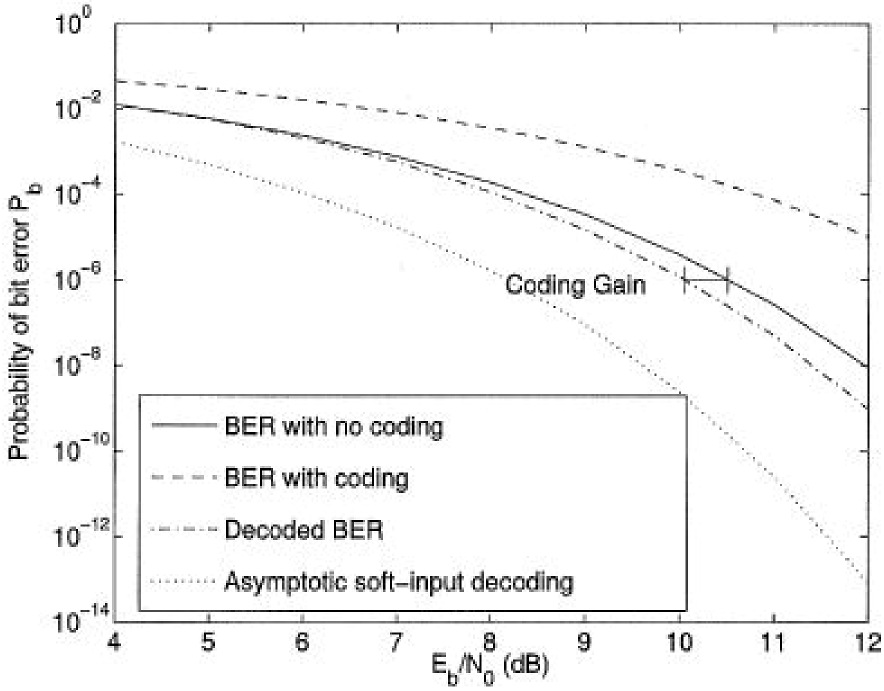


Figure 7.33: E_b/N_0 versus *BER* for different data rates, including coding effects [SLR 0718]

Figure 7.33 illustrates the performance of a (7,4) Hamming code, the specifics of which are not important here. The uncoded *BER* versus E_b/N_0 can again be seen, as well as the graph for a coded signal without being decoded. Also added then is the resulting *decoded BER* versus E_b/N_0, which indeed can be seen to introduce a certain *coding gain* with respect to the original signal. A maximum theoretical performance for soft-input decoding is given; soft-input or soft-decision decoding uses a demodulator that is not limited to binary data, instead a value that is closer to a '1' or a '0' is taken along in the decoding process. The indicated performance is theoretical as it indicates the result for an infinite amount of bits taken along in the decision-making process.

In conclusion, it can be seen that the addition of FEC bits lowers the total E_b/N_0 required for a certain *BER* to be achieved. As such, more bits can be transmitted with the same *BER*, *thereby effectively not introducing error correction to the end-to-end product*. Of course, the error correction capability can also just be used to decrease the overall *BER*. Generally, FEC is simply used to increase data rates at the cost of increased complexity. Depending on the type of error correcting code (if no estimation algorithms are used, such as with simple block codes) these codes *also allow for error detection* if more errors are induced. Otherwise, low-complexity error detection fields can still be added to detect errors. An acceptable *BER* is determined at mission definition, in case of Delfi-n3Xt 10^{-5}.

Error correction is a very broad and especially complicated field, which is further complicated by active on-going research and the existence of intellectual property and patents. As such this chapter gives an introduction to the field, including definitions and techniques used, and presenting typical performances. Decoding approaches and algorithms are generally not discussed, books as [SLR 0573] and [SLR 0718] can be consulted for this. Highest performances are currently achieved with turbo codes and concatenated coding schemes, so more time is taken to explain these coding schemes.

General code classes are first explained in subsection 7.8.1. Subsections 7.8.2 through 7.8.5 discuss typical techniques applied in coding schemes. Using this knowledge, both turbo codes and a more specific type of concatenated coding scheme designed within the radio amateur community are introduced in subsections 7.8.6 and 7.8.7. Finally, coding performances are listed and compared in subsection 7.8.8.

7.8.1 Code classes

Channel codes can be classified into two broad categories:

- Block codes
- Convolutional codes

Block codes are the simplest to visualize. A number of *k* bits is processed by a coding algorithm to produce a number of coding bits. These coding bits are added to the *k* bits to produce a number of *n* bits. Knowledge of the used algorithm when decoding will provide the ability to detect and usually correct a certain number of errors. *Convolutional codes* provide continuous coding data on a data stream, in other words they have memory. Based on the current bit and a number of previous bits certain coding bits are produced and added to the data stream. One can also distinguish between *systematic codes* and *non-systematic codes*. Block codes are per definition systematic codes; the original data bits are embedded in the resulting bit stream. Convolutional codes do not necessarily have to be systematic; the coded bits can replace original bits. The systematic coding property is important for turbo codes as is assessed below.

Convolutional codes with little or moderate memory are the easiest to implement on the encoder side; these however increase the *code rate*, as more bits are added overall. Block codes are neither very complex to implement, but require a certain buffer size; usually limited to one or two hundred bits. The code rate for block codes is usually moderate, typically between 0.70 and 0.90, whereas for convolutional codes they tend to be lower. The best-known block codes are those of the BCH (Bose and Ray-Chaudhuri) class, of which a subtype is that of the Reed-Solomon (RS) codes. RS codes add to the general BCH code the ability to correct also a number of burst errors instead of just single errors.

Most complexity is incurred in the decoding process. A complete discussion of the techniques required for decoding falls out of the scope of this document, and is the topic of many (large) books. Nevertheless, for standard codes the decoding algorithms are readily available in programmed packages. The performances of codes are introduced in subsection 7.8.8 below.

7.8.2 Viterbi algorithm

A very important decoding algorithm for convolutional codes is the Viterbi algorithm. The Viterbi algorithm is a maximum likelihood sequence estimation algorithm. Having knowledge of the encoding process, the algorithm calculates the most likely original data stream by considering all possible state transitions at the encoder. For a detailed explanation, see [SLR 0573].

An important conclusion with regard to the Viterbi algorithm is that it increases system complexity only at the decoding side; a regular convolutional encoder is used in encoding the data.

Viterbi decoding can in theory also be applied on block codes, but as the block size increases the number of decoding states increases exponentially. Other maximum likelihood algorithms such as the Chase algorithm can deal with block codes in a smarter way, nevertheless block sizes should be minimized, while larger block sizes already yield higher coding gains reducing the need for complex decoding. Having knowledge of the allowed complexity a trade-off can be performed.

Finally the Viterbi algorithm can also be applied directly to the demodulation process, without added FEC coding. This causes the demodulator to decide on the value of the received bit, taking into account also a number of other received values representing bits. If soft-decision demodulation is applied however the same information that would be used in direct Viterbi demodulation is passed on to the decoder.

7.8.3 Soft- and hard-decision decoding

An important distinction can be made between hard-decision and soft-decision decoding; this is especially important when applying Viterbi decoding, but soft-decision decoded block coded and convolutional codes (without Viterbi decoding) have also been applied. A usual demodulator will yield bit values, in other words 1's and 0's, indicated by high and low bit values. The nature of demodulation however, be it phase, amplitude or frequency shifted (see chapter 7), will never present clear values at bit boundaries. For example in the case of PSK the phase will, due to channel noise and the bit precision point, never be exactly +1 or -1. If the demodulator outputs not 1's and 0's or +1 and -1 values, but a number of intermediate values, the decoding process can achieve much better performance. In terms of E_b/N_0 (assessed in more detail in chapter 7), the difference between hard-decision and soft-decision decoding is an approximate 2 dB.

7.8.4 Interleaving

A technique frequently used in coding to circumvent *burst errors* is interleaving. An interleaver takes a sequence of data bits and changes its order in a predetermined way. The result of this is that burst errors, or errors that occur on a number of consecutive bits, are in the deinterleaving process spaced apart and as such better correctable. Burst errors usually occur due to fading effects, as discussed in chapter 7, as well as due to spin effects. The latter refers to a satellite that spins and consequently produces a fading and growing signal as its antenna changes orientation. The application of interleaving is demonstrated in Figure 7.34 below.

After the FEC coding, the order of the bit stream is changed by inversing the rows and columns of a matrix of predefined size. In the reception process the same inversion takes place after demodulation and before FEC decoding, to again inverse columns and rows. It can be envisioned that in this manner, burst errors occurred during transmission are spaced apart and therefore more easily detected.

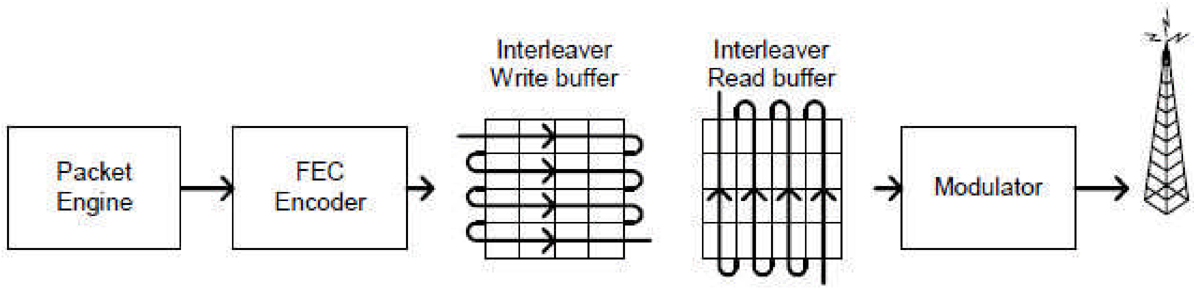


Figure 7.34: Interleaving applied during the transmission process [adapted from SLR 0556]

An important application of interleaving is the use of turbo coding.

7.8.5 Concatenation

Concatenated codes are a means of effectively obtaining a longer code, while retaining modest decoding complexity. A basic concatenated coding scheme is shown in Figure 7.35 below. Two codes, an *inner* and an *outer* code, are combined and executed in series.

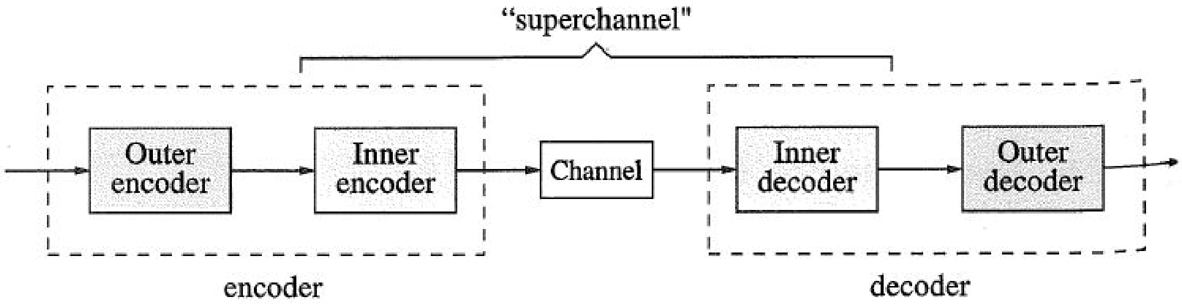


Figure 7.35: A concatenated coding scheme [SLR 0718]

Generally, the outer code is a relatively simple RS code, involving low complexity but being applied for its ability to correct burst errors. The inner code then usually is a convolutional code, which is optimally decoded by a Viterbi decoder. If a Viterbi decoder makes a mistake, it typically involves a few consequent bits. The burst-error correction capability of the RS decoding, taking place after the Viterbi decoding, can then handle this.

7.8.6 Turbo Codes

A number of techniques introduced above, such as interleaving, soft-decision decoding and Viterbi decoding are combined in the application of *turbo codes*. Turbo codes are the first line of codes which perform close to the maximum channel capacity. The general turbo encoder is graphically illustrated in Figure 7.36.

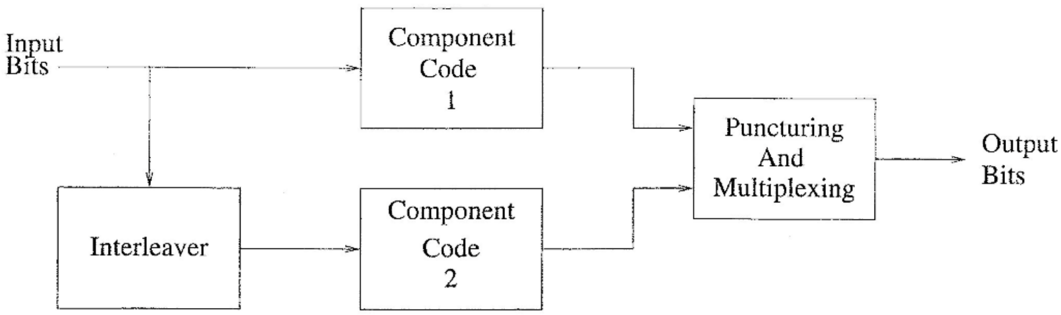


Figure 7.36: Turbo encoder schematic [SLR 0573]

As can be seen in Figure 7.36 the information sequence is splitted in two similar streams, of which one of the two is interleaved to make the two approximately statistically independent. Consequently, they are encoded twice by two component codes, but with the two being in *parallel*. Turbo codes do not apply a concatenated coding scheme with two codes in series. In turbo coding schemes, these two codes are then logically equal; this has to do with the *puncturing* taking place afterwards.

The combining of the two streams is simply called *multiplexing*. When identical and systematic codes are used as component codes, both codes produce the original bits and half can be discarded. The usual approach is to consequently *puncture* the parity bits from both encoders; only half the parity bits are kept.

The general turbo decoder is shown in Figure 7.37.

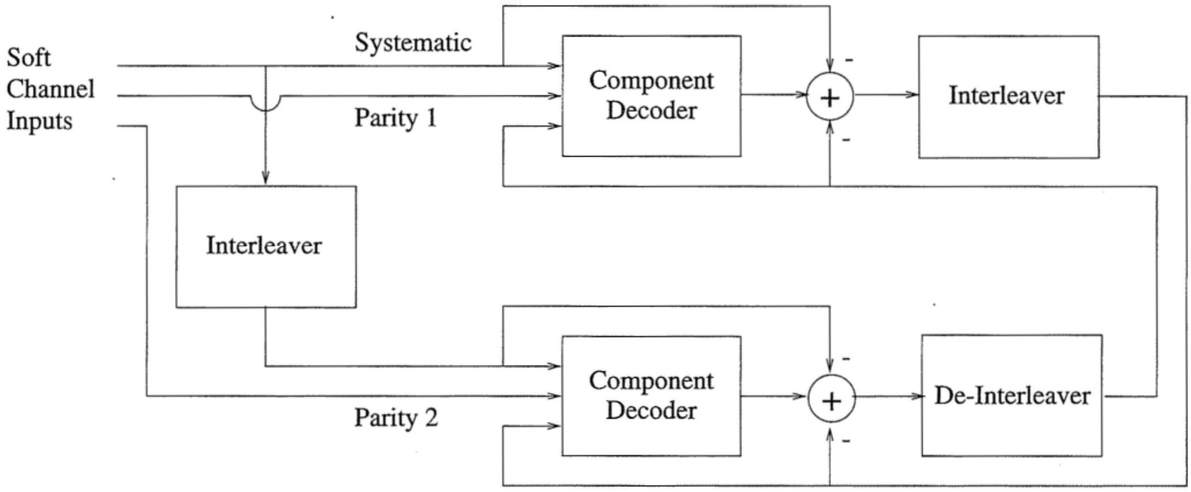


Figure 7.37: General turbo decoder schematic [SLR 0573]

The decoder as shown is the ideal decoder, as it is the most advanced. The modulator is assumed to have produced soft(-decision) channel data, and the input to Figure 7.37 has already been demultiplexed. The decoder shows the inverse of the encoding process, with an extra introduced iteration process. The top component decoder first processes the systematic bits along with the first set of parity bits, after which the data is interleaved and sent to the second component decoder. The second decoder uses not only the systematic bits and the second set of parity bits, but also the assumed decoded bit stream as decoded by the first component decoder. This completes the first iteration, after which the first decoder can take into account the results from the second decoder to reexamine the expected bit stream with more information on the plausible original. The component decoders can apply maximum likelihood sequence estimation algorithms such as the Viterbi algorithm, next to the required component decoder algorithms.

It can be seen that the decoding complexity is much higher than the encoding complexity, due to the possible inclusion of iteration, use of algorithms such as the Viterbi algorithm, and soft demodulation data.

[SLR 0573] provides a detailed review of the performance changes of a turbo coded BPSK data signal sent over a Gaussian channel, related to its parameters. Space can be modeled very well as an AWGN channel. For a detailed study on the design of a turbo coding scheme one is referred to this book. The main variables in a turbo coding scheme, which are also assessed are listed:

- The component encoding / decoding algorithm used
- The puncturing approach
- The number of decoding iterations used
- The frame length of the input data
- The interleaver design

The performance of turbo coding is evaluated below in subsection 7.8.8.

A final aspect important to mention on the topic of turbo codes, is their constrained use in terms of *intellectual property* (IP). Turbo codes have not been developed until 1993 onwards, which means that 20-year patents are still valid. As such, types of component codes as well as puncturing approaches are in many cases still tied up in legal restrictions. The radio amateur community has expressed the desire for a high-performance FEC scheme, without such legal problems, and has therefore developed the AO-40 FEC coding scheme, introduced next.

7.8.7 AO-40 FEC coding

The AO-40 FEC coding scheme is one that approaches performances of turbo codes, but with techniques that have been developed in the '60s and '70s. In fact it is based on the coding scheme applied on Voyager I and II missions of 1977. Many digital broadcast satellites have a coding scheme based on the latter. As such, no IP problems exist. At that, the radio amateur community is one that aids others in the development of useful techniques, and shares this knowledge freely in order to advance their potential. As such, all details and design choices of the coding scheme are worked out, and preprogrammed encoding and decoding packages even exist. The following details follow from three articles published on websites: [SLR 0731, 0732 and 0733].

The AO-40 FEC encoder and decoder are shown in Figure 7.38.

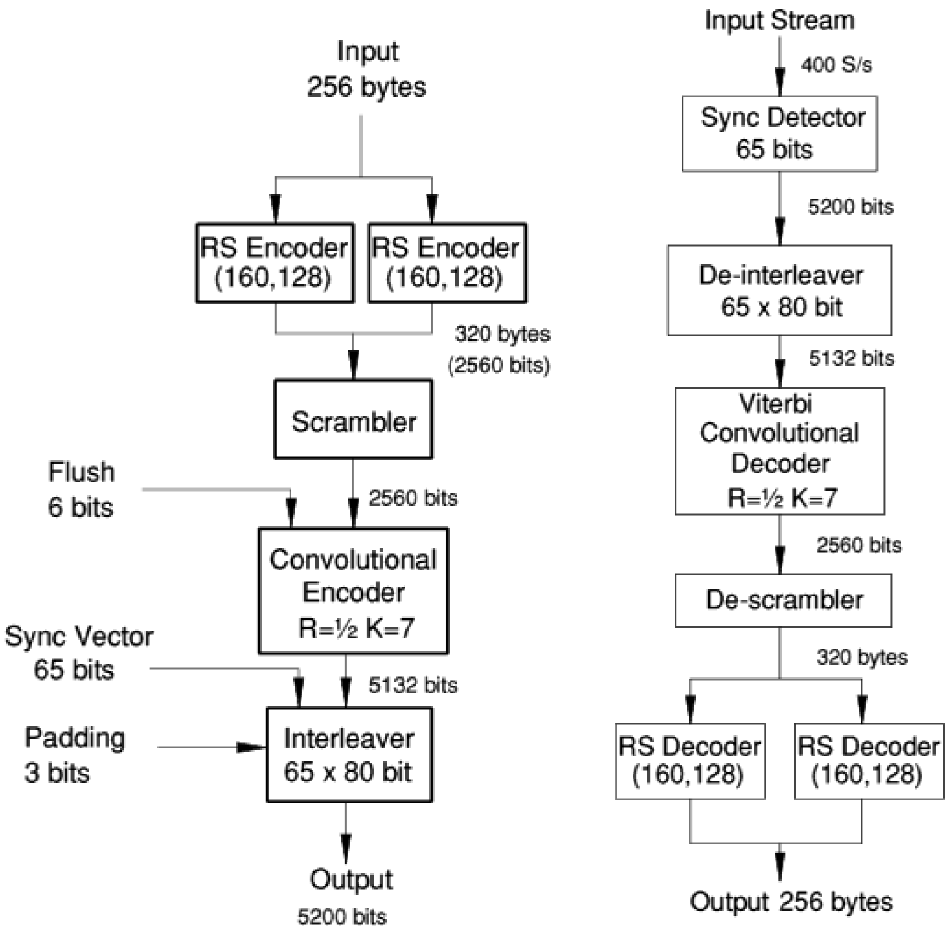


Figure 7.38: AO-40 coding scheme [SLR 0731]

All different fields of the encoder are quickly discussed here, with the number of bits at that point in the chain indicated:

- **Input (2048 bits)**
  A predetermined message size has been set, which works out with all other coding lengths and inserted bit values.
- **RS Encoders (2560 bits)**
  Two RS encoders are used, with even-numbered bits encoded by the first encoder, and odd-numbered bits encoded by the second. The details of the code can be found in [SLR 0731]. The two streams of data are then joined simply in an odd/even fashion again.
- **Scrambler (2560 bits)**
  Scrambling is actually applied to insert sufficient transition density in the data stream; required for DBSPK demodulation. In case of MSK demodulation (see S-band communication link definition, chapter 12), this might actually not be required. In any case, the procedure here is to XOR the data with a generator, not unlike CRC calculation. The generator suggested in this case is $x^8+x^7+x^5+x^3+1$.
- **Flush (2566 bits)**
  Six 0's are added in front of the data stream in order to make the convolutional encoder start in an 'all 0's state'. This is likely beneficial to know when decoding.
- **Convolutional Encoder (5132 bits)**
  This encoder applies a convolutional code with a code rate of ½ and a memory size K of 7 bits. Again, code specifics are presented in [SLR 0731]. Nevertheless, the purpose of combining an RS code with a convolutional code is to remove errors that might occur in the Viterbi decoding process of the convolutional code.
- **Sync Vector (5197 bits)**
  A sync(hronization) vector is added to the data for correct data reception. The sync vector is matched to the row size of the consequent Interleaver, and adds one sync bit to each column. It is generated by using the x^7+x^3+1 generator, repeated to form a 65-bit bit stream. This vector should allow a receiver to recognize the data while scanning, and to position the start of the package. In fact, this is a *spread spectrum technique* as introduced in Appendix A. As [SLR 0731] explains, a correlator that looks for a 65-bit vector has 10*log(65) = 16 dB more signal strength than the message in which the vector is contained, which is argued to be some -1 dB in case of bad reception. 16-1=15 dB which is sufficient to guarantee near-always reception.
- **Padding (5200 bits)**
  Padding (of 0's or 1's) is quite simply applied to round-off the total number of bits to 5200, to simplify the interleaver.
- **Interleaver (5200 bits)**
  This interleaver is, as explained before, applied as an anti-fading measure.

The resulting message can be decoded by logically inversing the main operations. An input stream is suggested (400 symbols/s) but is not restricted to low data rates. Viterbi decoding is logically applied in the decoding of the convolutional decoder. Soft-decision demodulation is advised with 3-4 bit values [SLR 0732]. 256 bytes should result again at the end of the cycle.

The performance of this code is described below, and compared to other code performances.

Software ('C' language) for the implementation of this FEC coding scheme has been written, and can be found here (last accessed 13-06-2010): http://www.ka9q.net/ao40/

7.8.8 Coding performances

As introduced above, the most prominent class of block codes is that of Bose-Chaudhuri-Hocquenghem (BCH) codes, and a subclass of it Reed-Solomon codes. These codes are usually expressed by for example BCH(n,k,t), where n is the number of bits per total block, k is the number of information bits and t is the maximum number of correctable bit errors. Due to its complexity the block size of BCH codes rarely is above 127. Consequently the BCH(127,71,19) code yields best performance, with a gain of almost 4 dB for a *BER* of 10^{-5} over uncoded BPSK. Conversely, a smaller block codes can be used with a Viterbi decoder; a soft-Viterbi decoded BCH(31,16,7) code can yield a similar gain of a little over 4 dB. [SLR 0730] yields more BCH codes and their performance, with gain values between 1.7 and 3.0 dB.

RS codes perform slightly worse for random errors, but they are slightly better in dealing with burst errors. As such these codes are used in compact-discs (CDs).

At the cost of more complexity, convolutional coding schemes can be pushed to have better performances. A number of space missions [SLR 0575] have applied sizeable convolutional coding schemes. The planetary standard code, with a rate of ½, a memory size of 7 bits and soft-decision Viterbi decoding achieves a gain of 5.1 dB. Pioneer missions 10 and 11 to Jupiter and Saturn applied as well a rate ½ coding but with a memory size of 32 bits, using a sub-optimal decoding scheme called sequential decoding as Viterbi decoding could not be applied due to the large memory size, achieving a gain of 6.9 dB. The aforementioned Voyager I and II missions, as well as the 1998 DirecTV satellite system use again a variation of the planetary standard code. Using the latter as the inner code, and a combination of a convolutional code with RS(255,223,33) block coding with Viterbi soft-decision decoding, produces a gain of 7.1 dB.

The AO-40 FEC coding scheme is heavily based on this Voyager scheme, and [SLR 0732] mentions a very similar performance; 2.6 dB E_b/N_0 is noted to be required for BPSK modulation, normally requiring some 9.6 dB. As such, 7 dB coding gain is noted, of course assuming soft-decision demodulation.

Finally, turbo codes yield the best theoretical performance. Using turbo codes as assessed above, a coding gain of more than 8 dB is relatively easily achieved over the required E_b/N_0 of uncoded BPSK with a required E_b/N_0 of 9.6 dB. Whereas this requires at least 4 to 8 iterations at the decoder, the encoder complexity is still relatively low. A frame length of at least 4000 bits is preferable. The maximum gain to be achieved using the current architecture is almost 9 dB, however using an interleaver of size 2^{16} and 18 iterations at the decoder.

In summary, typical coding performances are listed in Table 7.3 below; values are generally taken from [SLR 0443]. A data rate capability has also been indicated, as a measure of complexity. All values mentioned above can also be found in the table. AO-40 FEC coding is in fact of the type *'concatenated RS and convolution (Viterbi decoding)'*.

Table 7.3: Typical coding performances

Coding technique	Coding gain (dB) for *BER* = 10^{-5}	Data rate capability
Ideal coding	11.2	
Turbo code	8.0-8.8	Moderate
Concatenated RS and convolution (Viterbi decoding)	6.5-7.5	Moderate
Convolutional with regular decoding (soft decisions)	6.0-7.0	Moderate
Block codes (soft decisions)	5.0-6.0	Moderate
Convolutional with Viterbi decoding (hard decisions)	4.5-5.0	High
Convolutional with regular decoding (hard decisions)	1.5-5.0	High
Block codes (hard decisions)	1.5-4.0	High

7.9 Selecting the data rate for optimum data volume

[SLR 0106] introduces the link budget. The link budget takes into account all losses and gains along a transmission link, and basically results in final signal strength. Having knowledge of the minimum required signal strength then allows to determine a maximum achievable data rate, given a certain bit error rate requirement to be met.

As is introduced in the link budget, a number of parameters vary with the elevation angle. These are:

- Atmosphere loss
- Receiver noise temperature
- Path loss

The atmosphere loss in fact varies somewhere between 0 and 1 dB between elevation angles of 90° and 10° respectively. The contribution is therefore generally minor. A notable exception however occurs at lower elevation angles, at 5° elevation the atmosphere loss becomes some 2.1 dB, whereas moving towards 0° increased the atmosphere loss to over 10 dB [SLR 0219].

Noise temperature is presented in [SLR 0106], and is composed of a *receiver noise temperature* and an *antenna noise temperature*. The former is dependent on the equipment used, whereas the latter depends on the received noise through the antenna. The antenna temperature is not actually a physical temperature. It is a parameter that is hard to determine, and is usually estimated based on the frequency band and the ground station location and isolation. Or, it is measured. At low elevation angles, the antenna receives more noise from terrestrial (transmitting) sources. At very low elevation angles the antenna noise can therefore be much higher than at slightly higher ones. Again, its effects are most significant at low elevation angles.

Then there is path loss. Path loss L_p is given by the following relation [SLR 0316]:

$$L_p = \left(\frac{\lambda}{4\pi \cdot S} \right)^2 \tag{7-11}$$

Where λ is the (constant) signal wavelength and S stands for slant range, in turn given by the following equation, assuming a circular orbit:

$$S = R\left(\sqrt{\frac{(R+h)^2}{R^2} - \cos^2 \varepsilon} - \sin \varepsilon \right) \tag{7-12}$$

Where R is Earth radius, h equals orbit altitude and ε the *satellite elevation (angle)*.

Maximum achievable data rates $R_{\max}$ can be directly correlated to path loss, as a higher path loss (counterintuitively defined to be a smaller actual loss) directly increases the achievable data rate. In other words:

$$R_{\max} \sim L_p \tag{7-13}$$

Then, $R_{\max}$ is indirectly given by relation 7-11 via the following relation, assuming all other losses are constant with elevation angle ε:

$$R_{\max}[\varepsilon] \sim C \cdot L_p[\varepsilon] \tag{7-14}$$

Concluding from equation 7-12, given an orbital altitude of 600 km; the total distance between the GS and the satellite changes from 1932 km to 600 km with an elevation angle between 10 and 90°. According to equation 7-11, path loss then shows a whopping reduction from -141.40 dB to -131.24 dB. This difference of 10 dB equals a factor 10; in other words a data rate of a factor 10 more can be reached when transmitting at maximum data rate when the satellite is directly overhead!

According to relation 7-14, the instantaneously achievable data rate $R_{\max}$ is given as a function of elevation angle ε. The resulting graph for different orbit altitudes (corresponding to the Delfi-n3Xt range of orbit altitudes) can be seen as Figure 7.39 below. The achievable data rates are normalized with respect to the highest value.

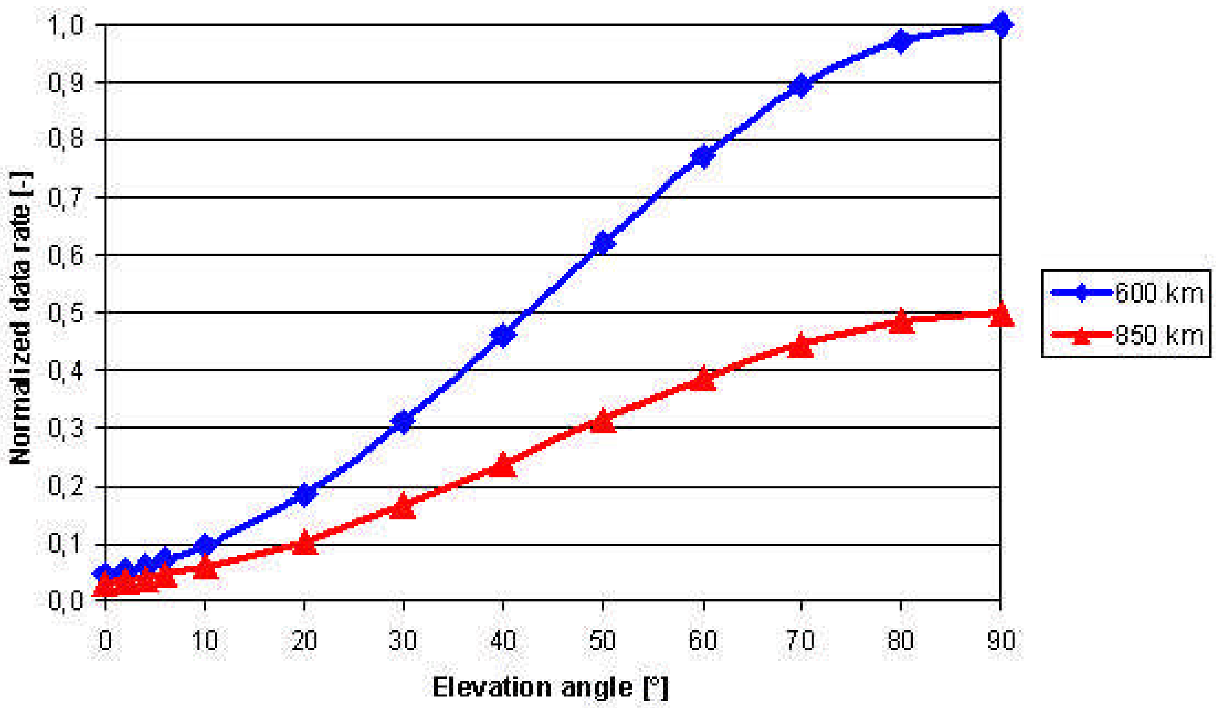


Figure 7.39: Normalized achievable data rate versus elevation angle for the Delfi-n3Xt range of orbits

As the horizon is usually obstructed by buildings, and as explained atmosphere loss and antenna noise can increase abruptly moving to an elevation angle of 0°, the elevation angle for start of communication is usually taken to be slightly higher. A usual value is 10°, or possibly 5°, but rarely less. This then influences the total useful contact time, as a larger minimum elevation angle decreases the total ground station visibility zone. The data rate and the contact time together determine the total data volume that can be transferred during one pass, following the integral:

$$D_{\text{pass}} = \int_0^{T_{pass}} R_{data}(t) \cdot dt \qquad (7\text{-}15)$$

Where D_{pass} is the total data volume that can be transmitted during a pass, T_{pass} is the contact time during a pass and $R_{data}(t)$ is the data rate over time.

Generally, the data rate in satellite missions is constant throughout the mission. This is due to the complexity involved with generating a variable stream of data, changing the data rate on-board the satellite and properly communicating a changing data rate between the satellite and the GS. Also, the data rate is generally set to have a proper communication link during the entire pass; from a small elevation angle onwards.

This section discusses the effects of path loss on the achievable data rate. This will show that an approach where the data rate that is based on the minimum elevation angle when communication is possible, will in fact not always maximize total downloadable data volume, depending on the mission constraints. Using the design target of *maximizing this downloadable data volume*, different scenarios are discusses in this section.

The scenario first described is that of a single direct overhead pass but applying a constant data rate; this is done in subsection 7.9.1. Afterwards, in subsection 7.9.2 the case shall be enlarged to include all passes, but still keeping the constraint of having a constant data rate throughout all passes. Subsections 7.9.3 and 7.9.4 respectively relax and remove the constraints of having a constant data rate. The former subsection describes the case where data rates can be changed *between passes* and the latter describes a scenario where data rates can be changed even *during passes*. The latter allows approaching the maximum (theoretical) possible data volume download.

7.9.1 Constant data rate, zenith pass

The most simple analysis case for a satellite pass is that of zenith pass. In this case, the pass is basically *one-dimensional*. This section shows what the optimal data rate approach would be in case of a perfect zenith, or overhead pass.

A zenith pass with a certain orbit altitude yields one contact time value. As any pass that is not directly overhead will have a lower contact time, the zenith contact time is always the maximum contact time achievable given a certain orbit altitude. Thus considering only zenith passes:

$$T_{\text{pass}} = T_{\max} \tag{7-16}$$

Given also the fact that data rate $R_{data}(t)$ is constant in the case assumed in this section, relation 7-15 can be reduced to:

$$D_{\text{pass}} = T_{\max} \cdot R_{data} \tag{7-17}$$

For simplicity, the effects of a non-spherical and rotating Earth will be ignored, as these effects are minor; the flattening of the Earth is less than 0.5% and the Earth rotation is less than 5° for a <20 minute pass.

The following relations follow from [SLR 0316]. The maximum contact time for a satellite in a circular orbit is given by:

$$T_{\max} = P\frac{\lambda_{\max}}{\pi} \tag{7-18}$$

Where P is the orbital period and $\lambda_{\max}$ is the maximum Earth central angle. The first parameter, the orbital period P, is for a circular orbit given by:

$$P = 2\pi\sqrt{\frac{(R+h)^3}{\mu}} \tag{7-19}$$

Where R is Earth radius, h equals orbit altitude and μ is the Earth gravitational parameter. The second parameter, the Earth central angle $\lambda_{\max}$ is for a circular orbit given by:

$$\lambda_{\max} = \frac{\pi}{2} - \varepsilon_{\min} - \sin^{-1}\left(\frac{R}{R+h}\cos\varepsilon_{\min}\right) \tag{7-20}$$

Where $\varepsilon_{\min}$ is the minimum elevation angle. Substituting relations 7-18 and 7-19 into 7-17 gives:

$$T_{\max} = \sqrt{\frac{(R+h)^3}{\mu}}\left(\pi - 2\varepsilon_{\min} - 2\sin^{-1}\left[\frac{R}{R+h}\cos\varepsilon_{\min}\right]\right) \tag{7-21}$$

A plot of the resulting maximum communication time is shown in Figure 7.40 below.

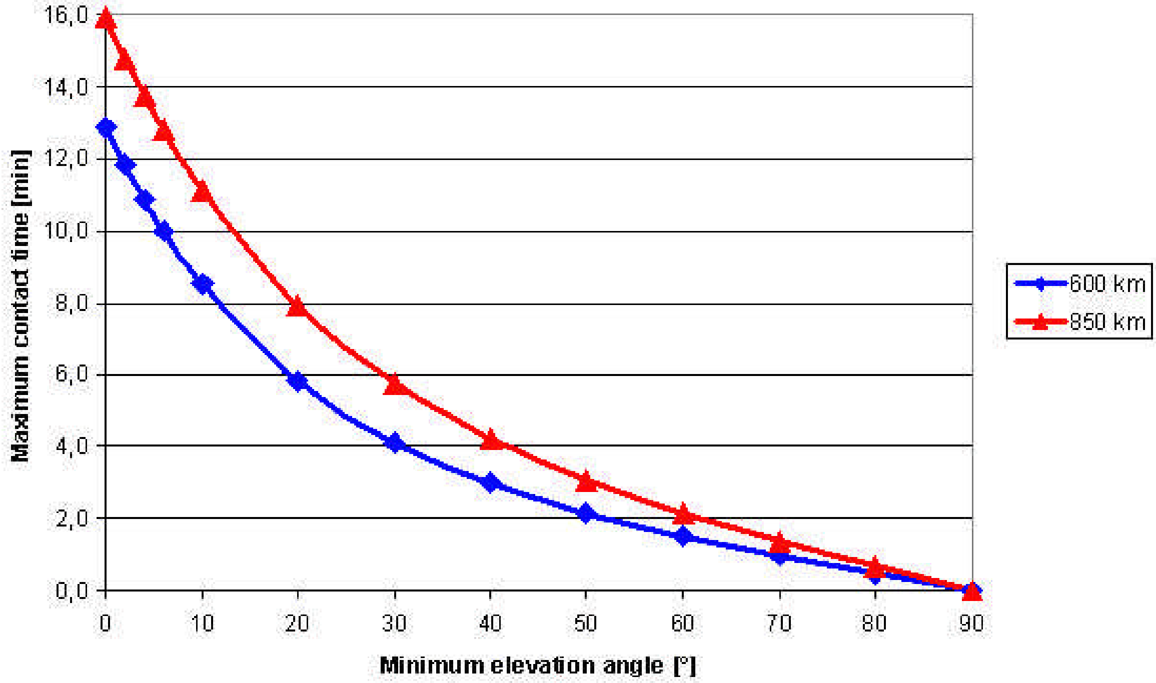


Figure 7.40: Maximum contact time versus minimum elevation angle for the Delfi-n3Xt range of orbits

The above plot of the maximum communication time and the data rate versus elevation angle plot can be combined according to relation 7-15 to yield the downloadable data volume. The resulting figure is shown in Figure 7.41 below. The downloadable data volume is normalized with respect to the highest value.

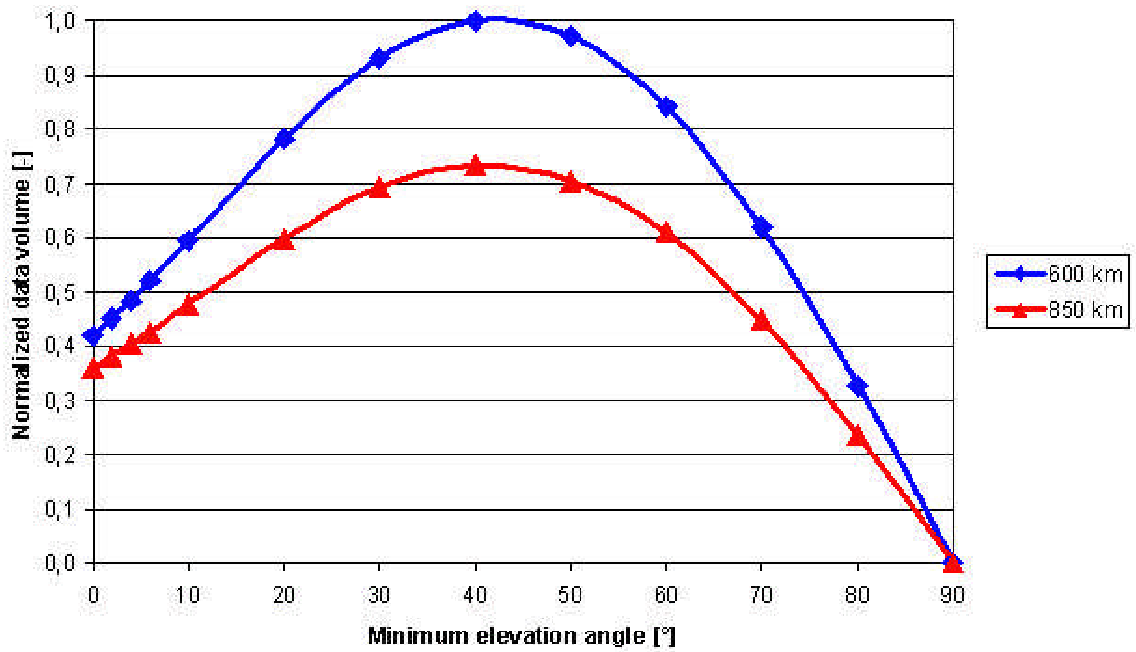


Figure 7.41: Normalized downloadable data volume versus elevation angle for the Delfi-n3Xt range of orbits

The latter figure yields an interesting conclusion. The figure shows that there is a minimum elevation angle and respective constant data rate for which the total data volume per pass is maximal. The minimum elevation angle for which the communication link is then designed to yield a maximum data volume download will be called the ***optimum elevation angle***, or ε_{opt}.

The optimum minimum elevation angle shown by Figure 7.41 is clearly larger than the 5° or 10° of minimum elevation that is usually assumed for satellite communication.

[SLR 0739] continues on this conclusion, and better specifies the effects of altitude on the optimum elevation angle. It presents the following relation between the optimal elevation angle ε_{opt} and orbit altitude h:

$$\varepsilon_{opt} \approx 45[°] - 0.00446[°] \cdot h[km] \quad (7\text{-}22)$$

For Delfi-n3Xt, the orbit altitude range defined to be of interest is 600-850 km. Using relation 7-22 this gives optimal elevation angles of 42.5° and 41.2° for the lower and higher boundary value respectively.

Of course, the scenario described is for a one-dimensional perfect overhead pass, which, in fact almost never occurs. Therefore, the following three sections will look at two-dimensional cases, including next to the along-track direction also an off-track direction.

7.9.2 Constant data rate, all in-range passes

A satellite in a sun-synchronous orbit will in fact pass over completely different Earth locations in a time period of days or even weeks. Whereas one of these passes might be or might be close to a zenith pass, most will be more 'to the side'. The next section will introduce an *off-zenith* angle to describe this angular distance.

When the communication link is designed for a certain minimum elevation angle, the visibility zone of the ground station (GS) is determined. While a higher minimum elevation angle will decrease overall along-track pass time, it will also decrease off-track field-of-view. This then restricts the amount of satellite passes that can be seen by the GS.

The difference between the number of daily passes that can consequently be 'seen' for high and low minimum elevation angles is drastic. This is illustrated by Figure 7.42 below, generated using Satellite Tool Kit or STK. Whereas 5-6 passes daily are within range in case of a 10° minimum elevation angle, only 1-2 passes are seen in case of a 42.5° minimum elevation angle. The latter would be the optimum elevation angle for maximized downloadable data volume, in case of a 600 km zenith pass, as shown in the last section.

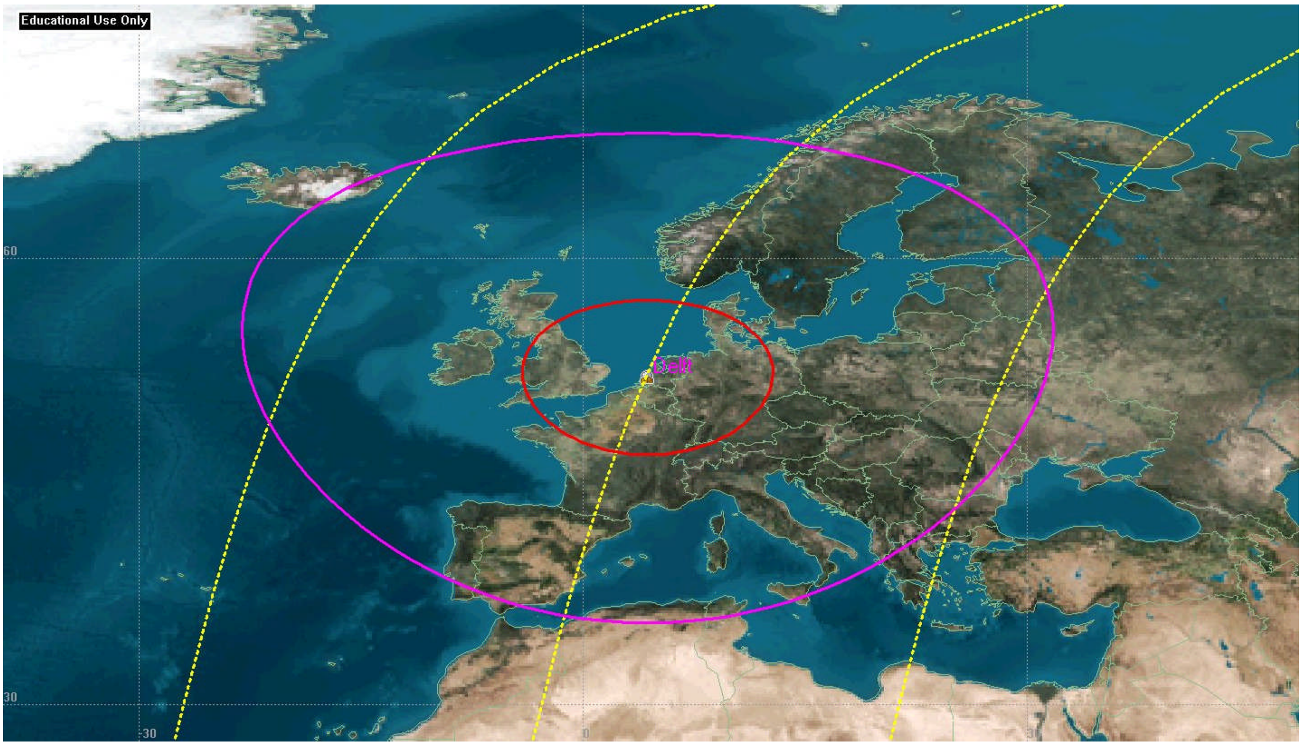


Figure 7.42: 3 consequent passes projected on the Earth surface with GS visibility shown

The figure above clearly shows the much greater visibility zone in case a 10° elevation angle is designed for, compared to the 42.5° case; both are indicated by the colored oval shapes, purple and red respectively. Directly obvious is the very sizeable difference in area enclosed by the two shapes.

The figure also shows three consequent passes for a typical sun-synchronous orbit. Whereas one pass is directly overhead of the Delft GS in this case, the other two miss the small visibility circle completely, but are still in range in case of the larger visibility zone. Concluding from the spacing between the passes, all passes can actually miss the small visibility circle.

But, as the satellite has a consequent number of passes on each 'side' of the globe (once top-to-bottom, and once bottom-to-top), the satellite will always have at least one pass inside of the 42.5° circle per day. This can be during nighttime depending on the orbit chosen. The 10° circle will contain at least 5 passes daily, consistent with the value given above.

Whereas the previous section has concluded that path loss quickly decreases with increasing elevation angles ε_{min}, it can thus be seen that total visibility decreases quickly with increasing ε_{min} as well. This factor will more than counteract for the decreasing path loss, as is shown below.

If one takes into account all passes within a visibility zone, the contact time per pass T_{pass} can be replaced by an overall coverage value, $T_{coverage}$. The latter is defined as the *time period that the satellite is within proper communication range of the GS.* Again assuming a constant $R_{data}(t)$, allows to simply equation 7-15 to:

$$D_{\text{pass}} = T_{coverage} \cdot R_{data} \tag{7-23}$$

It is very difficult to obtain analytical values for the total coverage of a day, or the average coverage during a day. STK however can be used to friendly yield numerical approximations of the coverage time, or more precisely the coverage percentage; even taking into account Earth shape and rotation. This coverage can then be multiplied by the length of a day to give a daily data yield. Generating coverage percentages for all elevation angles, for both a 600 and 850 km orbit yields the following plot, see Figure 7.43. These values are overall averages for a sun-synchronous orbit with the given orbit altitude, taken over a two year period.

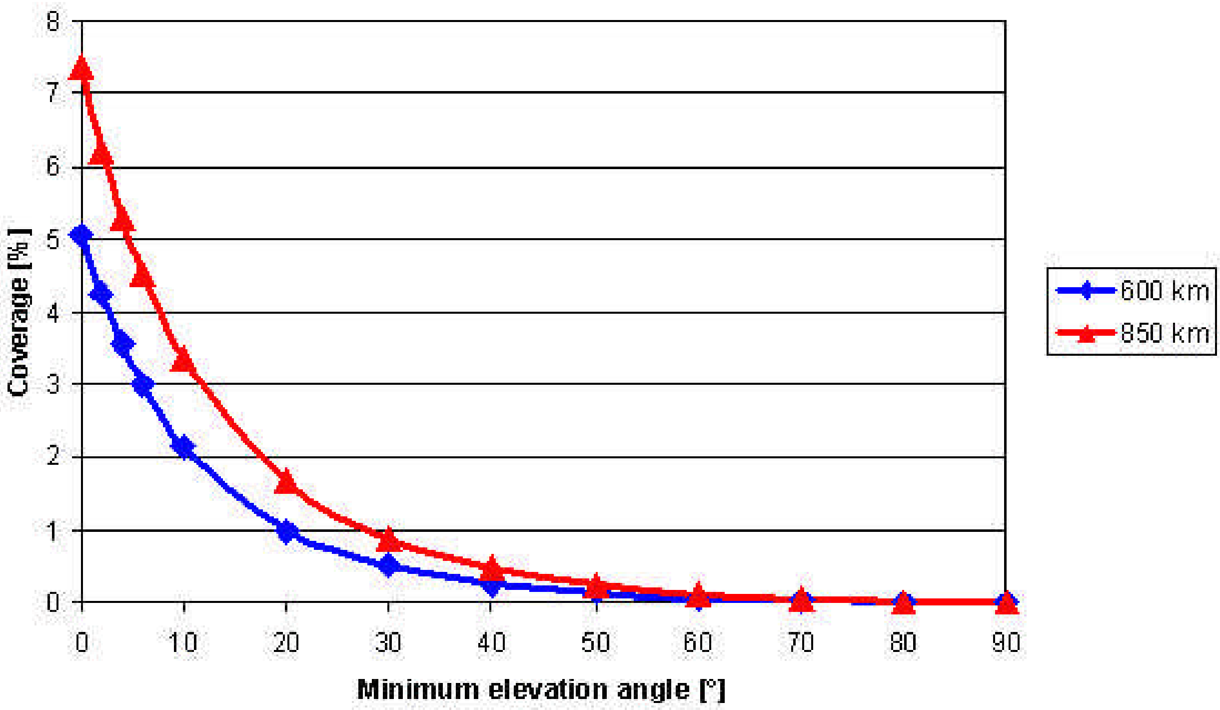


Figure 7.43: Overall ground station coverage versus elevation angle for which the link budget closes

The above figure should be compared to Figure 7.40; the latter showing only some factor 4 difference in maximum contact time between a 0° and 40° elevation angle case. If however the total coverage is taken into account, as done in Figure 7.43, a factor 15-20 difference exists between the two cases depending on the orbit altitude. Figure 7.43 and Figure 7.39 can again be combined according to relation 7-22 to yield the achievable downloaded data volume given a minimum elevation angle, as is shown in Figure 7.44 below.

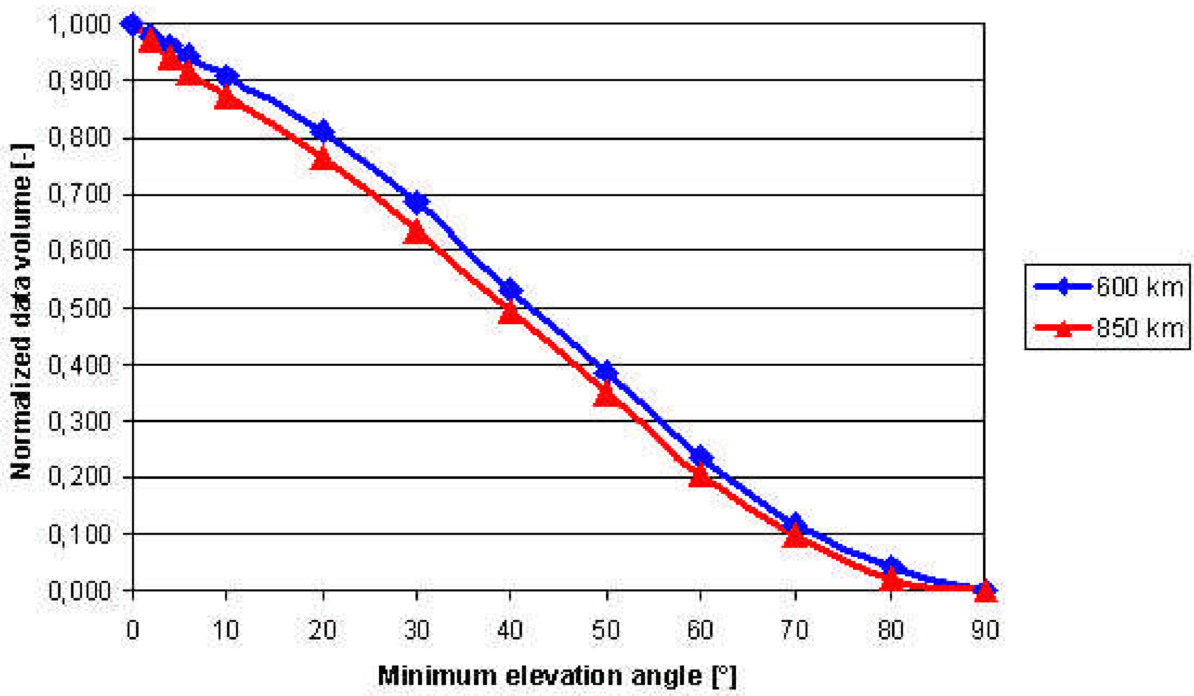


Figure 7.44: Normalized achievable data volume versus elevation angle for the Delfi-n3Xt range of orbits

This figure clearly shows that if a constant data rate is to be used *throughout the mission*, and a satellite can be tracked during every pass, it merits doing so with a link budget that closes for elevation angles as low as possible. *There is no minimum elevation angle that performs better than another, except for those being lower*. In other words, a lower minimum elevation angle is *always* better, *if* all passes can be used for communication. However, as was already presented before, at very low elevation angles atmosphere loss increases drastically, antenna noise rises, and obstructions can be in place. As such, a value of 5° to 10° should then be used in the link budget.

Finally, it is interesting to note that taking into account coverage minimizes the differences in downloadable data volume between 600 and 850 km orbit altitudes, only a few percent difference remains whereas for a single zenith pass the difference is a factor 1.4.

The number of passes related to ground station visibility was in fact not previously taken into account when the theory was investigated by Martijn de Milliano. It was argued that an average value for the pass time could be used, one that is presented in [SLR 0316]. Whereas the latter source does not give any comments on the relation to orbital altitude and visibility, it mentions that the average pass within the field-of-view of a GS is approximately 80% of the maximum pass time. Using this value to correct for the fact that off-zenith angles are in fact incurred in actual passes, neglects to take into account the fact that the visibility of the ground station becomes much less, and thus much less passes are seen.

7.9.3 Changing data rates between passes

The previous section has shown that given the constraint of a constant data rate throughout a mission, the communication link can best me designed based on a near-zero minimum elevation angle. However, if the constraint of having a constant data rate is relaxed or even removed, higher data rates can be useful. This and the next section investigate these options. In order to provide limited complexity, this section describes the case where data rates can be changed *only between passes*.

A zenith pass of a satellite for an orbit altitude between 600 and 850 km leads to an optimum minimum elevation angle ε_{opt} of 41.2° to 42.5°. Logically, this minimum elevation angle goes down for passes with an introduced off-zenith angle; the latter causes the pass time to be reduced and the maximum achievable data rate to be lowered. The exact impact is calculated and shown in this subsection. As such, *the optimal minimum elevation angle shall be presented as a function of the off-zenith angle θ, to be defined below*. Again, Earth rotation and flattening shall be ignored.

In order to calculate the effects of an introduced off-zenith angle, only the reduced contact time has to be taken into account. To further clarify, one can look at Figure 7.42. During a pass a certain data rate is allowed, specified by the minimum elevation angle that is designed for. This minimum elevation angle in turn specifies a visibility circle. When a pass occurs that is not a zenith pass, this circle radius does not change, but the satellite passes more to the right or left of the GS. In result, only its *time within in visibility circle reduces*.

As $R_{data}(t)$ is again constant in this scenario, equation 7-15 can again be used and simplified to:

$$D_{\text{pass}} = T_{\text{pass}} \cdot R_{data} \tag{7-24}$$

T_{pass} then remains to be specified, and can be written according to a more general form of equation 7.180 as presented in [SLR 0316]:

$$T_{pass} = \frac{P}{\pi} \cdot \cos^{-1}\left(\frac{\cos(\lambda_{\max})}{\cos(\lambda_{\min})}\right) \tag{7-25}$$

Where $\lambda_{\min}$ is the minimum Earth central angle, and could be set to zero in case of a zenith pass. This angle specifies in fact the off-zenith angle, but *with respect to the center of the Earth*. Therefore, in order to make the above relation useful, the minimum Earth central angle should be expressed in terms of the *off-zenith angle*, being the angle between the line perpendicular to the Earth surface at the location of the GS, and the line between satellite and GS at minimum pass distance. The relation between the two can be determined using Figure 7.45 below. Also, the off-zenith angle θ can more conventionally be described as:

$$\theta = 90° - \varepsilon_{\max} \tag{7-26}$$

Figure 7.45 shows the scenario of a satellite pass at the point of minimum pass distance. In this case, the Earth central angle λ becomes the minimum Earth central angle $\lambda_{\min}$, and the angle θ can be seen to indicate the off-zenith angle. These two angles are in the same plane. The maximum elevation angle $\varepsilon_{\max}$ has also been indicated in the figure.

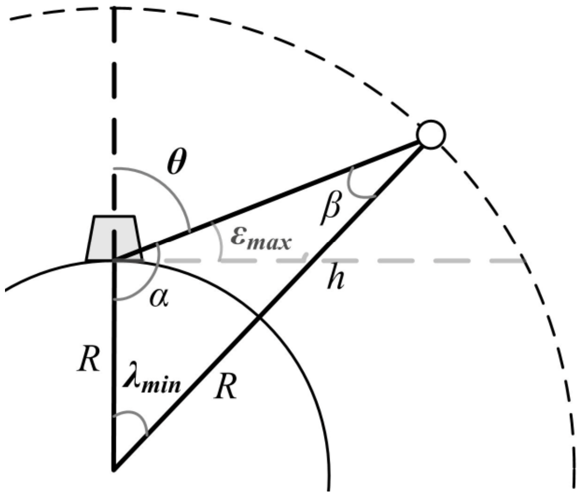


Figure 7.45: Pass geometry showing the relation between off-track angle and Earth central angle

A number of relations can be stated using geometry:

Applying the sine rule: $$\frac{(R+h)}{\sin\alpha}=\frac{R}{\sin\beta} \quad\Rightarrow\quad \beta=\sin^{-1}\left(\frac{R\cdot\sin\alpha}{R+h}\right) \tag{7-27}$$

Property of a triangle: $$\pi=\lambda_{\min}+\alpha+\beta \quad\Rightarrow\quad \lambda_{\min}=\pi-\alpha-\beta \tag{7-28}$$

Property of a line: $$\pi=\theta+\alpha \quad\Rightarrow\quad \alpha=\pi-\theta \tag{7-29}$$

Substituting equation 7-29 into 7-27, and 7-27 and 7-29 into 7-28, as well as realizing that $\sin(\pi-\theta)=-\sin\theta$ yields for the minimum Earth central angle:

$$\lambda_{\min}=\theta+\sin^{-1}\left(\frac{R\cdot\sin\theta}{R+h}\right) \tag{7-30}$$

$\lambda_{\max}$ has already been expressed in terms of the minimum elevation angle $\varepsilon_{\min}$ in relation 7-19, and P has been given by relation 7-20, so that T_{pass} can consequently be expressed directly in terms of variables off-zenith angle θ, minimum elevation angle $\varepsilon_{\min}$ and orbit altitude h:

$$T_{pass}=2\sqrt{\frac{(R+h)^3}{\mu}}\cdot\cos^{-1}\left[\frac{\cos\left\{\frac{\pi}{2}-\varepsilon_{\min}-\sin^{-1}\left(\frac{R}{R+h}\cos\varepsilon_{\min}\right)\right\}}{\cos\left\{o-\sin^{-1}\left(\frac{R\cdot\sin(\pi-o)}{R+h}\right)\right\}}\right] \tag{7-31}$$

The results of equation 7-31 are visualized in different manners below. The overall 3D plot including both the variables off-zenith angle θ and minimum elevation angle $\varepsilon_{\min}$, but with h constrained to be 600 km is shown as Figure 7.46 below.

In Figure 7.46 the case of a zenith pass, thus when θ = 0, can clearly be recognized against the 'back wall', being identical to Figure 7.41 above; the same bell shape is present. Moving along the axis of off-zenith angles, the bell shape is being flattened, and its peak lowered.

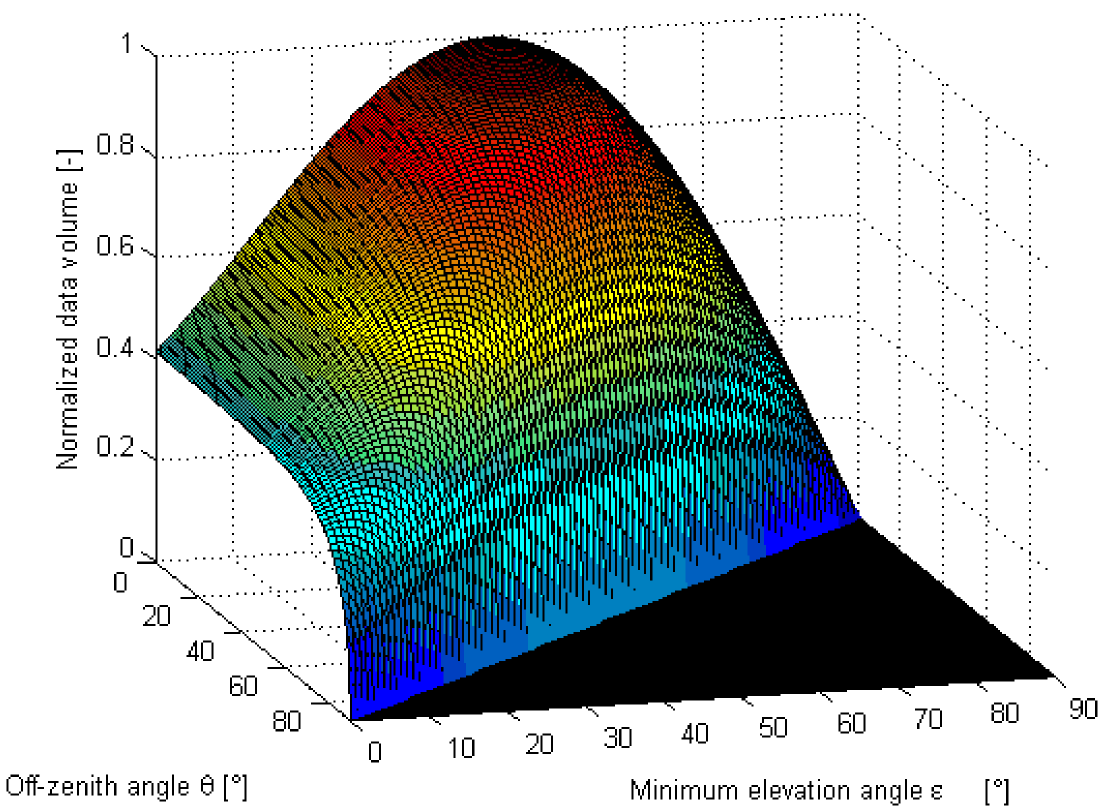


Figure 7.46: Normalized data volume vs. off-track angle and minimum elevation angle, *h* = 600 km

The behaviour of the 3D plot in Figure 7.46 is better demonstrated by Figure 7.47 and Figure 7.48 below.

Figure 7.47 shows a contour plot of Figure 7.46. It can be clearly seen that off-zenith angle θ plus minimum elevation angle $\varepsilon_{\min}$ can never be more than 90°; a 90° off-zenith pass can only be seen at the edge of a 0° elevation GS visibility circle.

Figure 7.48 better demonstrates the effect of an introduced off-zenith angle on downloadable data volume; in other words the decrease of size of the bell shape. This curve can again be compared to Figure 7.41, of which the 600 km case is actually included in the figure below as the red (top) line. Again, orbit altitude is set at 600 km; this plot demonstrates behavior and not absolute values.

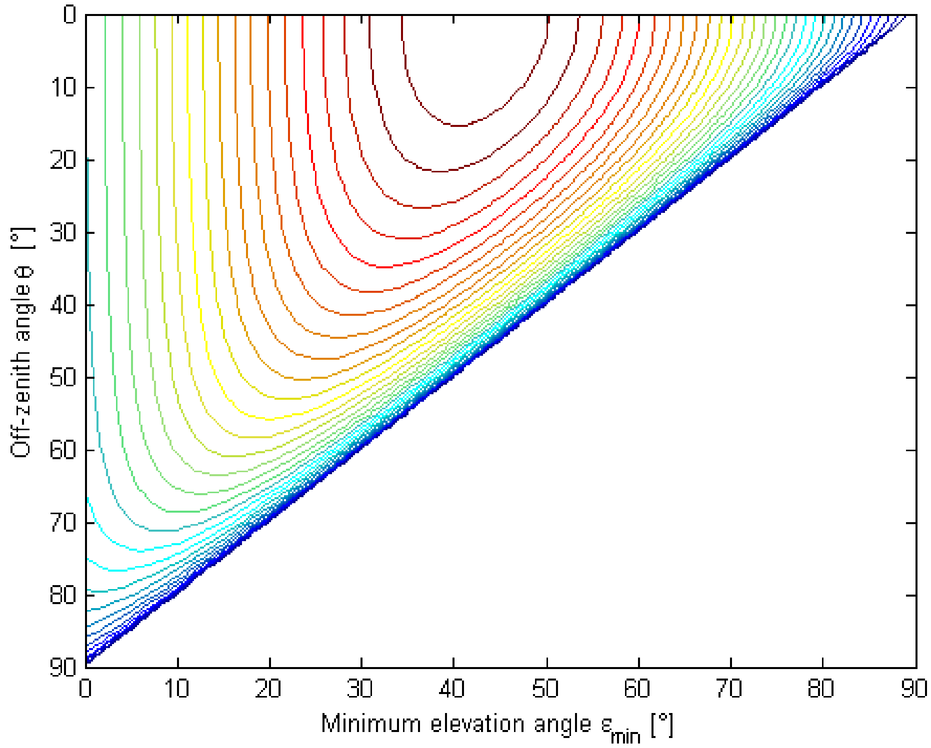


Figure 7.47: Normalized data volume vs. off-track angle and minimum elevation angle, *h* = 600 km

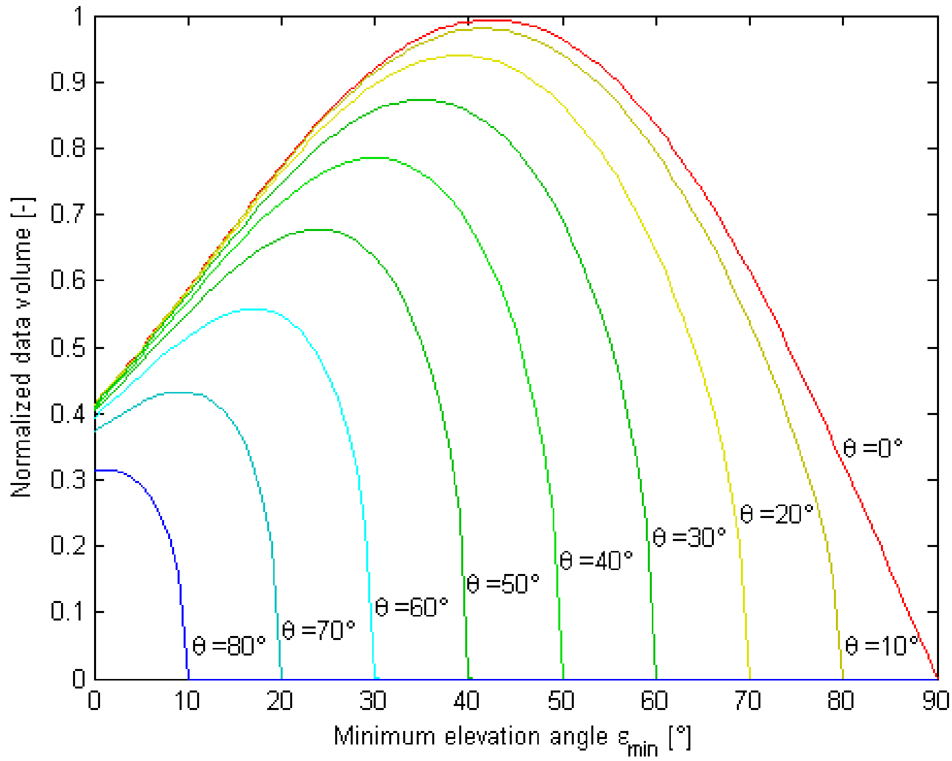


Figure 7.48: Normalized data volume vs. minimum elevation angle for different off-track angles, *h* = 600 km

In Figure 7.48 above, as well as in the contour and 3D plots it can be seen clearly that ε_{opt} decreases for higher values of θ. The exact relation between optimum elevation angle ε_{opt} and off-zenith angle θ is shown in Figure 7.49 below, for the possible range of Delfi-n3Xt orbit altitudes.

In fact, the curve produced is rather simple, with an optimum elevation angle decreasing with time in a slightly accelerating fashion. Also, in off-zenith angles up to 20°, ε_{opt} changes only by some 5°. It is interesting also to note that at some point, before 90° off-zenith angles are reached, the optimum elevation angle already becomes 0°; the higher the orbit the sooner it does.

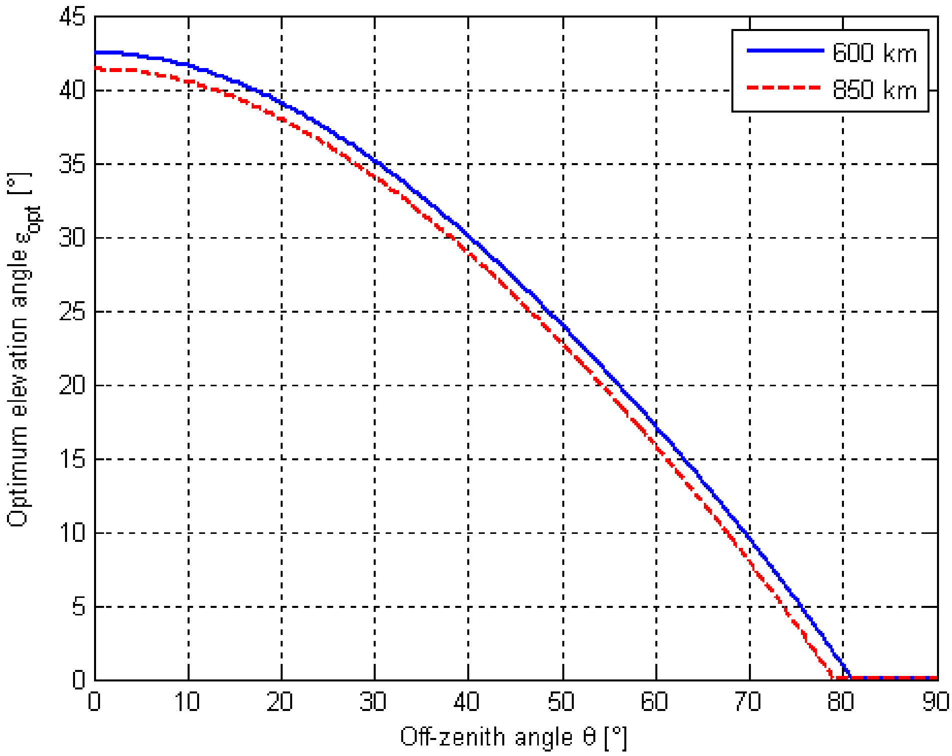


Figure 7.49: Optimum elevation angle vs. off-track angle for the Delfi-n3Xt range of orbits

In section 7.9.1 an approximization relation (equation 7-22) was presented for the optimum elevation angle ε_{opt}, given the orbit altitude h. This relation was appealing because it turns out to be a linear one. In the case off-zenith angles are considered, a similar approximation can be made but unfortunately not a linear one. However, it turns out that given orbit altitudes above a certain value ($\sim h = 200[km]$), the only approximate effect of a different altitude is a downwards translation of all values of ε_{opt} versus θ; again linearly. Therefore, it then remains to find an approximization relation for the above plot of ε_{opt} versus θ. In fact, it can be approximated relatively well by a third-degree polynomial.

The resulting approximization relation can be given by:

$$\varepsilon_{opt} \approx 3.3\cdot10^{-5}\cdot\theta^3[°]-8.9\cdot10^{-3}\cdot\theta^2[°]-1.25\cdot10^{-2}\cdot\theta[°]-0.00446\cdot h[km] \qquad (7\text{-}32)$$

Equation 7-32 is only valid for values of ε_{opt} above 0°, for negative results the optimum is logically 0°. It gives an accuracy of ~1° for useful orbit altitudes *h* above ~200 km. Orbit altitudes above ~10,000 km no longer have a nonzero optimum elevation angle.

The accuracy of the above equation is further illustrated by the following Figure 7.50.

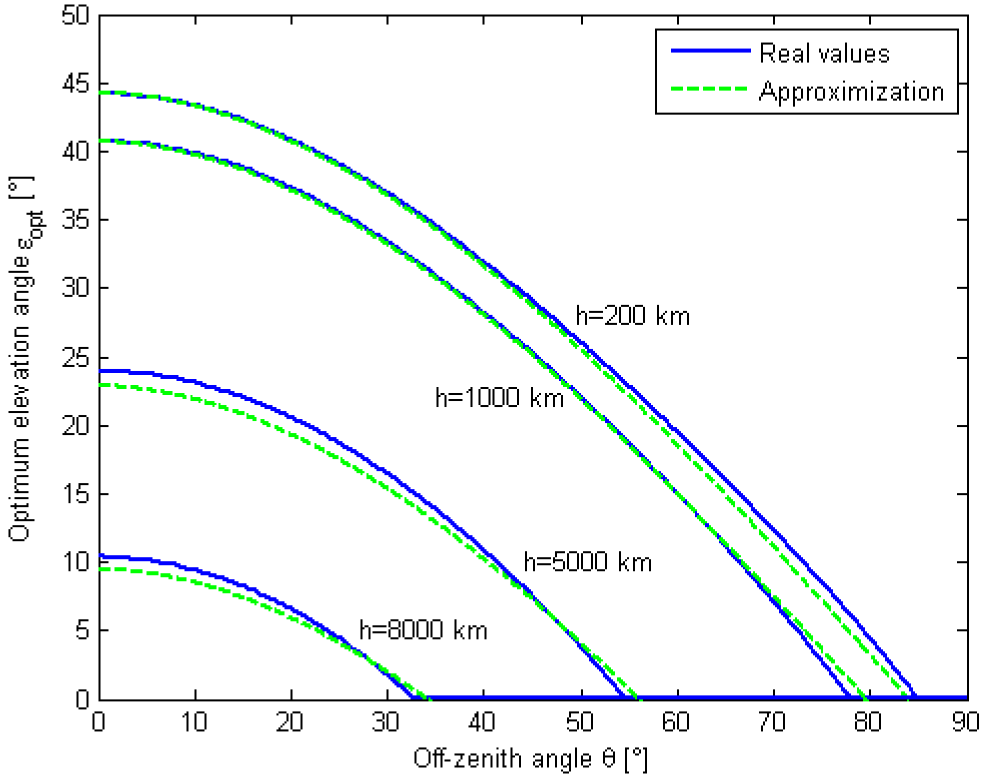


Figure 7.50: Illustration of the accuracy of the approximization relation for ε_{opt}

The following two figures better demonstrate the data volumes that can be obtained given certain off-zenith angles θ.

Figure 7.51 shows total data volumes that can be obtained given a certain off-zenith angle, if the minimum elevation angle $\varepsilon_{\min}$ *is allowed to be changed* between passes to match the optimum elevation angle ε_{opt} for that off-zenith angle. The plot is normalized to the maximum downloadable data volume in case of a zenith pass of 600 km altitude, being the minimum Delfi-n3Xt altitude.

Figure 7.52 on the contrary, shows the (normalized) maximum downloadable data volume versus off-track angle for a 600 km orbit when $\varepsilon_{\min}$ *is not allowed to be changed* and is thus kept constant, independent of off-zenith angle. For reasons of clarity, the subscript of $\varepsilon_{\min}$ has been dropped in this figure.

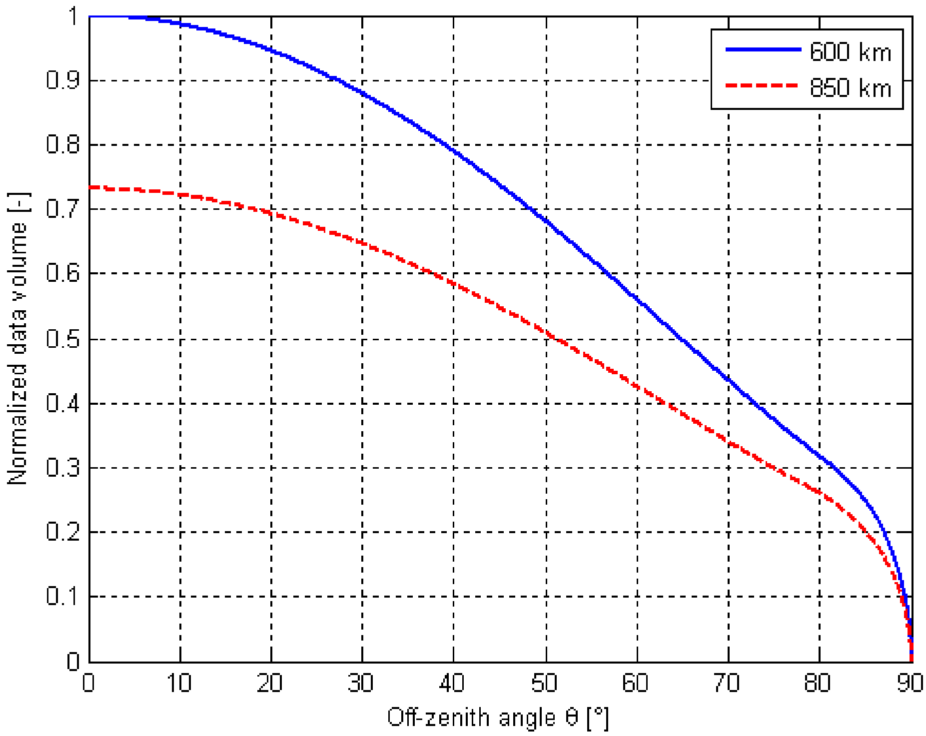


Figure 7.51: Normalized and optimized data volume vs. off-track angle for the Delfi-n3Xt range of orbits

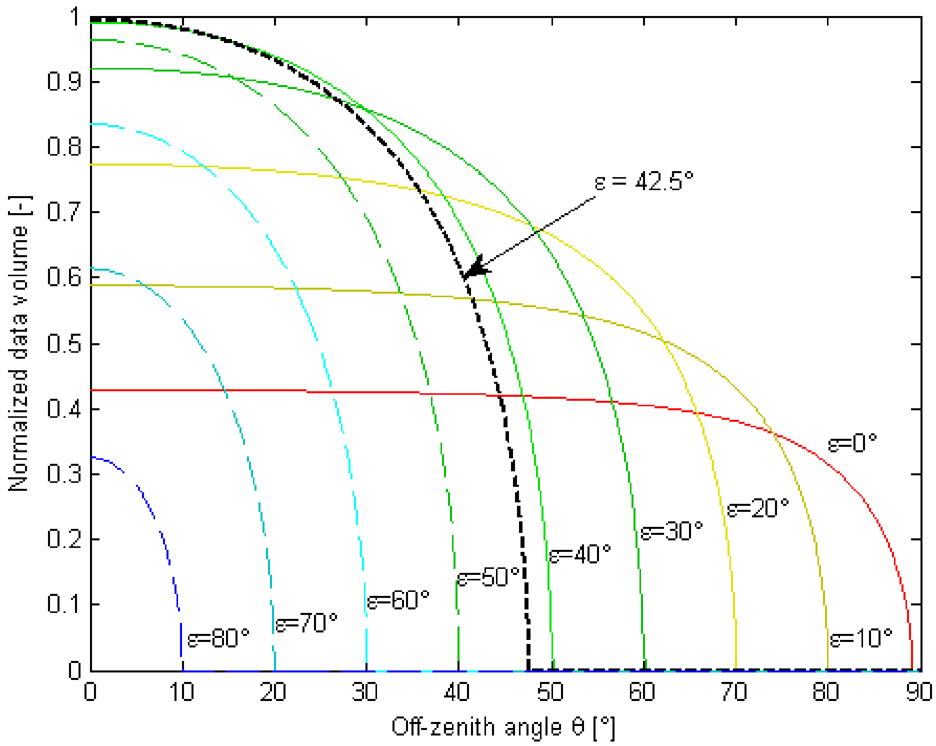


Figure 7.52: Normalized data volume vs. off-track angle for the Delfi-n3Xt range of orbits ($\varepsilon = \varepsilon_{\min}$)

The plots in Figure 7.51 and Figure 7.52 show the difference in performance between choosing a fixed data rate throughout the mission and what can be achieved using an optimized minimum elevation angle in the link budget design. The line indicated by $\varepsilon_{\min}$ = 0° in the bottom figure gives achievable download volumes for a *conventional constant data rate approach* based on a (very) low elevation. At low zenith angles, it can be seen to be outperformed by a link budget design based on ε_{opt}, by a factor of *more than 2*.

However, the red line also shows a constant, positive data volume value over a large range of values. Whereas the optimum elevation angle of 42.5° yields the highest possible data yield, its productivity drops off quickly for increasing off-zenith angles. This clearly shows why a lower data rate is favored if, and only if, a constant data rate is used during the entire mission.

Finally, Figure 7.52 also shows again that minimum elevation angles higher than the value of 42.5° should never be designed for; the curves belonging to these higher elevation angles are completely dominated by the 42.5° curve. All points of maximum performances for the curves shown in Figure 7.52 in turn make up the blue plot shown in Figure 7.51, belonging to an orbit altitude of 600 km.

Possible applications

The presented theory in this subsection and the last has shown a preference for a certain momentarily fixed data rate, depending on the mission constraints. It had already been shown that if in fact *all passes* can be monitored, and a constant transmission data rate is used during the mission, the best approach would be to design this data rate for the link budget to close at very low elevation angles. However, the application of a data rate higher than this minimum value becomes interesting in case of a:

- **Constant mission data rate**, but a smaller number of monitored passes
- **Changing data rate**, but using only a limited number of data rate values
- **Changing and truly variable data rate**, but constant during a pass

Constant mission data rate

The first option basically becomes useful in case of:
- Power or pointing constraints on-board of the satellite
- Limited operational resources on-ground (man-power or costly equipment)

The first argument indicates that if only a small number of transmissions is supported by the power system on-board of the satellite or the pointing system, it is more beneficial to use a few close passes only, with a higher data rate than would be possible otherwise. The second argument states that if ground station operations are or tracking is expensive, or simply not possible due to shared schedules, it might be beneficial again to focus on a small number of passes, with a higher data rate.

If this approach would be chosen, a number of passes should be determined that is to be monitored, and based on this a maximum overall data rate can be chosen. As an incurred off-zenith angle will specify a optimal elevation angle ε_{opt} that is less than that would hold for a zenith pass, the data rate that should be chosen should also be lower than that advised for a zenith pass. The specific value will depend on the number of passes and maximum pass off-zenith angle θ.

An added benefit to choosing a (constant) data rate higher than the minimum data rate possible is that the link budget is generally based on worst-case values. As a result, the actual contact time might turn out to be higher, resulting in a higher actual downloadable data volume during the pass. In case of a lower minimum elevation angle, a lower 'unexpected' gain in data volume can be obtained due to this over-budgeting and the impact is lower or even non-existent if the horizon is obstructed.

Disadvantages or challenges that have to be taken into account when increasing the data rate at the cost of less communication time are:

- The available time becomes more valuable due to the higher data rates:
  - Signal acquisition should be accounted for
  - Transmission disturbances during the pass have a larger effect
- A higher data rate increases transmitter and receiver complexity, as well as required transmission bandwidth
- In case of Delfi-n3Xt, radio amateurs with lower antenna gain might not be able to receive a signals as the mid-pass link margin is reduced

Changing data rates

The second and third option of *changing data rates* are more desirable, and in all cases win over the conventional approach of having a constant data rate over the entire mission based on a low minimum elevation angle. They do however require a certain increase of complexity, necessitating:

- Accurate on-board orbit determination or accurate time-scheduling using telecommands
- A transmitter with a variable transmission data rate, including proper hardware (i.e. filters, amplifiers) and software (data protocol, internal communication)
- Varying data generation from the satellite bus, or data storage in the transmitter and intelligent software
- A receiver capable of receiving different data rates, possibly through auto-detection of the data rate

An interesting means of switching on the transmitter autonomously can be done by means of transmitted Two-Line Elements or TLEs; the satellite OBC can base its transmission times on TLEs that are updated daily or even weekly be means of telecommands. The satellite then does not need to have active orbit determination itself; this requires only simplified algorithms.

The difference between the second and third aforementioned options is that in the second case the complexity is limited by using a restricted number of discrete data rates. This can simplify for example internal operational modes and data generation, data packaging and filtering.

In case the *second option* is applied, concrete steps should be formulated in terms of maximum off-zenith angles. Afterwards Figure 7.49 or relation 7-34 can be used to determine the optimum elevation angle that should be used to calculate the data rate per step (keeping in mind a certain *inaccurate* link margin in the link budget). Due to the presented reasons of coverage and the resulting increased number of passes, the data rate should then always be designed to match the *maximum off-zenith angle of the pass zone*. This could result in for example a data rate presets based on an θ = 20, 50 and 80°, resulting in $\varepsilon_{\min}$ values of approximately 40, 25, 10°.

The *third option* on the contrary could in the best case result in the maximum performance as shown in Figure 7.51. In this case, the data rate should basically be completely variable. An added benefit of continuously being able to adjust the data rate of transmission is that any errors (margins) in the link budget can be corrected for, and overall performance can be maximized.

7.9.4 Changing data rates during a pass

The previous sections have shown that given the constraint of having a constant data rate during the mission or even between passes, leads to a preferred data rate if a maximum downloadable data volume is to be designed for. Of course, the maximum downloadable data volumes can be achieved if data rates are not only changed between passes, but in fact also *during passes*.

In this case, the data rates can be chosen to follow the evolution of path loss during a pass, as shown by Figure 7.39, and the limiting factor becomes the complexity involved with changing data rates during a pass. The following figure nicely illustrates the downloadable data volumes achievable when constant data rates are used, but also illustrates the effects if several *discrete steps* are used.

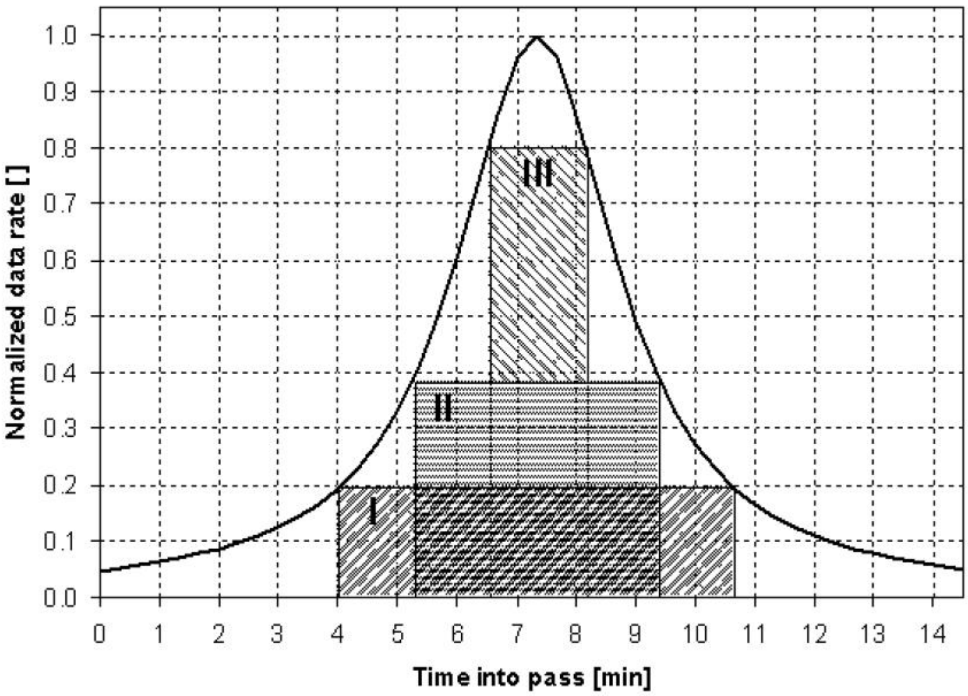


Figure 7.53: Normalized data rate vs. time into pass with data volumes (I, II & III) [SLR 0739]

Figure 7.53 demonstrates the data volumes that can be downloaded when transmission is started at different (pass) times, using again an optimized data rate (given the path loss) at the time of start of transmission. Cases I and III yield equal data volumes, whereas II yields a maximized data volume for a direct overhead pass. If however the data rates can be changed following discrete steps, a data volume equal to the area covered by I, II and III together can be reached. In this manner, the maximum theoretical downloadable data volume can be approached, which in turn would require *an infinite amount of possible steps*.

Expression 7-15 can be rewritten to signify this maximum downloadable data volume $D_{\max}$ for a given off-zenith angle θ:

$$D_{\max} = \int_0^{T_{\max}} R_{\max}(t,\theta) \cdot dt \qquad (7\text{-}33)$$

Equation 7-31 can be used to yield $T_{\max}$ given $\varepsilon_{\min} = 0°$. However, equation 7-14 only specifies $R_{\max}$ as a function of ε. As such, a relation for $R_{\max}(t,\theta)$ would be required, or a relation for $\varepsilon(t,\theta)$ to multiply $R_{\max}(\varepsilon)$ with, and the resulting equation should be integrated. As such, a more simple numerical solution is proposed by writing equation 7-33 as a sum:

$$D_{\max} = 2 \cdot \sum_{\varepsilon=0}^{90} \left[R_{data}(\varepsilon,\theta) \cdot \Delta t \right] \tag{7-35}$$

This sum can be evaluated for any $\theta \in (0,90)$:

$$D_{\max} = 2 \cdot \sum_{i=0}^{89} \left[R_{\max}(\varepsilon = i+0.5, \theta = x) \cdot 0.5 \cdot \left\{ T_{pass}(\varepsilon = i, \theta = x) - T_{pass}(\varepsilon = i+1, \theta = x) \right\} \right] \tag{7-36}$$

Using as mentioned above relations 7-31 to obtain T_{pass} and 7-14 to obtain $R_{\max}$.

The precision of the above relation can of course be improved by increasing the step size. The resulting plot of the maximum theoretical downloadable data volume per off-zenith angle θ is given in Figure 7.54 below. Also, the case where a data rate is optimally matched to the off-zenith angle is shown, as well as the more conventional approach of having a constant data rate.

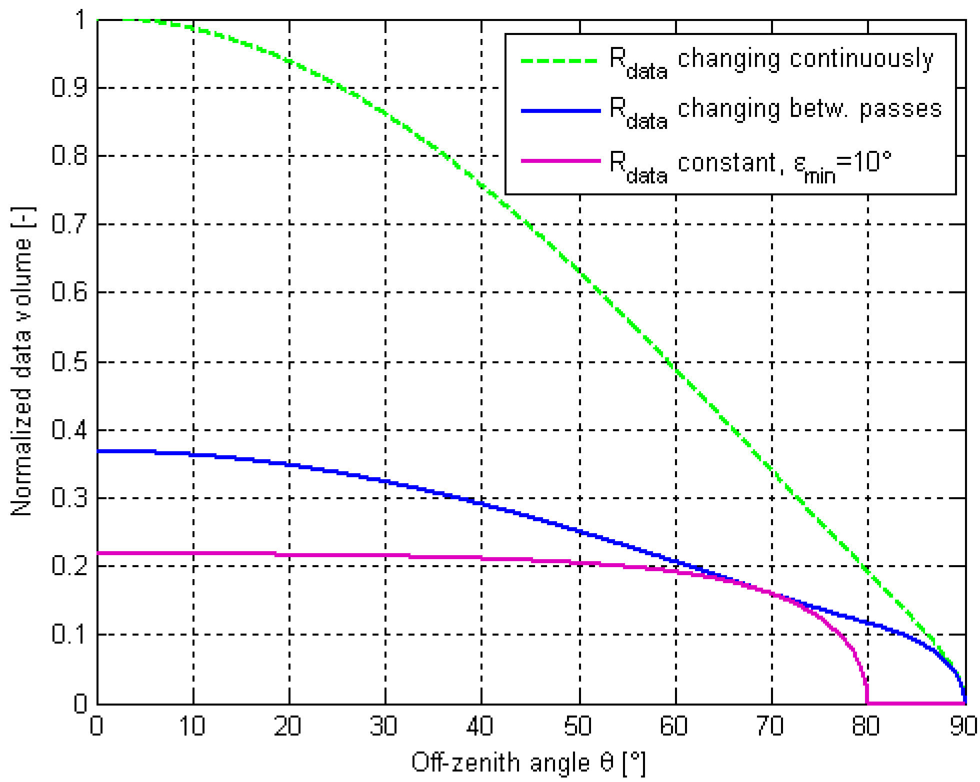


Figure 7.54: Maximum downloadable data volume versus off-zenith angle, plotted with two other scenarios

In the above figure it can clearly be seen that where large gains in data volume can be achieved by changing the data rate between passes, more significant gains can be achieved still; but requiring a more advanced scenario where data rates are changed, if possible almost continuously, during a pass.

Implementation complexity:

Of course, having a data rate that changes even between passes is constrained by complexity. All complexity issues introduced in the previous subsection still hold, but become even stricter:

- On-board orbit determination or time-scheduling should be accurate enough for pass information in the order of seconds, possible tenths of seconds
- Data preparation and resulting data rates should be adjusted almost instantaneously
- Auto-detection of the data rates becomes a near-requirement, or a complicated (failure-sensitive) feedback process using the uplink should be incorporated. Auto-detection of the data rate is already challenging, but can be incorporated most notably in software; its complexity however increasing vastly with each possible data rate.

But in fact, the differences in complexity between a truly variable data rate between passes and a data rate that can switch several times during a pass can be said to be limited. As a matter of fact, a truly variable data rate between passes might be more complex than one that changes along several discrete steps. These discrete steps than may be predefined data rates, such as those presented in the last subsection, or the steps might be based on the actual pass and its off-zenith angle θ.

In terms of performance, a change of data rate will likely need some 'digital hand-shake' procedure in order to establish the new data rate, which will involve telecommanding. However, the lost time is quickly made up for by the increased data rate afterwards. The resulting gain in performance can, as shown by Figure 7.54, be enormous.

8 The UHF/VHF communication link

Using the theoretical background presented in the previous chapter, this chapter specifies the Delfi-n3Xt communication down/uplink over the UHF and VHF bands. Afterwards the radios making use of this link, the PTRX and ITRX are further introduced. A later chapter will similarly establish the S-band communication link.

The *primary use* of both the PTRX and ITRX is nominal up- and downlink communication; sending telecommands and receiving telemetry. *Secondary uses* of the UHF and VHF bands are the transponder up- and downlink, the OLFAR downlink, and strictly speaking the ITRX experiments. The transponder and OLFAR are analogue functionalities, and no signal content processing is performed. The ITRX experiments are not of our concern, given the payload status. These functionalities are therefore *not further discussed* in this chapter. The transponder/OLFAR bandwidth however is shortly presented.

This chapter introduces the functionalities required to specify the primary telecommand/telemetry link divided over the different layers.

The primary communication requires both first and second layer functionalities, as introduced in the previous chapter. Multiple *scenarios* can also be recognized, such as transmitting TCs or TM, these can be said to form another layer. All required *high-level* functionalities are determined in this chapter.

The **boundary conditions** of the link are established first in sections 8.1 and 8.2, presenting:

- The transmission frequencies and bandwidth available
- The power constraints following from the link budget

Afterwards, the **physical layer functionalities** are determined in sections 8.3 and 8.4:

- The modulation scheme of the VHF downlink
- The modulation scheme of the UHF uplink

Finally, all **data link layer functionalities** are presented in sections 8.5 and 8.6:

- The data protocol applied to both the UHF and VHF communication links
- The operations and data involved with the different *scenarios* as introduced above

8.1 Transmission frequencies and bandwidth

In case of the (double) primary radio communication system, consisting of PTRX and ITRX, transmission shall take place within the UHF and VHF bands. More specifically, the 144-148 MHz VHF radio amateur band is to be used for downlink, whereas the 420-450 MHz UHF radio amateur band is to be used for uplink signal transmission. As Delfi-n3Xt will have two transceivers in these bands, the PTRX and the ITRX, in total there are *4 transmission frequencies* required for Delfi-n3Xt. Also, an *up- and downlink bandwidth* should be available for the transponder, the downlink bandwidth of which can also be used for the OLFAR experiment.

In discussion with Graham Shirville of AMSAT, it has been decided that Delfi-n3Xt can in principle reuse the Delfi-C^3 frequencies; also 4 in total as well as the transponder bandwidth. Of course, final frequency filing should confirm this. These frequencies are illustrated in Table 15.1 and Figure 8.1 below.

Table 8.1: UHF/VHF transmission frequencies

Link type	***Band***	***Specific frequencies***	***Comments***
Telemetry downlink	VHF radio amateur band (144-146 MHz)	145.870 MHz 145.930 MHz	Primary Delfi-C^3 frequency Backup Delfi-C^3 frequency
Telecommand uplink	UHF radio amateur band (430-440 MHz)	See [SLR 0309]* See [SLR 0309]*	
Transponder / OLFAR downlink	VHF radio amateur band (144-146 MHz)	145.880-145.920 MHz	
Transponder uplink	UHF radio amateur band (430-440 MHz)	435.530-435.570 MHz	
Transponder beacon	VHF radio amateur band (144-146 MHz)	145.870 MHz *or* 145.930 MHz	Same as PTRX telemetry downlink frequency

*Due to reasons of security the specific uplink frequencies are not published in general project documentation.

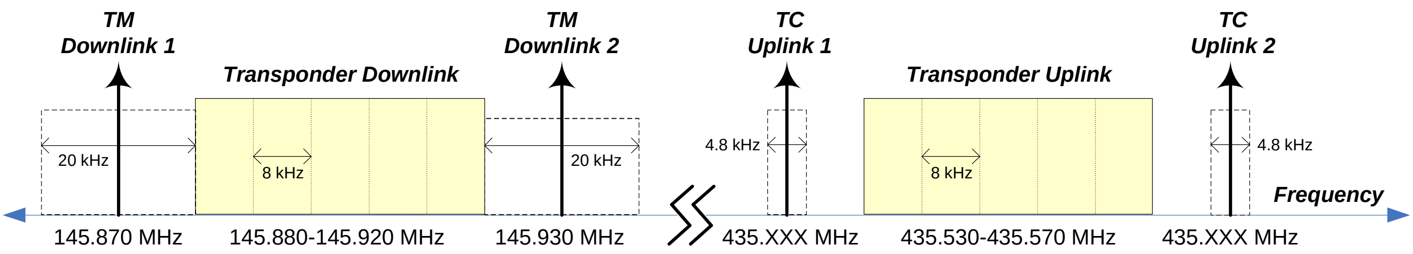


Figure 8.1: Transmission frequencies in the UHF and VHF band

In Figure 8.1 three aspects are important:

- Transmission frequencies op the up and downlink
- Allocated bandwidth per transmission frequency
- Transponder bandwidth

These three aspects are addressed in sections 8.1.1 through 8.1.3 below.

8.1.1 Transmission frequencies

Table 15.1 states the specific transmission frequencies clearly. The question that remains is which of the frequencies will be allocated to the PTRX and which to the ITRX. Currently the primary Delfi-C^3 RAP is used, in other words, downlink operations are nominal. The possibility for a disturbance following both satellites having the same frequencies in use is small, having posed an argument for frequency reuse in the first place together with the minimal required design changes. Nevertheless, it makes sense to allocate the back-up Delfi-C^3 frequency to the effective primary Delfi-n3Xt radio. While the PTRX is designed to be the most secure radio, the ITRX might turn out to be more power-efficient. As such, the ITRX might turn out to be the primary radio in operations. This question therefore remains to be answered. For the uplink, it is less of an issue, as Delfi-C^3 and Delfi-n3Xt will never be commanded at the same time and they cannot respond to each others commands.

There is in fact also an Indian satellite, VUsat, or named by its radio amateur distinction OSCAR-52, which has complete overlap with the Delfi-C^3 frequencies. Its transponder frequency range overlaps with the whole of Delfi-C^3; 145.870-145.930 MHz. This theoretically would allow for clashes of communication momentarily. But, sharing of the (free) resources is a necessary evil.

To be determined:

- Which downlink/uplink frequencies will be allocated to respectively the PTRX and ITRX?

8.1.2 Transmission bandwidth

The transmission frequencies establish the *center* of the transmission bandwidth. As explained in the previous chapter, all data transmission requires a certain transmission bandwidth.

Bandwidth and its allocation can be better explained by looking at Figure 8.2 below.

Figure 8.2 shows a certain *necessary band*, required for actual data transmission and more importantly its correct reception. The rest is actually not required for correct interpretation, but the signal bandwidth cannot be reduced without affecting also the signal content within the necessary bandwidth. Therefore, unwanted emissions are a necessary evil.

The 'end' of the bandwidth will be defined in terms of a required power reduction with respect to the main transmission signal power, or *center frequency power*. But, as a certain band is required for power levels to drop, a certain *out-of-band domain* is automatically required. Anything outside of this band will then be called the *spurious domain*.

The total allotted bandwidth within a certain frequency band is then a *combination of the necessary bandwidth and out-of-band bandwidth.*

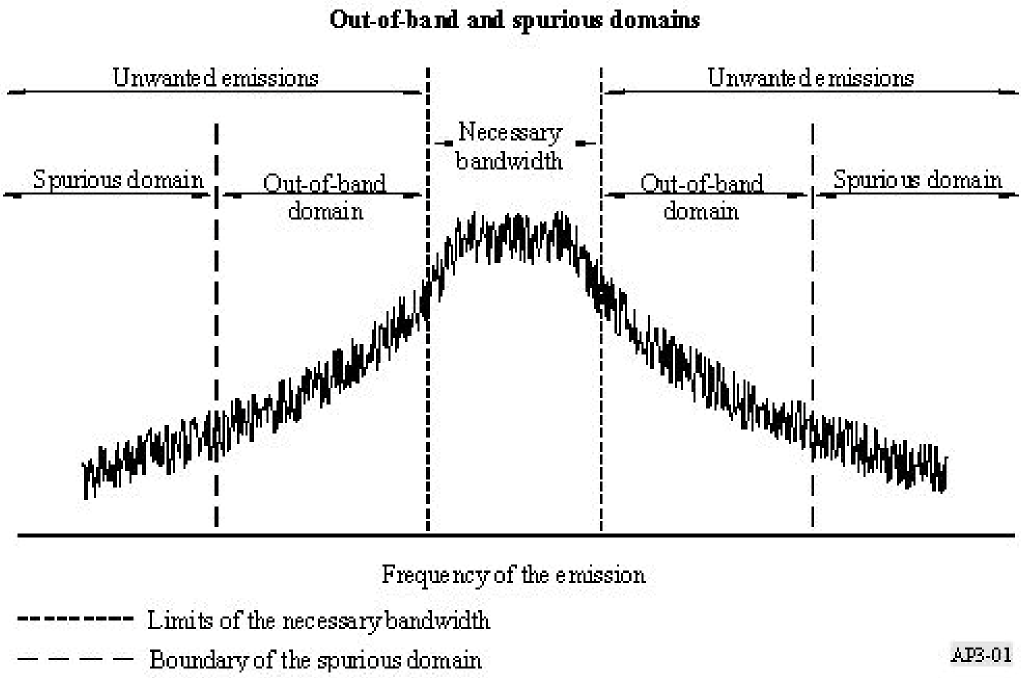


Figure 8.2: Definition of the bandwidth domains [ITU regulations]

In case of the VHF and UHF radio amateur bands, available transmission bandwidth is severely limited. This is illustrated by the fact that frequencies can even be allocated to multiple satellites at the same time. Therefore, *the general consensus is that no data rates higher than 9.6 kb/s are supported*. This holds for both the VHF and UHF bands.

Based on the specified data rate, a bandwidth shall be allocated. Usually, this bandwidth is more or less twice the symbol rate. In case of Manchester coding for example, the symbol rate is twice the bit rate. For Delfi-C^3 therefore, the specified bandwidth was **2.4 kHz**, for both the up- and downlink.

Higher data rates then would require a larger bandwidth. In fact, there seem to be no objections from the radio amateur community (according to Graham Shirville) to increase the allocated bandwidth of Delfi-n3Xt to support higher data rates. As will be explained in sections 8.3 and 8.4 below, there is good reason for the downlink to aim for a 9.6 kb/s data rate. For the uplink, 1.2 kb/s or 2.4 kBaud is set as a maximum. Similarly the ITRX aims to transmit data at a data rate of 9.6 kb/s, and will be able to receive data at 1.2 kb/s. As such, logical bandwidth values coupled to data rates of 9.6 kb/s and 2.4 kBaud are **20 kHz** and **4.8 kHz**. These values have been indicated in Figure 8.1 above.

In any case, there are no repercussions if a larger bandwidth is requested and the higher data rates are not achieved due to technical difficulties. As such, the *bandwidths specified above should be filed for*.

For satellite transmissions, the minimal signal reduction out-of-bandwidth, thus at the start of the *spurious domain* as introduced above, is stipulated by given ITU regulations, see requirement SAT.2.3-P.01. The constraint states that spurious emissions for the satellite and ground transmitter should have a maximum strength of -43 - 10 log (*P*) in dBc (which indicates the amount of dB with respect to the main transmission frequency or center frequency), where *P* is the transmitted power, *or* -60 dBc, whichever value is larger.

Given the anticipated transmitter power on the RAP and designed to be similar on the PTRX, being 0.2 W:

- *Spurious emissions on the downlink should be limited to* ***-36.0 dBc***

For the uplink transmission, the HLV-550, which is the final amplifier, can theoretically deliver some 550W transmission power. In fact, it is limited in operational use to 250 W. This means the calculated value easily surpasses -60 dBc. As such:

- *Spurious emissions on the uplink should be limited to* ***-60 dBc***

Finally, it can be said that according the ITU regulations, this reduction is measured over a *4 kHz bandwidth*, independent of frequency band, the *reference bandwidth*. In fact this means that from the start of the spurious domain, an average power value over the next 4 kHz is taken. One can realize that this means that a slightly higher power at the actual boundary of the spurious and out-of-band domains can thus be tolerated. When low total bandwidths are used, such as the 2.4 kHz bandwidth used for Delfi-C^3 or the Delfi-n3Xt uplink, the effect of this can actually be sizeable.

There is an important note to be made. It has been explained that following the bandwidth requirements, transmission data rates are limited. The exact transmission bandwidth however, is less so. It is recognized that the power reductions stated above are hard to achieve in practice within the stated bandwidths, given limited complexity. Therefore, it is not essential for typical radio amateur satellite transmissions with *low power emissions*. At the same time, ground station transmissions do have high signal strength, but only *transmit sporadically* and with a *very high directivity* towards space. In other words, the actual bandwidth requirements within the radio amateur community are more indicative than anything else.

8.1.3 Transponder bandwidth

The transponder bandwidth has been indicated in Figure 8.1, and is 40 kHz broad. There is no specific transmission frequency belonging to it. Multiple users can make use of the transponder bandwidth at the same time, using different transmission frequencies.

The total bandwidth was determined originally in discussion with the radio amateur community, preferring a transponder service basically as broad as possible. Due to circuitry design complexity and most notable linear amplifier bandwidth and power the total bandwidth was limited to **40 kHz**. Albeit not officially constrained, this allows for 5 individual slots of usual sizes of 3 kHz, with 5 kHz in between slots for signal damping [SLR 0244]. These slot sizes are per usual radio amateur convention. From a design point-of-view, these slots are not important.

The transponder bandwidth can also be used for the OLFAR experiment.

8.2 Link budget

The link budget [SLR 0106] of all transmission links gives the available power for transmission of data. The link budget takes as an input the transmission power at the transmitter output and takes into account all losses and gains on the link to finally in combination with the modulation scheme be able to conclude on achievable data rate. At the same time it gives the transmission margin left over when a certain data rate is selected.

The orbit altitude of Delfi-n3Xt is to be determined, but is slated to be between 600 and 850 km. The minimum elevation applied determines the received E_b/N_0, as path loss changes significantly with elevation angle. In turn this allows drawing conclusions on data rate. A worst case situation, power wise, of designing for a 10° elevation angle is used below to give received E_b/N_0 values; this choice is explained in the next section.

Table 8.2 summarizes the results of the current link budget [SLR 0106]. The modulation scheme is identical to that of Delfi-C^3. Minimum data rates, identical to those of Delfi-C^3 are shown.

Table 8.2: Link margins of the UHF/VHF communication links for the Delfi-n3Xt range of orbit altitudes

Link direction	***Orbit altitude***	***Data rate***	***Received E_b/N_0***	***Required E_b/N_0 incl. min. margin***	***Link margin***
Down (VHF)	600 km	1200 Baud	26.45 dB	13.30 dB	*13.15 dB*
	850 km	1200 Baud	24.23 dB	13.30 dB	*11.02 dB*
Up (UHF)	600 km	1200 Baud	42.27 dB	16.30 dB	*25.97 dB*
	850 km	1200 Baud	40.14 dB	16.30 dB	*23.84 dB*

In the last section it has been explained that a maximum symbol rate of 9600 b/s is allowed over the radio amateur bandwidths. For the uplink, the current bit rate is 600 b/s, but the symbol rate is 1200 b/s or 1200 *Baud* as will be explained in this chapter. In both cases then, an increase to 9600 Baud would come at the cost of **9 dB**. *The maximum data rate can then be achieved for both the up- and downlink according to the link budget*.

Another option that will be discussed in chapter 9 on the PTRX is to *lower the gain of the power amplifier on the downlink*. In this manner, the total output power of the PTRX can be reduced, logically at the gain of having less power consumption. The above link budget shows that a -2 dB link budget would still close for data rates of up to 9.6 kBaud, for all satellite altitudes. For lower data rates, even larger reductions are possible.

A small note regarding the transponder as well as OLFAR; no complete link budget can be given as the receiving stations are not determined as well as the signal content. Using automatic gain control and the transmitter power amplifier, the output power of the resulting signal (or the power of the peak component in the bandwidth) shall be a constant 0.2 W, or -7 dBW. As said before, the following sections will no longer take into account transponder and OLFAR communication.

8.3 Physical layer: downlink modulation scheme and data rate

As addressed in chapter 7 physical layer activities range from modulating (in turn necessitating line coding and/or pulse shaping) to frequency conversion and amplification. Frequency conversion and amplification do not change the information content of the signal. Establishing the physical layer then results in a question of *selecting modulation scheme*. At the same time, line coding and pulse shaping approaches should be selected.

There are two cases relevant for the choice of modulation methods; *downlink and uplink*. These cases will be addressed separately, in two separate sections, in order to keep a clear header structure. This section, section 8.3, discusses the downlink modulation scheme, whereas section 8.4 handles the uplink modulation scheme. The introduction given directly below applies to both sections.

As presented in section 7.2, there are three general criteria for selecting modulation methods:

- **Bandwidth efficiency**
- **Power efficiency**
- **System complexity**

The bandwidth and power constraints resulting from the choice of frequency and the link budget have been addressed above. System complexity of modulation schemes has been addressed in the previous chapter, but more important is *heritage*; a system already available or partly available has (much) less added complexity and poses the default choice. Also, when proposing changes the availability of COTS components again reduces implementation complexity.

Related to the three criteria is achievable data rate; whereas a data rate as high as possible is in principle preferable; it is limited by power, bandwidth and complexity arguments. Data rate is concluded upon below as well, and using theory presented before the approach of fixing or changing the data rate, and to what amount(s) is discussed.

In the two following sections six aspects are discussed in order:

1. Selection of modulation scheme of Delfi-C^3
2. Possible improvements of the modulation scheme
3. Scheduled implementation of the modulator
4. Scheduled implementation of the demodulator
5. Determination of the data rate approach
6. Selection of modulation scheme of Delfi-n3Xt

The (de)modulator implementation sections show the currently envisioned hardware implementation. The designs are based on the RAP implementation, but in some cases updates have already been made in terms of components, such as the microprocessors and software. This allows drawing conclusions on what improvements are possible and at what cost.

The six aspects are discussed in subsections 8.3.1 through 8.3.6 below.

8.3.1 Selection of modulation scheme of Delfi-C^3

The trade-off for the Delfi-C^3 downlink modulation scheme was performed in [SLR 0248]. It was performed by looking at modulation schemes (generally) applied within the radio amateur community, in order to ensure compatibility with radio amateur equipment. The schemes introduced were BPSK and BFSK, with variants for the first two applying filters (raised cosine and a filter type called G3RUH; the latter is a rather complicated scheme which limits the bandwidth). Also, an analogue modulation scheme with very bad error performance (20 dB for a *BER* of 10^{-5}), but historically often used (AFSK or Audio FSK) is taken into account. Finally, and rather incidentally, a non-radio-amateur-standard, MSK, was compared.

The resulting option used for the downlink of Delfi-C^3 was:

> **Delfi-C^3 Modulation scheme**: BPSK (1200 b/s)
> **Delfi-C^3 Pulse shaping**: (Time domain) raised cosine (roll-off factor of 1)
> **Delfi-C^3 Line coding**: NRZ-I, or more specifically NRZ-S (0 causes a transition)

BPSK has, of the proposed options, the best error performance, together with MSK. MSK however has better bandwidth efficiency (see Figure A-17). Modulation complexity for both is very low, while demodulation complexity neither is very high, albeit slightly in favorite of MSK; the latter can be easily demodulated at the cost of not even 1 dB of gain. However, under the argument of being an amateur standard, BPSK was chosen. The used frequency spectrum was to be limited using a (time domain) raised cosine filter before modulation, and the problem of phase ambiguity was to be solved by adding NRZ-I encoding.

The total approach of the trade-off performed in [SLR 0248] does not feel very complete, as it pays no mind to system complexity, it only describes bandwidth efficiency by "good" or "poor" (not quantitatively, such as by using out-of-band power), and finally, with the exception of MSK, it presents the argument of being a radio standard as a definitive criterion.

Nevertheless, the result of the trade-off was the fact that raised cosine BPSK was selected and successfully applied on Delfi-C^3. The consequent effect is that the relative complexity of the chosen modulation scheme is very low as successful heritage is available. In other words, it becomes the default choice for Delfi-n3Xt. Using general and relative complexity, as well as bandwidth efficiency and power efficiency arguments, conclusions on the preferred Delfi-n3Xt modulation scheme are drawn below.

8.3.2 Possible improvements of the modulation scheme

The previous section shows that bit rates of 9.6 kb/s are achievable given the current link budget. Higher data rates than 9.6 kb/s are not allowed within the radio amateur VHF band, but 9.6 kb/s itself is. The RAP however only performed at 1.2 kb/s. In conclusion, the system was:

- **Complexity limited**

For data rates above 9.6 kb/s, the system would become both *power limited* and *bandwidth limited*.

However, it has been stated that official bandwidth regulations are actually not met, although this is (unofficially) accepted by radio amateur standards. Nevertheless, it would be desirable to reduce transmitted frequency spectra. As such, the system can also be said to be generally bandwidth limited.

Concluding from the above, the modulation scheme does not limit performance. It is in fact the complexity that has limited higher data rates. Also, the current modulation scheme has been selected largely for reasons of minimum complexity. As such, the logical direction of advancement is:

1. Increasing the data rate with the same modulation scheme

   An increased data rate is preferable for Delfi-n3Xt, because of the increased number of payloads/experiments as well as satellite bus complexity (resulting in more housekeeping data) with respect to Delfi-C^3. Also, a proven transmitter which delivers a higher data rate increases the (future) potential of the satellite bus.

It has however been stated also that a reduced bandwidth spectrum would be desirable. Chapter 7 has presented alternatives of BPSK which have *higher bandwidth efficiency*. With minimum added complexity and equal power efficiency, the alternatives are the higher MPSK schemes and MSK. For $M \geq 8$, the system would become more power inefficient. This would not be acceptable given the link budget. Therefore, **QPSK** and **MSK** are the two options.

While the bandwidth spectrum of MSK is narrower than that of QPSK, QPSK can easily be shaped by a (time domain) raised cosine filter. In both case more advanced pulse shaping schemes would also be possible. Also, the bandwidth efficiency of QPSK can be increased by implementing an extra control loop to avoid 180° phase shifts; in other words by applying π/4-QPSK. Also, MSK and QPSK are closely related as they can be derived from each other by applying pulse shaping.

The real benefit of QPSK lies in the fact that the current BPSK modulator, the *RF 2713* [SLR 0741], that replaces the BPSK modulator and mixer of the old RAP on the PTRX, is in fact a QPSK modulator. Currently, it is connected as a BPSK modulator. However, at a later stage it can be tested when operated as a real QPSK modulator. Therefore the second option of advancement is:

2. Applying QPSK to increase bandwidth efficiency at no power cost

   QPSK would theoretically half the bandwidth produced by modulation

The resulting incurred complexity in the satellite and on-ground for both advancements is commented on below in a discussion on modulator and demodulator.

Pulse shaping

In case of Delfi-C^3, no 'real' raised cosine filtering was applied, but a more easily implementable time-domain raised cosine filtering. This type of filtering has extensively been addressed in chapter 7. Without its application, required bandwidth for a BPSK signal would enlarge to an unreasonable extent; chapter 7 indicates BPSK requires 4 times the data rate to reduce signal power with only -20 dBc. Also, no additional filters are in place after modulation in the PTRX system design to limit the resulting bandwidth. Filters for these narrow bandwidths are practically not available.

In chapter 7, the raised cosine filter has been introduced, including both its time domain and frequency domain implementations. For Delfi-C^3, it has been shown that the downlink transmission actually occupies *9 kHz* given the power reductions required, of which *5 kHz* is above noise levels on-ground. For higher data rates, this bandwidth is expected to scale approximately linearly with the data rate, so that 40 kHz might be received on-ground above noise levels. Although not very satistactionary, it gives acceptable performance.

This filter should therefore certainly be reapplied in Delfi-n3Xt, also given its limited complexity. A desirable imporvement however would become:

3. Implementing a (real) raised cosine filter
   A simple implementation of this type of filter could limit the frequency spectrum to 2 times the data rate, with required *spreading period* of only 5-7 symbols most likely. A small power cost is incurred, but this amount should be available according to the link budget. The frequency spectrum would then likely adhere to all requirements. The filter should be completely implemented *digitally*.

Line code

The default line code is unipolar NRZ, which is the output of any microprocessor. For the modulation process this line code should be converted to bipolar NRZ in case of BPSK. NRZ has the minimum bandwidth as explained in chapter 7, as well as the maximum error performance. At the same time it does not guarantee a zero dc-component or a large number of transitions necessary for symbol synchronization. The former has not proven to be a problem in Delfi-C^3 signal reception, as the signal is demodulated in software. Sufficient transitions are introduced due to the bitstuffing feature of the data protocol.

It is interesting to note that an RZ code would effectively create an OOK scheme, as the carrier is switched on and off. RZ codes require a 3 dB higher E_b/N_0 for correct demodulation. And, indeed, OOK has 3 dB worse error performance than BPSK. Also, more complicated codes such as AMI codes require a higher E_b/N_0 for demodulation, which is not desirable. AMI-NRZ, requires 3 dB more, at the gain of error detection. It is however more oversightful and less complex to keep error detection functionality at the data link layer, also costing less bandwidth.

As with BPSK the phase is not determined to signify either 0 or 1, but the phase is shifted in case of a transition from a 0 to 1 or vice versa, a signal at any given time can signify both a 0 and a 1. As such, *differential coding should be applied*, as is further explained in the previous chapter. A differential version of NRZ, NRZ-S shall therefore be used for the downlink. The fact that 0's are mapped as transitions as opposed to 1's, is a matter of definition and is used in the data protocol.

The use of a differential code does introduce a slightly reduced error performance of 0.8 dB, so that 10.3 dB E_b/N_0 is required to demodulate the signal. This is due to the demodulation dependence on a previous bit, which might be erroneous.

No changes of line code are suggested.

8.3.3 Scheduled implementation of the modulator

In case of the downlink modulation scheme, it is important to have little complexity on-board, as opposed to on-ground. Indeed BPSK delivers a very simple modulation scheme; in effect only an oscillator should be mixed with a bipolar NRZ bit stream, as illustrated by Figure A-11. In functional components, the entire modulation can then by illustrated by the following set-up (Figure 8.3). Details are limited to the components that play a role in the modulation scheme, but the general interfaces and system boundaries are shown.

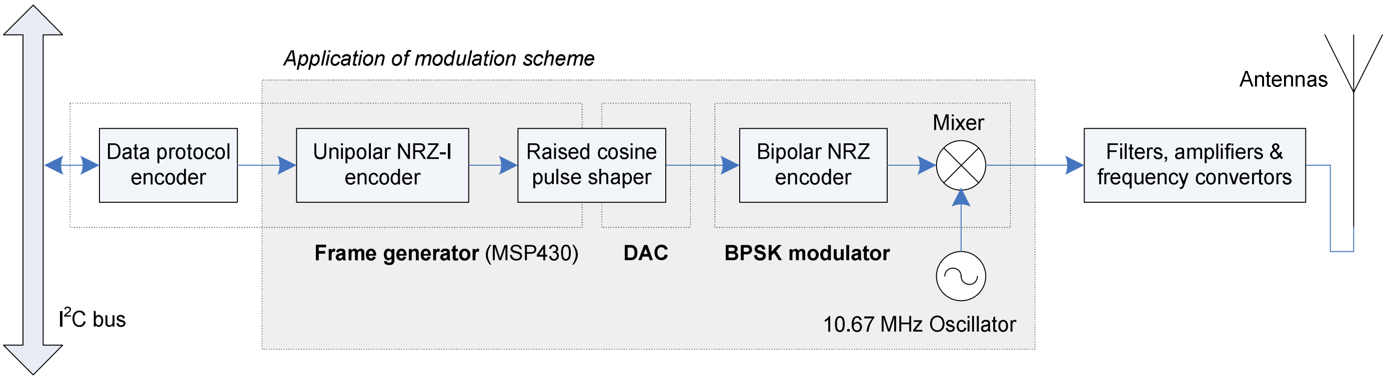


Figure 8.3: Default integration of the Delfi-n3Xt downlink modulator

Data received over the I^{2}C bus is coded according to the data protocol, and the Unipolar NRZ-I line code. Along with pulse shaping all these functionalities are taken care of by one microprocessor, being an MSP430 processor (this will be further explained in the next chapter). It does however involve a DAC (digital-to-analogue converter) to convert a digitally sampled raised cosine shape to an actual analogue cosine signal. In Delfi-n3Xt the combination of microprocessor and DAC is to be called the *Frame Generator* in case of the transmitter.

Bipolar NRZ coding is consequently created, quite simply by subtracting half the voltage used to indicate a unipolar 1, using an offset of 1.65V. This functionality together with the mixing functionality is taken care of by a BPSK modulator. The oscillator and mixer bring the signal to a (first) intermediate frequency (IF1) of 10.67 MHz.

A system of filters, cascaded amplifiers and several frequency conversion steps follow before the signal is transmitted by the antennas.

The possibility of implementation of the improvements suggested before, with the current hardware, and the impact of its change is indicated below. A short description is given as well as a color code for illustrational purposes. A final judgment of the feasibility for Delfi-n3Xt is given in subsection 8.3.6.

Green indicates little to no changes with a good idea of the impacts
Blue indicates moderate changes with some unknown impacts
Red indicates a serious change with unknown specific impacts

1 Increasing the data rate with the same modulation scheme

A higher bit rate would have the following effects on the components above.

- ***Microprocessor (MSP430)***: 9.6 kb/s is likely acceptable
  The microprocessor, next to applying NRZ-I encoding and outputting the digital raised cosine shape with 16 samples per bit, is also tasked with the application of the data protocol and I^2c communication. A preliminary test for a bit rate of 2400 b/s has been tested in terms of microprocessor capacity and was positive [SLR 0722]. Higher bit rates should still be tested, but are likely acceptable. As an illustration; the actions currently most load-consuming are pulse shaping and as a reasonable second, the calculation of the data protocol CRC (see section 8.5).
  - Every sample requires 5 *operations* for it to be pulse shaped. 16 samples gives 80 operations per bit. A data rate of 9600 b/s requires 77 kHz, which is far below the maximum processor frequency of 8 MHz.
  - The entire CRC for a 1200 bit package is calculated in (only) the time it takes to prepare 4 pulse shaped bits.
- ***DAC***: Acceptable without changes
  The digital-to-analogue converter is in charge of converting an input bit value to a certain output voltage. The current clock speed is some 7.4 MHz, and it takes 13 clock cycles to process one bit value. As such, up to 500 kb/s is theoretically possibly.
- ***Oscillator frequency***: Acceptable without changes
  The IF carrier frequency is largely sufficient for and independent of large bit rates, as long as the IF frequency (10.67 MHz) is much larger than the bit rate.
- ***Analogue components***: Acceptable without changes
  The bandwidth over the entire process, most notably from the frequency converter (mixer) onwards, is increased. As the next chapter will show, no narrow-band filters are in place, thus it should pose no problem.

2 Applying QPSK to increase bandwidth efficiency at no power cost

As explained before, the BPSK modulator is in fact a QPSK modulator. The only affected component then is the:

- ***Microprocessor (MSP430)***: Acceptable with changes to software and electrical integration
  The *interfaces* between the processor (incl. DAC) and the QPSK modulator should be adapted. The processor should output signals to an amplitude (*I*) and quadrature (*Q*) output. Software should therefore be adapted and processor load is slightly increased. Electrical interfaces should be changed and tests are required.

3 Implementing a (real) raised cosine filter

As the raised cosine filter is a *digital filter*, the only affected component is:

- ***Microprocessor (MSP430)***: Likely requires a new processor
  The difficulty with a frequency domain raised cosine filter lies with the memory required; an analogue wave should be formed based on a number of bits, likely 3, 5 or more. At the same time, the amount of samples per bit is important; in fact the number of samples per bit times the number of bit periods gives the required complexity. Complexity thereby increases exponentially. As such, integration of this functionality within the microprocessor should be seriously tested. In any case the filtering requires a software implementation.

8.3.4 Scheduled implementation of the demodulator

Demodulation takes place on-ground and as such complexity is less of an issue. The entire ground station set-up in terms of hardware is described in chapter 15.

Figure 8.4 below describes the block diagram of the demodulation process, procured in software, as well as its context.

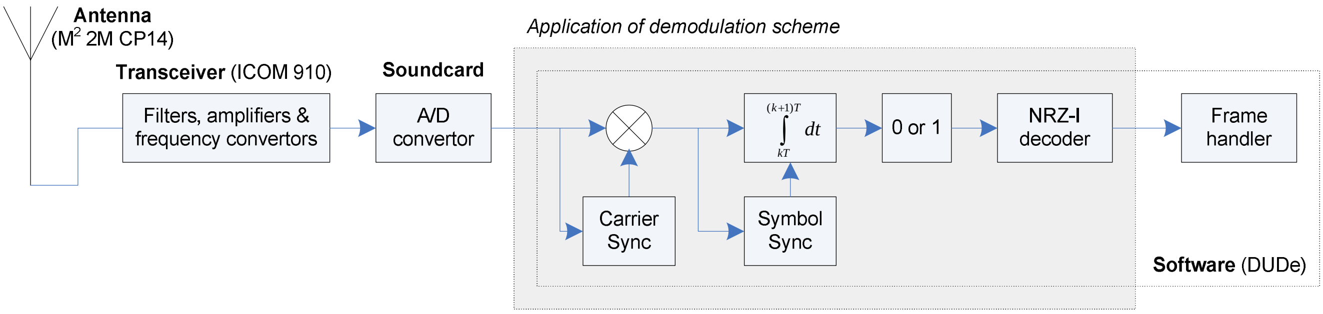


Figure 8.4: Default integration of the Delfi-n3Xt downlink demodulator

In case of Delfi-C^3 operations, the RF signal is received by an antenna, and forwarded to the ICOM 910 where the signal is filtered, amplified and converted to a 1600 Hz signal. The signal can consequently be demodulated. As on-ground this is done in software, the 1600 Hz analogue signal is fed to a sound card, which samples the analogue wave with 38.4 kHz, saving 32-bit values at each point. This sampled signal is fed to the software.

The software program, DUDe, takes cares of the entire demodulation process. The software in case of the Delfi-C^3 mission was RASCAL, with its demodulation core called Warbler. DUDe should be an improved and more modular update. The functionalities performed are exactly those shown in Figure A-13, but with an added symbol synchronizer which in fact is also necessary. Carrier synchronization is taken care of using a Costas loop. In fact the 1600 Hz value was chosen based of convenience due to the Costas loop and the ICOM 910; for demodulation the requirement is that it is more than the bit rate. The symbol synchronizer takes the signal after carrier synchronization has been achieved, and extracts the clock signal from it. This clock signal is used to determine the bit period *T*, to perform the integration. Afterwards it can be decided whether a 0 ort 1 is being received, and NRZ-I coding can be removed.

Frames are consequently handled and send to a database for storage.

1 Increasing the data rate with the same modulation scheme

In case downlink data rate is increased, three components are affected:

- ***The ICOM 910***: Requires an available replacement

  Both filtering and frequency conversion are dependant on the bit rate. The ICOM 910 has a 2.7 kHz channel bandwidth for BPSK reception, which is sufficient for the reception of a 1.2 kb/s data signal. For higher data rates however, the ICOM 910 filter bandwidth will limit performance. As such, a replacement is required. However, as further discussed in chapter 15, ISIS provides a transceiver that will be able to support much higher data rates on the downlink. Consequently, a specific center frequency for the resulting signal should be selected and can be set in coherence with the demodulation software.
- ***The sound card***: Requires no changes or an available replacement

  The current sampling rate of the video card integrated in the ground station computer is 38.4 kHz. Nyquist theorem states that a wave being sampled should be sampled with at least two times the maximum frequency that is to be recuperated from the signal. Practical values are however much higher, also as demodulation involves integration the resulting signal. The received signal frequency in current settings is 1600 Hz, yielding 24 samples per frequency period, or 32 per bit period at a data rate of 1200 b/s. Then, the sampling rate is likely sufficient for data rates of 2.4 kb/s. Higher data rates however require a higher sampling rate; (cheap) commercially available video cards with sampling rates up until 192 kHz exist and as these are plug-and-play components in modern PC hardware an easy update can be performed. *Also, the transceiver that ISIS can provide should have integrated sampling rates up to several MHz.*
- ***The software, DUDe***: Requires minor changes

  The demodulation core should be easily reconfigured to allow for a higher bit rate. Primarily, the expected bit period should be changed, in the actual demodulation process and in the synchronization loops. At the same time the expected input signal frequency should be changed. This is all a matter of proper settings. Finally, computer processing power and process memory should be sufficient for the increased calculation load; assuming moderate data rate increments this should not be a problem for a relatively powerful desktop computer.

2 Applying QPSK to increase bandwidth efficiency at no power cost

On-ground, the actual demodulation takes place in software. As such, the only affected component is:

- ***The software, DUDe***: Requires changes to the demodulation algorithms

  The demodulation core should be configured to allow for a higher bit rate. Primarily, the expected bit period should be changed, in the actual demodulation process and in the synchronization loops. At the same time the expected input signal frequency should be changed. This is all a matter of proper settings. Finally, computer processing power and process memory should be sufficient for the increased calculation load; assuming moderate data rate increments this should not be a problem for a relatively powerful desktop computer.

3 Implementing 'real' raised cosine filtering

No changes are required on-ground.

8.3.5 Determination of the data rate approach

In section 7.9 theory was presented on maximizing the downloadable data volume. In case of the UHF/VHF communication link, a dedicated ground station is in place, which is able to track Delfi-n3Xt. Also, it is automated. As such, it at least makes no sense to, if a constant mission data rate is used, select one based on a minimum elevation angle no higher than 10°. The 10° in this case is based on:

- **Increasing atmosphere loss and antenna noise at low elevations**
  See [SLR 0106] or section 7.9.
- **Obstructions on the horizon**
  These are present up to several degrees in every direction, but there are more profound obstructions up to 10° elevation between 140 and 160° azimuth direction (Delft GS, [SLR 0243]).
- **A (slight) over budgeting in the link budget**
  This is due to the calculation with worst-case values as well as a link margin; therefore the link budget can reasonably be assumed to remain closed for smaller elevation angles.

Then the only relevant decision becomes to determine whether the data rate will be changed during the mission, either between passes or during passes. More downloadable data volume can always be useful, even if it might not strictly be required. However, there are a couple of arguments against either option:

- **The PTRX, and similarly the ITRX, is not power limited**
  Even at 10° minimum elevation the power is available to go up to 9.6 kb/s. Higher bit rates are not allowed due to bandwidth constraints.
- **Generation of data with a varying data rate is difficult**
  For a constant data rate, the CDHS needs just to deliver and *acquire* a constant amount of data following a fixed schedule. For multiple data rates, different data acquisition schemes should be incorporated introducing also more complexity for the CDHS, as well as data generating systems. Due to CDHS being the main project-constraint at this moment, again this complexity is not desired. If on the other hand data acquisition is kept constant, some kind of data buffer should be introduced to be able to send stored data, again at the cost of extra complexity.
- **Either orbit determination or scheduling via telecommands should take place**
  Orbit determination should not be assumed. Scheduling again would increase complexity at the CDHS, and put extra emphasis on the availability of an accurate on-board clock.
- **On-ground systems should be prepared to receive different data rates**
  Currently, the ground system is not prepared for this. Whereas the transceiver able to deal with higher data rates should be able to receive multiple data rates, it will not be able to auto-detect these data rates, and similarly the demodulation software (DUDe) is not prepared for this. Again, this would require a significant time-investment.

In conclusion, changing data rates for the PTRX is not deemed useful as the system is not power limited. In any case, the addition of this feature would come at the cost of added complexity in at least PTRX, CDHS and GS. At the same time, the UHF/VHF communication link is of course mission-critical and experimental features at the cost of extra complexity are not desirable. A possible STX would be more open to these kinds of features. In result, the link budget for the VHF downlink shall indeed be based on a 10° minimum elevation angle, with a constant data rate throughout the mission.

8.3.6 Selection of modulation scheme of Delfi-n3Xt:

Three options of improvement have been suggested. The conclusions regarding implementational difficulties are summarized below, and the resulting modulation scheme and data rate for Delfi-n3Xt are presented.

1 Increasing the data rate with the same modulation scheme

Increasing the data rate to 9.6 kb/s is likely possible. Even if not, data rates (much) higher than 1200 b/s are. Functional tests should be performed, and hopefully this can be done soon given the transmitter section finalized by Maurits Schaap and the Frame Generator software prepared by Christiaan Hartman. No additional changes to the satellite hardware are required.

On-ground, the transceiver should be replaced and the DUDe software core should be updated, but only in the form of simple operational parameters. The transceiver can be delivered by ISIS.

If proven, it should still be decided whether 9.6 kb/s will be functionally used by the CDHS. In other words: whether 9.6 kb of data will be generated every second. If not:

- Frames can simply be repeated a number of times, possibly with repeats spaced apart by a given amount of time. This would require a certain buffer size implemented in the Frame Generator, but would allow full in-orbit testing of the maximum data rate.
- The transmitter can work at a lower speed in nominal operations, but 9.6 kb/s operation can be integrated as test functionality, as it only involves a software setting

The latter suggestion can also be used if the high data rate will not be used in nominal operations *and* the power consumption of the PTRX will be reduced by lowering the gain. This option is discussed in chapter 9, but will likely imply a lower nominal data rate to be used on the PTRX than the maximum allowable 9.6 kb/s.

2 Applying QPSK to increase bandwidth efficiency at no power cost

QPSK can be integrated with minimal on-satellite complexity, given the current inclusion of a QPSK modulator. The processor should however be configured and connected differently. The exact impacts are to a certain extent unknown, so tests might be performed if time allows for it.

On-ground however, the change to QPSK requires a sizeable update of the DUDe demodulation software core. Most notably, this would require a significant effort to understand the current demodulation software and adjust it accordingly. Also, this would require making significant adjustments to a very critical and currently proven software core.

Nevertheless, optimally this functionality would be integrated in the PTRX *as test functionality*. The software core can be updated at a later stage as this requires on-ground complexity. For now however, QPSK implementation is considered an optional nice-to-have.

3 Implementing 'real' raised cosine filtering

This option infers sizeable extra processor load and complexity, likely requiring a new processor. This would require moving away from the standard Delfi-n3Xt processor, and rewriting part of the software already written for the MSP430. On-ground however no changes are incurred.

Due to its on-satellite implications, it is not assumed this functionality is necessary on Delfi-n3Xt given the level of its impact and the mildness of radio amateur bandwidth restrictions.

As a result:

Delfi-n3Xt:

Modulation scheme: BPSK
Pulse shaping: (Time domain) raised cosine (roll-off factor of 1)
Line coding: NRZ-I, or more specifically NRZ-S (0 causes a transition)
Bit rate: up to 9600 b/s
Requested bandwidth: 20 kHz

Alternative names for the combination of modulation scheme and pulse shaping:

- RC-BPSK (raised cosine BPSK)
- SSB-SC (single side band, suppressed carrier)

Alternative names for the combination of modulation scheme, line coding and demodulation method:

- Coherent DEBPSK

8.4 Physical layer: uplink modulation scheme and data rate

As was done in the previous section for the VHF downlink, this section discusses the UHF uplink modulation scheme. It follows the same structure, moving from Delfi-C^3 modulation scheme selection to Delfi-n3Xt modulation scheme in six subsections, being subsections 8.4.1 through 8.4.6 below.

8.4.1 Selection of modulation scheme of Delfi-C^3

The selection for the uplink modulation scheme is again described in [SLR 0248]. FSK, or in fact Manchester encoded FSK was chosen for the simple reason that it requires the simplest architecture on-board of the satellite. Its reduced power efficiency is of little concern because large (on-ground) amplifiers can be used for the uplink. Being an amateur standard neither was of concern, as the uplink transmission would not be public. Manchester coding was added for easy clock synchronization, although it comes at cost of a halved effective data rate. The resulting option used for the uplink of Delfi-C^3 was:

Delfi-C^3 Modulation scheme: FSK (600 b/s)
Delfi-C^3 Line coding: NRZ-I (NRZ-S) followed by Manchester coding
Delfi-C^3 Pulse shaping: None

This scheme again poses the default scheme for Delfi-n3Xt.

8.4.2 Possible improvements of the modulation scheme

Following the rationale of the radio amateur bandwidths, 9.6 kb/s or actually 9.6 kBaud would be acceptable from a bandwidth point-of-view. The Delfi-C^3 uplink bit rate was 0.6 kb/s, and 1.2 kBaud. At the same, the link budget shows a very large link margin. In other words, the Delfi-C^3 uplink was:

- **Complexity limited**

As for the downlink, it is however desirable to reduce the transmitted spectrum to approach regulatory values. Nevertheless, in case of the uplink the impact is very minor as the uplink is not continuous and highly directed.

Compared to other modulation techniques, FSK, or more specifically BFSK is neither very bandwidth efficient nor power efficient. Both aspects however, as argued above, do not pose a serious argument for change. FSK was chosen as it yields the simplest possible demodulation architecture, inherent to FSK schemes. Even so, the system is still complexity limited. Therefore changing away from a FSK type scheme does not seem desirable, or necessary. Only a 600 b/s bit rate was achieved, but the transmission of TCs does not require a higher data rate, thus a higher complexity.

However, a potential future software upload feature, or other more complex uplink signals, might. In that case, the primary direction of advancement would be:

1. <u>Increasing the data rate with the same modulation scheme</u>
   For Delfi-n3Xt higher data rates might be useful, but in any case it would create a more powerful (future) satellite bus.

Although it might not be directly required, a number of changes can be made which would adjust for the wasteful aspects of the current modulation scheme. First of all, FSK is not very power efficient. Second of all, Manchester coding is very bandwidth efficient.

The first option can be addressed by:

2. <u>Changing the modulation scheme from (B)FSK to MSK</u>
   It has been explained in chapter 7 that MSK can actually be derived from BFSK by applying NRZ-I coding and choosing two frequencies with a distance of h = 0.5. As NRZ-I, as explained below, is also performed, only the frequency distance should be changed. In case of a changing data rate, this is automatically done in fact as h depends on data rate.

MSK yields an increased error performance of 3 dB; theoretically not directly required but a worthwhile improvement if at minimal cost. An added benefit of MSK is that the produced bandwidth is strongly reduced. The bandwidth inefficiency of Manchester coding is addressed below.

<u>Line Coding</u>

As explained for the downlink, the default line code is unipolar NRZ. FSK modulation does not require this to be converted to bipolar NRZ. For design facility however, NRZ-I can be applied; in that case (COTS) modulation and demodulation hardware both do not need to agree upon which frequency represents which bit (0 or 1). The same holds for the couple of processor and (de)modulator in the receiver and transmitter. Furthermore, NRZ-I encoding on both the downlink and uplink allows the flags of the data protocol used to be the same. As such, the downside of 0.8 dB reduced error performance and slightly increased complexity are acceptable; the former because of the huge link margin compared to only 0.8 dB loss, and the latter because of its proven unimportance, as is was applied in the Delfi-C^3 uplink.

In order to be as simple and resilient as possible, Manchester coding can be applied. Manchester code, or Bi-Φ-L as introduced in section 7.3, is a very simple code to program, guaranteeing *a transition per bit* and *an overall zero dc-component per bit*. It does come at the cost of a halved effective bit rate, as it is a 1B2B code.

- *<u>Guaranteed transitions</u>*: not only does a high transition density allow maintaining symbol synchronization, a *guaranteed* transition per bit allows achieving this synchronization even more simply. No bit duration estimation based on for example a Costas loop is required, as every bit allows for a unique and new identification of the bit length, independent of Doppler shift or other impacts. A transition can be expected within a preset bit length, adjusted for the maximum offset. As Doppler shift in the transmission frequency in the current ground station set-up is updated once per second, the bit length change can be rather abrupt.
- *<u>Zero dc-component</u>*: if a dc-component exists, for example, a string of 1's can become interpreted as 0's because no centre dc level has been defined to signify a 0. The importance of this to the demodulator is dependent on the electronic design.

While particularly simple, Manchester coding is also quite wasteful. Therefore an optional improvement would be:

3. <u>Removing Manchester coding for an automatic doubling of the bit rate</u>
   Manchester code has a bandwidth of twice that of regular NRZ. As such, a 600 b/s bit rate requires a 1200 Baud symbol rate. When removing the Manchester code, changes are only occurred within a select number of components as most components are prepared for a symbol rate of 1200 b/s. At the same time however, a doubled bit rate is obtained.

Some comments can be made with respects to the impact of removing the Manchester code

Transitions: currently (in AX.25 and the DelfiX protocol, see next section) a flag includes a small period of no bit transitions, so that knowledge of the Manchester code allows locking onto the only other transitions; those in the middle of bits. If on the contrary the flag would include forced bit transitions, similar locking could take place, but focussed not on the middle but the edges of bits. After the flag, transitions are no longer guaranteed so the processor should estimate the length of a bit, and control this estimation by regularly recurring transitions. This latter process is in indeed generally applied for symbol synchronization. The main downside of this approach is the added complexity of these control loops. If therefore the Manchester code would be removed, more processor complexity is required, along with the time investment of developing software. However, as on Delfi-n3Xt MSP430s will be applied as opposed to much less powerful PIC controllers, it might indeed be feasible.

DC-component: as assessed above, the existence of a dc-component might be of concern. This then might increase implementation complexity, with implications on all electronics. A possible solution could then be offered by more complex coding schemes, such as other mB(m+1)B line codes. For example a 3B4B Griffiths code has 75% data efficiency, as opposed to 1B2B Manchester with 50% efficiency. Even better would be a 5B6B Griffiths code with 82.3% efficiency, while still limiting the number of 0's or 1's to 6-in-a-row. However, these codes introduce more substantial processor load.

The resulting incurred complexity of the possible changes in the satellite and on-ground again is commented on below in a discussion on modulator and demodulator.

<u>Pulse shaping</u>

Figure A-4 shows that for coherently modulated FSK, the signal dampens relatively quickly. Nevertheless, the required reduction of -60 dBc is not reached within a bandwidth shown. In fact, even for MSK this limit is not reached within a bandwidth of five times the data rate. However, pulse shaping applied on FSK schemes requires more complicated modulators, due to the required generated frequencies between the two main transmission frequencies.

The bandwidth performance of a transmitted Delfi-C^3 telecommand has been measured and is shown in Figure 8.5 below. The frequency span illustrated is 50 kHz. The decimals of the transmission frequency have been censored for security reasons.

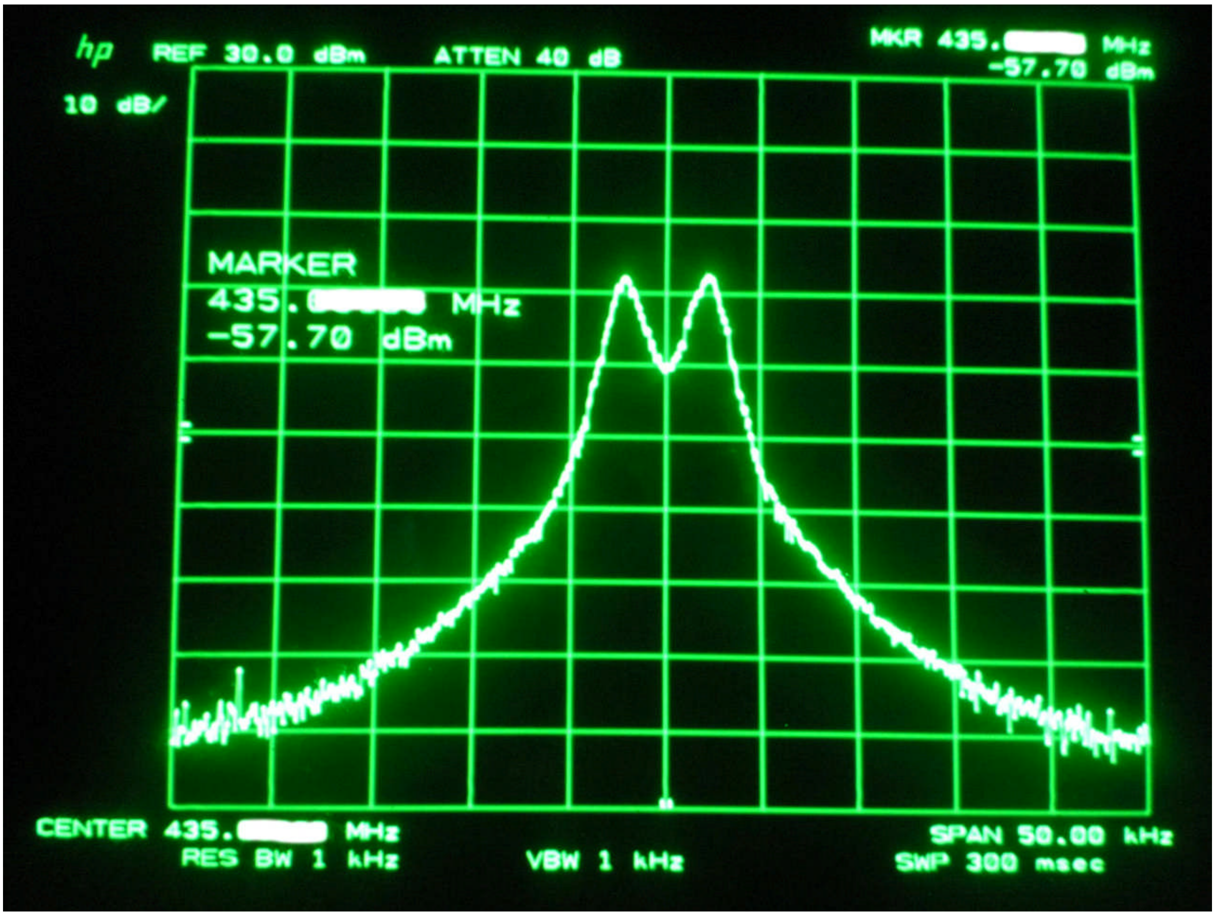


Figure 8.5: Produced frequency spectrum of a Delfi-C^3 telecommand transmission over a 50 kHz span

First of all, the two peaks belonging to both modulated frequencies in a BFSK scheme can clearly be seen. What is however interesting to note, is that the peaks are *4.8 kHz* apart. The frequency of the 1200 b/s symbol stream is however only 600 Hz; this would require only a separation of 0.6 kHz. In other words, *the FSK modulation applied is not ideally matched to the transmission data rate*. In terms of a parameter introduced in chapter 7; h = 8 instead of 1. The reason for this is quite simple; the applied COTS equipment has been prepared for a maximum data rate allowable within the radio amateur bands. The effect of this however is that produced spectrum is larger than required.

Second of all, the overall produced spectrum can be seen. Although the spacing between the peaks contributes to a larger spectrum, the main effect results just from the broadness of the modulation scheme applied. To reach the regulatory -60 dBc, about **40 kHz** is required. This is again *far* more than the 2.4 kHz allotted. Although filtering is applied within the transmitter, *this filtering is again prepared for a data rate of 9.6 kb/s*. Even if it were not, the regulatory values would not be reached.

Nevertheless, it has been said before and will be said again, the actual regulations on the radio amateur bands are not that strict. Even less so for an uplink, which has only very (geographically) local impacts. To this it can be added that pulse shaping applied to an FSK scheme is difficult.

In conclusion, no pulse shaping on the uplink is required.

8.4.3 Scheduled implementation of the modulator

As opposed to the downlink, modulator complexity if of less importance to the uplink as it takes place on ground. But in fact, modulator complexity is also very low, as BFSK modulation can be performed by the simple architecture sketched in Figure A-2. The implementation of the modulation scheme on ground will then be as follows:

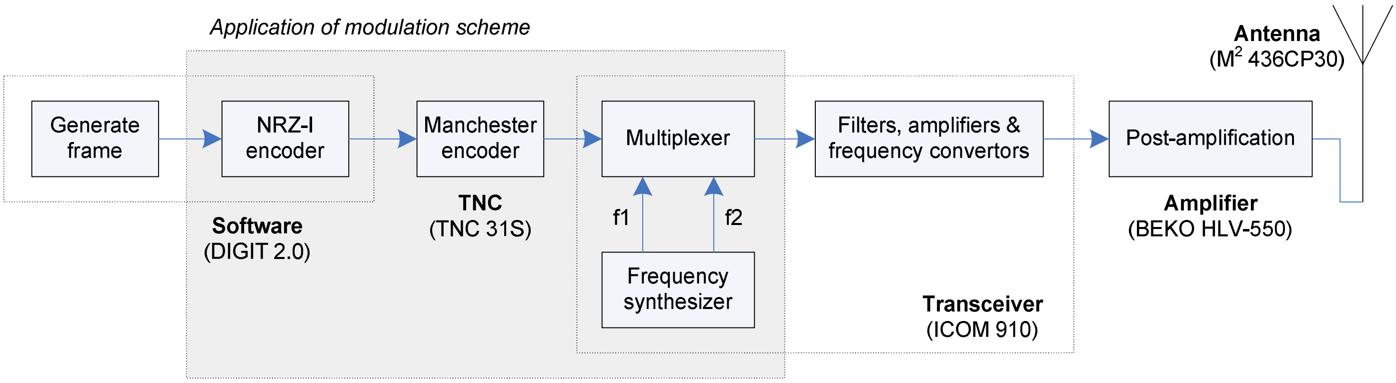


Figure 8.6: Default integration of the Delfi-n3Xt uplink modulator

Figure 8.6 demonstrates that a new program, suggestively called DIGIT 2.0, should basically prepare a stream of data (according to the data protocol), encoded with NRZ-I. The original DIGIT (1.0) prepared Delfi-C^3 telecommands so it should at least be updated.

However, in case of Delfi-C^3 the TNC31S took on part of data protocol application; it received data frames according to the KISS protocol, and replaced this protocol by AX.25; including the NRZ-I encoding. As these are software functionalities, the execution of them can relatively easily be shifted. Changes to the data protocol (see next section) leads to prefer integration in DIGIT, also circumventing the use of the extra KISS data protocol. Software on the TNC31S can also be updated.

The TNC31S should in any case take care of timing the bits, as a data stream coming from a desktop PC is unreliable in timing due to delays and queuing. Manchester encoding is also performed by the TNC hardware.

Consequent frequency modulation of the signal, as well as filtering, amplifying and up-converting of the signal is performed by the ICOM 910. An additional amplifier, the BEKO HLV-550, is generally used to further amplify the signal, as explained in chapter 15.

The proposed improvements are discussed below. In the cases where the effective data rate is increased, the software should also allow a larger data rate output, but this has little impact given the assumed large on-ground processor power. The TNC is as said concerned with consequent timing of the bits, resulting in a constant bit period.

The color codes used to illustrate the impact of implementation of the suggested advancements are the same as in the previous section. For clarity:

> Green indicates little to no changes with a good idea of the impacts
> Blue indicates moderate changes with some unknown impacts
> Red indicates a serious change with unknown specific impacts

1 Increasing the data rate with the same modulation scheme
An increase of data rate affects the following components:

- ***TNC31S***. Requires an available replacement
  The current TNC31S supports only a 1200 Baud signal, in other words a 600 b/s Manchester encoded signal, and should thus be changed. The TNC currently incorporates a specific modem that applies the Manchester coding to a specific bit rate; this modem can be replaced by a modem available at ISIS allowing up to 115 kBaud.
- ***ICOM 910***. Acceptable without changes
  The ICOM 910 can actually support up to 9.6 kBaud FSK signals, with a maximum filter bandwidth of 15 kHz. Therefore, up to 4.8 kb/s Manchester encoded data signals would be acceptable.

2 Changing the modulation scheme from (B)FSK to MSK
This change would require changes in one component:

- ***ICOM 910***. Requires a unidentified component update
  The value of *h*, the distance between the FSK frequencies, depends on the bit rate, and it determines whether one deals with 'regular' FSK or FSK with *h* = 0.5, MSK. Therefore a component should be found that creates an MSK signal for a given bit rate, instead of the current ICOM 910. This application does not seem unique and a COTS component replacement should be available. Nevertheless, no such component has been identified yet.

3 Removing Manchester coding and doubling the data rate
The option of removing Manchester *and* doubling the data rate would have one effect:

- ***TNC31S***: Requires a unidentified component update
  The modem which is integrated in the TNC should be replaced, as the current hardware applies Manchester encoding.

8.4.4 Scheduled implementation of the demodulator

The demodulator should be as simple as possible. Part of the solution is to use COTS components. However, just as with the modulator circuit of the downlink, the frequency conversion system is custom designed. The integration of the total demodulator and its (internal) functionalities can be seen in Figure 8.7 below.

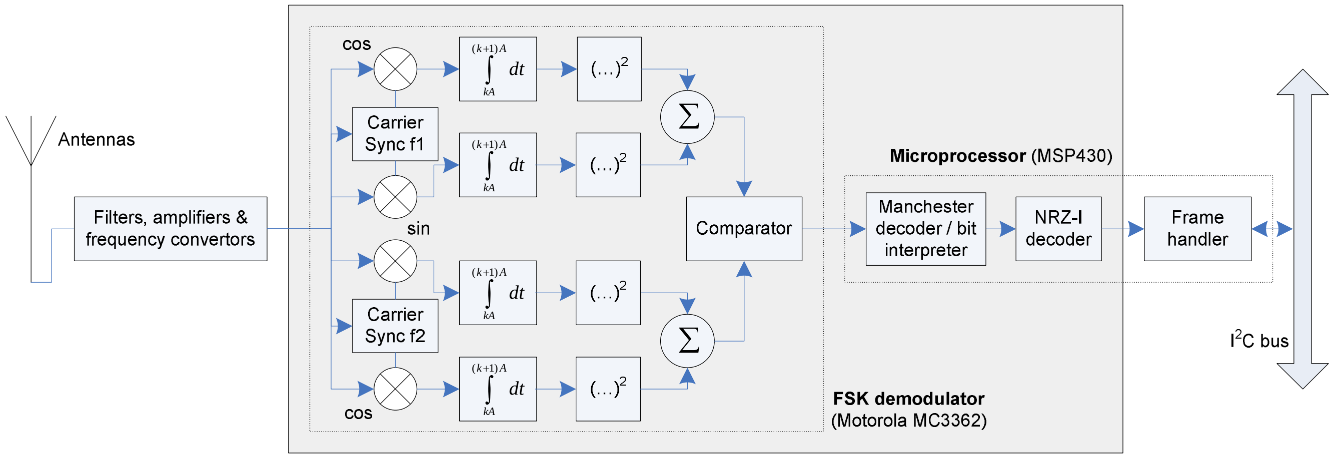


Figure 8.7: Default integration of the Delfi-n3Xt uplink demodulator

After filtering, amplifying and frequency conversion the signal is passed on to the Motorola MC3362. This chip is in charge of extracting the carrier frequencies from the signal (only in terms of frequency, not phase). It yields an output at each moment consisting of a *low* or *high* signal; not to be confused with digital data as the period or *clock* is not defined. This system is explained in detail in Appendix A.

The consequent high/low signal should represent a Manchester encoded bit stream, which is decoded by the *Frame Interpreter* in the form of an MSP430 microprocessor. This same processor then removes the NRZ-I code, handles the frames and communicates with the I^2c bus.

Whereas the demodulation seems really complex, even much more so than that of BPSK reviewed above, it is in fact not so. Whereas Doppler shift requires carrier synchronization, the FSK demodulator allows having only carrier frequency synchronization. Consequently, the use of Manchester coding renders complex symbol synchronization unnecessary. Finally, as can be seen in Figure 8.7, most complexity is integrated in one COTS chip.

1 Increasing the data rate with the same modulation scheme

An increase of the data rate would involve the following components:

- ***Motorola MC3362***. Requires replacement *but* one that is most likely already necessary
  Due to a change of data rate, the required carrier frequencies of the FSK signal change (*h* relation). As such, it is not automatically true that this chip would support these frequencies. In fact, its datasheet [SLR 0199] mentions that a data rate of 1200 Baud, due to the Manchester coding, 600 b/s, is the typical limit to ensure data integrity. This can also be related to the fact that the chip bases its integration period not on the bit length, but on a period presumably much smaller than that. A halved bit period might complicate this approach. As such, this component requires replacing. *However*, this component is already out of production, so it is likely that it should be replaced in any case.
- ***Analogue components***: Requires replacement *but* one that is most likely already necessary
  The relevant bandwidth contained within the signal received by the receiver is doubled by a doubled data rate, more so for higher data rates. The last filter is only 5.5 kHz broad, so might unnecessarily decrease received signal quality for data rates at least above 1.2 kb/s. However, this filter is part of the Motorola MC3362 so should be replaced in any case. Also, this filter might in fact not be required in the first place.
- ***Frame Interpreter***: Data rate of 1.2 kb/s likely acceptable, possibly higher
  The microprocessor, interpreting the Manchester code and removing it to yield bits, and consequently removing the NRZ-I code, should do this functionality twice as fast for a doubled bit rate. At the same time, it should still perform its frame extraction functionality, most notably including *decryption* and the calculation of *two CRCs*. It capabilities should be tested. *Nevertheless*, on Delfi-C^3 the OBC, which was the same MSP430 processor as the Frame Interpreter is, was in charge of these taks *as well as other tasks*. It interpreted the bit signal from two (RAP1 and RAP2) radios at the same time. As such, it seems likely that at least a doubled bit rate is achievable.

2 Changing the modulation scheme from (B)FSK to MSK

When implementing MSK the following components have to be thought of:

- ***Analogue components***. Acceptable without changes
  This optional change is actually one of luxury; it can be decided to limit the filter bandwidths as MSK requires less transmission bandwidth as shown by Table A-1.
- ***Motorola MC3362***. Requires replacement *but* one that is most likely already necessary
  The data sheet of the chip gives no answer as to what values of *h* it supports, only that it is fabricated for FSK modulation. As such, possibly it might be useable. But, given the fact that the model is already out of production, it is preferred to check for this specification in a successor.

3 Removing Manchester coding and doubling the data rate

Removing Manchester would require a new method of symbol synchronization. Furthermore, as an unknown and slightly varying dc-component arises this should be designed for.

- ***Motorola MC3362***. Requires feasibility test
  The specifications for a resulting dc-component in the demodulated signal are unclear, as the chip is a 'black-box'. It should therefore be tested, possibly in its successor.
- ***Frame Interpreter***. Requires feasibility test
  Without Manchester coding, a cleverer symbol synchronization method should be integrated in software (most logically), this comes at the cost of complexity and thus required processor power. Again this would come on top of decryption and CRC calculation, and the doubled data rate.

Also required for the last option:

- ***Changing the flags of the data protocol of the uplink signal***. Requires minor changes

  Albeit not a physical layer functionality; important not to forget. In order to have a well-functioning symbol synchronizing circuit, this improvement would require an adapted flag, with a sufficient number of transitions. More about this in the next section on data link layer functionalities.

8.4.5 Determination of the data rate approach

In case of the UHF uplink, power efficiency if of no concern. As such, changing data rates with elevation angle would not be useful. The system is, as concluded above, *complexity limited*.

As a result, following the same arguments as for the VHF downlink, a minimum elevation angle of 10° is designed for in the link budget.

8.4.6 Selection of modulation scheme of Delfi-n3Xt

Again three options of improvement have been suggested.

1 Increasing the data rate with the same modulation scheme

Increasing the uplink data rate to above 600 b/s requires, on-ground, a change of the TNC integrated modem, which is no problem. On-board however the receiver requires a new a new demodulator chip as well as possibly a changed filter approach and minor software changes. Also, it should be tested whether the processor is able to handle increased speeds, however speeds of 1.2 kb/s are likely possible.

As such, it can be said that a possible data rate increase depends mainly on the electrical integration complexity in the receiver. Currently no one performs work on the receiver section of the PTRX. As such, the possible replacement of the demodulator remains to be commented on.

It can however be said that an increase of the uplink data rate is *not* required for the transmission of telecommands. It might be useful for a future software upload capability and would thus create a more powerful radio. For Delfi-C^3, a bit rate of 1.2 kb/s was in fact envisioned, but in the end was not realized due to reasons of complexity. For Delfi-n3Xt it therefore does seem desirable to increase the data rate as an improvement, but no significant resources should be devoted to it. As such, a data rate of 1.2 kb/s is set as a target. At the same time, the 'improved RAP' that ISIS has made works at 1.2 kb/s (not more). As such, this seems a reasonable goal. *The final data rate currently shall be assumed to be either 0.6 or 1.2 kb/s*.

2 Changing the modulation scheme from (B)FSK to MSK

The implementation of MSK requires a change of the ICOM 910 and the demodulator chip, having an unknown impact. The ICOM 910 does have to be replaced already in case of higher downlink data rates.

However, in the best case this upgrade gives an additional link margin of 3 dB, next to the link margin of *more than 20 dB* already available. Therefore this update is considered a luxury option, but is not to be actively pursued.

3 Removing Manchester coding and doubling the data rate

The implementation of no Manchester coding along with a doubled bit rate would require a change of the demodulator, as well as increased complexity of and time investment in the Frame Interpreter. As this is a mission-critical element, added (un-proven) complexity is not much preferred. Certainly as the increase of data rate to 2.4 kb/s is of unknown relevance to Delfi-n3Xt.

Given the limited project occupancy within the Delfi-n3Xt programme, and the fact that the primary functionality (transmitting TCs) can easily be performed using the same modulation scheme as Delfi-C^3, most proposed changes above are, for now, of more academic importance.

However, an increased bit rate should be further investigated in cooperation with a future student in charge of the PTRX (transmitter) as its potential value to the Delfi programme is recognized.

In conclusion:

Delfi-n3Xt:

Modulation scheme: FSK (h = 1)
Line coding: NRZ-I (NRZ-S) followed by Manchester coding
Pulse shaping: None
Bit rate: 600 b/s or 1200 b/s
Indicated bandwidth: 2.4 kHz or 4.8 kHz

Alternative names for the modulation scheme:

- Orthogonal FSK (*h* equals an integer number)
- Sunde's FSK (*h* equals 1)

Alternative names for the combination of modulation scheme and pulse shaping:

- Manchester FSK

8.5 Data link layer: data protocol

Data link layer functionalities deal with bits and are necessary to establish a link between two stations. The required feature is the data protocol; it supplements the modulation scheme by offering synchronization capabilities and it adds addressing and possibly error detection or correction.

In case of the physical layer there is a functional distinction between downlink and uplink. In case of the data frames and operation, there logically is a difference between not only up and down, but the type of communications as well; telemetry downlink, telecommand (TC) uplink, telemetry requests and possibly software updates. These cases will be referred to as *scenarios*. These functionalities are also strictly no longer *data link layer* functionalities, but those belonging to higher layers as they are not strictly required for establishing a link between two stations.

What will be defined as the data protocol *shall be equal for all these scenarios*. This section introduces this data protocol. The next section then addresses scenario-dependent data functionalities.

The AX.25 protocol is extensively used for communication, and has successfully been applied in Delfi-C^3. It has been introduced in detail in chapter 7. It is usable for a wide range of applications. Due to this multi-functionality however the overhead created is rather large. For example, it incorporates very large address fields allowing for both a sender and a receiver, and a PID and a Control field which, as demonstrated by Delfi-C^3 have no use. Even more so, in case of Delfi-C^3 RASCAL was used and required for the forwarding of telemetry; as such no standardized equipment allowing for AX.25 decoding was preferred as this did not allow for telemetry forwarding. Without RASCAL frames could be received, but not interpreted. Therefore it seems the protocol can be adjusted without much loss of functionality. The idea suggested here is therefore to base a new design on AX.25, which is a proven protocol, but to improve its efficiency for the mission at hand. In fact, the resulting protocol is one that is useful for all similar satellite missions.

The newly derived protocol, the DelfiX protocol, will be derived in two steps. First, the AX.25 protocol is reduced without changing the transmission characteristics, to form the *conservative DelfiX protocol*. Afterwards the *final DelfiX protocol* is presented, taking into account another improvement. Figure 8.8 below demonstrates the conservative DelfiX frame.

Flag *1 byte*	**Sender Address Field** *3 bytes*						**Frame Type** *1 byte*	**Information Field** *n bits*	**FCS** *2 bytes*	**Flag** *1 byte*
01111110	1	7-bit ASCII character	1	7-bit ASCII character	1	7-bit ASCII character	8-bit number		16-bit CRC value	01111110

Figure 8.8: DelfiX frame, conservative version

In general, the characteristic of having byte-length fields is retained due to 8-bit processor calculations and therefore modularity. Nevertheless, the information field, which is not further discussed below, can have any content and any number of bits as the protocol itself is not dependent on an integer number of bytes. Also, *all fields are filled most-significant bits first as opposed to with the AX.25 protocol*; in other words all bits and fields are read logically, left-to-right.

The result is a *saving of 96 bits* with respect to the AX.25 protocol, keeping its essential functionalities and adding a Frame Type field in a total of 64 bits. Also, using the same CRC as AX.25, single bit error correction can possibly be achieved. *However*, a large amount of bitstuffing is required due to the small flag size. This aspect is improved upon in the *final DelfiX frame*; by extending the flag to two bytes instead of one, maximum required bit stuffing is reduced from 1 in 6 to 1 in 14 bits. This is illustrated by Figure 8.9 below.

Flag 2 bytes		Sender Address Field 3 bytes						Frame Type 1 byte	Information Field n bits	FCS 2 bytes	Flag 2 bytes	
01111111	11111110	1	7-bit ASCII character	1	7-bit ASCII character	1	7-bit ASCII character	8-bit number		16-bit CRC value	01111111	11111110

Figure 8.9: DelfiX frame, final version

The characteristics of the DelfiX protocol and the changes with respect to AX.25 are elaborated on in the subsections 8.5.1 through 8.5.5 below.

Figure 7.2 below illustrates how the both frame types should be constructed, from the point of view of the transmitter; in case of Delfi-n3Xt the microprocessor on the PTRX, or the ITRX. The OBC then provides Information Field plus a Frame Type label, which depends on the information content (it is still part of the DelfiX protocol as it is present for *all scenarios* introduced below). In an order similar to that of the AX.25 protocol, fields are added and bitstuffing is applied. A note on the effects of the moment of bitstuffing can be found in subsection 8.5.6. A receiver should apply the process shown in Figure 8.10 in reverse.

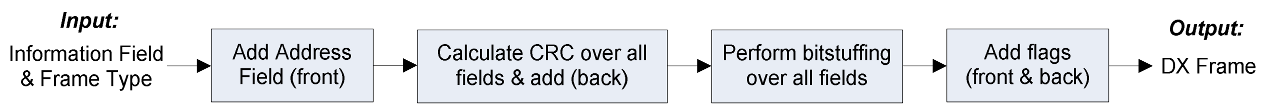


Figure 8.10: DelfiX frame construction

In subsection 8.5.7 throughput efficiency of the DelfiX protocol is commented upon, yielding conclusions on the optimum frame size, depending on bit error rate.

8.5.1 Flags

The flags have two functions:

- Indicate the start and end of a frame
- Allow for carrier and/or symbol synchronization, depending on the modulation scheme

The flags are thus used not only for interpreting data, but also for synchronization required as part of demodulation. The start flag 1's yield a perfect sinus when combined with BPSK modulation as no transitions occur, which allows for carrier synchronization (this sequence is unaltered by NRZ-I encoding, as this changes the stream of 1's to 0's). In case of Manchester encoded FSK modulation it yields continuous transitions, but only in the middle of bit periods (due to the Manchester code and not due to bit transitions) which allow for symbol synchronization.

An end flag is not necessarily required if packed length is known. Nevertheless, bit stuffing introduces some variability in length, or its definition should be changed to insert a default number of bits as no flag has to be prevented in the frame content if no end flag is used. In any case, bit stuffing itself is required for keeping symbol synchronization, as addressed below. The end flag can also be replaced by a frame length field. However, the demodulation software is simplified and data protocol modularity is increased by keeping it.

Furthermore, in the AX.25 protocol it was allowed to combine the end flag of one frame with the start flag of another frame. This will, at least in case of Delfi-n3Xt, *not* be applied. There is no good reason to do so given the transmission scenarios; due to a varying amount of bitstuffing per frame length frames will not seamlessly cross over.

Due to the function of the synchronization sequence as start and end flag, bit stuffing is applied to make sure it does not reappear in the frame; after 5 1's a 0 is inserted *in the AX.25 protocol*. At the same time however, the insertion of '0' bits creates a transition, following the NRZ-I encoding. Therefore, transitions are realized every 6 bits maximum, allowing for *keeping symbol synchronization in the downlink signal*. (The uplink signal does in fact not require this feature due to the Manchester coding. Nevertheless, as the primary uplink signal of TCs consists of small packages, it has no significant effect.)

This exact functionality has been copied in the conservative version of the DelfiX protocol, as *bit stuffing affects also signal demodulation*. However, (digital) symbol synchronization does not seem to require such a large transition density, as confirmed by multiple internet sources. Therefore, it is suggested that the flag length be doubled to a frame of 14 1's, flanked by 0's, creating the *final DelfiX protocol*.

Using a custom-written Matlab modulation program, applying NRZ-I coding and BPSK modulation to a 1600 Hz carrier as that being used to demodulate the signal in RASCAL/DUDe, the effects of streams of 1's have been tested. Matlab was used to generate a digital audio file that could consequently be played. Using an audio cable the signal was monitored by RASCAL/Dude, in a way identical to how the ground station listens to the satellite. A continuously increasing stream of 1's, alternated by a number of flags for re-synchronization was played.

The results:

- The program showed remarkably little problems with streams of up to 30 1's in-a-row.
- The same held true when a Doppler shift of 50 Hz was added, which is about twice the maximum Doppler shift rate per second; Doppler shift is corrected for every second by the receiver on-ground.
- Finally, as signal power was decreased (by adjusting the volume of the transmission, simulating a reduced signal-to-noise) the seamless recovery ability was slightly reduced but not by more than 5 bits in-a-row. At lower power the entire stream became unreadable. The realistic signal-to-noise ratio should than have laid somewhere in between.

As such, preliminary tests show that a doubled flag length, to form a continuous stream of 13 1's in the worst case, should likely be no big problem. Of course, the same test should again be done as the actual analogue section of the transmitter is finished, to confirm the implementation of the final DelfiX protocol.

Due to the constant data generation of Delfi-n3Xt, downlink frame length is constant. Bit stuffing then yields a number of bits to be reserved for bitstuffing, *assuming the worst case scenario*.

Increasing the header length to two bytes comes at the cost of two bytes in the complete frame, at the *gain of more than half the reserved bit stuffing bits*. In a frame length of 1200 bits, this decreases the bit stuffing bits from 200 to 86. An enlargement of another 8 bits for the flags would only save another 30 bits, while costing 16 bits. Of course this effect would be bigger for larger data frames, but the above test results show that larger streams of 1's could result in demodulation problems. Also, moderate room assumed for bit stuffing, if not required in a certain frame with plenty transitions, can be used for appending flags useful for signal synchronization.

8.5.2 Address field

Two changes with respect to the AX.25 address field are made:

- There is no longer a sender *and* receiver field
- The total length is reduced to three bytes.

In case of simple satellite mission, there is only one ground station and one satellite. As such, a satellite is logically linked to the ground station and stipulating only one of the two gives enough information. In case of Delfi-n3Xt, as with Delfi-C^3, there is only one true ground station, but many receivers. Nevertheless, the addressing is always the same and on-ground software (DUDe and RASCAL respectively) takes care of the distribution. As such, no receiver field is proposed.

The AX.25 protocol allows many users and as such many address bytes are required. 6 ASCII characters are allowed per sender/receiver, followed by a SSID or Secondary Station ID. In fact, this means 7 bytes, or 42 bits are used for an address. This field can be minimized by using a numbered approach; 8 bits can allow for 256 'address numbers'. It is here that a trade-off should be made between minimizing bits used, and modularity.

A small number of bits is not preferred, as a bit sequence or a (hexadecimal) number should be remembered for a satellite (or satellite-ground station combination). This gives little oversight when multiple satellites use the same protocol, be it future Delfi satellites or even other satellites. Therefore, the use of ASCII characters is deemed useful. To make meaningful acronyms, yet give a large number of options, a 3 byte address field is selected; allowing for not only letters but also numbers and symbols. It is then suggested to use:

- **DNX** for Delfi-n3Xt; indicating a downlink signal
- **DGS** for Delfi ground station; indicating an uplink signal

As flags are similar to AX.25 frames, as well as the CRC, and the address field starts off similar to that of the AX.25 frame, it makes sense to guarantee that the difference between the two is clear; an AX.25 frame receiver should reject these frames, and a DelfiX frame receiver should reject AX.25 frames. A dual receiver could switch modes when it recognizes the data protocol. A clear marker of difference is integrated *with the 1st bit of the each address byte*; in case of the AX.25 frame these bits are set to '0' to indicate that the address field continues, with a mandatory number of 14 bytes minimum. As such, it is proposed to set all 1st bits to '1'. In fact, in AX.25 frames the field contents are inverted ('least significant' or last bit first, per byte). The inversion is deemed not useful, and is not applied in the DelfiX protocol.

8.5.3 No PID and Control fields

Both the PID and Control fields in the AX.25 protocol were of no use to Delfi-C^3, as only the UI-frames were used and no higher layer protocols were specified. Nevertheless they took up 16 bits and as such they are removed in the DelfiX protocol.

8.5.4 Frame Type indicator

As was mentioned above, a number of different scenarios exist, yielding different frame compositions when the info field is taken into account. These scenarios are (currently) telecommand, telemetry, telemetry request and software update. More can be added when required. Different telemetry types such as housekeeping data and payload data can also be indicated and even combinations of payload data if not all payload data will fit in singular frames. Therefore 4 bits are deemed necessary. In order to stick to an integer number of bytes, it is extended to 8 bits; 4 bits can be used to indicate the generic frame type (scenario) and another 4 bits can indicate the payload frame type in case of telemetry data, for example. In the current data budget these two counters exist and are separated in location. *This is to be further specified by the CDHS subsystem engineer who is in charge of determining frame content.*

8.5.5 Frame Check Field / Sequence (FCS)

In case of an AX.25 frame, a CRC check is applied, which is a form of error detection. Single, double and triple error can be detected within a frame of a maximum of 32,768 bits, as well as burst errors up to 15 bits long, using a 16-bit CRC. Section 7.7 has given an introduction to data security measures. As any frame to be transmitted using the DelfiX protocol will be subjected to a certain bit error rate, it makes sense to add certain measures to make sure corrupted data is not accepted in the database. Hash and encryption measures are not designed for this purpose, and neither are advanced error correction schemes. The latter yields large overheads which are not preferable in a general data protocol standard. Therefore, *cyclic redundancy checks or CRCs are suitable*.

The total *BER* for which Delfi-n3Xt transmission links are designed is 10^{-5}. This means that, depending on the frame size, a certain number of frames is slated to have errors in them. A single error will occur on average every 100,000 bits, or roughly every 40 frames with 2400-bit length, or every 20 frames for 4800-bit frames. This is a sizeable number. Double errors are however slated to occur some 100,000 less often, as presented in section 7.7. Nevertheless, with one frame per second, one such frame will exist every 50 days or 25 days on average. This suggests single error detection is at least necessary, but a reasonable argument is posed for having larger error detection of at least 2 bits, next to resulting in a more general sense of safety and system with some margin for failure.

In reality, the *BER* will not be equal to 10^{-5}, but higher in nominal operations. There is at least a small link margin on the VHF downlink (in addition to the minimum link margin of 3 dB), and a huge link margin on the UHF uplink. Nevertheless, if margins turn out to be lower (most noticeably on the downlink), the system should not fail on error detection capability. At the same time, many amateurs might receive and forward telemetry frames, using less powerful or dedicated equipment. To not underestimate this feature, it makes sense *to slightly over-design the error verification feature*.

In section 7.7 it has been shown that only single errors can be detected using up to 12 bits for the desired message lengths. This is the same result then as a 1 bit parity check, although burst errors can then also be detected. A small increase to a 16-bit CRC then gives triple error detection capability as well as 15-bit burst error detection capability. Therefore, the increase to triple error as opposed to single error detection seems worthwhile, *and a CRC of at least 16 bits is advised*.

If the frame message length will be less than 4097 bits, a 24-bit CRC can yield *HD*=5 (allowing 4 independent errors to be detected), but at larger lengths only *HD*=4 is feasible, and using a different CRC. Slightly higher results can be obtained by 32-bit CRC schemes, but these are optimal for (large) frame lengths not likely for Delfi-n3Xt or similar satellites for that matter. Given the above, there seems to be no good argument to increase the CRC length to be able to detect 4 errors, or slightly more.

In conclusion:

Choice of CRC: CCITT-16 (16 bits) with a generator of $x^{16}+x^{12}+x^5+1$ or 1 00010000 00100001.

Additional argumentation is given below.

The given generator is a general standard as it is applied in AX.25, and because of this it has successful Delfi heritage. Section 7.7 gives two better generators, being those labelled as *200433* and *321353*. The first yields *HD*=6 up until 115 bits total length, while having a maximum *HD*=4 until 28,658 bits. The second dominates CCITT-16 in terms of singular errors as it yields *HD*=6 up until 109 bits total length with the same maximum length of 32,767 maximum total length for *HD*=4. [SLR 0685] does suggest however that the error detection performance for more than 3 errors is better for CCITT.

Also, a maximum achievable 122 bits with a *HD* above 4, 5 in this case, and at the same time giving *HD*=4 for message lengths above 1024 bits, is given by the 0xBAAD polynomial. Its maximum message length is not clear however for *HD*=4.

While a CRC generator change should have no other implications than changing the simple bit sequence in the packaging and depackaging software, it does involve reverting from a proven standard to a more theoretically presented one. Furthermore, as is shown in the next section, the smallest frame, a TC, will not fit the 115 bits total length, adding no extra functionality. The added benefit of any general 16-bit CRC is that it allows for larger (future) data frames involved with higher data rates and it creates a modular and future-proof standard, even though 32 kb will not be reached as it would be better to break up frames at these lengths.

Error correction

An added advantage of a scheme of at least triple error detection capability is that *it can in theory also correct for single error detections*. Three requirements have been presented in section 7.7, which are met; triple errors are very unlikely given the *BER*, the message lengths are not variable during transmission and the generator degree is sufficient.

The incurred complexity of the error correction process is of course completely on-ground. Even if the calculation intensity is too high for real-time correction to be applied, it makes sense that between passes there is easily sufficient processor power for it to be performed. In that case, it should be a feature of DUDe, as erroneous frames are normally discarded in the data interpretation process. These frames can, instead of being discarded, simply be set aside to correct, directly or when processor power is available. As such, the relevant incurred extra complexity is involved with preparing DUDe to apply error correction. As such, *error correction is certainly suggested if project development constraints allow for it*.

Message length

It has been concluded that CRC features heavily depend on the message length. As such, it can be questioned whether a smaller package would not merit a different CRC. In other words, why should the CRC be a data protocol specific feature and not a scenario specific feature? The answer is that it simply is clearer in this manner, and that there is no large incentive to have different CRCs. In case of Delfi-n3Xt the only frame of significantly different length is that of the TC. The TC will in fact, as argued below, have an additional CRC targeted at a small message length. The size inefficiency of having two CRCs is not important as total size is not an issue for TCs at those scales.

8.5.6 Bitstuffing bit errors

As introduced above, bitstuffing under the DelfiX protocol, such as with the AX.25 protocol, takes place after the calculation of the CRC. The effect of this is that bit errors, thus bit flips of bitstuffing bits, or those surrounding them, can have unpredictable effects. Three situations can occur, involving 1 bit error:

- ***An inserted 0 can flip to a 1, being followed by a 0***
  The resulting pattern is recognized as a flag, and the current frame (before and after the 'flag') will (very likely) fail as not all protocols elements are present in the frame.
- ***An inserted 0 can flip to a 1, being followed by a 1***
  The first 0 to follow this sequence of 1's will be removed. The result is that a sequence of bits will be shifted one position to the right, with an erroneous 1 inserted, and a 0 removed at the end. The CRC capability to detect a burst error of maximum length 15 will detect this error. An additional error in the frame however might possibly yield the errors undetectable, but this is very unlikely.
- ***A 1 before (in any of the 5 or 13 spots) an inserted 0 flips to a 0***
  The 0 is not removed and the frame length is increased by 1. The entire bit sequence is shifted to the right, and two 0's have replaced a 1. As explained by the operation of a CRC in subsection 7.7.1, this causes 1 more bit modulo-2 addition to take place then otherwise; the result is that this type of error is always detected. An additional error in the frame however might possibly yield the errors undetectable, but this is again very unlikely.

It has been shown that the CRC takes care of bit errors related to bitstuffing in a sufficient manner. A very small chance exists that two bit flips occur, possibly with one of the cases above, causing a bit error to pass through. Double bit flips have a very small chance of occurrence, whereas in turn the problem combinations have an even smaller chance of occurring, and even against those the CRCs detection capabilities are far from powerless. And, no such error that could take place would and should have a catastrophic effect.

If, in case of a lower *BER* for example, it would be required to protect against these scenarios, bitstuffing can be performed *before* calculating the CRC. This would mean that a larger and unknown number of bits is controlled by the CRC, which is not a problem and comes at a small cost of increased processor power. The CRC should however also be bitstuffed afterwards, and for example a 0 has to be inserted to make sure that the end of the message in combination with the start of the CRC will not incur a flag (as opposed to checking for this scenario). Therefore also a small additional loop should be integrated in software to take care of the extra CRC check. Although this implication is not very difficult, it might lead to confusion as processes have to be reversed. Also, users familiar to the AX.25 order might be confused which can lead to errors. In general it is concluded that *bitstuffing can just be performed after CRC calculation at minimum extra incurred risk*.

8.5.7 Throughput efficiency and optimum frame length

Section 7.6 has presented a discussion on data protocol throughput efficiency, given frame sizes and *BER* values. This section compares the throughput efficiency of the DelfiX protocol to that of the AX.25 protocol and comments on its optimum frame length.

Throughput efficiency

As explained in section 7.6, Delfi AX.25 indicates the use of AX.25 frames with the addition of a frame counter. The final DelfiX frame has been presented above, as well as the conservative DelfiX frame. The throughput efficiencies of the three frame types for different frame sizes are given in Figure 8.11 below.

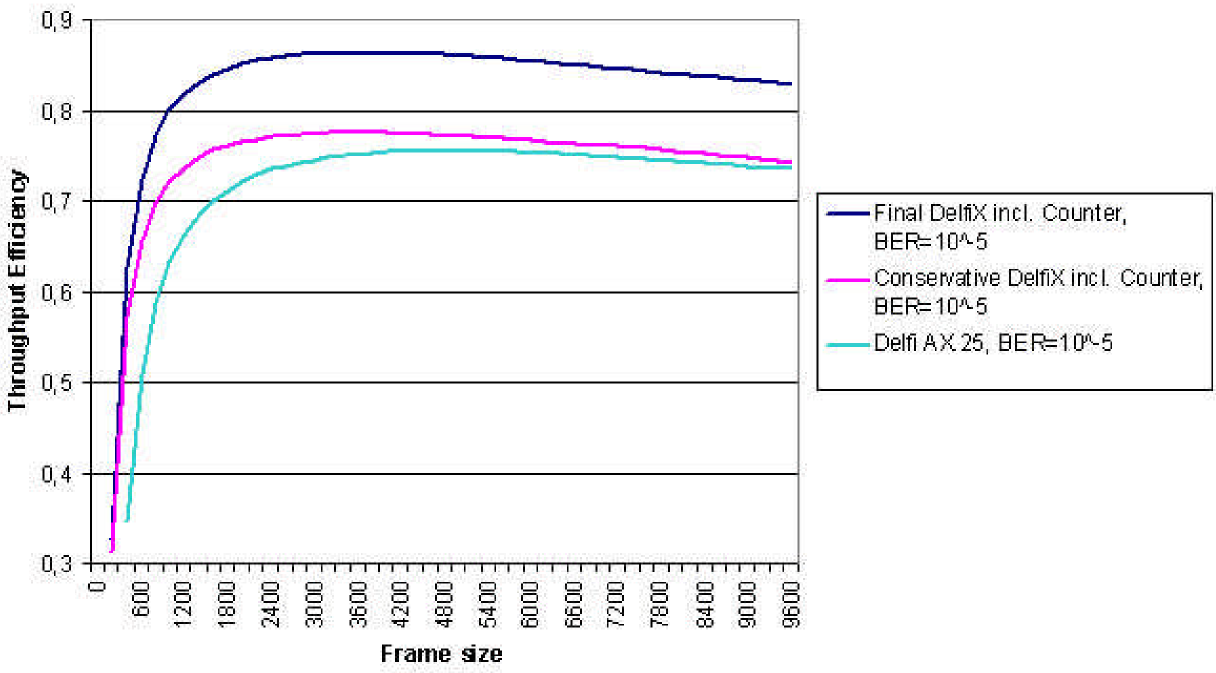


Figure 8.11: Throughput efficiencies of DelfiX versus Delfi AX.25

In conclusion it can be seen that the conservative DelfiX frames outperform Delfi AX.25 frames especially at low frame sizes, with a difference of 8% (74% to 66%) at 1200 bits, 4% (77.5% to 73.5%) at 2400 bits and only 1% (77% to 76%) at 4800 bits. The final DelfiX frames however show very different absolute values, with respectively 82%, 86% and 86% efficiency at 1200, 2400 and 4800 bits per frame.

As such, the final DelfiX protocol can realize some *10-15% better throughput efficiency* than the AX.25 version applied on Delfi-C^3. This can be translated to some 100 received bits per 1000 transmitted bits *extra*, or simply ***10-15% more received payload and housekeeping data during the entire mission***.

Optimum frame length

As can be seen from the slopes in Figure 8.11, at low frame lengths throughput efficiencies are very low, and at a certain point maximum throughput efficiency is reached. The following Figure 8.12 zooms in on the area around this maximum. This allows determining an *optimum frame length*. However, as indicated before, a decreased *BER* increases this optimum frame length. And, as a link budget margin usually remains, a higher optimum frame length is more usual during transmission. As such, no one point is easily classifiable. Nevertheless, it can clearly be concluded that certain low frame sizes yield relatively low efficiencies.

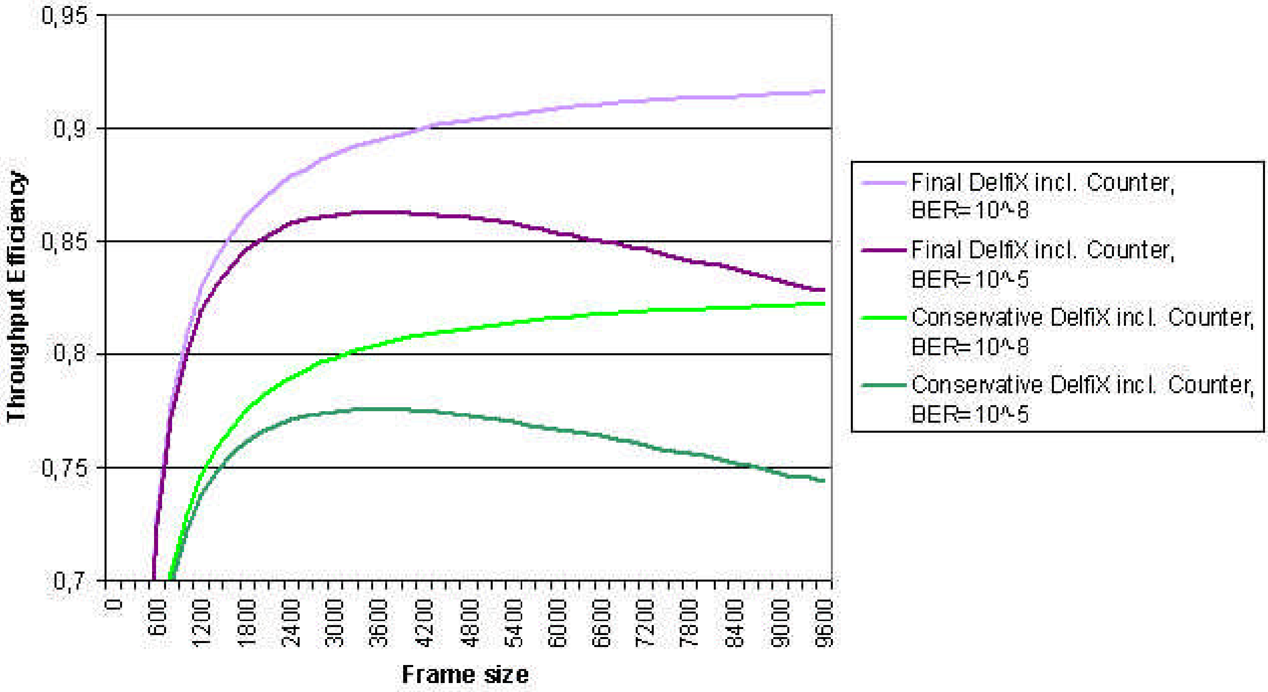


Figure 8.12: Throughput efficiencies of DelfiX with *BER* values of 10^{-5} and 10^{-8}

In case of a *BER* of 10^{-5}, the optimum frame length for both DelfiX versions is 3600 bits, with both 2400 and 4800 bits being within 0.5% distance. Frame lengths up until 9600 bits cause some 3% efficiency to be lost. Data frames of 1200 bits cause some 4-5% efficiency to be lost. As such, *frame sizes of 2400-4800 are preferred*.

In case of a *BER* of 10^{-8}, the optimum frame length for both DelfiX versions surpasses some 200,000 bits. Nevertheless, the increase beyond 9600 bits is actually only some 1%. Decreasing to 4800 bits again causes some 1% loss, and some 2% is lost when moving from 4800 to 2400 bits. *As such, frames sizes of 4800-9600+ are preferred*.

As in reality the *BER* will usually be lower than 10^{-5}, given that this is a worst case value, a wide range of frame sizes can be argued for. As the *BER* varies between 10^{-5} and 10^{-8}, *frame sizes of 2400-9600 have good performance*. 1200 bits is not preferred.

8.5.8 Implementation of a new data protocol

While DelfiX clearly outperforms AX.25 for the Delfi mission, its actual implementation (effort) should also be considered. In case of the UHF/VHF transmission link on Delfi-n3Xt there are two radios that transmit and receive data, and there is the ground station network that is to receive and transmit data. The components that should be considered when changing the data protocol are shown in Figure 8.13 below. The indicated components are commented on afterwards.

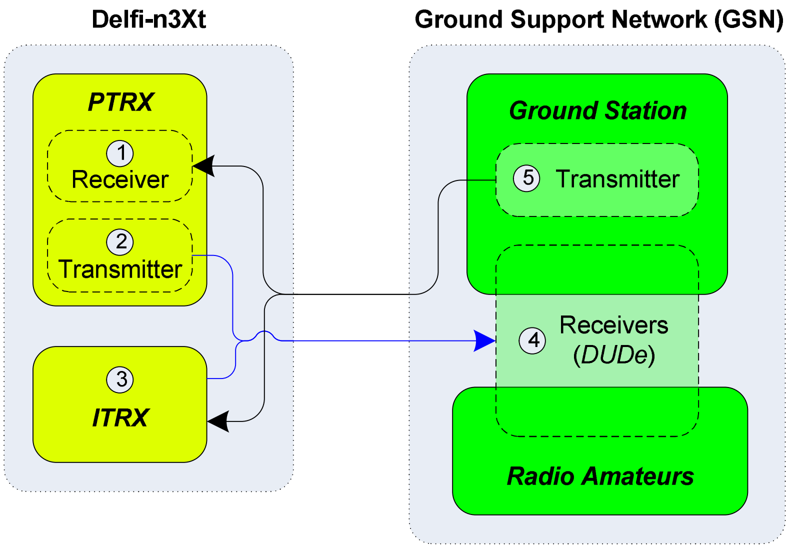


Figure 8.13: Components which are impacted by a change of data protocol

The indicated components:

1. **PTRX receiver**: *updated*.
   As will be introduced in the next chapter, PTRX receiver and transmitter sections are separated functionally. As such, both components should be updated with the new data protocol DelfiX, if indeed both the down- and uplink apply it. The software of the TX and RX sections has already been prepared for the protocol.
2. **PTRX transmitter**: *updated*
   See receiver.
3. **ITRX**: *will be updated*
   The ITRX is formally a payload and to be treated as a black box. It has been discussed with ISIS and this protocol can be integrated without problems.
4. **All ground receivers**: *updated*
   The ground station(s) as well as radio amateur receivers (all further introduced in chapter 15), are slated to use the same software for demodulation and data interpretation: DUDe. The possibility to receive the DelfiX protocol has been integrated in this new program.

5. **Ground station transmitters**: *to be done*
   The ground station(s) hold(s) the capability to transmit telecommands to the satellite. Soft- and hardware is in place to generate telecommands. In case of Delfi-C^3, two components were important: DIGIT, and the TNC31S.

   - **DIGIT**: This software program runs on a PC, and generates telecommands. It should in any case be updated to generate Delfi-n3Xt telecommands. At the same time, the data protocol can be updated. However, currently it generates KISS frames, a data protocol used to communicate with the TNC31S.
   - **TNC31S**: This piece of hardware with integrated software expects KISS frames, and generates AX.25 frames. Thereby it takes away part of the functionality which can be done by DIGIT, however requiring a second data protocol to be used.

   For Delfi-n3Xt it has been suggested already above that software intelligence be shifted to the PC, or to DIGIT 2.0. Then the KISS protocol can be replaced by the DelfiX protocol. Software on the TNC can easily be updated, more easily as functionalities only have to be removed. A sizeable work package still exists in updating the ground station transmitter software, but only a minor addition is required by the implementation of a new data protocol.

It can be concluded that for the downlink, for which the update is most important given the improved effective data rates, the implementation of the DelfiX protocol has no implementation drawbacks (anymore). In case of the uplink, some work is to be done, but it should pose no barrier. However, if for some reason it does, the AX.25 protocol can still be used; only for the uplink.

8.6 Data link layer: different scenarios

As announced above, different scenarios are said to exist. Per scenario, different aspects are important:

- Length per frame and frame rate
- Amount of preamble
- Additional processes such as encryption
- Additional procedures such as digital handshakes
- Additional subfields within the info field

All of these aspects are related to proper communication. The actual (complete) content of the information field is to be determined *by the CDHS systems engineer*.

The telecommand frame is discussed first in subsection 8.6.1, as it determines part of the content of the telemetry downlink (the TC confirmation). Afterwards telemetry frames are discussed as well as its counterpart, the telemetry request frame, in subsection 8.6.2 and 8.6.3. Finally a possible future software update frame is discussed in subsection 8.6.4.

8.6.1 Telecommand frame (uplink)

Telecommands frames or *TC frames* are the shortest frames and the simplest to visualise. Their main information content is a telecommand, portrayed by a number of bits. Delfi-C^3 applied the following approach [SLR 0043].

In total 8 bytes were used per TC:
- A first byte specified a total of 7 types of TC, based on functionality.
- Another byte signified the destination within the satellite, which in fact also gave the parameter to be changed.
- The final 6 bytes allowed for the definition of values, but only 4 bytes were effectively used.

The first two bytes for type and destination have a certain 'minimum distance' between them, so that in case of a *bit flip* no other command results. This holds for data bus errors only, as the encryption (introduced below) will scramble the effects of bit flips during transmission. Transmission error detection is also taken care of by the FCS field.

For Delfi-n3Xt, the exact content is to be determined and again by the CDHS systems engineer. Nevertheless an anticipated configuration shall be given here, to conclude on aspects such as frame length. 8 bytes was largely sufficient for Delfi-C^3, and can likely be reduced. Nevertheless, as the TC frames are by no means large frames there seems to be no necessity to reduce frame length further. Therefore an 8 byte actual telecommand is assumed.

Telecommands are essential to the success of the mission, or at least to the practical success of it. Third parties should not be allowed to send TCs in name of the ground station, thereby influencing your mission. Also, a transmitted TC should not be interpreted as another TC. Thus, security is required. A first step of protection is keeping the transmission frequency secret. Nevertheless, this frequency is decided upon by a third party and is therefore possibly not very secure. In the past, this frequency was even found online. Also, the transmitted signal could theoretically be measured.

Delfi-C^3 applied three other measures of security:

- **Encryption**: The overall message was encrypted; using an ARC4 software cipher.
- **TC verification**: A standard TC was sent to Delfi-C^3, to be followed by a downwards confirmation of the received TC. Afterwards the TC would be confirmed by the ground station by a TC confirmation and executed. Blind TCs were also possible, inferring immediate execution, most notably for emergency purposes. In fact, blinds commands are now often used as the radio in fact switches off sporadically, thereby removing the option for a downwards confirmation.
- **Extra error detection**: An elaborate scheme of multiple additional CRCs, calculated over the TC fields depending on the transmitted message (standard TC, acknowledgement or blind TC) was applied. Three different CRCs were used; one for blind TCs, and two different sequentially applied ones for the standard TC and acknowledgement.

These measures and their reimplementation are commented on below for Delfi-n3Xt.

Encryption

Section 7.7 has introduced the topic of encryption. In Delfi-C^3, RC4 or ARC4 encryption was employed, using one unique 25-character-long key. For Delfi-n3Xt it makes sense to have similar encryption; it proved its functionality without problems while discouraging possible wrongdoers. While it is by no means the most secure encryption, it adheres to complexity restrictions in Delfi satellites and requires a dedicated attack by means of constant monitoring to be cracked. A more detailed research is therefore not deemed necessary. The key can be determined at a later point. Possibly, it can also be made updatable via telecommand. The encryption should be applied over the *actual telecommand*, but not over any of the other fields required for a proper connection. This better illustrated below. The CRC should logically be calculated over the field afterwards.

Also, an *offset* can be applied to the key stream. In Delfi-C^3, the boot counter was used, as the decoder had knowledge of this number via the downlink, and the OBC was in charge of decrypting thus logically had the same information. Nevertheless, this gives little guarantee that the two have the same number all the time, and the boot counter has been stopped after little operation time, thereby being a static number. More importantly however; in case of Delfi-n3Xt the PTRX will decrypt data (see chapter 9) so that its processors do not automatically have knowledge of the telemetry content. Therefore, it is suggested that *the TC code as introduced below is used as the offset*.

This number is not to be encrypted, so that it can simply be read out at the reception of the frame. In result, encryption is calculated *only* over the *telecommand* as introduced below.

TC verification

The sequence of confirmations in Delfi-C^3 avoids satellite-hijacking even more, as it prevents that anyone records the signal sent and re-transmits it. As such knowledge of operations would be necessary and the recording of the two signals transmitted. At the same time it protects against wrongly interpreted TCs. It does however require two TCs upwards and a downwards confirmation. As such for Delfi-n3Xt an improvement is suggested.

Each TC is equipped with a *TC code*. This TC code is simply an 8-bit number. At any time the satellite expects a certain number, which is send down with telemetry. When this number is received in a TC, the satellite sends a next expected number down by means of confirmation. The *next* TC then requires carrying this new TC code. This system gives both a confirmation of the correct reception of a TC, and it prevents that a TC signal can be copied and retransmitted. It does not however protect against the incorrect reception of a TC, nevertheless the error detection field also protects against this, and an *additional CRC* is proposed below.

In order to confirm proper execution, *the last TC processed is also sent down, along with the new TC code.*

A means of blind telecommanding could be the following: a *golden TC code* can be defined. This one TC code from the same set of numbers is not allowed as a regular TC code, and will trigger action without consequently demanding a new golden TC code. This allows for consecutive commands without immediate confirmation, in case of emergencies. A later TC can then alter the golden TC code to a new number.

A special case arises when the transponder or OLFAR is switched on; in this case no TM frames are transmitted. The suggested approach in that case is rather simple and in line with the above scheme. Either the last transmitted TC code, if received, is used, or the golden TC code is used. If the last transmitted TC code is used to switch the transponder off or OLFAR, a new TC code will be transmitted over the downlink as usual afterwards.

Additional CRC

As explained, Delfi-C^3 applied multiple additional CRCs. In fact, other than to create confusion outside of and within the team, there is no good reason to have multiple extra CRCs. To keep functionalities well-separated, it is therefore proposed to have one additional CRC *just to ensure proper reception of the TC*. The CRCs of Delfi-C^3 have as hexadecimal polynomial generators: 0x1021, 0x8005 and 0x1957. All these CRCs are of 16-bit length. Their source is unknown, but as mentioned by the writer of the DIGIT software, Gerard Aalbers, they are based on well-known CRC polynomials.

Chapter 7 has provided an in-depth discussion of available CRC codes and their performances, and gives an optimum alternative. The maximum amount of detectable errors for a field size of 36 to 151 bits is 5, given a Hamming distance of 6. This field size optimally matches the TC content, of 72 bits, including the 64-bit actual TC and an 8-bit TC code. Chapter 7 presents the optimal code for this range. It has the generator polynomial *236535*. The article presenting this code uses a polynomial code expression of three octal products.

It generator polynomial can then also be given by the product:

- $(x^4+x^2+x^1+1)\ (x^6+1)\ (x^5+x^3+x^2+1)$ or 1 00000110 00101011

It is proposed that this code be used as an extra CRC. In this manner 5 errors can be detected by the first CRC, and another 3 by the second CRC, which is the general DelfiX CRC. Also, the latter should not be used twice as its mathematical similarity would remove its added effect. In both CRC cases it holds that more than 99% of larger combinations of errors are also detected, as with general 16-bit CRCs. Then, it becomes reasonably safe to say that no corrupt TCs will be processed by the OBC in the limited lifetime of Delfi-n3Xt.

The smaller CRC introduced above should be calculated *before* the regular CRC.

The resulting telecommand frame is demonstrated in Figure 8.14 below, illustrating the use of a *final DelfiX* frame. Although this frame has been optimized for large frames, the small loss of bits is not very important. Given the fact that a first Frame Check Sequence results from the data protocol, the additional FCS is labelled FCS 2. Both CRCs are logically calculated over all fields before it, except for any flags.

				← *Encryption* →			
Flag *2 bytes*	**Address Field** *3 bytes*	**Frame Type** *1 byte*	**TC Code** *1 byte*	**Telecommand** *8 bytes*	**FCS 2** *2 byte*	**FCS** *2 bytes*	**Flag** *2 bytes*
01111111 11111110	3 ASCII Characters	8-bit number	8-bit number	64-bit value	16-bit CRC value	16-bit CRC value	01111111 11111110

Figure 8.14: DelfiX TC frame construction

The result is a TC frame of 168 bits. Bit stuffing can increase this number to a maximum of 180 bits. Additional comments:

- **Amount of preamble**
  It makes sense to add a number of flags to the start of the frame to ensure both carrier synchronization by the Motorola MC3362, and consequent symbol synchronization by the Frame Interpreter. In case of a proper link no large amount of flags should be required by the demodulator, but as bitrate is not stringent it is better to be on the safe side. In Delfi-C^3 20 were added before every TC, and 4 after. The prelude flags are undoubtedly excessive, but a similar amount of flags is suggested for Delfi-n3Xt. 15 flags create packages of more or less 300 bits, taking up either half or a quarter of a second with a bit rate of 600 or 1200 b/s respectively. More than one end flag after a frame however is practically useless, so this functionality is removed for Delfi-n3Xt.

- **TC repetition**
  It makes sense to repeat a TC a number of times to ensure proper arrival. Due to the inserted TC code a TC will then not be executed more than once. Also, a safety feature can be included in the OBC that for most TCs no two identical TCs can be executed. This repeat value can be set in DIGIT (or its successor, version 2.0), to be changed on transmission or set to a default number, possibly 2 or 4 for a 1-second transmission. Doppler updates of the transmission frequency also occur every second by the ground station, which might upset the receiver for the duration of one TC.

- **Frame rate**
  Due to the low number of TC bits per frame, TC definition is independent of data rate.

8.6.2 Telemetry frame (downlink)

Telemetry by means of *TM frames* is meant to provide a continuous downlink signal over the PTRX (or ITRX). Also, its reception is best-effort, it is not guaranteed. Data rate is a limiting factor; content is determined based on what amount of space there is. In case however the data rate turns out to be high enough so that the content cannot be filled usefully, consecutive frames can be repeated, possible spaced apart by a certain amount of time if allowed for by the processor buffer.

For the TM frames the frame rate as well as amount of preamble should be determined. At the same time, some subfields of the information field can be specified, related to communication aspects. The final implementation of these subfields is again to be determined *by the CDHS systems engineer*. The above aspects are discussed below.

Frame rate

As mentioned above it makes sense to maximize the amount of data sent by increasing effective data rate. This can be done by:

- Increasing the actual data rate
- Increasing the average throughput efficiency

As shown in subsection 8.5.6, the DelfiX protocol increases the throughput efficiency by some 10-15%. In case of Delfi-C^3 with a bitrate of 1200 b/s, 1 frame per second was transmitted on the downlink. Section 7.6 has shown that this was on the high side, a *2-second frame* would have increased data throughput by 8%. This would then decrease the amount of frame per second to 0.5.

Nevertheless, frame duration is also limited by time-varying communication aspects such as changing Doppler shift rate and satellite rotation. Satellite rotation rates however turn out to be not very significant on this time scale (an estimated 0.2° per second for Delfi-C^3), and neither are Doppler shift rates on the VHF downlink compared to the frame duration. The choice of having one frame per second for Delfi-C^3 was rather arbitrary. However, it has proven to work nicely.

Satellite rotation will assumably be of the same order for Delfi-n3Xt, although dependent on the final ADCS implementation. So the frame rate should not have to be reduced. Subsection 8.5.6 has shown that frame sizes of 2400-4800 are preferred for *BER* values of 10^{-5}, and frame sizes of 4800-9600+ are preferred for *BER* values of some 10^{-8}.

In conclusion, frame sizes of 2400-9600 are all acceptable. 4800 bits would be on the safe side for various *BER* values. For a data rate of 2400 however this would involve spreading a frame over two seconds, whereas total overhead per frame is relatively small with the DelfiX protocol. As such, the following frame sizes and frame rates are suggested for different final bit rates:

- **2400 b/s**: frame size of *2400 bits → frame rate of 1 frame/s*
- **4800 b/s**: frame size of *4800 bits → frame rate of 1 frame/s*
- **9600 b/s**: frame size of *4800 bits → frame rate of 2 frame/s*

Amount of preamble

In case of Delfi-C^3 the effective amount of frames per second was only 4/5, as one second (thus one frame) was filled with flags (*preamble*) after the transmission of every 4 TM frames (in Science mode). This second was simply required by satellite processing speeds. An added benefit was that it gave a second for radio amateurs to clearly spot the transmission frequency, but in fact the frequency spectrum can also easily be located without this constant frequency. Similarly, simple tests performed by playing back a recorded Delfi-C^3 telemetry sample, without flag frame, show a seamless demodulation in RASCAL/DUDe. *Therefore, no such period is planned for with Delfi-n3Xt.*

The TM signal can be assumed to be continuous; there is no defined beginning and end of transmission other than those of the frames. As bit stuffing will almost never be maximal, frames of less than the defined frame size will be supplemented by flags, and some 1's to reach an integer number of bytes. *This will add some preamble in between every TM frame to allow for synchronization*.

Unique frame ID

Two types of labels are useful for telemetry data; one is a ***time stamp***, and one is a ***frame counter***. The counter can yield information on reception completeness, whereas the time stamp can give information on the moment of frame generation. In delfi-C^3, no time stamp was used as *no real time clock was integrated*. A counter was used, but had to solve the problem of uniqueness; when a reset occurred the counter should remain unique. Storing each counter however in nonvolatile memory would quickly use up the allowed memory-write cycles. Therefore, a boot counter and a cycle (temporary) counter were incorporated. Nevertheless, after countless resets the boot counter was switched off in fear of using up the memory. Whereas a counter can be seen as part of the protocol, is it also interesting for communications as *it allows for requesting data if storage is incorporated in Delfi-n3Xt*, such as on the STX.

In Delfi-n3Xt continuous operation is assumed, due to the incorporation of a battery. The previously proposed solution of adding IDs to frames [SLR 0282] is to add a 39-bit RTC (Real Time Clock), which are then unique markers. Furthermore, in case of a payload frame another RTC is added at the exact moment of payload data generation. Whether this last one is indeed necessary and useful, is part of the CDHS data assessment. In any case, no counter is present, just the clock data. The problem with this set-up is twofold:

- ***It does not give data on completeness***
  A frame is not per definition generated every second in case of failures or non-nominal operations.

- ***It does not automatically allow for unique data requests***
  The RTC counts in tenths of seconds and data frame generation by the OBC is not constricted to a specific moment in a second. Simply ignoring the tenth seconds is no option either as possibly a first frame is generated, say, in the early 29^{th} second, and a second one in the late 29^{th} second.

Furthermore, as will be shown below, 39 bits is excessive if the time is expressed in tenths of seconds from a certain moment in time onwards. Therefore, two options of improvement are suggested. The general solution is to *add a (small) counter to the RTC added*. This counter will be stored in volatile memory and will just restart after its maximum length is reached or after a power loss.

Two propositions:

- **Maximized completeness, good requestability**
  In this case, the counter cycle is completed: it runs its full cycle and restarts afterwards. It is not linked to the RTC, but the combined RTC allows concluding on when the cycle took place. Requestability is acceptable through requesting numbered frames belonging to a certain period. An RTC of 32 bits allows for 13.6 years to be counted in tenths of seconds, which with an 8-bit counter yields cycles of 256 numbers, totalling 40 ID bits, or 5 bytes.

- **Maximized requestability, good completeness**
  In the case, the counter is linked to the RTC and does not run a full or predefined cycle. It is suggested that the RTC counts in minutes, where 24 bits allow for 31.9 years. Seconds and tenths of seconds are then counted by a 10-bit counter. Every minute then a new 4-bit counter starts, which can run to 59, 60 or 61 depending on the frame generation speed within a certain minute. In this manner, each frame is allocated a unique number which can be anticipated in advance or retrospect, and requested with maximum ease. It might however not be immediately clear whether a 60th or 61st frame exists within a minute, assuming a frame generation speed of 1 per second. Also, this approach requires an active link between clock and counter generation. This system is shown schematically in Figure 8.15 below; from left to right a complete counter can be obtained, either a total time counter or a unique frame counter.

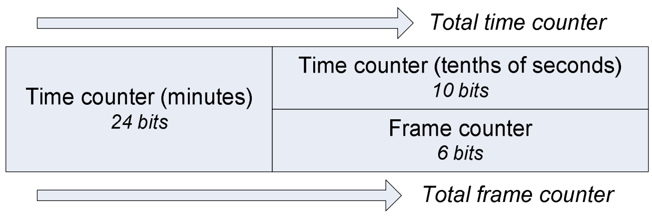


Figure 8.15: Suggestion for a practical unique frame ID

The result of the suggestions above is a *40 bit unique ID*.

In the end, CDHS complexity and reliability should be taken along to make the final decision. Of course, a large (unique) counter can also be added but this would require more bits; the same holds for another dedicated boot counter. Then again, large framer counters might not end up to be a problem if the downlink data rate of 9.6 kb/s is reached. And still, this would require the problem of required storage to be solved in case the satellite is not constantly powered.

For the clock it holds that it can be assumed to work constantly, depending on its implementation (including a small battery for example). If it does not, a small overlap can be accepted (revert in time). In case of communications this can be monitored by adding another time stamp to the received telemetry, on-ground.

Data security

The first line of defense against corrupted data is the CRC. As presented above it allows discovering up to 3 independent bit errors, or more in specific patterns. With dedicated processing even single bit errors can be corrected. This then protects sufficiently against transmission errors, given the *BER* of 10^{-5}.

A tweaked recording of TM data (send by a possible wrongdoer) can however circumvent the CRC. Consequently, data transmitted by RASCAL, or data received by RASCAL, does not necessarily have to be the original data. Section 7.7 has introduced techniques to circumvent this.

Encryption is not allowed on the downlink over a radio amateur frequency. Nevertheless, a unique hash can be inserted. As this will come at the cost of many bits, it is not preferred. An encrypted hash can be used, as it reduces the bit load. Nevertheless, it increases complexity.

As it is not assumed that Delfi-n3Xt will be the target or database spamming attacks, and every frame submission will be accompanied by a user name, *no additional security is deemed required on the downlink*.

TM info field

As introduced in the last section, the telemetry downlink frames should also include the *TC code*, as well as the *last executed TC*. The latter can simply be done with sufficient accuracy by transmitting just the 8-bit TC type and destination. The TC code field can in case of a received golden TC code be set to 00000000 or 11111111. This section has suggested a *40-bit counter*. Both the TC code and last TC fields can be said to be part of the CDHS housekeeping data, as a matter of convention.

The total (minimum) CDHS HK field length is then 24 bits.

The resulting suggestion for (part of) the content of the TM frame is demonstrated in Figure 8.16 below. If high data rates are reached and more space is available, of course a more complete repetition of the last executed TC can be sent down. In that case, even *a short history of executed TCs* can be added.

Information Field			
Unique ID	*CDHS HK Field*		
RTC + Counter *5 bytes*	TC Code *1 byte*	Last TC *2 bytes*	

Figure 8.16: TM Information Field content suggestion

8.6.3 Telemetry request frame (downlink)

Assuming some form of storage will be available on Delfi-n3Xt, be it on the STX or even directly connected to the OBC, stored-telemetry requests will be useful. Even if not, telemetry packages can be requested with specific data via telecommand. A telecommand will therefore have to be defined for these cases, and a telemetry request frame will have to be defined, the *TM-request frame*. The proposed *frame type* field which is part of the DelfiX protocol can simply be used to indicate a special type of telemetry frame, even if it means the body of the frame is identical to the originally transmitted telemetry frame.

8.6.4 Software update frame (uplink)

Software updating is a possible addition to Delfi-n3Xt, if feasible. Especially when Delfi-n3Xt might become a test bed for ACDS purposes, software updating can be very useful. Whether software updating is feasible from a CDHS side should be investigated, but it should be no problem for communication reasons. A telecommand should announce the start of a software update transmission, which will then last for as long as there is data to be transmitted. As this is surely more than one usual TC frame length, a special frame type should be defined; the *software update frame*.

Different aspects are quickly discussed below.

Frame rate: The software update frame would most likely resemble the TM frame, in the sense that data content exceeds data frame length. Therefore the maximum data rate should be used, and a default frame length of 600 bits can be used, given a 600 b/s bit rate. If however the bitrate indeed is only 600 b/s, a 6% increase in data throughput can be achieved by *lowering the frame rate to 0.5 frames/s*. A further halving would yield only 2% increase. Nevertheless, if the upload program is not excessive, transmission time is not critical and safety can be chosen over speed. The frame rate can be changed by means of experiment, if indeed the buffer of the PTRX or ITRX receiver is sufficiently large, but it does not make much sense to do this for a dangerous procedure such as software uploading.

Data fields: The *frame type* field can again identify the software update frame. The *info field* can be filled with any defined information, most likely preceded by a *software package frame number* and a *total number*.

Security: As any upload procedure, software updates can influence mission success. As such, some form of security is recommended, and allowed as we are dealing with an uplink signal. Most simply, the TC code can again be used as a simple 'password'. Nevertheless, as it does not make sense to change the TC code per software update frame, a single password for the complete session or the program to be uploaded can be used. This code can simply be the one delivered as a response to the TC announcing the software upload. Furthermore, encryption can again be applied; possibly with a different key than regular TCs. Nevertheless, in this case it should be tested whether the OBC can keep up with real-time decryption.

Amount of preamble: As software upload sizes can be significant, it does make sense to limit the number of flags transmitted before a frame. On the other hand, it is important that <u>every</u> frame should be received properly in contrary to the downlink. At the same time, a large link margin is available. As such, *an amount of 3-5 flags between frames should be sufficient*. The first software upload frame however can again be preceded by the same number of flags as a telecommand to ensure proper arrival, such as 20 flags.

9 PTRX

The **p**rimary transceiver (TRX) on Delfi-n3Xt is logically called PTRX. It is a complicated system as seen from a top-level point of view, due to the reason that it is based on the RAP, on which many people have worked and last-minute tweaking have extensively been performed without extensive documentation. At the same time a radio requires many people with different competences to work together, with work varying from top-level conceptual to low-level electrical integration.

This chapter describes the top-level conceptual design, but also discusses lower-level component integration. The previous chapter has described the techniques and components involved with the UHF/VHF link; on the satellite this link is operated primarily by the PTRX.

A functional description of the PTRX is used to define operational modes in section 9.1. This leads to a number of CIs, as presented in section 9.2. The resulting hardware integration of the PTRX CIs in presented in section 9.3, and its interfaces with the local EPS and CDHS components and external components are shown in section 9.4.

Using knowledge of the hardware components allows determining the housekeeping data that is to be gathered and transmitted to the OBC in section 9.5. The latter is to be done by the PTRX processors, which at the same time have a rather large number of functionalities to take care of; this is explained in section 9.6. Related to the processors is the choice of having telecommand interpretation functionality integrated in the PTRX RX, discussed in section 9.7.

The linear transponder, which functions as the return service to the radio amateur community, and its implementation are presented in section 9.8. The option of reducing the power consumption of the PTRX by lowering the gain of its PA is discussed in section 9.9. Consequently, two experimental features on a mission-level, software uploading and the variable voltage bus, are discussed in section 9.10. Finally, the status of the PTRX, the overall changes with respect to its predecessor and potential future improvements are discussed in sections 9.11, 9.12 and 9.13 respectively.

9.1 Functional description and modes

Based on the set of requirements on the PTRX specified in chapter 6, the main tasks of the PTRX are to:

- Receive telecommands (TCs)
- Transmit telemetry (TM)
- Provide linear transponder functionality
- Generate housekeeping data

Also, being a secondary mission objective, it should if possible:

- Facilitate the OLFAR experiment

These different functionalities of the PTRX gives rise to multiple *operational modes*. Telecommands can be received in *all* modes. Housekeeping data is also generated in *all* modes. The other functionalities cannot be executed at the same time, due to power reasons and reasons of shared bandwidth. The resulting mutually exclusive modes are:

- **TM mode**
  This is the default mode. Telemetry is transmitted, which is internally received from the OBC over the I^2C bus. If however no telemetry is received, *a continuous stream of flags is transmitted*.
- **Transponder mode**
  The transponder is activated. Any signal received within the transponder bandwidth is converted to transmission frequency, amplified and transmitted.
- **OLFAR mode**
  The OLFAR experiment is activated. Low frequency signals (<30 MHZ) are transformed to the transmission frequency, amplified and transmitted.
- **TX OFF mode**
  As regulations prescribe that it should be possible that no signal is transmitted (requirement SAT-F.03), all transmissions should have an OFF switch. The processor however does not require knowledge of this mode; TX and RX sections are separated. Power is supplied, or not. When power is received in the TX section, TM mode will be activated.

The functionalities are further discussed below, as well as the required subfunctionalities, in subsections 9.1.1 through 9.1.4. The following section can then derive a number of configuration items (CIs) required on the PTRX, after which section 9.3 can comment on the final hardware implementation.

9.1.1 Receive telecommands

The functionality of receiving telecommands can be functionally described by Figure 9.1 below.

This figure as well as those that will be introduced below, are conceptually very similar to those introduced in chapter 7 on transmission techniques. Nevertheless, these are more expanded and allow demonstrating overlap in components. Physical or Data Link Layer functionalities are separately indicated by a small header and different colours.

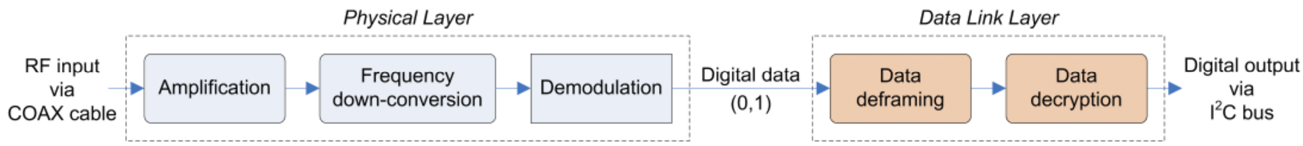


Figure 9.1: PTRX functionalities required for receiving telecommands.

An RF signal is received via a COAX cable, after which the physical layer functionalities as amplification, filtering and frequency conversion are performed, as introduced clearly in the previous chapter. *Amplification* is taken as a separate block, as a primary RF amplifier is required, which is a more important component that simpler IF amplifiers. *Frequency down-conversion* is a conceptually important step, required to deliver the proper signal to the demodulator. In the process of frequency down-conversion a number of filters and IF amplifiers is necessary. Actually multiple down-conversion steps are applied. These are collectively indicated as 'frequency down-conversion'. Afterwards *demodulation* can take place, and a digital signal results.

Data link functionalities are different in that they require a *processor*, and *software*. The most important step taken care of by the PTRX processors is the application and removal of data frames. In case of uplink packages however, data encryption is applied and it is suggested that the PTRX also removes data encryption. This is not done similarly on Delfi-C^3, however it is logically integrated with other data stream functionalities on the PTRX, and it takes away complexity from the OBC.

9.1.2 Transmit telemetry

The functionality of transmitting telemetry is again best shown by a figure, being Figure 9.2 below.

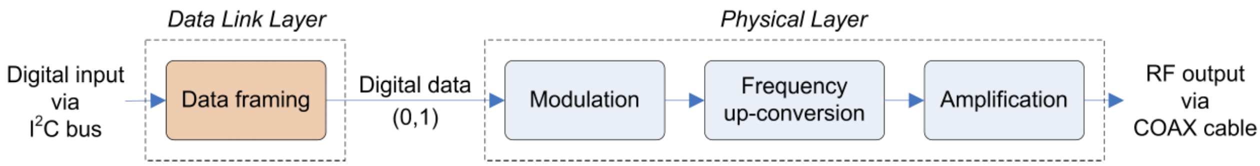


Figure 9.2: PTRX functionalities required for transmitting telemetry.

The functionalities shown above are very similar to those of Figure 9.1, but in reverse. No encryption is applied on the downlink. The (final) amplification step is again shown separately as the power amplifier required is an important separate component, which could also be changed.

9.1.3 Provide linear transponder functionality

The linear transponder subfunctionalities are shown in Figure 9.3 below.

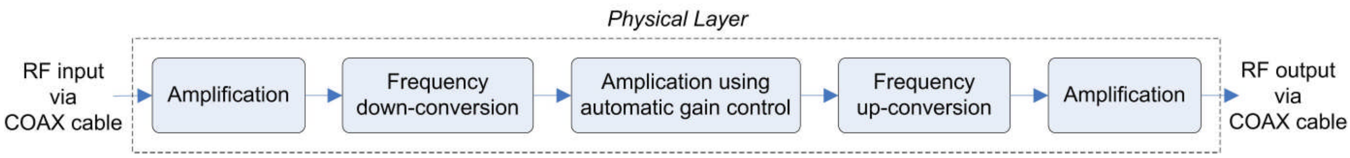


Figure 9.3: PTRX functionalities required for providing linear transponder functionality.

The linear transponder functionality is conceptually very simple, as no data interpretation has to take place. An RF signal is first received, amplified and frequency down-converted. It is brought back to an IF of 10.67 MHz, at which it is amplified using automatic gain control in order to give the signal a defined peak power. Afterwards the signal is again frequency converted, but this time to a higher frequency, and afterwards amplified for transmission.

The benefit of converting to a low IF is that simple COTS electronics can be used for automatic gain control, but more importantly part of the transmitter and receiver architecture required on the PTRX can be reused.

9.1.4 Generate housekeeping data

The final main PTRX functionality required is to generate housekeeping data. This functionality has no direct external communication aspect, as data is gathered within the PTRX, and transmitted to the OBC. The OBC then joins housekeeping data and payload data in telemetry frames, to be returned to the PTRX for transmission.

The functionalities involved are shown in Figure 9.4.

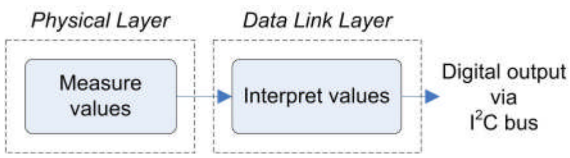


Figure 9.4: PTRX functionalities required to provide housekeeping data.

Different housekeeping values will be produced at different locations on the PTRX. In case these are analogue values, an analogue-to-digital converter integrated in the PTRX processor can be used to interpret the values. In case sensors are digital, the data is directly read out by the processor. The processor can then forward the data to the OBC.

In case of the RAP, values such as amplifier temperature, Doppler shift current, forwarded and reflected power were gathered. The approach for the PTRX is further assessed in section 9.6 on the PTRX processors.

9.1.5 Facilitating the OLFAR experiment

The OLFAR experiment is not considered to be an actual part of the PTRX, and in fact should add little to the PTRX architecture. On the PTRX it would require the following functionalities (Figure 9.5):

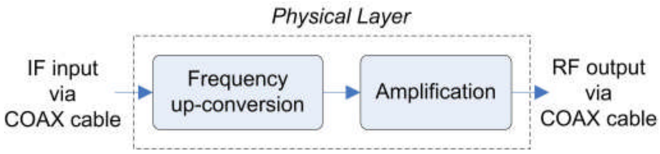


Figure 9.5: Functionalities required on the PTRX to integrate the OLFAR experiment.

The majority of OLFAR components should be located on the DAB, or Deployment and Antenna Board, being part of the Antenna System. This is further explained in chapter 11 on OLFAR. The entire OLFAR experiment is explained in [SLR 0170]. A signal received via the antennas is converted to an IF of 10.67 MHz, after which it is guided to the PTRX by means of a COAX cable. Consequently, the signal is transformed to RF, amplified for transmission and forwarded to the antennas.

9.2 Configuration items

In the previous section all required subfunctionalities of the PTRX have been presented, following from the main functionalities of the PTRX. It has also been shown that different functionalities require similar low-level components, which leads to having shared components. Receiver, transmitter, transponder and OLFAR functionalities can be combined as shown in the following Figure 9.6. Data link layer functionalities have been indicated in red, and, in contrary to the other functionalities, these can be combined in software. As such only one physical component is strictly necessary; a processor.

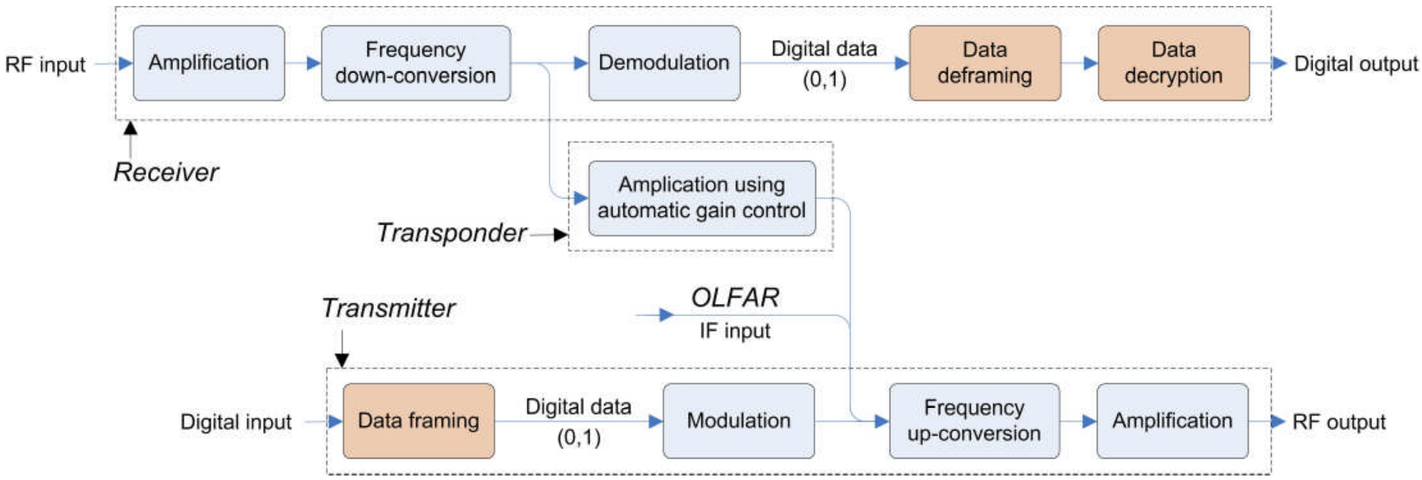


Figure 9.6: Combined functionalities required on the PTRX.

This diagram allows determining all required lower-level components on the PTRX. The benefit of doing so is that concrete work-packages are formed on which people can perform work, and over which people can have responsibility. As in case of the PTRX many electrical engineers are involved, it is and has been very important to establish boundaries and interfaces.

The subfunctionalities in Figure 9.6 above can be translated into the CIs below, see Figure 9.7 below. The context of the PTRX CIs can also be seen, indicated by faded boxes. The components of the PTRX are shortly addressed below.

PTRX RX & TX: It can be seen that the three physical layer functionalities required for receiving and transmitting are translated directly into three lowest-level CIs, belonging to the higher-level PTRX RX and PTRX TX, forming the first two branches. The functionalities are *(de)modulation, frequency conversion* and *amplification*.

PTRX Transponder: The transponder forms another branch, but introduces two components other than those already presented; joining two power paths requires dedicated components. These are chosen to be part of the transponder CI, as it is the addition of the latter that requires these components.

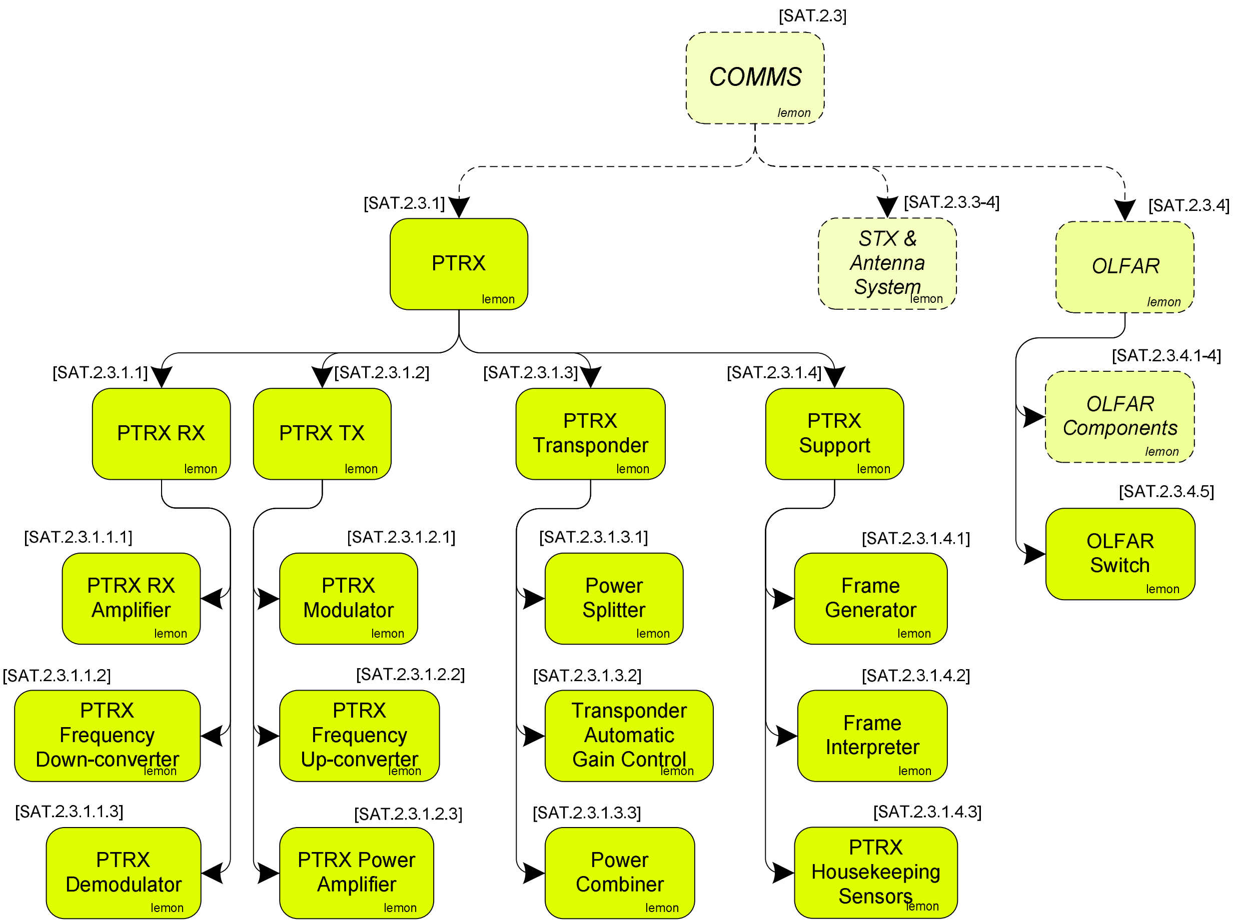


Figure 9.7: Configuration Items to be implemented on the PTRX.

PTRX Support: Both a *Frame Generator* and a *Frame Interpreter* can be seen, these are the processors (aided by oscillators and ADC/DACs) of respectively the TX and RX sections. The two processors are branched under PTRX Support and not under PTRX TX and RX. The reason for this is that in this manner the PTRX RX and PTRX TX CIs are completely analogue. The processors on the other hand involve software, and therefore different integration expertise. Also, the processors are charged with preparing housekeeping data. The final CI of the support branch shows these Housekeeping Sensors, which should be placed on the PTRX and should be read out by the processors.

Finally, ***OLFAR*** is included as a separate branch. This is mainly because the OLFAR system is spread over the PTRX and the antenna board. The latter is part of the Antenna System. In fact, it should almost completely be integrated within the latter. But, as OLFAR requires the use of part of the transmitter section, the OLFAR signal from the antenna board is led to the PTRX, in order to be combined with the transmitter section. This is done through the use of a switch, in order to have full control over the possible reception of an OLFAR signal on the PTRX.

9.3 Hardware implementation

As was already suggested before, the actual low-level implementation of the PTRX is slightly more complex than suggested by figures such as Figure 9.6 on the combined functionalities. The reason is that multiple frequency conversion steps are used as well as multiple amplifiers and a number of filters. The actual implementation of the PTRX on the component level is based on the design of the RAP, but has evolved since that time, and is shown in Figure 9.8 below.

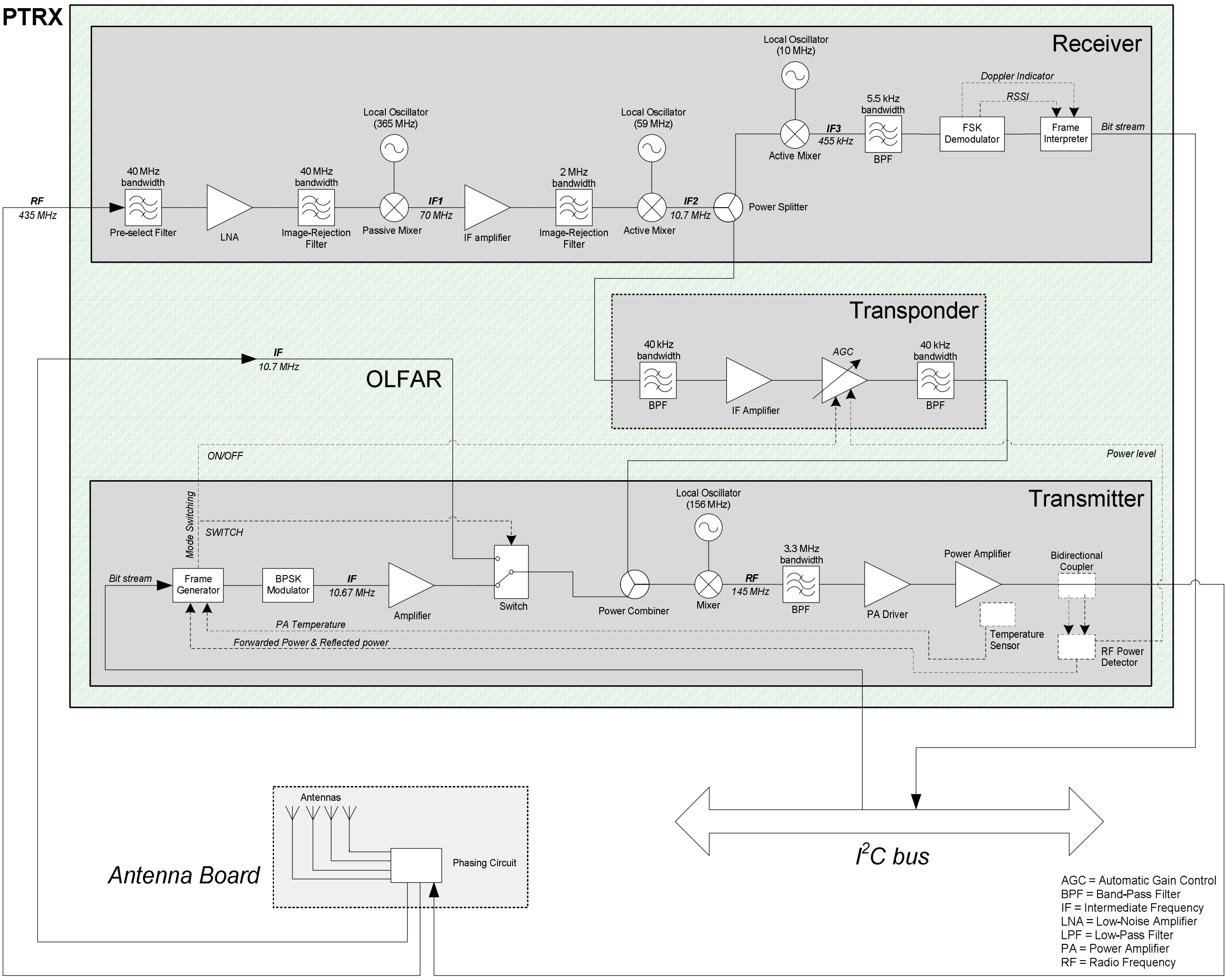


Figure 9.8: Hardware components of the PTRX and its interfaces.

The design behind Figure 9.8 is a result of many design steps, originating with the RAP, but with the last design stages of the PTRX described in [SLR 0281] and others, see chapter 5 for the complete list. The process of preparing an analogue signal for transmission or reception is clearly seen in hardware blocks, with transmission signal flow indicated by solid lines and generally moving from left to right. Some operational support connections are indicated by dashed lines.

The main characteristics shown in Figure 9.8:

- ***PTRX functional blocks***
  Three clear functional blocks can be seen: receiver, transmitter and transponder. OLFAR is a fourth functionality that has no clear functional blocks integrated on the PTRX except for the mandatory switch, being implemented in the transmitter path.

- ***External interfaces***
  Two bit stream interfaces with the I^2C bus can be seen, as well as *three* radio frequency interfaces with COAX cable connections to the Antenna Board. The latter include the two logical RF signal connections to receiver and transmitter, as well as an OLFAR IF signal generated on the antenna board. A final interface, that supplying power, is not indicated. For clarity reasons, the I^2C interfaces are indicated as being unidirectional.

- ***Superheterodyne principle***
  This principle applied on receiver, transmitter and transponder section can clearly be recognized, applying frequency converters, image-rejection filters, bandwidth filters and IF amplifiers. Some deviating names are used for different amplifiers in the design, such as LNA for low-noise amplifier, PA driver for Power Amplifier driver or pre-amplifier, and AGC for automatic gain control.

- ***IFs and RFs and oscillator frequencies***
  All oscillator frequencies required for the frequency conversion steps are indicated in the figure, as well as the resulting or incoming IFs or RFs. Components operating on certain IF logically have this IF as centre frequency.

- ***Filter bandwidths***
  Filter bandwidths as opposed to centre frequencies have been indicated. These bandwidths are important for top-level conclusions on transmission bandwidths. The downlink shows *no* practical bandwidth filters, only image-rejection filters. The final filter has 3.3 MHz bandwidth. In case of the receiver the signal is reduced to have only 5.5 kHz bandwidth, which does limit the actual ideal transmitted signal as an FSK signal of 1200 symbols/s is broader than 5.5 kHz.

- ***Processors and (de)modulators***
  Most components are relatively standard components as is, but processors involve a lot more functionalities in software, and the (de)modulator integrates more hardware functionalities as already has been introduced in the previous chapter. In the figure they have been logically indicated as *single components*. The processors are introduced in section 9.6 below.

- ***Processor interfaces***
  Dashed connections are indicated to show processor interfaces. These connections allow for *mode switching* and the *generation of housekeeping data.* The connections between the Frame Generator and respectively Transponder AGC and Transmitter Switch allow switching between PTRX modes. The four other dashed connections deliver housekeeping data to the two PTRX processors.

9.4 PTRX interfaces

The components introduced in the previous sections are all combined onto one PCB. These PCBs are stacked inside the structural frame of Delfi-n3Xt, and have interfaces to the *power bus* and *data bus*. In order to properly provide these interfaces, components belonging to the electrical power subsystems (EPS) and command and data handling subsystem (CDHS) are integrated on each PCB. These components are jointly called the *local EPS* and *local CDHS* respectively. Also, in case of the PTRX, coax cables are connected to the PTRX PCB and communicate analogue signals between the PTRX and the Antenna Board. As such the PTRX PCB and its interfaces can be shown by Figure 9.9 below. A more detailed explanation on the local EPS and CDHS components can be found in the respective top-level documents, [SLR 0019] and [SLR 0029].

In fact, the illustrated interface diagram shown below is outdated. The final design of the Standard System Bus (including power and data interfaces) is still to be determined.

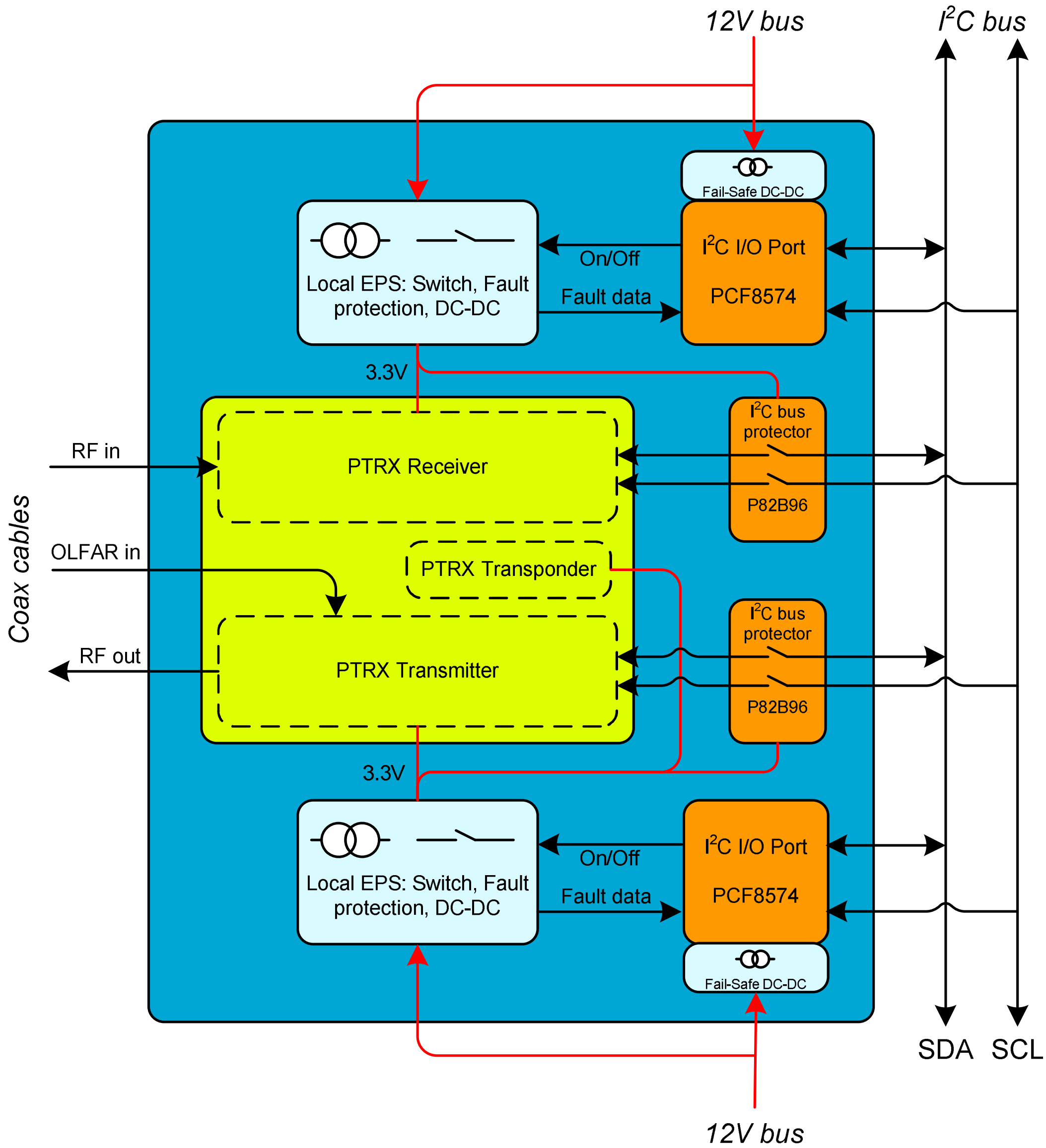


Figure 9.9: PTRX interfaces and the implementation of the local EPS/CDHS.

Some comments on Figure 9.9:

- ***3 types of external interfaces between the PTRX PCB and other systems***
  As explained above, a power and data bus interface connect to the PTRX, as well as the analogue communication lines. 3 coax cables, as suggested by Figure 9.8 above, are required for analogue signals. A 12V bus supplies power for the PTRX and the local CDHS system. Digital data in communicated via the SDA line, or *serial data line*, and is supported by the SCL line, or *serial clock line*.

- ***Double power and data bus architecture for both PTRX processors***
  In the previous section it has been stated that two processors will be implemented on the PTRX, one in charge of receiver functionalities and one in charge of transmitter functionalities. This decision is further explained in section 9.6 below. As the transmitter should be switched off if necessary due to reasons of interference, but the receiver should in principle never be switched off, the local circuits should be doubled for both the Frame Generator and Frame Interpreter. This also allows failures in either transmitter or receiver not to force a transceiver shut off. *In fact, it can be decided to remove the possibility of switching the receiver off altogether*. In this case, the I/O port can be removed, and possibly a simple PTC (positive thermal conductor) could be used as a short-circuit protection. The same approach has been used in Delfi-C^3; this choice depends on the final system bus integration.

- ***Local EPS***
  A local EPS consists of DC-DC convertors to convert from 12V to 3.3V, and protection circuits. The local EPS powers both the PTRX core and the local CDHS. The transponder is powered by the transmitter power supply as it is also controlled by the transmitter processor.

- ***Local CDHS***
  A local CDHS consists of an I^2C I/O Port and an I^2C bus protector. The former basically switches on the power supply of the PTRX via an I^2C command originating from the OBC, whereas the latter shuts off the PTRX from the I^2C bus when not powered.

9.5 Housekeeping data

To monitor the performance of the PTRX, or any subsystem or PCB, the regular gathering of housekeeping data by the OBC and consequent transmission of it is required. The types of housekeeping data that are interesting from an electrical point of view are generally:

- Current consumption values
- Signal strength indicators
- Temperature values

Other values that can be thought of are *digital values related to processor operations*. The different measurables per category are commented on below. Most values have been measured similarly on the RAP. Other, new suggestions have been indicated as being *optional*.

On the RAP all values were measured with 10-bit precision; the reason for this was quite simply that the ADC converters of the PICs had this precision. Those of the MSP430s to be used on the PTRX have a 12-bit accuracy. Nevertheless, in all cases an 8-bit accuracy seems to be enough; as in all cases the values are actually measured to spot irregular values, which can be spotted without large precision. A possible exception is information on Doppler shift; nevertheless the *Doppler indicator* introduced below value quite simply does not give a very absolute accurate Doppler shift value. The amount of bits per value is indicated below for clarity and explained if the value is (possibly) different from 8.

All values except for the current consumption values are to be measured by the PTRX processors.

Current consumption values:

- ***Receiver current (8 bits)***: In order to get an idea of whether the receiver is operating without serious electrical problems, the current consumption can be monitored to check for nominal values.
- ***Transmitter current (8 bits)***: The same holds for the transmitter current.

In general also the total PTRX current can be monitored. However, as explained below, two processors will allow or even require a separate power supply. Consumption is logically measured at the power supply, by what will be introduced below as the *local EPS*. Current consumption monitoring of individual components does not seem necessary.

Signal strength indicators:

- ***Forwarded power (8 bits)***: The forwarded power, together with the reflected power, indicates the total power forwarded to the antenna system by the transmitter. In case of an antenna failure, these two values allow to notice and draw conclusions on the failure that has occurred. This value is measured in an analogue fashion using a power detector and an ADC port of the MSP430.
- ***Reflected power (8 bits)***: See forwarded power.
- ***Received Signal Strength Indicator (RSSI) (8 bits)***: To have an indication of the signal strength received in the PTRX, this value can be used. This value is delivered by the FSK Demodulator and to be digitized by an MSP430.

Temperature values:

- ***Power amplifier (PA) (8 bits)***: The power amplifier is the only component that significantly heats up during nominal operations. With an efficiency of only 16%, more than 1 W is effectively used to heat the amplifier and its environment. Therefore it is essential to monitor this heating. A digital temperature sensor is used that yields a digital value. The 8-bit precision of the value can be increased to produce a better thermal model.

One of the downsides observed of Delfi-C^3 is that there were not enough temperature measurements done to validate the thermal model. As such, it might make sense to increase the amount of temperature measurements. Optional:

- ***Processors (Frame Generator and Frame Interpreter) (8-12 bits)***: MSP430 processors have an integrated temperature sensor. As such, these temperature values can easily be communicated to the OBC and transmitted to the ground if the data budget allows for it. As MSP430 processors are to be implemented everywhere on the satellite, a general approach should be chosen. 12 bits is the precision of the value provided by the MSP430 processors, but 8 bits might be enough.
- ***PCB (8-12 bits)***: One or multiple PCB temperature values can be measured to get a good idea of the temperature distribution on the PCB and its components. This would however require dedicated temperature sensors at strategic locations. This is a choice that should be made on the satellite level, as the same argumentation holds for all PCBs, albeit with more or less activity. Again 8-12 bits can be used.

Other analogue data:

- ***Doppler indicator (8 bits)***: The chip used for signal demodulation constantly outputs a value to indicate the amount of frequency correction required to demodulate the signal. This amount of correction is an indication of the amount of Doppler shift that has occurred. It does not give an absolute Doppler shift value, as aging and temperature effects change the oscillator specifics.

Processor (digital) values:

A number of values can be thought of that can be provided by the processor as it performs its functionalities. Processing times can be measured on-ground and are assumed not to change during operation. Transmission data can generally be analyzed by the OBC itself, as the data originates there. Therefore options are most likely limited to reception data. Two optional suggestions:

- ***Received erroneous frame counter (8 bits)***: It is interesting from a transmission point of view to get a good idea of how many transmission errors actually occur. The most logical way of counting would be to keep track of how many times an erroneous CRC is calculated. It is assumed that no large numbers of erroneous frames are received, therefore an 8 bits counter should be sufficient, but more is possible. If the counter is 'full' it can simply be reset to 0. In order to have an idea of the percentage of erroneous frames of all received frames, *a received frame counter can also be kept*:
- ***Received frame counter (16 bits)***: This counter will more quickly grow to large numbers. It does not however have to be a unique number, it can reset after a certain value; most likely together with the received erroneous frame counter. It should then however be a lot larger than the latter, as many more correct frames than incorrect frames are anticipated. The ratio of 8 to 16 bits is rather arbitrary; another ratio can be chosen pending the data budget.

All values shall in principle be gathered once per second. The OBC will pick-up the data every second, or less often.

9.6 PTRX Processors

The processors on the PTRX have been chosen to differ in type and in terms of functionalities from those of the RAP. This section argues for the changes and introduces the activities to be performed by the PTRX processors, being the Frame Generator and Frame Interpreter. The technical note dealing with the implementation of the functionalities on these processors is [SLR 0722]. Subsection 9.6.1 below introduces the processors and their interfaces and functionalities of the RAP of Delfi-C[3]. Subsection 9.6.2 consequently decides on the approach for the Delfi-n3Xt PTRX, and specifies all functionalities.

9.6.1 RAP processors and their functionalities

On the RAP, there were three processors; being Peripheral Interface Controllers, or PICs with rather low processing power. These were the Control PIC, the Bitshaping PIC and the EPS PIC. The latter in fact had a function similar to the Local EPS included on the PTRX board, as introduced in section 9.4. As such, two microprocessors were basically used for the operation of the RAP. The Bitshaping PIC was involved with transmission only, whereas the Control PIC was in charge of support. The OBC supplied frame content to the Bitshaping PIC, but it received TC data *directly* from the receiver section of a separate bus line. The implementation of the two RAP operation microprocessors and their functionalities are illustrated in Figure 9.10 below.

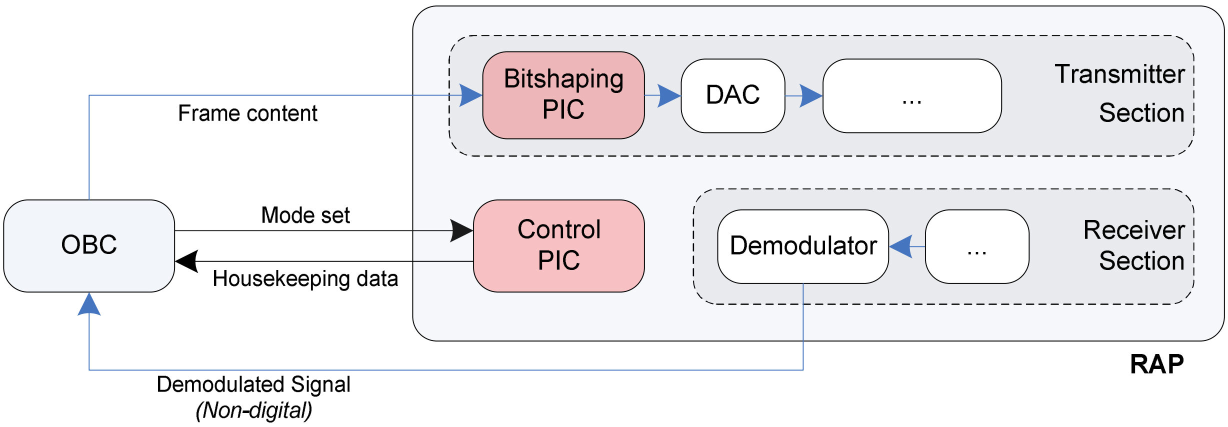


Figure 9.10: Integration of the RAP processors

From the figure the performed functionalities can be derived. These functionalities are presented in Table 9.1 below. In case of the OBC only the functionalities that are related to communication are shown. *Four types* of functionalities can be seen:

- **I^{2}C communication**
- **Housekeeping data generation**
- **Mode switching**
- **Data preparation (transmission and reception)**

The latter functionality can be subdivided in multiple subfunctionalities, which together form all digital actions between having interpretable frame content and a modulated / demodulated signal.

Table 9.1: RAP transmission functionalities per processor

Bitshaping PIC	Control PIC	OBC
• I²C communication • Data preparation (transmission) o AX.25 frame application o NRZ-I encoding o Raised cosine PWM*	• I²C communication • Housekeeping data generation • Mode switching	• Data preparation (reception) o Bit interpreting (removing Manchester) o NRZ-I decoding o AX.25 frame removal o TC decryption

**PWM: Pulse Width Modulation, which indicates approximating an analogue wave by using many digital values at a specific sample rate to indicate the amplitude of the analogue signal. In this case the DAC afterwards converts the PWM signal to an analogue signal.*

It can be seen that the OBC has significant functionality. In fact, the bit interpreting capability required a sampling rate of 25 kHz, requiring a halt of operations every 1/50,000 seconds as two receivers were on-board. The NRZ-I encoding as well as later CRC checking and decryption required a calculation intensive bit-per-bit operation, and at the same time the AX.25 frame removal requires searching for flags and recognizing proper frames. Finally the system required a separate connection as the signal forwarded to the OBC is not yet digital. The system is therefore not very modular.

9.6.2 PTRX processors and their functionalities

There are several good arguments to move all transmission functionalities to the PTRX:

- ***Reducing the processing load on the OBC***.
  In case of the RAP significant actions had to be executed by the OBC. This posed significant load on the OBC. As Delfi-n3Xt is a satellite significantly more ambitious than Delfi-C³, it makes sense to assume that the OBC will have more tasks to perform as well, but a similar MSP430 processor is slated to be used.
- ***Simplifying the software of the OBC***.
  In Delfi-C³ the OBC is interrupted every 1/50,000 second for sampling the signal on one of the receiver lines (the sample rate is 25 kHz for each receiver). These interrupts increased the complexity of the software.
- ***Increasing reliability.***
  The interrupts occurring every 1/50,000 second may well have been the reason for synchronization problems that occurred during testing.
- ***Removing one line on the bus***.
  All communications can take place over the I²C bus as bits are interpreted by the PTRX itself.
- ***Increasing the modularity of the system***.
  The radio can receive and transmit digital from and to the bus without extra processing required. The OBC needs no knowledge of transmission aspects. Also, the number of transceivers connected to the OBC is no longer important. Finally, the PTRX transceiver can be tested separately from the OBC. The radio thereby becomes a *modular system*.
- ***More powerful MSP430 processors have been selected for subsystems.***
  Per default these processors are to be used on all subsystems. As these are more powerful, the transition of functionalities seems feasible.

The main disadvantages of including the all communication functionalities on the PTRX are:

- ***It requires a significant change in the design of the RAP****.*

  Requirement SAT.2.3-C.03 states that changes with respect to the RAP design shall be minimized.

- ***The Delfi-C^3 system has proven to work reliably in space****.*

  Moving part of the functionality to the radio board requires a change of a mission critical part.

The disadvantages given above relate to the proven nature of Delfi-C^3 and its RAP. At the same time however, stability issues of the OBC are known to be related to the previous implementation. Furthermore, significant benefits in terms of implementation and testing due to increased modularity are foreseen. As such, *it is decided to include all transmission functionalities on the PTRX.*

When deciding to move all transmission functionalities to the PTRX, it remains to decide which processor will take care of this load. As on Delfi-n3Xt MSP430s will be used, processor load is less of a problem. Nevertheless, it does seem rather heavy to integrate all functionalities on one processor. Also, the following requirements state:

SAT.REQ.O.000 – The first priority of the satellite shall always be to enable commanding of the satellite.
SAT.REQ.F.001 – The satellite shall have a means of switching off any transmitter on the satellite.
SAT.REQ.F.003 – There shall be no provision for switching off a telecommand receiver other than an over-current protection.

Whereas the transmitter should be able to be switched off, most logically by removing the power of the entire section, the receiver should not be able to be switched off other than by an over-current protection. As such, a more elegant solution is the following:

Two processors are used, which each take care of all functionalities involved with reception and transmission respectively. This means the generation of housekeeping data generation is shared.

In summary, the main benefits of this approach follow from reasons of *security, regulations and load sharing*:

- A failure of the receiver or transmitter would not automatically cause failure of the other
- The transmitter must be shut off in case of disturbances and can so be shut off by removing power without shutting off the receiver
- Processing load is shared between the two processors

The resulting architecture can then be seen in Figure 9.11 below.

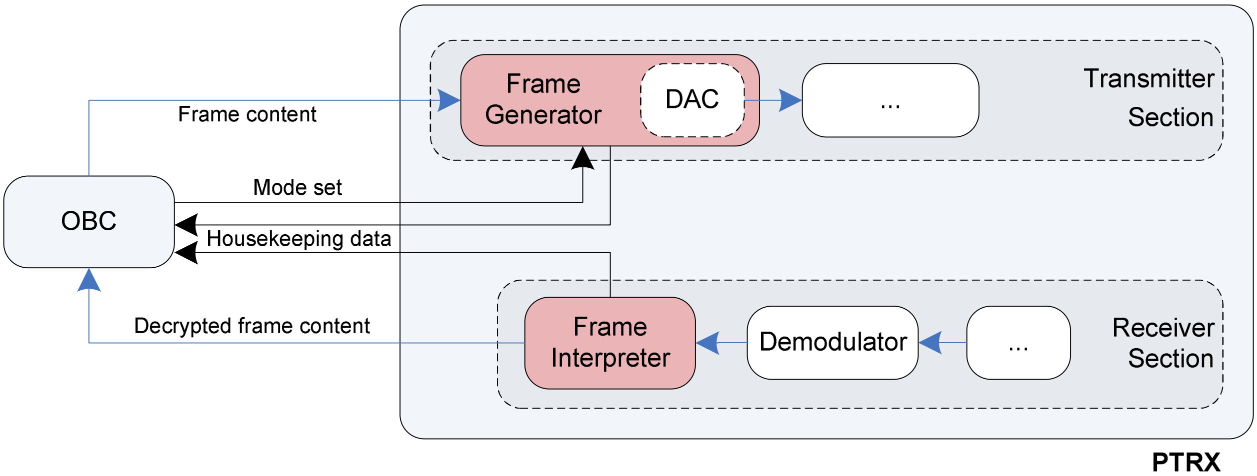


Figure 9.11: Integration of the PTRX processors.

The Frame Generator and Frame Interpreter are the names given to the processors aided by oscillators and ADCs and DACs. These processors are MSP430 processors, most likely of the type MSP430F1612. The resulting functionalities are summed up in Table 9.2 below.

Table 9.2: PTRX processor functionalities

Frame Generator	Frame Interpreter
• I^2C communication • Housekeeping data generation • Mode switching • Data preparation (transmission) o AX.25 frame application o NRZ-I encoding o Raised cosine PWM	• I^2C communication • Housekeeping data generation • Data preparation (reception) o Bit interpreting (removing Manchester) o NRZ-I decoding o AX.25 frame removal o TC decryption o *Putting the TC in the buffer and await pick-up by the OBC*

The same four different functionalities that have been introduced before are divided over the two processors. These four functionalities are shortly commented on below in three subsections.

I^2C communication

Both processors should be appointed an I^2C address, and should be able to properly communicate over the I^2C bus. The protocols to be followed are equal to those of other subsystems on-board of Delfi-n3Xt. These and a number of general requirements thereby apply which apply to all subsystems that are connected to the data bus; see [SLR 0029].

Housekeeping data generation

A number of housekeeping data variables have been introduced in section 9.5 above. The current consumption values are to be read out and provided to the OBC by the local CDHS. The other values are to be read out by the *PTRX processors.*

The housekeeping data generation is divided over the Frame Generator and Frame Interpreter, respectively being part of the transmitter and receiver sections. All values should be read out once per second after reception of a command from the OBC. This is important as it means that sensors are read out at approximately the same point in time. A consequent command will request the values from the processors.

The following Table 9.3 shows the values to be read out and provided to the OBC by both PTRX processors.

Table 9.3: PTRX housekeeping values per processor

Frame Generator	**Frame Interpreter**
Certain: • Forwarded power (8 bits) • Reflected power (8 bits) • Power amplifier (PA) temperature (8 bits) Optional: • Frame Generator temperature • PCB temperature (read-out by both processors or just by the Frame Generator)	Certain: • Received Signal Strength Indicator (RSSI) (8 bits) • Doppler indicator (8 bits) Optional: • Frame Interpreter temperature • PCB temperature (read out by both processors or just by the Frame Interpreter) • Received erroneous frame counter (8 bits) • Received frame counter (16 bits)

Mode switching and data preparation (transmission and reception)

This subsection presents the last two functionalities, as switching between modes yields certain data preparation actions.

Mode switching only applies to the Frame Generator. The Frame Receiver only has one mode, the default 'receiver' mode. The four modes of the Frame Generator are *equal to the four PTRX modes*. The Frame Generator is in charge of activating and deactivating the Transponder AGC and switching the OLFAR switch; this has been illustrated by Figure 9.8 above.

The following Table 9.4 shows the possible PTRX modes including the peripherals, two of which are to be changed by the Frame Generator itself. The consequent required data preparation actions are demonstrated. Similarly, the more evident Frame Interpreter mode and its required actions are shown in Table 9.5. A more detailed description of all data preparation activities has been given in chapter 8 on the UHF/VHF transmission link.

Table 9.4: Frame Generator modes and consequent data preparation actions

PTRX mode	Peripherals	Frame Generator Data preparation actions
TM mode	• PTRX TX power supply on • Transponder AGC off • OLFAR switch set to TX	• If data is received from the OBC: o AX.25 frame application • If no data is received from the OBC: o Generation of continuous flags • In both cases: o NRZ-I encoding o Raised cosine PWM
Transponder mode	• PTRX TX power supply on • Transponder AGC <u>on</u> • OLFAR switch set to TX	• Generate Morse signal (see section 9.8) o Generate Morse text message by means of PWM
OLFAR mode	• PTRX TX power supply on • Transponder AGC off • OLFAR switch set to <u>OLFAR</u>	*None*
TX OFF mode	• PTRX TX power supply <u>off</u>	*None*

Table 9.5: Frame Interpreter modes and consequent data preparation actions

PTRX mode	Peripherals	Frame Generator Data preparation actions
Any	• PTRX RX power supply on	• When a signal is received: o Bit interpreting (removing Manchester) o NRZ-I decoding o AX.25 frame removal o TC decryption o *Putting the TC in the buffer and await pick-up by the OBC*

In the last table it can be seen that the consequence of moving the frame interpretation functionality to the PTRX is that a buffer has to be used. The OBC will consequently poll for TCs periodically as the PTRX processors will functions as slaves.

9.7 Telecommand interpretation on the PTRX

A functionality that has been suggested to incorporate in the PTRX has been *the ability to interpret telecommands*. In case of a failure of the OBC, this would mean the PTRX could put received telecommands on the bus directly. More specifically, this would concern the *receiver section* of the PTRX and its Frame Interpreter.

At the very least this would involve:

- **Required master switching functionality**. All subsystems should be able to receive commands from both the PTRX and the OBC. Currently, the PTRX RX is only integrated as a slave.
- **Required OBC failure verification.** A system should likely protect the bus from PTRX intervention in case no failure has occurred for security reasons. This could be done by letting the PTRX monitor the I^2C bus constantly to see if the OBC fails. In any case, the PTRX should provide the clock signal if the OBC no longer does it.
- **Increased software complexity and processor load**: Monitoring of the system bus as well as the functionalities required to properly address and send commands to systems increase software complexity as well as processor loading.
- **Increased overall risk.** Increased complexity logically gives rise to increases risk. This holds for all systems that should be able to receive data from two masters, but also for the mission-critical PTRX.
- **Decreased modularity**: As functionalities will overlap between different systems, overall modularity of Delfi-n3Xt would be decreased

The above scenario would only involve incorporating on the PTRX the ability to put telecommands directly on the bus. As such, the CDHS *would not be single-point-of-failure free* as a failure of the OBC will still drastically reduce mission capabilities. Therefore, more OBC capabilities can be placed with the PTRX. Of course, this would come at the cost of yet more increased complexity, more failure cases and thus increased risk.

In order to circumvent a single-point-of-failure free system without moving all OBC functionalities to the PTRX, it has been decided to design a system with two redundant OBCs on the same PCB [SLR 0105]. In this case, the question becomes *whether additional TC functionality on the PTRX would be worthwhile*.

This would however still incur the above disadvantages. The only extra failures cases that can in this case be resisted are:

- Software bugs and errors
- Common mode failures in the OBC hardware design and circuitry
- Specific second-degree failures

The latter failure case is however not part of the design-philosophy of Delfi-n3Xt; second-degree failures may jeopardize the mission. Given future results of the testing and integration of the (redundant) OBC, these and possibly additional failure cases should be assessed. If then in turns out that some additional TC interpretation functionality on the PTRX would be useful, it should be reassessed. Nevertheless, the above disadvantages should be taken into account seriously.

For now the conclusion is that, given the redundant incorporation of the OBC, *added TC interpretation functionality on the PTRX is not required*, and it would involve only limited extra functionalities at the cost of increased complexity and thus risk.

9.8 Linear transponder

As introduced in chapter 5, the main return service to the radio amateur community is the inclusion of a linear transponder. As it shares hardware architecture with the PTRX receiver and transmitter, it is logically included completely on the PTRX board. This section introduces the functioning of the linear transponder.

When the PTRX is set to transponder mode, it will stop transmitting telemetry, but at no time it will stop being able to receive telecommands. The transponder functionality is explained below in section 9.8.1. At the time the transponder is activated, a signal will be continuously transmitted, or broadcasted to announce the active transponder. This signal, *the transponder beacon*, and its contents are explained in section 9.8.2. A final application worth mentioning is that of ranging, commented on in section 9.8.3.

9.8.1 Transponder functionality

The hardware architecture of receiver and transmitter is shared with the PTRX. This automatically means that the UHF band will be used for the transponder uplink, and the VHF band for the transponder downlink; the transponder can be said to be of *mode-UV*.

A signal received in the uplink band will be transformed to the downlink and *inverted*. The inversion means that a high frequency in the uplink (i.e. 435.570 MHZ of .530-.570) will be transformed to a low frequency in the downlink (in that case 145.880 MHz of .880-.920). The benefit of the inversion is that downlink Doppler shift can be subtracted from uplink Doppler shift. Also, as explained before, *linear amplification* is required over the whole bandwidth to not distort received signals and therefore allow a wide range of signals to be forwarded.

The transponder functionality only involves a physical layer; data is not interpreted. Independent of the significance of the signal that is transmitted, it will be returned. As such, the depth of discussion in this top-level link establishment is low. The transmission frequency will be transformed and the signal will be amplified. These functionalities have been shown in Figure 9.3 above.

The linear transponder was first implemented in the RAP of Delfi-C^3. It has flown and operated successfully, for a reasonable amount of time. It is a secondary functionality, and as such possible improvements to the design are not of high priority. Therefore it is suggested to keep the design identical to that of Delfi-C^3. This conclusion goes hand-in-hand with the suggestion from AMSAT to keep the transmission frequencies the same.

For a more detailed design description of the transponder, also reviewing electrical figures, one is referred to [SLR 0281] where the PTRX is further discussed.

9.8.2 Transponder beacon

As introduced above, the function of the transponder beacon is to signal the activation of the transponder mode (to radio amateurs). Its function is therefore not complicated, and no strict requirements exist as to how it should be done. For Delfi-C^3 it was therefore chosen to send a simple Morse code, using a technique known as continuous wave or CW transmission. The name of the latter does not properly explain its application, but it includes switching on and off a continuous wave signal. The language used to wrap information into the transmission is that of Morse code. The idea is that the signal can be received even by the simplest receiver set-ups.

CW transmission is in fact a variation to a type of digital modulation explained in chapter 7, being amplitude shift keying or ASK, or more specifically on-off keying (the case in which one of the two amplitude values is zero). While Morse code does involve a primarily binary alphabet, of 'short' and 'long' characters, it also includes spaces and the length of each character is different. As such, *CW is actually a type of the more general amplitude modulation or AM*.

What is interesting about using a CW transmission is that it can in fact be generated by the BPSK modulator. In effect, the most simple approach would be transmitting a string of 1's *or* 0's to create a constant signal, which combined with the 10.67 MHz IF carrier would yield a perfect sinusoid; consequently turning this transmission on and off allows for a Morse beacon with two signal frequencies at IF; 10.67 MHZ and 0 MHz. Nevertheless, a slightly more advanced integration is chosen.

The 'off' Morse code before modulation will indeed simply be portrayed by a string of 0's, resulting in no transmitted signal. The 'on' Morse code however will be portrayed *by a sinus of 1200 Hz*, generated by the Frame Generator (digitally and followed by a D-A conversion). The benefits of this are:

- **Reduced transmission bandwidth**: no abrupt signal changes take place
- **Increased testability**: the 1200 Hz tone is an audible tone and the Morse generator can then be tested before frequency up-conversion even takes place.

The PTRX Frame Generator can generate this stream of data itself independent of the OBC when put into transponder mode, and it can bypass operations as NRZ-I encoding the data and pulse shaping. As transmitter power is to be shared between transponder and beacon, the beacon signal should have lower signal strength than a normal BPSK signal. As such, the stream of 1's that will be transmitted (as a matter of convention) is transmitted with a signal power value of 10 dB less than normal. The pulse delivered to the BPSK modulator is then 5 times less in amplitude -> a halved amplitude gives a quarter of the power. Also, 10 dB equals a factor of 10.

The transponder beacon transmission frequency options are given in chapter 8, and depend on what frequency will be allocated to the PTRX.

The Morse alphabet

The Morse code is defined by a number of simple rules. As said above and generally known, the alphabet consists of a 'short' character, a 'long' character and a space or silence. The combination of the three characters can be used to form letters, which are the general letters of the international alphabet. Definitions have also been established for general symbols such as the hyphen (-) in Delfi-n3Xt. This leaves open for interpretation the relative length of the characters.

The rules are:

- A dash is equal to tree dots
- The space between parts of the same letter is equal to one dot
- The space between two letters is equal to three dots
- The space between two words is equal to seven dots

The absolute length of the characters however thereby is not yet determined, as everything has been described in terms of *the length of a dot*. To give a rather extensive description; transmission speed is generally denoted by average amount of words per minute. As word lengths again differ in size, a general reference is the word PARIS as it has average length in terms of characters, being 50 dots. For a common speed of 20 words per minute this yields 60 milliseconds per dot. In terms of the Delfi-n3Xt application of generating a 1200 Hz tone, this means *one dot equals 72 complete sinuses*.

The transmitted message

In case of Delfi-C^3 the transmitted message was:

"HI HI DE DELFI-C3 DELFI-C3", followed by ten dots

The message is based on an international standard of broadcasting one's presence; 'de' signifying the French 'of'. As such, the Delfi-n3Xt transponder message is proposed to logically be:

"HI HI DE DELFI-N3XT DELFI-N3XT" followed by ten dots

9.8.3 Ranging

The communication system, or more specifically the transponder system, can be used for performing satellite ranging. Ranging is a method of determining the distance between objects. Ranging is not a required feature, as two-line elements or TLEs of Delfi-n3Xt can easily be acquired as the satellite is automatically tracked by NORAD, a US defence initiative.

Both one-way ranging and two-way ranging can be performed. One-way ranging requires high-precision components on the spacecraft [SLR 0371], whereas two-way ranging does not. For two-way ranging, the satellite needs to have some form of a transponder that relays the signal transmitted to the satellite by the tracking station back to the Earth. On Delfi-n3Xt, a suitable transponder is of course present.

As such, optionally ranging experiments can be done with Delfi-n3Xt using the PTRX transponder in order to determine the orbit. For these experiments to be accurate, tests are required on-ground to characterize the linear transponder in terms of time delay, phase noise and bandwidth. The ITRX also has ranging functionality; this is one of the experiments that will be conducted with the ITRX.

The ranging method can be code tracking (for coarse measurements) or carrier phase tracking (for fine measurements), or both. The accuracy that can be achieved depends on the properties of the linear transponder.

In conclusion, if time allows for it, ranging experiments can be performed. A spare model of the transponder can be used to characterize it; as such, the experiments can be prepared even after launch of Delfi-n3Xt.

9.9 Decreasing the gain of the PTRX Power Amplifier

Due to the 'redefinition of Delfi-n3Xt', the anticipated available power will be much lower than first assumed. Delfi-n3Xt currently is slated to use more power than Delfi-C^3 in nominal operations and it is desired to have continuous operation throughout the eclipses using the battery. Therefore this does not simply allow for reapplying the Delfi-C^3 solar panel configuration without additional power saving measures. Even if the solar panel configuration is changed, these measures might be required.

One of the measures that have been suggested is to reduce the power consumption of the PTRX. The PTRX is necessary in all nominal operations, as it is the primary transceiver. Even if an ITRX or even STX might be able to deliver the same downlink data transmission at lower power consumption, the power budget should be designed to support the PTRX.

The PA used on the PTRX allows for decreasing its gain and thus supply power relatively easily by tweaking its resistor network. As has been shown in chapter 8, even for data rates of 9.6 kb/s and 850 km orbit altitude, a reduction of 2 dB is acceptable according to the link budget. For lower final data rates, even more link margin remains. It might not seem very desirable to give up the entire margin to save on power, nevertheless, a suitable reduction can be considered if required from a power point-of-view.

9.10 Possible experimental features of Delfi-n3Xt

Two experimental features that might be implemented on Delfi-n3Xt are *software upload capability* and a *variable voltage bus*. This section quickly discusses their impact on or relation to the PTRX in subsections 9.10.1 and 9.10.2 below.

9.10.1 Software upload capability

This feature was actually already introduced in the last chapter on the UHF/VHF transmission link with respect to the data protocol. It can be very useful to be able to correct for malfunctioning software or to adjust software capabilities to actual in-orbit Delfi-n3Xt performance. In case of the ADCS it could allow for different experiments to be defined later on in the mission.

In fact, the PTRX would be little impacted other than in software. As such, the only requirement is that the communication protocol between the CDHS and PTRX is well-defined in case a larger number of continuous packages is received. In order to do so, the buffer of the PTRX microcontrollers or more specifically that of the Frame Interpreter should be (temporarily) enlarged. Nevertheless, this should not be a problem. In other words, this experiment is completely feasible as far as the PTRX is concerned.

9.10.2 Variable voltage bus

The variable voltage bus is a bus that might be implemented to basically present all left-over power to any system. In theory, COMMS could use this power for an increase of transmission power or data rate. It is however not preferred to connect this bus to the PTRX, for the following reasons:

- **The PTRX amplifier is designed to work on a predetermined supply voltage**
  A high-efficiency class C amplifier is however being designed, primarily for the ITRX. This type of amplifier can accept a variable supply voltage. Even if however this amplifier would be implemented on the PTRX, the electrical design of a system that would guarantee to provide the minimum amount of power required by the PTRX, but possibly more by means of a variable voltage bus, is difficult to design and implement.

- **The PTRX is a mission-critical design**
  Any additional changes might only serve to decrease the reliability of the system, at the cost of more required project resources. Sporadically increased transmission power is also not required for primary functionalities.

- **The variability of the variable voltage bus is, at least at this point, unknown**
  If it would be very high with fast transitions, unwanted effects might be incurred as the transmitted signal might become too difficult to properly receive.

9.11 Status

Currently the PTRX is still under development. Three main sections have been the target of development:

- **PTRX processors**
  The general software for the PTRX processors has been written. The DelfiX protocol has been integrated, and the software has been prepared for 2400 b/s downlink and 600 b/s uplink. I^2c protocols and telecommand definitions should still be integrated. A higher bit rate can be integrated relatively simply in software. The software has also been prepared for the proper read-out of housekeeping sensors and mode switching.

- **PTRX transmitter**
  Currently, all analogue components of the transmitter section are selected, and problems with the BPSK modulator have been solved. All analogue components have not yet been interfaced to form a working analogue transmitter section and also the Frame Generator has not been connected to the analogue transmitter section.

- **PTRX receiver**
  An updated design of the receiver has been made, but no working hardware is available yet. The option to increase the data rate is also still to be assessed in terms of components.

The required consequent steps:
1) As soon as working analogue transmitter and receiver sections are available, they should be interfaced to respectively the Frame Generator and Frame Interpreter and functional tests should be performed.
2) Consequently, the transponder section can be incorporated based on the RAP design, and complete functional tests can be performed.
3) Afterwards the PCB design can be made, including also all housekeeping sensors. The PCB can be fabricated and complete functional testing can be performed. Possibly this might already include the standard system bus as to be implemented on Delfi-n3Xt.
4) Finally, processor software can be upgraded to include all required functionalities.

9.12 Changes compared to RAP

There have been a number of (serious) changes, or *improvements* to the RAP design. These are:

- **The system has only digital data interfaces; it is therefore more modular**
  The RAP had a separate signal line from the demodulator on the RAP to the OBC, which was charged with data interpretation
- **The RX and TX sections are fully separated**
  By including two complete system bus interfaces, the TX sections can be switched off in hardware while the RX sections remains active. This is useful due to the large power constraints of the TX sections, and the essential functionality of the RX section. At that, it provides independent failure protection.
- **The PTRX transmitter can support higher data rates, possibly up to 9.6 kb/s**
  The Frame Generator is programmed to support higher data rates, 9.6 kb/s might even be allowed and used in operations. All other sections of the PTRX transmitter are unaffected by higher data rates.
- **The PTRX receiver can possibly support a higher data rate of up to 1.2 kb/s**
  Depending on its final design, 1.2 kb/s might be supported.

- **The data protocol has been customized for better performance**
  The DelfiX protocol allows up to 10-15% more data to be sent compared to the AX.25 protocol, in the same transmission time using the same data rate.
- **The overall electrical design has been reviewed and improved**
  The entire design has been reviewed and still is under development, but a number of components has already been changed. This includes the BPSK modulator, filters and crystals. In turn this improves on stability, performance and areal requirements, amongst others. In fact, part of the changes were required to replicate a working TRX on-ground, as last-minute changes to the RAP were not documented.
- **The OLFAR experiment is most likely connected with the PTRX board**
  This results in an extra switch on the PCB.

9.13 Future improvements

From a top-level point-of-view, a number of possible improvements can still be made:

- **A more efficient PA can be integrated in the PTRX TX**
  A switching amplifier is under development by Robin Kearey, and is slated to be integrated on the ITRX. This PA is currently specified to provide some 60% efficiency, which for similar transmitter power would decrease the PTRX power requirements by almost 1 W.
- **The downlink modulation scheme can be changed to QPSK**
  The current BPSK modulator in fact is a QPSK modulator, electrically connected to pose as a BPSK modulator. If it can be connected as a QPSK modulator, and on-ground demodulation software is adjusted, the required transmission bandwidth can be halved.
- **Frequency domain raised cosine filtering can be applied in the PTRX TX**
  Currently, raised cosine filtering in the time-domain is used. As chapter 7 has shown, the resulting frequency spectrum is in fact not compliant with regulations. Therefore, a more complicated Finite Impulse Response filter should be implemented, which is a digital filter or pulse shaper.
- **A look-up table can be used for quick CRC calculation in the PTRX TX**
  While requiring a sizeable memory size, this can free up essential processor power requires for higher data rates or FIR filtering.
- **Uplink data rates can be increased**
  This can be done by increasing the actual data rate or by removing the (need for) Manchester coding. Even if the data rate is increased to 1.2 kb/s, maximum data rates of 9.6 kBaud are allowed on the UHF radio amateur bands.
- **The uplink modulation scheme can be changed to MSK**
  This allows for a quick improvement of 3 dB in the link budget.

10 ITRX

The secondary transceiver on-board of Delfi-n3Xt is the ITRX, or ISIS transceiver. It is also a payload, manufactured and provided by ISIS BV. The primary functionality for Delfi-n3Xt is to serve as a back-up radio for the PTRX. However, as a payload it is supposed to demonstrate and test its correct performance in orbit. Also, it will be used to perform ranging experiments.

The ITRX will have characteristics similar to the PTRX. The main differences between ITRX and PTRX are:

- The ITRX has a high-efficiency power amplifier.
- The ITRX has a variable downlink data rate between 1200 and 9600 bit/s.
- The ITRX does not have a linear transponder.
- The ITRX has ranging functionality as well as a large number of other transmission modes.

As the ITRX is in fact a payload, it is not designed by the TU Delft or any part of the Delfi-n3Xt team. Therefore, there is only one CI as far as the team is concerned, see Figure 10.1 below.

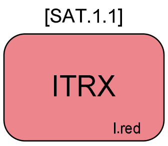


Figure 10.1: The ITRX CI.

The ITRX shall be treated as a black box. No detailed subfunctionalities are to be derived by the Delfi-n3Xt team and no components will be designed or selected. The operation and interfaces however are important. This chapter will focus on the definition of these.

The chapter follows the same structure as the chapter on the PTRX, but then more limited in content. The functional description is presented first in section 10.1, along with the operating modes. The external interfaces are explained in section 10.2. Afterwards, the housekeeping data produced is presented in section 10.3. The feasibility of some Delfi-n3Xt experiments in relation to the ITRX is then shortly discussed in section 10.4. Finally, the status of the ITRX is commented on in section 10.5.

This chapter actually gives a small summary of the communication aspects covered in the more general interface control document, or ICD, see [SLR 0182]. However, some aspects as the support of the DelfiX protocol and its interfaces are not yet included in the last version of the latter document.

10.1 Functional description and operating modes

As the ITRX is slated to serve as a back-up for the PTRX, it should at least provide the main functionalities of the PTRX:

- Receive telecommands (TCs)
- Transmit telemetry (TM)
- Generate housekeeping data

It is not required to have a second linear transponder. The ITRX is however a payload, and as such it has a number of other functionalities. These will be commonly named as the ability to:

- Perform ITRX tests

These main functionalities as well as the ability to perform ITRX tests leads to a large number of *operational modes*. As with the PTRX, telecommands can be received in *all* modes. Housekeeping data is also generated in *all* modes. The resulting mutually exclusive modes are presented in [SLR 0182] and are:

Essential modes for Delfi-n3Xt:

- **Receive only mode**: This is the default mode. It is basically the TX OFF mode of the PTRX.
- **DelfiX telemetry mode**: This mode is the normal mode for transmitting telemetry, which is also used by the PTRX. The ITRX will include the DelfiX data protocol in its operations.

ITRX test modes:

- **Morse beacon mode:** A beacon is transmitted continuously with a fixed Morse signal (CW).
- **Morse telemetry mode:** Similar to the Morse beacon mode, but instead of a fixed signal, telemetry is transmitted in Morse.
- **AX.25 telemetry mode**: This mode is the normal mode for transmitting telemetry, which is also used by the PTRX and the RAP of Delfi-C^3.
- **AX.25 beacon mode:** In this mode a signal is transmitted that consists of AX.25 flags.
- **ISIX beacon mode:** This mode is comparable to AX.25 beacon mode, but instead of AX.25 the ISIS protocol ISIX is used.
- **ISIX telemetry:** This mode is comparable to AX.25 telemetry mode, but instead of AX.25 the ISIX protocol is used.
- **Loopback mode:** In this mode any signal received by the receiver is transmitted directly. This is done for the purpose of performing ranging experiments.

All modes other than those essential for Delfi-n3Xt are only activated by means of TC and above Delft so that their functionality can be verified. As such, the only incurred complexity of these modes other than the usual ones is the required definition of an internal telecommand as well as a method for switching these modes off. The latter can be done by means of switching to another mode via telecommand, or using an automated timer in OBC or ITRX, after which regular TM transmission is resumed.

10.2 ITRX interfaces

The interfaces of the ITRX adhere to the standard system bus interface as defined for Delfi-n3Xt. However, depending on the actual definition of this interface, it is to be determined whether the ITRX will have an interface lay-out similar to the PTRX, having separated RX and TX sections on the bus, or not.

The ITRX will in fact be based on the *TRXVU* (VHF/UHF Transceiver – which in turn is based on the original RAP). The TRXVU has an I/O pin that can be used to switch on or off the power supply of the transmitter section. Therefore the transmitter section can in fact be switched on relatively simply, but in a manner not standard given Delfi-n3Xt interface definitions. An adjustment to the Local EPS can be made to be able to switch off the TX section directly, but as this involves extra complexity in a non-modular manner, it is not preferred. Possibly, the option of separated sections can also be ignored, given the fact that the ITRX itself is already a back-up radio and does not require an independent receiver section. In the end, the final implementation choice will lie with ISIS, given the input of the Delfi team. After the standard system bus has been completely defined, ISIS has promised to review the matter and discuss the approach.

10.3 Housekeeping data

The housekeeping data produced by the ITRX again strongly resembles that produced by the PTRX. The following Table 10.1 shows the resulting values. It should be noted that more data is produced during the ITRX test modes; this data is however referred to as payload data. Also, it is transmitted by the ITRX itself, so not important for Delfi-n3Xt data budgets.

Table 10.1: Housekeeping data generated during receive only mode

Telemetry value	Size [bits]	Minimum sampling rate [Hz]	Average bit rate [b/s]
PA temperature	8	0.2	1.6
PCB temperature	8	0.2	1.6
Forwarded power	8	0.2	1.6
Reflected power	8	0.2	1.6
Doppler voltage	8	1	8
Received signal strength	8	1	8
Receiver supply current	8	0.2	1.6
Transmitter supply current	8	0.2	1.6
Bus voltage	8	0.2	1.6
Total	**72**	**N/A**	**27.2**

It can be seen that a sampling rate is given, as in fact housekeeping data will not be produced every second. Two extra values are introduced with respect to those of the PTRX, being the *PCB temperature* and the *bus voltage*. The use of the former was commented on in the previous chapter, whereas the bus voltage is in fact implemented because of the modular implementation design of the ITRX (for customers). It makes sense for the ITRX to provide housekeeping data on both power reception and power consumption, but, in fact, in case of Delfi-n3Xt the local EPS already takes care of measuring the incoming power. If necessary, it can therefore be required to drop this variable.

10.4 Delfi-n3Xt experiments on the ITRX

Two experiments have been discussed in case of the PTRX, *software upload capability* and the *variable voltage bus*. It was concluded that the latter is not required or feasible in relation to the PTRX. The same holds for the ITRX, even more so as the ITRX is a payload and such experiments are not preferred by the payload partner and have not been incorporated in any agreements. The capability to upload software however, as explained before, comes down to a (small) software update and a proper definition of the interface. As such, would it be decided that software uploading is feasible, it can easily be discussed with the payload partner whether the ITRX will or will not support this feature.

10.5 Status

The ITRX is currently under development. Within ISIS sufficient radio experts are available, although Wouter Jan is unavailable for a prolonged period. The resources (money) required for the actual implementation of all presented features (improvements) have not been secured however. Nevertheless, the ITRX is as said before based on the TRXVU, which for Delfi-n3Xt purposes would be sufficient. As such, there is no direct reason to assume a working payload, at the same time being a redundant radio, will not be delivered on time.

For Delfi-n3Xt, the ITRX can theoretically also be replaced by a second PTRX with little implications.

11 OLFAR

OLFAR has been introduced in chapter 5, being a *technology demonstration experiment* which is part of the communication subsystem. It can be defined to be a radio, but it is not designed to receive human-made signals. Using the antenna system of Delfi-n3Xt, a naturally present signal in the bandwidth of 0.3 – 30 MHz will be received and amplified, and converted to an intermediate frequency. Sharing PTRX architecture, it should consequently transmit the signal over the transponder transmission bandwidth, to be received on-ground.

OLFAR has no high priority and is not allowed to have mission-slowing impacts. Therefore, human resources should be found to finalize the OLFAR experiment that would otherwise not logically provide mission-critical functionalities. OLFAR and its technical design and design considerations are introduced in detail in [SLR 0170]. This chapter gives only a functional and hardware introduction, in order to specify the systems engineering impacts of the OLFAR experiment on Delfi-n3Xt.

Following the same structure as the PTRX chapter for as far as possible, the OLFAR functional description is first given in section 11.1, followed by the definition of the CIs as well as the hardware components in sections 11.2 and 11.3 respectively. Section 11.4 then describes possible OLFAR integrations and section 11.5 finishes off by giving the current status of OLFAR.

11.1 Functional description

OLFAR has only one simple function, which is to:

- Receive signals in the frequency range 0.3 – 30 MHz and prepare these for retransmission

The PTRX will consequently retransmit these signals. Also, it is envisioned that no housekeeping data is generated by the OLFAR experiment as only relatively simple and low-power electronics are applied.

A number of subfunctionalities can be distinguished in order to perform the main task, given the interfaces with the antennas and the PTRX transmitter section. These subfunctionalities are shown in Figure 11.1 below.

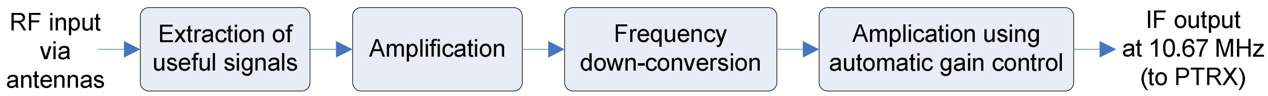


Figure 11.1: Extra functionalities required to facilitate the OLFAR experiment.

Signals (that are automatically received on the UHF/VHF antennas) should first be extracted, and other components should be filtered out. The resulting signal should be amplified, and converted to the right frequency in order to join the PTRX transmitter path with the proper IF of 10.67 MHz. Finally, the signal is brought to a predetermined power value, similarly as is done in the transponder.

11.2 Configuration Items

The subfunctionalities can be used to deduce a number of CIs. These CIs can be seen in Figure 11.2 below. Most CIs and their meaning are assumed to be straight-forward. The OLFAR Phasing Circuit should serve to extract useful signals and filter unwanted signals. The OLFAR Switch at the end was already introduced in the chapter on the PTRX and is required to join the OLFAR and PTRX paths, and to switch between modes.

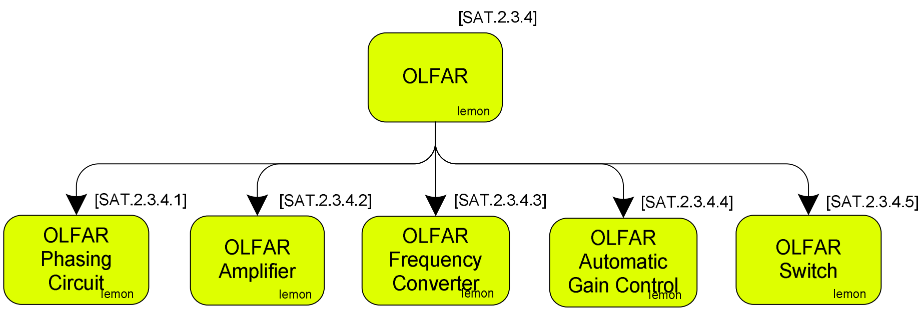


Figure 11.2: OLFAR Configuration Items.

11.3 Hardware components

Given the required subfunctionalities, a component design can be set-up. This diagram is shown as Figure 11.3 below. The design rationale will be given in [SLR 0170]. A short description of the functioning of this circuit is given below the figure.

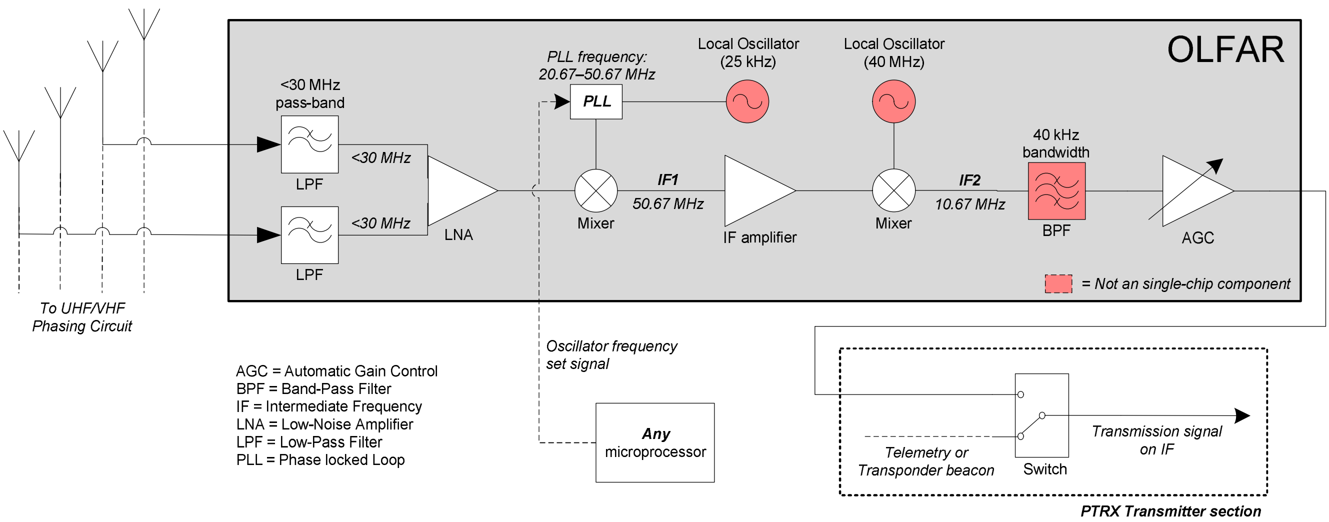


Figure 11.3: OLFAR hardware components.

The basic idea of the OLFAR design is that the desired signals are extracted from the signal received automatically by the UHF/VHF antennas. However, as these antennas are the same ones that transmit VHF signals, including the OLFAR signal, the transmitted signal should be suppressed. This signal is not centred in the 0-30 MHz range, but as its power even at low frequencies is much larger than naturally present low-frequency signals, harmonics and intermodulation products pose a serious disturbance.

In order to suppress the transmission signal, OLFAR extracts signals from the first and third antennas, which individually transmit a signal with a 180 degree phase shift with respect to each other. When summing these signals, the transmitted signal is completely suppressed. Two low-pass filters filter out high-frequency components so that only the low-frequency components remain, of frequencies 0 to 30 MHz.

Consequently a low-noise amplifier combines the signals and increases the power content. Afterwards frequency conversion will take place. This consists of two steps; a frequency up-conversion and frequency down-conversion. This is because the final IF lies within the range of 0-30 MHz; frequencies of interest can either have a frequency lower or higher than the IF. If the entire signal would not first be up-converted, the frequency conversion step would at times have to down-convert, and at times have to up-convert. Therefore the entire 0-30 MHz band is lifted over the 10.67 MHz. A phase locked loop (PLL) integrates a VCO (voltage-controlled oscillator) to change the conversion frequency.

As the final transponder bandwidth is only 40 kHz broad, the 40 kHz of interest was effectively selected in this last phase. The consequence is that there is a need for a processor that supplies a VCO input value, originating in the OBC. In other words; ***a processor is required to control OLFAR operations***. This is an important conclusion that is further evaluated below.

After up-conversion the frequency is brought (back) down to 10.67 MHz, the IF of the transmitter section of the PTRX, and the signal bandwidth is brought back to 40 kHz. An AGC sets the signal's peak power to a predetermined value, and the signal is forwarded to the PTRX transmitter section, joined together by means of a switch.

It is important to note that all components within the OLFAR box, with the exception of the three components marked red, are *to be integrated in one chip*. As such, a small chip (some 1 cm^2) with 24 interface pins results, supported by two stable crystal oscillators and a sharp bandwidth limiting filter. To give a simple value for the total integration space required, the four components can be estimated to require 4 cm^2.

11.4 Possible OLFAR integrations

In order to establish the impacts of the OLFAR experiment, this section evaluates the OLFAR integration options. Three options have been suggested throughout time, which will be introduced below in subsections 11.4.1 through 11.4.3, and traded off in section 11.4.4. All options take into account complete integration on either or both the PTRX PCB and Phasing Board. In theory another PCB could house the components required, but as the input comes from the Phasing Board and the output should be presented at the PTRX PCB it makes no sense to place them somewhere else. This in turn would introduce even more cables and complexity. As almost all components are integrated in one chip, the integration is either mainly on one PCB, or the other.

11.4.1 Completely integrated on PTRX

The first idea at the conception of OLFAR was to basically have an implementation similar to that of the transponder, on the PTRX. OLFAR mode could be switched on, at which point the OLFAR signals would be extracted from the RF signal received on the PTRX PCB via a preexisting connection. The required impact would be limited to:

- Area on the PTRX PCB
- Power from the Local EPS of the PTRX
- A VCO interface with the PTRX Processor, and slightly increased complexity for the latter

This option however is no longer possible due to the integration of the UHF and VHF antennas to a combined set of UVF/VHF antennas (see [SLR 0036]). This is due to the interference on OLFAR signal reception resulting from data transmission. It might already have been difficult in case of separated antennas. Therefore, this solution is basically not an option.

11.4.2 Mainly integrated on PTRX

A second option, solving the problem of the previous solution, is to retain having the most components on the PTRX, but moving the functionality of 'extraction of useful signals' to the Antenna System, or more specifically the Deployment and Antenna Board (DAB). The latter houses the Modular Antenna Boxes (MABs), their deployment systems connected to a standard system bus and the UHF/VHF Phasing Circuit (as further explained in [SLR 0036]). All other components would however be kept on the PTRX. Moving this functionality would basically mean having two signal 'taps' connected to two of the four antennas. Total required impacts:

- Area on the PTRX PCB, and some on the DAB
- Power from the Local EPS of the PTRX
- A VCO interface with the PTRX Processor, and slightly increased complexity for the latter
- 2 COAX cables between the Phasing Board and the PTRX

However, one problem is created by this type of set-up. The OLFAR entrance filters have a very high-impedance. Indeed this high impedance is required as no signal of the PTRX, reception or transmission, should be lost to OLFAR. However, this requires that the filters be connected to the regular signal lines with minimum extra incurred distance. If however a sideline is created, which by means of a cable should transport a signal to the PTRX *before* being filtered, the high impedance is lost.

In effect, the addition of a transmission line creates what is called an *open stub*. Using different cable lengths the impedance can be tweaked, but [SLR 0715] shows that it is not a matter of a few centimeters. It can however be resolved by introducing extra circuitry. Therefore, this second option would also require:

- Additional phasing circuitry on DAB or PTRX that is yet to be designed

11.4.3 Mainly integrated on Phasing Board

A third option then is to move the overwhelming amount of components to the DAB, and have only a switch on the PTRX to join the paths. This removes the problem of having a need for additional circuit complexity such as in the second option, but at least requires space on the DAB. The consequent required impacts:

- Area on the Phasing Board, and some on the PTRX PCB
- Power from the Local EPS of the DAB
- A VCO interface with the DAB processor, and slightly increased complexity for the latter
- 1 IF cable between the Phasing Board and the PTRX; this cable can be less thick and protected than those required for RF signal transportation

It can be seen that this third option shifts occupancy and complexity to the DAB, but requires a simpler cable to transport the IF signal.

11.4.4 Conclusion

The first option, as explained, is no longer feasible. As such, the second and third options remain.

In both cases power and a processor interface is required, on either the PTRX PCB or the DAB. This is not a significant impact. Processors are in both cases in place and logically connected to the standard system bus.

Both solutions require cabling from the DAB to the PTRX, but in case of the latter only one, being a thin one. Similarly, the third option does not require additional circuit complexity as the second option would require. This favors the third option.

Finally there is the question of space. Both PCBs are not lightly packed.

- **In case of the PTRX**, space has always assumed to be available but no significant blank spots can be spotted on the RAP. However, COAX connections are assumed to be decreased in size. At the same time, the bus interface has been changed with unknown precise impacts as the design has not been finalized.
- **In case of the DAB**, space will be created by the removal of the old connector which in case of Delfi-C^3 was placed in the middle of the board. Also, in case the solar panel deployment systems will all be or all be operated from one side, the deployment mechanisms can be removed from the DAB. However, a slightly more complicated Phasing Circuit is necessary due to the integration of the antennas.

In both cases it holds however that only a small amount of space is required.

In conclusion, the space argument is still inconclusive, giving no security for space on either PCB, but likely space on both. Implementation on a different PCB would have far larger consequences and is certainly not preferred. The cable impact as well as the requirement of additional complexity and time investment of the second option would prefer the third option. As such, this option shall be assumed as the implementation for OLFAR. It should be considered at a later stage if OLFAR indeed has advanced far enough to merit the additional impact on the mission.

11.5 Status

The OLFAR hardware and its chip integration are under development by Edwin Wiek. The end-result of his work is a finalized chip design. Production and testing will remain to be done. As such, the experiment can in any case only be implemented if more resources are found for OLFAR testing and post-development.

12 S-band communication link

Similar to how chapter 8 did for the UHF/VHF communication link; this chapter specifies the S-band communication link. This link in fact is one that is only between the STX and the S-band ground segment; it is therefore not linked to the other communication link or its radios. Also, there is just one transmitter and just one receiver; it is a unidirectional link from satellite to ground.

As the STX is a new radio, this link is in fact not based on Delfi heritage. As such, more is unknown and the order of this chapter will be slightly different from that of chapter 8. The first topic to be discussed is again that of transmission frequencies, in section 12.1. Section 12.2 will consequently select the most preferable modulation scheme. No data rate is determined yet, as this requires the link to be further defined. Sections 12.3 through 12.5 deal with data link layer functionalities, being respectively forward error correction, data protocol and data compression. Finally section 12.6 uses all information presented before to determine a data rate approach and give estimated values, using knowledge from the technical note describing the STX design [SLR 0387] and the link budget [SLR 0106].

12.1 Transmission frequencies and bandwidth

As was already stated, the STX will use the S-band, and more specifically the radio amateur portion of the S-band. This band overlaps with the larger ISM band, used for industrial, scientific and medical applications, which ranges from 2400 to 2500 MHz. The latter has a centre frequency set at the logical centre of 2450 MHz, which means that lowest interference should in principal occur near the 2400 MHz frequency of the radio amateur band. This is due to the fact that many applications in this band, such as wireless internet, have varying data rates and thus varying overall spectrum occupancy; still, they are centred towards the middle to remain in the band even for high data rates.

A preliminary discussion has taken place with Graham Shirville, which has lead to the conclusion that:

- Frequency allocation should indeed be as close to the bottom of the band, but not within 2400 to 2400.1, to allow for a maximum Doppler shift. This Doppler shift, as calculated in chapter 3, is actually not more than 60 kHz.
- The resulting maximum data rate determines the required bandwidth and exact transmission frequency.

There seem to be no objections to the operation of the STX in the radio amateur band, and similarly no strong constraints exist as to what data rates are acceptable. This is due to the sharing of resources inherent to the S-band radio amateur and ISM band. The allocated bandwidth should be reasonable given the data rates however.

The transmitted power should be limited at the boundary of the allocated bandwidth. The same regulations apply for the S-band as for UHF/VHF frequencies. In other words, the minimal signal reduction out-of-bandwidth is as stated by requirement SAT.2.3-P.01 at maximum -43 - 10 log (*P*) dBc, where *P* is the transmitted power, *or* 60 dBc, whichever value is smaller.

In case of the STX, an amplifier is suggested in [SLR 0387], the SST12LP00, which would have an output value of 0.315 W. In this case:

- *Spurious emissions on the downlink should be limited to **-38 dBc***

As is explained in [SLR 0387], currently the *CC2500* is assumed to be used as modulator, and is a COTS chip produced by Texas Instruments. This chip has a maximum datarate of 500 kBaud. As explained later in this chapter, given the pass characteristics, this data rate could be reached momentarily. The CC2500 data sheet [SLR 0556] indicates that a default digital channel filter of 812 kHz is applied at 500 kBaud. It also indicates that using this filter, at the 'next channel', outside of a 1 MHz bandwidth, -21 dBc is reached. However, no bandwidth value is given for the required -38 dBc requirement.

As the CC2500 is available for testing, its bandwidth can be measured. The problem however, is that knowledge of the specific software is required to set the CC2500 to both 'MSK' and '500 kBaud', and finally to let it output a constant stream of data. Nevertheless, it can deliver 250 KBaud, MSK and its temperature sensor data once per second. This will not generate a constant stream of data, but its envelope can be monitored. The resulting figure can be seen in Figure 12.1 below.

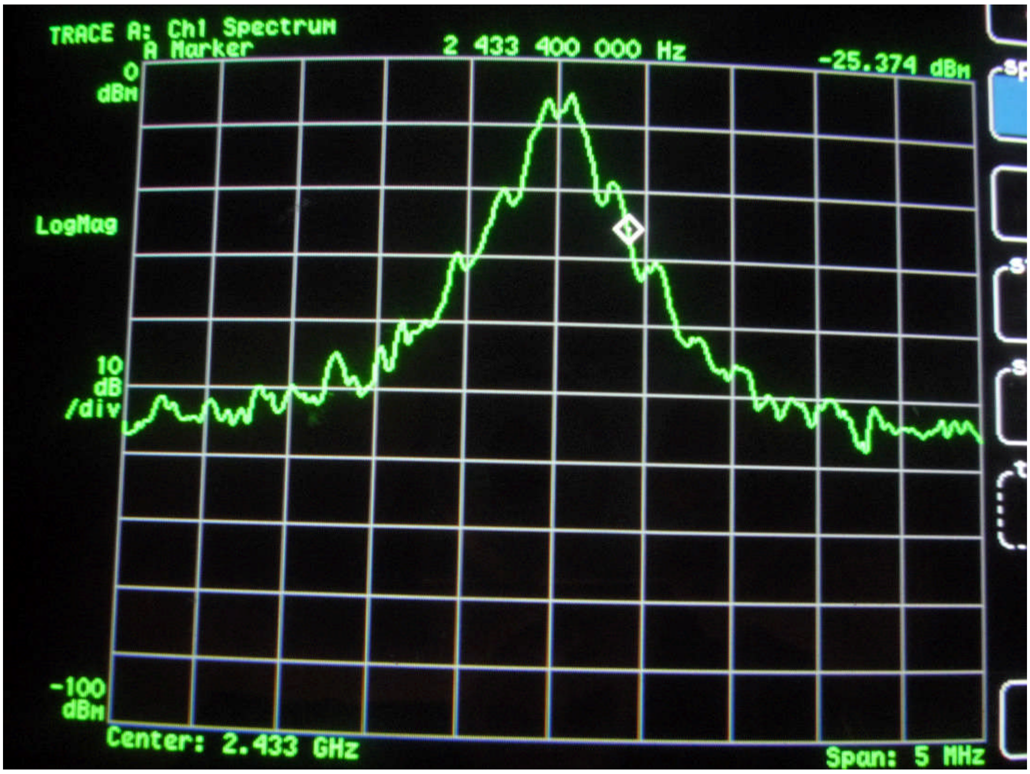


Figure 12.1: Measured frequency spectrum of CC2500 transmission (2500 kBaud, MSK) over a 5 MHz span

In the CC2500 datasheet [SLR 0556], -21 dBc is stated to be reached within 750 kHz for a 250 kBaud data rate. Also, this setting applies a digital filter bandwidth of 540 kHz. The figure above shows that, given the 5 MHz span, indeed 750 kHz is the approximate bandwidth for signals with a power value above -21 dBc. The required -38 dBc reduction then requires a bandwidth of *some 1.75 MHz*.

As for a data rate 500 kBaud a -21 reduction is stated to occur within 1 MHz, as opposed to 0.75 MHz for 250 kBaud, it is concluded that bandwidth does not scale linearly with data rate. This can have to do with the different digital channel filter bandwidth values of 540 and 812 kHz respectively for 250 kBaud and 500 kBaud settings. Therefore, it is concluded that some 2.5 MHz would be necessary to reduce the power to the necessary values, assuming MSK, 500 kBaud and an 812 kHz filter bandwidth.

In fact, the 812 kHz bandwidth can still be reduced. This could introduce ISI, likely at the cost of a measurable reduction in power performance. This can later be tested, if further bandwidth reduction is important.

Although the bandwidth of maximally 2.5 MHz is officially required, it becomes the question whether it is useful to file for such a large frequency bandwidth. This is due to the low power transmission. It *is* however important that the reduction of -38 dBc is reached at the boundary of the radio amateur and ISM bands.

As such, the centre frequency can be determined to be at 2.5/2+2400.1 MHz, in other words 2401.3 MHz. To take a little safety margin, **2401.5 MHz** is taken. This centre frequency can be lowered following measurements of the actual performance later on. Nevertheless, if frequency filing is commenced before that time, the above centre frequency can just be used; the difference in disturbances will be rather minor around within several tenths of a MHz.

The results of the performance measurements have again been discussed with Graham Shirville, after which he has suggested to *use the expected 2.5 MHz in the frequency allocation, but indicating that this is the bandwidth required to reduce the signal strength at the boundaries to -38 dBc.*

The results are summarized in Table 12.1 and Figure 12.2 below.

Table 12.1: S-band transmission frequency

Link type	***Band***	***Specific frequency***	***Comments***
STX downlink	Radio amateur S-band (2400-2450 MHz)	~2401.500 MHz	To be formalized

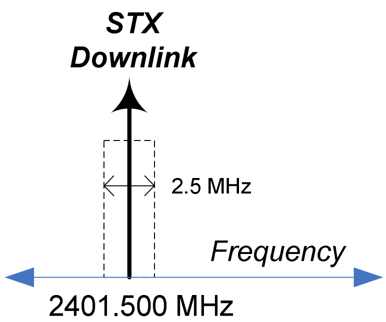


Figure 12.2: STX downlink frequency (bandwidth)

The exact bandwidth performance of the CC2500 at maximum operations should of course be tested. However as explained, as this requires some specific software to be written, the test has not been performed yet. When this test will be performed, attention will have to be paid to set the device to '**500 kBaud**' and '**MSK**'. Consequently, the effects on bandwidth and demodulation performance can be tested when the digital filter bandwidth is reduced.

12.2 Physical Layer: modulation scheme

As explained above, the STX is not a bandwidth limited system. On the other hand, the goal of the STX is to have *data rates as high as possible*. The system is therefore both:

- **Power limited and complexity limited**

Power requirements and complexity can be traded off against each other, for example by using forward error correction coding and directional antenna systems. As the STX is a system that should be designed from scratch, and at the same time project constraints limit the available resources, a suitable middle-ground should be found.

Chapter 7 has presented the different available modulation schemes. From a power point-of-view, it follows that the preferred modulation techniques are the higher order FSK methods, as they become gradually more power efficient at the expense of bandwidth efficiency. The simpler schemes such as MSK, BPSK and QPSK have good power efficiency as well, while at the same time requiring relatively low complexity.

Finally, there is one important aspect related to the power amplifier; as explained in [SLR 0387] a constant envelope scheme is preferred as this allows for amplifiers to be used that have a high efficiency but are non-linear, lowering their complexity.

In the following table the most relevant modulation methods are summarized with their power and bandwidth efficiencies indicated, as well as whether they have a constant envelope or not. The indicated bandwidth efficiency is the Nyquist bandwidth efficiency as is explained in chapter 7.

Table 12-2: Power and bandwidth efficiencies of relevant modulation schemes, *BER*=10^{-5}

Modulation scheme	***Power efficiency (required E_b/N_0)***	***Bandwidth efficiency (bit/s/Hz)***	***Constant envelope***
Coherent QPSK with pulse shaping	9.6	2	no
Coherent MSK	***9.6***	***1***	***yes***
Coherent BPSK with pulse shaping	9.6	1	no
Coherent OOK	12.6	1	no
Noncoherent BFSK	13.3	0.5	yes
Coherent BFSK	12.7	0.5	yes
GFSK	12.0	0.5	yes
Noncoherent 4FSK	10.7	0.5	yes
Coherent 4FSK	9.8	0.5	yes
Noncoherent 8FSK	***9.2***	***0.375***	***yes***
Coherent 8FSK	***8.5***	***0.375***	***yes***
Noncoherent 16FSK	***8.2***	***0.25***	***yes***
Coherent 16FSK	***7.5***	***0.25***	***yes***

The 'default' E_b/N_0 required for the simplest schemes is 9.6 dB. Power efficiency is almost halved (~3 dB difference) for the low order FSK schemes as well as for OOK. Therefore these latter schemes are not desirable. QPSK has no constant envelope. BPSK in fact does, however it is has unacceptably poor bandwidth efficiency for high data rates due to the occurrence of large sidebands. Therefore, it would require pulse shaping, which in turn renders the envelope non-constant.

The best-performing modulation schemes in terms of power efficiency are the higher order FSK schemes, as indicated in ***bold italics***. They do however become quite wasteful in terms of bandwidth and at the same time require increasing complexity. Also, power efficiency does not increase rapidly for increasing values of *M*. 4FSK however does not outperform MSK in terms of power efficiency and complexity, so is no good option.

MSK, also indicated, finally forms a solid option with acceptable power efficiency and low relative complexity. It has the added benefit that MSK is frequently used, also at S-band frequencies. As such, standard equipment can be assumed to exist; at least with higher certainty than for the higher order FSK schemes. In case of the STX system, the project constraints lead to a preference for COTS equipment. [SLR 0387] further expands on the argument of available COTS equipment, and indeed presents an available MSK modulator (the CC2500).

As detailed electrical integration has not started, the choice for this modulator and thus modulation scheme should not be final. Nevertheless, as power efficiency is good, a useable COTS component exists and has been purchased, and bandwidth occupancy seems acceptable:

- ***The preferred modulation scheme on the S-band communication link is MSK***.

Pulse shaping

As introduced in chapter 7, pulse shaping reduces transmitted bandwidth. Two filter types have been introduced in more detail due to their frequent application, *Gaussian filtering* and *raised cosine filtering*.

Raised cosine filtering leads to having no ISI. A frequency domain raised cosine implementation through the use of a FIR would quickly become very complex and undoable for high bit rates in the order of hundreds of kb/s. A simple time domain implementation would not give a very satisfying required bandwidth, and is quickly outperformed by a Gaussian filter. Gaussian filters come at the cost of only slight power performance, for low values of *BT*.

GMSK, thus Gaussian filtered MSK, has been introduced in and is used in the GSM system; mobile telephone communication. The frequencies in GSM are however lower, so no equipment is per definition automatically available. Nevertheless from its application in the GSM system some achievable performance characteristics can be taken. Similarly Bluetooth applies Gaussian filtering, but to a FSK scheme. From Figure 7.21 the bandwidth performance of Gaussian filters in MSK schemes can be seen for applied *BT* values:

- A *BT* value of 0.5 reduces the bandwidth to the required -43 dBc within ~1.1**R* [Hz]
- A *BT* value of 0.3 reduces the bandwidth to the required -43 dBc within ~0.75**R* [Hz]

The bandwidth of Gaussian filtered MSK signals is far superior to the spectrum generated by the CC2500. Nevertheless, bandwidth performance is said to be of limited importance within the radio amateur bands, given the large (shared) bandwidth available. Also, the CC2500 does not support Gaussian filtering (for high data rates and a scheme other than FSK).

Therefore, *pulse shaping will currently not be assumed to be required*.

Line coding

Line codes can serve to remove a dc-component, introduce extra timing information, add error correction or remove signal polarity. In general, this comes at the cost of power efficiency. As such, it is not desirable for the STX if not required.

Differential coding is actually already implied as MSK can be seen as differentially coded FSK. Error correction is more logically implemented on the data layer, and at the same time introducing more levels to the data signal would create a signal with a non-constant envelope. Finally, the removal of a dc-component and introducing extra timing information is not anticipated to be needed, if proper hardware is used. The CC2500 mentioned above, can transmit and in fact also receive MSK-modulated signals without requiring these extra line codes.

As such, *no additional line codes are suggested for the STX modulation scheme*.

12.3 Data link layer: Forward Error Correction

In case of the UHF/VHF communication link, it was not necessary to add FEC as the system was not power limited. Also, FEC comes at the cost of bandwidth efficiency and the UHF/VHF links are in fact bandwidth limited. For a link such as the S-band communication link, the addition of FEC then becomes a logical choice, if allowed for by the complexity limitations.

When it comes to introducing error correction, the STX has two other big advantages:

- There is no need for real-time or little-delay communication
- The STX only requires an encoder

Firstly, a lot of codes in use today have a direct need for little delay in either encoding or decoding, think of their use in telephone or data traffic, as well as in playing music. The STX will only transmit at given intervals, giving it has a long period to gather and prepare data. Also, at the ground station no requirements are stipulated for the speed at which information should be made available; decoding does not have to be real-time.

Secondly, it has been pointed out that advanced coding schemes require sizeable complexity at the receiver, however at the encoder it is a matter of applying one or two codes as well as possibly interleaving, multiplexing and scrambling. At the receiver this prescribes the use of iteration, maximum-likelihood processing and if possible soft-decision demodulation outputs. All except for the last aspect simply require more processing power, which is of little problem in modern day computers.

The field of error correcting codes is one that is very large and very complex. Given the limited resources it therefore makes sense to make use of proven concepts, and likely those being simple and straight-forward in their implementation. The radio amateur community is a good example for these kinds of concepts, as they have similar (satellite) applications, and in general they provide open information. As such, two good suggestions result from chapter 7:

- **A block code**

  Many block codes are well-documented, and are easily integrated in both encoder and decoder using pre-written software. As opposed to many convolutional codes, they simplify operations and data preparation as specified block sizes are used. The radio amateur FX.25 data protocol [SLR 0294] prescribes the application of block codes in satellite communication. A choice can be made between BCH and RS codes, the latter of which provide some burst error capability but less overall coding gain. Some concrete examples are given below. Gain values of up to 4 dB can be obtained. Possibly, soft-decision block decoding can be applied, optimally increasing the gain to 5-6 dB.

- **The AO-40 FEC coding scheme**

  This coding scheme approaches the maximum achievable performance, given the constraints of availability and intellectual property. It can yield some 7 dB coding gain, and software has been written and is in principle available.

A further option of minimum complexity is provided by the CC2500 modulation chip:

- **A convolutional code**:
  The aforementioned CC2500 chip provides certain FEC features in its packet handler functionality. It can be set to apply a rate ½ convolutional code with a memory size of 4 bits, also applying a 4 by 4 interleaver matrix if preferred. Hard-decision decoding (such as provided by the CC2500 integrated receiver itself) gives rise to an estimated 2-3 dB coding gain at best. This coding scheme is logically applied 'on-the-fly' or real-time by the CC2500. A possible (soft-decision) Viterbi decoder could give an estimated 1 dB increase of performance given the small code memory.

Table 12-3 below summarizes the selected options, giving an indication of the performance as well the integration complexity.

On-ground complexity is expressed given the required on-ground equipment. The CC2500 yields its own demodulator and decoder, however possibly giving worse performance (such as sensitivity) than larger equipment. Hard-decision decoders only require simple demodulation, also possibly done by the CC2500. Soft-decision demodulation requires dedicated equipment.

Table 12-3: Possible STX FEC coding schemes with their performance and integrated complexity

Code type	***Code***	***Coding gain***	***Applied by***	***Software status***	***On-ground requirements***
Convolutional	CC2500 preset *R=0.5, hard-decision*	2-3	CC2500	Written and integrated	little
	CC2500 preset *R=0.5, soft-decision*	*3-4*	CC2500	Written and integrated	large
Block	Reed-Solomon *hard-decision*	~3	STX Processor	Presumably available	moderate
	BCH(127,106,3) *R=0.84, hard-decision*	3	STX Processor	Presumably available	moderate
	BCH(127,71,19) *R=0.56, hard-decision*	4	STX Processor	Presumably available	moderate
	Any, soft-decision	5-6	STX Processor	unknown	large
Concatenated convolutional and block	AO-40 FEC scheme *R=0.4, soft-decision*	7	STX Processor	Available, but not integrated	large

A definite trade-off cannot be performed, quite simply because not all data is available; on actual performance, software status and demodulation and decoding equipment. More important however: Delfi-n3Xt project resources. Three concrete options:

1. **Preset CC2500 FEC coding scheme**
   This is the simplest option by far as it is integrated in the CC2500 pre-written software. It can be simply switched on for a minimum coding gain. Then, if hard-decision demodulation is used, the same CC2500 chip can be used-ground to handle signal demodulation and decoding.

2. **BCH block code**
   The two BCH options are concrete and have defined performances. Whereas one gives better coding performance, the bandwidth performance is less which might limit achievable data rates with the CC2500 given its maximum data rate of 500 kBaud. Soft-decision decoding is less documented and therefore not anticipated.

3. **The AO-40 FEC coding scheme**
   This scheme is the most desired, as it gives best overall performance. It should be achievable given the fact that software and required ground station equipment is available.

Next to the technical aspects given above, the AO-40 FEC coding scheme has been tested, applied and its software is readily available. Also, this scheme is at least in some form supported by the radio amateur community, which can increase download potential and can increase radio amateur support of Delfi-n3Xt. Therefore it is concluded that:

- *The implementation of the AO-40 FEC coding scheme is preferred and poses the design goal for the STX. However, the preset CC2500 FEC coding scheme, as the name implies, can be preset in CC2500 software to form an absolute minimum at no cost. If the A0-40 FEC coding scheme turns out not to be feasible, the BCH block code scheme can be considered, taking into account available resources and software. The BCH block code is preferred over the CC2500 coding scheme due to similar coding gains, but the wasteful code rate of the latter.*

During STX operations, it would be easy to switch between FEC schemes if any scheme other that the CC2500 scheme is also implemented. This would logically be done by means of telecommand.

One constraint regarding the A0-40 FEC coding scheme:

- The code has been designed for a 400 b/s BPSK transmission signal [SLR 0732]. The stated data rate leads to having a fixed and relatively small number of input bits of 2048, while the stated modulation scheme leads to synchronization capabilities having been optimized for BPSK signals and BPSK equipment.

The synchronization capabilities of the AO-40 FEC coding will in any case not be a disadvantage. The amount of input bits is similarly not a problem and is taken along in the discussion on the data protocol in the next section.

12.4 Data link layer: data protocol

Three suggested options of various impacts have been suggested in the last section. A preference was expressed for the AO-40 FEC coding scheme. For all three options, specific data protocol requirements are incurred, which are discussed below.

The STX should receive from the OBC the exact data that is send to the PTRX or ITRX, for operational simplicity (requirement SAT.2.3.2-P.01). From the fields established within the DelfiX protocol that means it will receive *frame content* in the form of **Information field + Frame Type ID**. Also, within the information field a *frame counter* will be included to be able to identify the frame. The frame content is of current undefined length, pending the confirmation of the final downlink data rate and an update of the data budget.

In the following, a *frame* will always refer to a Delfi-n3Xt frame. If coding content of predefined length is discussed, this will be referred to as a *block*.

As specified by the last section, FEC coding is applied on the frame content. However, additional fields should be added:

- Before encoding: required for frame interpretation
- After encoding: required for demodulation and decoding

These fields together again form the *data protocol*. The following individual fields can be distinguished:

- **Preamble (required for demodulation)**
  A certain preamble is required for the modulator to achieve *signal synchronization* (chapter 7). Demodulation takes place *before* decoding, thus preamble should be added *outside of* coding fields. The type of preamble required is independent of the type of FEC applied as it depends on the modulation scheme. Its specifics for the STX are introduced below.

- **Block boundary indication (required for decoding)**
  A FEC decoder based on some kind of block code should know where to begin; as such some indicator should be included at the start. If frame or block length is constrained, no end field is required. The difficulty in applying this indication is that it should be placed outside of encoded fields. *The indicator should therefore be able to be detected in a noisy environment.*

- **Frame boundary indication (required for frame interpretation)**
  The frame interpreter should know what amount of bits signifies a frame. In the DelfiX protocol, the *flags* are used for this purpose; they indicate start and end. To circumvent the signaling of false boundaries, bit stuffing is then also required. For the S-band communication link another approach is chosen, as 1) performing bit stuffing before encoding has no use for purposes of establishing a link and 2) additional bit stuffing itself is not required for MSK demodulation. Therefore it is suggested to make use of the:
  - *DelfiX flag*
    The flag is a simple 2-byte indication of the start or end of a frame. It can easily be scanned for, as it is a simple sequence. It cannot be assumed to be a unique marker, as the data stream can quite simply include the 16-bit sequence. Therefore the unique marker will be the *combination of the flag and the address field*.
  - *DelfiX address field*
    Whenever a flag is located, the next field should be the address field. This 3-byte indicator is known for STX data. The 5 combined bytes of flag and address field give a pseudo-unique stream that is not likely to reappear in the data stream.

The same combination of flag and address field can be used to indicate the end of a frame. As the frame length will be known, an end marker is not strictly necessary. However, having knowledge of the frame length *and* having an additional end marker allows concluding on the proper reception of the frame. If a decoded block that is part of a frame has been discarded, the length between two indicators will be too small. In case multiple frames are transmitted directly after each other, *the start indicator of the next frame can function as the end indicator for the previous frame*.

- **Frame content control (required for data interpretation)**
  In the DelfiX protocol, a CRC is added to check for the proper reception of all frame bits. In case of FEC coding, any resulting errors in the data field are not likely restricted to single, double or possibly triple errors, as they result from single, double or triple errors in the *encoded data stream*. Nevertheless, the addition of a CRC provides a good additional check. Also, 99% of larger other amounts of errors are still spotted. As such, the following field can be added:
  - *DelfiX Frame-check sequence*
    The FCS field will contain the CRC. The same CRC as applied in the DelfiX protocol is suggested, as it is proven in its use and gives good result for long frames.

In conclusion, the preliminary work that has to be performed by the *STX processor* on the data received from the OBC will be, *irrespective of the FEC coding scheme*, as indicated in Figure 12.3 below.

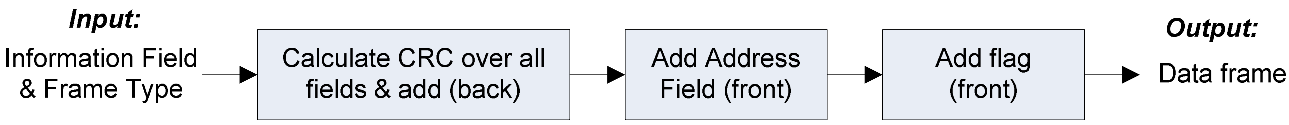


Figure 12.3: Primary data protocol operations

On-ground, the received data frames can be interpreted. The frame interpreter should perform the following steps:
1) Scan for a flag that indicates the start of a frame
2) Check whether a correct address field is placed next to the flag
3) Determine the next combination of flag and address field
4) Check whether the length of the frame is correct
5) Calculate the CRC as an extra check

If all checks are positive, the received data frame can be forwarded to the TM Database *as if it was received on the VHF/downlink*. It makes sense however to add a small marker to indicate the source.

For each of the FEC coding schemes, it remains to specify the:

- **Frame division**: The relation between the frame size and possible block size
- **Block boundary indication**: the indicator that signals the start of a coded block
- **Preamble amount and frequency**: The number of preamble bytes and how often they are added

These aspects are introduced for each of the three FEC coding schemes introduced in the previous section, in subsections 12.4.1 through 12.4.3 below. The data protocol steps that have to be taken to prepare the data for transmission, *in addition to those shown in Figure 12.3 above*, are then also stated.

Required preamble for MSK

In case of MSK modulation, carrier and symbol synchronization can be achieved rather simply by transmitting a series of 0's and 1's, interchangingly. This bit sequence creates a clear and consistent signal at the apparent centre frequency, and at the same time allows locking onto the symbol period. [SLR 0552] shows a dedicated circuit that when receiving this sequence can achieve both carrier and symbol synchronization. As such, it is suggested that *a preamble of alternating 0's and 1's is simply added to the start of the FEC-coded stream*.

The suggested modulator chip, the CC2500, in fact allows adding preamble in this exact form to a block of data. It also a suggestion to the amount required for normal operations, as described in [SLR 0556]. It mentions that for a Baud rate of 250 ksymbols, 4 bytes are required, but *for a Baud rate of 500 ksymbols, 8 bytes are suggested*, or 64 bits. This is further commented on below per coding scheme in subsections 12.4.1 through 12.4.3.

12.4.1 AO-40 FEC coding scheme

Frame division

The AO-40 FEC coding scheme requires 2048-bit sized blocks. The frame size has not yet been defined for Delfi-n3Xt. However, it seems unlikely that frames will fit seamlessly in AO-40 blocks. To illustrate this; the use of the final DelfiX protocol would yield a maximum frame content size of 1036 or 2148 bits for frame sizes of respectively 1200 and 2400 bits. A flag, address field and FCS should be added to the content. The 1036 bits would fit in a AO-40 block, but this would waste lots of space. *All other frame sizes of 2400 bits and higher, would simply not fit in AO-40 blocks*.

As explained before, frame sizes of 4800 bits are preferred. As such, *frame cutting will have to take place*.

In fact, this can be done rather transparently. Frames are received from the OBC, and the required flag, address and FCS fields are added. Assuming multiple frames to be queued, this *continuous stream of data* can simply be cut (per byte for example) to fit the 2048-bit blocks.

In this manner, the *data decoding layer and the data frame interpretation layer are completely separated functionally.*

It should be kept in mind that even when a block is decoded incorrectly, the RS codes of the AO-40 FEC coding scheme provide additional error correction. As such, incorrect AO-40 block are rejected, and do not end up in the decoded data stream. The 'frame length' check or the 'CRC' check as performed by the frame interpreter software will then reject the frame(s) to which this incorrectly decoded block belonged.

Frame boundary indication

A synchronization vector is added by the coding scheme for this exact purpose. This allows identifying the data sequence and synchronizing the blocks with respect to the decoder.

Preamble

Preamble can be added by the CC2500 chip or otherwise simply by the STX Processor. Operations will be noisier that the CC2500 default settings assume. Also, frames are not divided separately amongst blocks. As such, it is suggested to add not only 8 bytes, but *16 bytes to every block of 5200 encoded bits.*

The resulting data protocol operations that have to be performed on the data received the OBC, *in addition to those operations shown in Figure 12.3*, are illustrated in Figure 12.4 below.

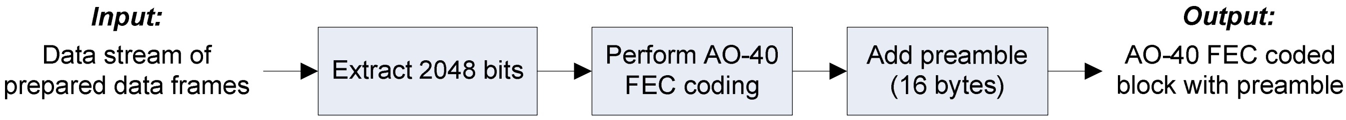


Figure 12.4: Secondary data protocol operations

12.4.2 Block coding scheme

Frame division

Similar to the AO-40 FEC coding scheme, a code block requires a certain bit size. This block size is limited to some 128 bits, of which only between 71 and 106 bits are information bits. Again, frames do logically not fit in coding blocks. However, since the block size is very small, it is not suggested to fully separate the frame sizes from the block size. It is suggested to:

1) Select a number of frames to be grouped
2) Add a combination of flag and address field to the end of the last data frame
3) Pad the bit sequence with 0's or 1's to make it divisible by the block information field size
4) Encode the bit sequence block-per-block

As this approach decouples the last frame from a subsequent frame, a *frame end marker* in the decoded data stream is required. It makes most sense to then repeat the combination of flag and address field at the end of the bit sequence as stated above.

The suggested approach gives a larger ***encoded frame package***; multiple encoded frames packaged together. The benefit of the suggested approach is that a *frame boundary indicator* and *preamble* are added only to complete frame sets. This makes sense as the reception of half a frame is useless.

As with the AO-40 FEC coding scheme, the block code itself provides additional error detection capabilities on-top of the error correction capabilities. This will provide a content check per coding block.

Frame boundary indication:

As explained before, the start indication should be error resistant. In case of the AO-40 scheme, an identification code is embedded in the interleaved code. In this case, another approach is suggested, taken from the design of the FX.25 data protocol [SLR 0294] as also introduced in chapter 7. This approach is to add a *correlation tag* to the encoded data stream.

The suggested correlation tag is a *Gold Code*, developed by Robert Gold. They are simple to generate and have favourable detection characteristics. A 64-bit size is suggested, providing good correlation at minimum bandwidth. A randomly taken Gold Code from [SLR 0294] which can be used is, in 64-bit hexadecimal notation:

- *0x41C246CB5DE62A7E*

Preamble

As again more coding gain is achieved with this coding scheme than with the standard CC2500 scheme, its indicated default preamble length is best increased. As such, it is suggested to again *add not only 8 bytes, but 16 bytes to every encoded frame package.*

The resulting data protocol operations that have to be performed on the data received from the OBC in addition to those shown in Figure 12.3 are illustrated in Figure 12.5 below.

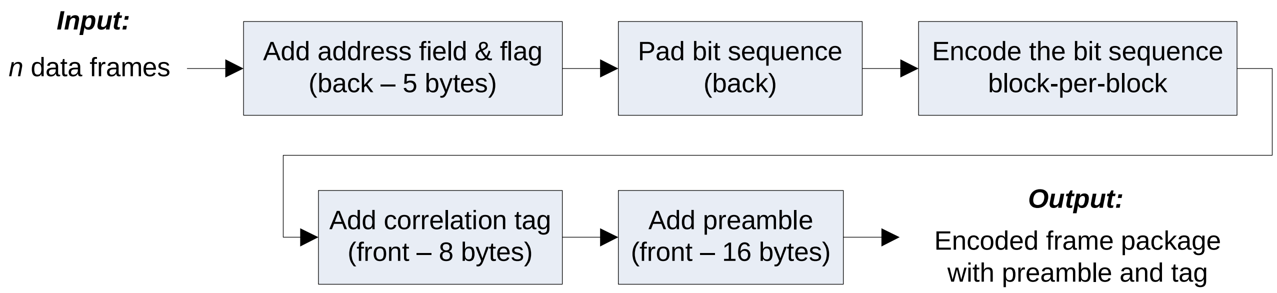


Figure 12.5: Secondary data protocol operations

The total number of frames per *encoded frame package* remains to be specified. As an illustration, the maximum incurred overhead by the above encoding process consists of:

- Addition of address field and flag: **5 bytes**
- Pad bit sequence: **16 bytes maximum**
- Correlation tag: **8 bytes**
- Preamble: **16 bytes**

This results in an additional overhead of at worst 45 bytes, or **360 bits**. If final frame sizes are indeed 4800 bits, yielding a frame content of some 4400 bits, some 8% extra overhead would be incurred. If two frames are joined together this overhead drops to 4%, and 2% for sets of n = 4. Then, n = 4 is assumed to be a good value.

Of course, in case of high data rates 2% overhead *per frame* can still be considered a lot if final data volume is considered. Therefore much higher values of n can also be chosen. The requirement of an uninterrupted data link during the transmission of an encoded frame package should then be taken into account properly.

12.4.3 CC2500 convolutional coding scheme

Frame division

A convolutional code works on a bit-per-bit basis. As such, an approach similar to that introduced in the previous section is suggested:

1) Select a number of frames to be grouped
2) Add a combination of flag and address field to the end of the last data frame
3) Encode the bit sequence bit-per-bit

Again, a *frame end marker* in the decoded data stream is required.

Preamble can be added to each *encoded frame package*.

Frame boundary indication

As a convolutional code works on a bit-per-bit basis, there is no block size defined. As such, it can simply *start decoding* (the small 4-bit coding memory is set to all 0's for a first bit). Therefore, no other frame boundary indication is required to signal the start of the coding block.

Preamble

The CC2500 itself suggests to *add 8 bytes to every encoded frame package.* Also, although not officially preamble, the CC2500 can apply *data whitening*, removing long strings of equal bits. This improves its own reception capability (see [SLR 0556]).

The resulting required additional data protocol operations in the STX can be summarized as shown in Figure 12.6. Possible data whitening is not included.

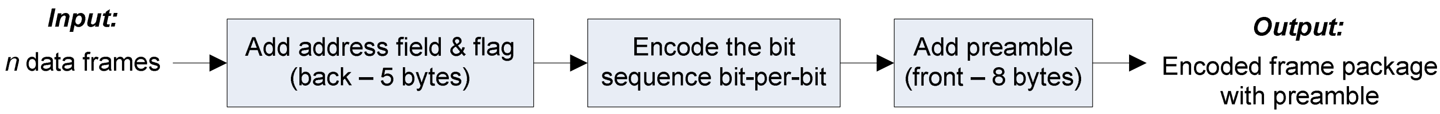


Figure 12.6: Secondary data protocol operations

Again, the total number of frames per encoded frame package should still be determined. In this case, the overhead incurred consists of:

- Addition of address field and flag: **5 bytes**
- Preamble: **8 bytes**

The incurred overhead is only 13 bytes, or **104 bits** per encoded frame package. Therefore single frames of 4400 bits can be grouped in an encoded frame package at the cost of only 2% overhead. Higher values of *n* can however again be chosen.

12.5 Data link layer: compress data (optionally)

An optional feature to implement on the S-band communication link is data compression. Data received from the CDHS can be compressed, before it is encoded or framed.

There are a number of good arguments for implementing data compression:

- Compression enlarges the effective data rate
- Data 'sits still' in storage at the STX processing core for a long time

The downsides of data compression however are that:

- Processing load of most importantly the compressor is increased
- Implementation complexity is increased, both at transmitter and receiver side
- Power consumption is increased
- Possibly more write cycles to the data storage are required

Implementation complexity indeed is increased, but relatively simple and well-known compression algorithms (such as ZIP) can be applied for which software exists and is easily available. The increase in power consumption is likely very minor; the MSP430 processor uses only 2 mW at maximum operation [SLR 0581]. The use of compression does also not automatically mean more write cycles to the data storage, requiring more power and using up the write cycles of the device.

The most important downside becomes the increased processor load, which will have an unknown impact. It is possibly that large periods of time exist during which the processor has excessive processing power available.

The effective data compression (compression ratio) achievable will be dependent on the compression algorithm and the TM data that is to be compressed. If a lot of similarities exist in the data (which consists of payload and housekeeping data) a high compression ratio can be reached. The similarities of course increase if larger time periods are grouped within one compressed file. Data compression therefore also requires dividing the received TM data in separate 'chunks' of multiple frames, not all too different from the application of FEC blocks as introduced in the previous section.

In conclusion, data compression can be a worthwhile addition to the STX. However, it will not increase the physical transmission capabilities of the STX. It is therefore not a primary requirement for the technology demonstration of the STX and the S-band communication link. *Data compression shall thus not be currently assumed for this link*.

Whenever software is developed however, and when the data budget has taken a more definite form, data compression can be reconsidered if its implementation complexity is minimal.

12.6 Data rate approach and determination

The main specifics of the S-band communication link have been specified in this chapter, except for the data rate. [SLR 0387] discusses the top-level design of the STX itself, and [SLR 0106] presents the link budget for the S-band communication link. Using the information on the link and the STX system design, the approach for the data rate can be established. This is done in subsection 12.6.1 below. Achievable data rates are then assessed in subsection 12.6.2.

12.6.1 Data rate approach

In section 7.9 theory has been presented that shows that the data rate should be changed between passes or even during a pass to realize maximum achievable download volumes. This is mainly due to the variation of path loss during a pass.

In case of the VHF downlink (PTRX) it was argued that (subsection 8.3.5):

- The system is not power limited
- Generation of a varying amount of data is difficult
- Scheduled link activation would be required
- On-ground systems should be prepared to receive a varying data rate

For the STX however it can be said that:

- The system is power limited
- The data storage creates a basically infinite pool of data and the CC250 chip allows setting data rate presets
- Scheduled link activation is already required (see chapter 14)
- On-ground systems should already be able to receive a varying bit rate.

As indicated by the above, the CC2500 can have a large number of internal data presets. Therefore it makes sense to integrate a large umber of presets, basically varying from minimum to maximum data rates. As the STX should be switched on by means of telecommand, the telecommand can at the time of activation specify the data rate, based on the link budget and satellite position. This allows to:

- Specify the data rate based on a calculation of optimum elevation angle, given the off-zenith angle
- Adjust the data rate for actual measured link (budget) performance, correcting for over-conservative or over-confident assumptions

Therefore it makes sense to, with the STX, aim for the following:

1) Establish a working link
2) Adjust link budget to actual performance
3) Perform experiments with changing data rates, between and during orbits

In this manner the STX will optimally yield an invaluable piece of technology for future missions, which can then be designed to operate exactly to provide maximized data volume downloads.

12.6.2 Link budget

The link budget is presented as [SLR 0106]. For the S-band communication link, in other words the STX downlink, it gives gains, losses and noise levels for a large number of elements in the chain. Using this link budget, achievable data rates can be determined, which is what will be done in this subsection.

A number of insecurities still exist for the STX and the S-band communication link, quite simple because:

- The STX has not been produced yet
- The S-band antenna system has not been produced and tested yet
- The S-band ground station is not in place yet

Although a large number of losses are based on assumptions in the link budget, the largest variables currently still are:

- **Ground antenna directivity**
  Optimally, a 3 m parabolic reflector will be installed on the Delft GS, with its structural and mechanical implications. The alternative would be a smaller one, with a minimum of some 0.7 m. The difference in directivity, corrected for pointing losses is huge, being some 10 dB.

- **Satellite antenna directivity pattern and assumed minimum directivity**
  The S-band antenna pattern will not be designed to rely on the pointing capability of Delfi-n3Xt. As such, an omnidirectional pattern will be designed for. The arguments and conclusions are presented in [SLR 0387] and [SLR 0036]. Two options are suggested:
    - A closed budget for all satellite attitudes, with a minimum gain value of -5 dB
    - A closed budget for ~80% of the time, with a minimum gain value of -2 dB

  Of course larger gain values can be chosen, resulting in less chance of having a proper satellite attitude; the maximum possible gain value is some 4 to 7 dB, depending on the final antenna patch design. Antenna pointing could also allow approaching these values.

- **Coding gain**
  Several options for coding schemes have been introduced above. Simple coding should surely be achievable (the CC2500 scheme or a single block code), but a concatenated coding scheme is preferred. Therefore, a value of 3 dB is assumed to be minimally achievable, but 7 dB is optimally achieved.

- **Orbit altitude**
  For Delfi-n3Xt the orbit altitude will be between 600 and 850 km, resulting in a combined path and atmosphere loss that varies some 3 dB.

Finally, as the data rate can and will be changed:

- **Pass characteristics**
  Depending on the off-zenith angle and possibly even the current elevation angle of the pass, the data rate can be adjusted.

The different variables can be tweaked to give reasonable data rates theoretically achievable with the STX and the S-band communication link. Two scenarios are presented below in Table 12.4 and Table 12.5 below:

- The STX is implemented with optimal coding gain, and ground antenna size. It does however employ a near-omnidirectional antenna system, according to the assumptions stated above. This is an ***optimistic scenario***.
- Both optimal coding gain and a large ground station antenna are not achieved. Due to this certain minimum satellite antenna directivity is required for the link budget to close reasonably; this value is taken to equal -2 dB. As a result, a larger range of satellite attitudes will not allow the link budget to close. This is a more ***pessimistic scenario***.

Table 12.4: Optimistic STX implementation, but omnidirectional pattern

Case description	Coding gain	7 dB		
	Satellite antenna directivity	-5 dB		
	Ground antenna directivity (ø = 3 m)	31.8 dB		
Orbit altitude	*h* = 600 km		*h* = 850 km	
Minimum elevation	10°	40°	10°	40°
Bit rate	93 kb/s	576 kb/s	57 kb/s	298 kb/s
Symbol rate	233 kBaud	1441 kBaud	143 kBaud	744 kBaud

Table 12.5: Pessimistic STX implementation, communication not for guaranteed for all satellite attitudes

Case description	Coding gain	3 dB		
	Satellite antenna directivity	-2 dB		
	Ground antenna directivity (ø = 0.7 m)	22.1 dB		
Orbit altitude	*h* = 600 km		*h* = 850 km	
Minimum elevation	10°	40°	10°	40°
Bit rate	4.0 kb/s	24.6 kb/s	5.1 kb/s	12.7 kb/s
Symbol rate	4.9 kBaud	29.5 kBaud	6.2 kBaud	15.3 kBaud

From the above figures it can be concluded that a very large insecurity in final achievable data rates still exists given the ~11 dB difference between the two scenarios. However, the main purpose of the STX is to *demonstrate the capability of the having an S-band transmitter*. In all cases above, the link budget allows for the STX to be at least tested sporadically. If low data rates can be accepted, the more pessimistic scenario would still allow sporadic testing of the STX in more than 50% of the time.

Consequently, if a proper pass takes place, data rates can be increased for the pass at hand, or during the pass. Two minimum elevation angles have been indicated in the above tables, which demonstrate that data rates can be increased by a factor of six for a 40° minimum elevation angle, in case of a 600 km orbit altitude. As explained in section 7.9, this requires increasing the data rate to compensate for the lower path loss at low slant ranges.

In the above tables the *coding rate* that is introduced by a certain coding gain can also be seen. The achievable bit rate is increased with a value of 1/0.40 and 1/0.83 for respectively Table 12.4 and Table 12.5. Now what is important to note is that *the transmission speed of the CC2500 is actually limited to 500 kBaud*. While in certain cases this would be a limiting factor, it would already be a sizeable accomplishment if these data rates can be reached momentarily.

Also, 26 kBaud is a minimum symbol rate if MSK is to be used on the MSK. If lower speeds are used, FSK should be applied, increasing the required signal-to-noise ratio on-ground by 3 dB. This therefore effectively means no data rates between 13 and 26 kBaud are supported, as 24 kBaud according to the above link budget would be equal to 12 kBaud on FSK. *Therefore, 26 kBaud should be aimed for as a minimum data rate*. In the pessimistic case, this would mean communication would only be possible for even a smaller percentage of the time, when satellite attitude is beneficial or at elevation angles even higher than 40°.

13 Integrating the COMMS on Delfi-n3Xt

All COMMS components, or CIs, are envisioned to be integrated on Printed Circuit Boards, or PCBs. These PCBs are stacked in Delfi-n3Xt, separated by certain distances. The exact structural rules are determined in the top-level STS document [SLR 0169]. The latter document also specifies the lay-out and exact areal space available on a PCB.

In case of the COMMS, *four* PCBs are reserved:

- The PTRX PCB
- The ITRX PCB
- The DAB
- The STX PCB

The DAB integrates the antennas for the UHF/VHF system. However, a number of cables for signal communication are also required. Finally, if the STX will be flown, it requires a separate antenna system, most likely in the form of two patch antennas, not integrated on a PCB but on the outer Delfi-n3Xt structure. Thus, in summary, also required are the [SLR 0387]:

- Communication cables (coax)
- S-band antenna system

All mentioned PCBs, as well as the main components of the Antenna System are shown in Figure 13.1 below. More specifically, four UHF/VHF antennas can be seen and a single Modular Antenna Box (MAB) is outlined. Two possible S-band antenna patch locations are indicated as well. The figure shows the bottom side of Delfi-n3Xt, per definition of [SLR 0169].

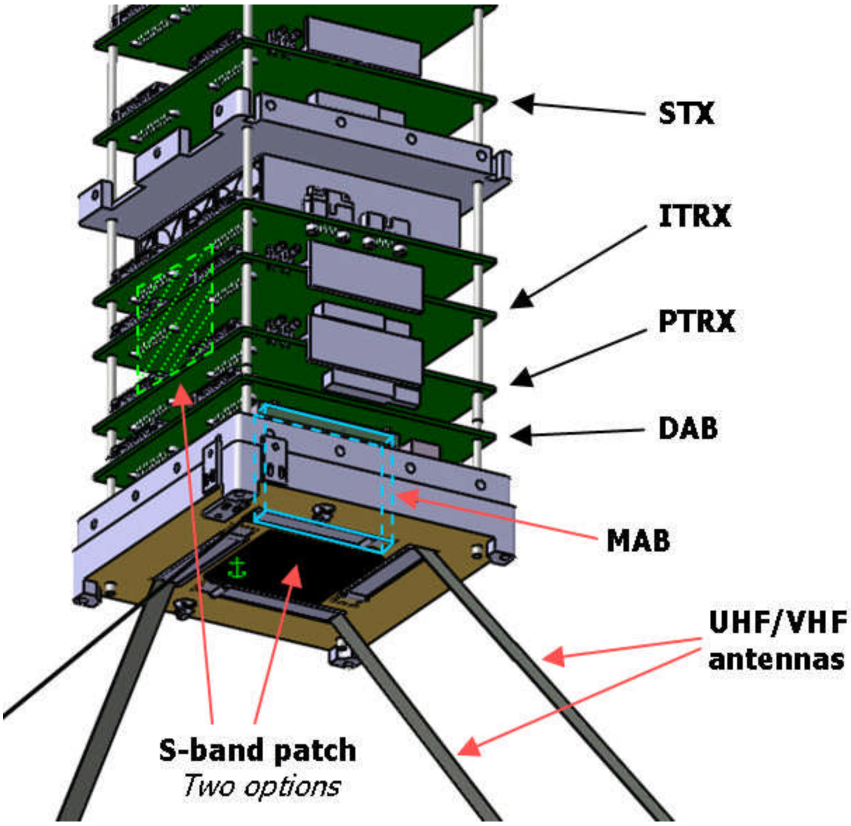


Figure 13.1: Structural integration of COMMS components [adapted from SLR 0169

In fact, the volume arrangement of Delfi-n3Xt still awaits to be changed after the removal of several payloads. *The above figure is just an illustration of the relevant COMMS components with their possible locations*.

The two integration options of the S-band patches are explained in [SLR 0387]. The options are to have two patches either at two opposite sides or at the top and bottom of Delfi-n3Xt. The BOP can be seen to have slits for the openings of the MABs. The MABs are placed on the DAB. Directly above this the primary radio is located, the PTRX. This arrangement allows for minimum cable losses for the latter. After the PTRX the ITRX can be placed, with similarly low cable losses. The STX is currently placed in the middle, which can be optimal if two antenna patches are placed on both the TOP and BOP; in that case cable lengths are equal to both patches. If the patch antennas are located at the sides, it is most beneficial if the STX is located at the same height as the patch antennas.

A number of general Delfi-n3Xt budgets are maintained. The three budgets of importance are the volume budget, the mass budget and the power budget. The data budget also exists on a satellite level, but will not be discussed here; the housekeeping data collected by the CDHS as well as the ultimate telemetry provided by the CDHS to the COMMS has already been commented on in this document. The three aforementioned budgets, volume, mass and power, are documented as [SLR 0303], [SLR 0018] and [SLR 0017] respectively.

The resulting values are summarized in Table 13.1 below.

Table 13.1: Mass, volume and power values for all COMMS components

Component	Mass	Volume	Power
PTRX	125.5 g	20 mm	240 mW (TX OFF mode) 1750 mW (All TX modes)
ITRX	132 g	20 mm	240 mW (TX OFF mode) 1750 mW (All TX modes)
STX	159 g	14 mm	110 mW (TX OFF mode) 799 mW (TM mode, 1 PA active) 1447 mW (TM mode, 2 PAs active)
DAB (incl. MABs, excl. OLFAR)	110.4 g	41.1 mm	N.A.
S-band patches	80 g	N.A.	N.A.
Connectors & cables	22 g	N.A	N.A.
OLFAR (incl. additional cable)	13 g	N.A.	49 mW

A couple of notes:

- **Mass** of the PTRX, ITRX and STX is in fact based on the RAP mass. The differences are due to included contingencies related to the development status.
- **Volume** envelopes were strictly limited before the MPS was removed. Currently volume restrictions can be loosened, if there would be reason for that. Only height in the PCB stack is taken into account for the volume budget.
- **Power** values for the PTRX and ITRX are also based on the RAP, with respective contingencies. The STX power values are as argued for in [SLR 0387]. Cables, connectors and antennas do not actively consume power. The required power consumption for deployment is not stated.
- *OLFAR* is indicated as a separate system. It should however be integrated on a PCB, logically the DAB if it will be flown. In that case, the weight and power required by only the additional components are shown above. This includes an extra cable, in the worst case being an RF cable.

14 Operations and performance

Given all the specific (sub)systems introduced in the previous chapters, this chapter aims to specify their implementation in the actual Delfi-n3Xt mission. That is to say, COMMS operations are defined and COMMS performance is analyzed. This chapter will make a division between the primary COMMS elements and the UHF/VHF communication link, and the S-band communication link and its STX.

The operations are in general specified by the different modes. In case of the UHF/VHF systems, these form the general *communication system modes*. These modes, their respective functions and their activation for the UHF/VHF systems are introduced in section 14.1 below. The operations of the STX are basically disconnected from other modes, as the STX itself is not strictly dependent on the operation of other COMMS systems. Therefore the operation of the STX is discussed in a later section, section 14.3.

The performance of the UHF/VHF transmission link will be discussed in terms of the data yield resulting on the downlink. For the uplink, TCs are transmitted with the only requirement that they are properly received. Whereas in case of Delfi-C^3 not all TCs are received, no clear data is kept on this percentage. Also, if TCs are only transmitted during passes, the theoretical reception should be 100%. For the downlink, per definition not all transmitted data will be received. VHF downlink performance is discussed in section 14.2. S-band downlink performance is discussed in section 14.4.

14.1 COMMS operations and modes

Primary COMMS operations are straight-forward and relatively simple. The UHF/VHF system is per default switched on, and set to transmit telemetry and receive telecommands. The system is not designed to require any orbital or attitude awareness, even if Delfi-n3Xt might have this, and thereby applies an omnidirectional antenna system. Transmitting TM and receiving TCs forms the default mode.

As two transceivers as well as systems such as the transponder and OLFAR (if possible) are to be flown on Delfi-n3Xt, a larger number of *communication system modes* is required. These are formed because of:

- The individual modes of the radios (the PTRX can transmit TM, function as a transponder *or* transmit the OLFAR signal)
- The power limitations, allowing not all systems to be on at the same time

As such, these global *COMMS modes* are combinations of PTRX and ITRX modes. The individual modes and the required actions have already been explained in the respective chapters on the PTRX and ITRX, chapters 9 and 10.

All modes are activated and deactivated by means of telecommand, except for the default mode. Upon reception of the TC, the OBS will then activate the respective PTRX and ITRS modes via the I^2C bus.

The resulting COMMS modes are listed in Table 14.1 below, listing also the respective PTRX and ITRX modes and the operational effects.

The possibly STX is not included in the table below; the operations of the STX are not linked to the other radios. It is to be decided via TC whether the STX will be activated, which in principle can be during any COMMS mode.

Table 14.1: COMMS modes and respective PTRX and ITRX modes

COMMS mode	Radio modes	Effects
PTRX mode (default)	• PTRX to "TM mode" • ITRX to "Receive only mode"	• TCs received by PTRX and ITRX • TM transmitted by PTRX • No transponder or OLFAR signal
ITRX mode (secondary)	• PTRX to "TX OFF mode" • ITRX to "DelfiX mode"	• TCs received by PTRX and ITRX • TM transmitted by <u>ITRX</u> • No transponder or OLFAR signal
Transponder mode	• PTRX to "Transponder mode" • ITRX to "Receive only mode"	• TCs received by PTRX and ITRX • <u>No TM transmitted</u> • No OLFAR signal • <u>Transponder activated</u>
ITRX test mode	• PTRX to "TX OFF mode" • ITRX to any mode other than "Receive only" and "DelfiX mode"	• TCs received by PTRX and ITRX • <u>No TM transmitted</u> • No transponder or OLFAR signal • <u>Various ITRX transmission</u>
OLFAR mode (experimental)	• PTRX to "OLFAR mode" • ITRX to "Receive only mode"	• TCs received by PTRX and ITRX • <u>No TM transmitted</u> • No transponder signal • <u>OLFAR activated</u>
All TX OFF	• PTRX to "TX OFF mode" • ITRX to "Receive only mode"	• TCs received by PTRX and ITRX • <u>No TM transmitted</u> • No transponder or OLFAR signal

14.2 VHF downlink performance

Any data received via the VHF transmission allows drawing conclusions on the success of the mission. If no data is received, nothing can be concluded on operations, and the mission is a failure. Whereas more data will not proportionally increase the value of the mission, more can in that case be concluded on the operations and payloads nonetheless. This section aims to give an idea of the expected data yield for the Delfi-n3Xt mission.

As the Delfi-n3Xt primary communication system is based on that of Delfi-C^3, in terms of its main radio, antenna system, ground station, and radio amateur involvement, a concrete example of performance exist: Delfi-C^3. A short description of the downlink data performance of the latter is given in subsection 14.2.1 below. Afterwards, the anticipated performance of Delfi-n3Xt is elaborated on in subsection 14.2.2, indicating the operational differences between Delfi-C^3 and Delfi-n3Xt, in other words the expected improvements.

14.2.1 Delfi-C^3 data yield performance

In case of Delf-C^3, the main communication system is very simple in operation. Once the satellite was first switched on, the satellite switched to Science Mode and data was generated, and transmitted to the ground. This data in Science Mode consists of a set of 5 repeating types of frames:

- **3 payload frames**
- **1 housekeeping frame**
- **1 flag frame**

The flag frame is not actually forwarded to the database as it serves to establish a good transmission link. Also, it was required internally (in Delfi-C^3) to process the other frames, and is thus in effect an empty frame. In total, the transmission of these 5 frames, of which 4 are actually interpreted on-ground, takes 5 seconds.

The satellite applies an omnidirectional antenna system, and has no orbit or attitude awareness. It simply transmits at every opportunity it has. As such, the only moments it does not transmit data is when:

- **A mode without data transmission is active**, such as Transponder mode. In Transponder mode no frames are transmitted, whereas in Basic Mode only housekeeping (HK) frames are transmitted. Only Science Mode and Basic Mode yield (digital) data transmission.
- **No power is available as the satellite is in eclipse**; in umbra or penumbra. Umbra means that the sun is completely blocked by the Earth with respect to the satellite, whereas in penumbra the sun is partially blocked and lower power is available. In fact, the period of penumbra is very small. A calculation of the time in umbra and penumbra is given below.

In case of Delfi-C^3, as will be the case for Delfi-n3Xt, radio amateur frequencies are used for data transmission, and at the same time radio amateurs were and are encouraged to receive data from the satellite and forward it to the TM Database using preconfigured software; RASCAL in case of Delfi-C^3. In fact, a total number of 339 radio amateurs worldwide have used RASCAL to subscribe, after which 184 have transmitted actual received data to the database, up until now.

The database is an SQL (Structured Query Language) database, and this section will use the data stored in the database to draw conclusions on the actual transmitted data performance of Delfi-C^3. Also, the theoretical contact times will be concluded upon using STK (Satellite Tool Kit), containing information on the orbital propagation of Delfi-C^3 over its first two years, from the *28th of April 2008 to the 27th of April 2010*.

Within the above period, STK can be used to give a more or less exact percentage of time the satellite was in direct sun, or in penumbra. The resulting values are:

- Percentage of time in direct sun: **66.27 %**
- Percentage of time in penumbra: **0.38 %**

As can be seen, the time in penumbra is small. The exact satellite operations are unclear in this period, as power generation might be insufficient. Therefore, penumbra shall be treated as umbra; no operations. Delfi-C^3 has no battery and switches off when no power is available. Upon exiting eclipse, the satellite boots and restarts transmitting data.

As in Science Mode a data frame is transmitted every second, except for the seconds a flag frame is transmitted, the amount of seconds the satellite has been in direct sun then gives a direct upper limit for the amount of data frames transmitted. Aspects that actually decrease this amount:

- The boot time required
- The time other modes have been activated
- The time the satellite software had crashed, or the RAP was switched off. In fact, in turns out Delfi-C^3 crashes at frequent intervals, and is not rebooted until an eclipse has passed. Similarly, the RAP can be automatically switched off even when a crash has not taken place. This causes a major and unpredictable loss of transmission time.

Still, the maximum amount of powered time will be used to calculate a total theoretical amount of data frames transmitted. Table 14.2 below compares this amount to different data resulting from the SQL database.

Table 14.2: Delfi-C^3 data yield from the 28th of April 2008 to the 27th of April 2010

Total data transmitted (theoretically)		**33,348,580 frames**
By type	Payload data	25,011,435 (75%)
	Housekeeping data	8,337,145 (25%)
Total *unique* data received		**618,091 frames (1.9% of theoretical total)**
By type	Payload data	409,148 (1.6% of total theoretical payload data)
	Housekeeping data	208,813 (2.5% of total theoretical HK data)
By receiver	Delft GS	190,757 (0.6% of theoretical total)
	Radio amateurs[1]	427,334 (1.3% of theoretical total)
Total *non-unique* data received		**940,966 frames (-)[2]**
By type	Payload data	641.137 (-)[3]
	Housekeeping data	305,316 (-)[3]
By receiver	Delft GS[4]	190,757 (0.6% of theoretical total)
	Radio amateurs	750,209 (-)

In the table above the percentages of the total theoretical values have been indicated. A number of notes have been indicated:

1. The amount given is actually the total amount of unique data received minus the unique data received by the Delft GS. If the Delft GS is not considered, more 'unique' data amongst radio amateurs can be said to be received. Nevertheless, in this manner the added value of radio amateurs is illustrated.
2. Non-unique data indicates a sheer number of frames, which might include double frames. Therefore, no percentage of a total can be given as the total would be infinite.
3. For reasons not understood, the sum of both payload and housekeeping data does not amount to the total, but differs slightly according to the database. The difference is however small.
4. This value is equal to the unique data received as the Delft GS cannot receive duplicate frames.

Conclusions:

- Almost **2%** of total theoretically generated data has been received over two years time.
- Radio amateurs have increased the received amount of unique data by **124%**, or when including non-unique data, even by **293%**.
- In both the unique and non-unique data cases, more housekeeping data has been received then ratio-wise expected compared to payload data, this should be due to the activation of Basic Mode.
- The reception of non-unique data has increased the total amount of received data by **52%**.

It is clear from the above data that the participation of radio amateurs has been a success, yielding a large increase of received data. What however is not instantly clear is how well the Delft GS performed, given the fact that it only has limited visibility. The resulting performance is shown in Table 14.3 below.

Table 14.3: Delfi-C^3 to Delft GS data yield from the 28^{th} of April 2008 to the 27^{th} of April 2010

GS range	**$\varepsilon = 0°$**	**$\varepsilon = 10°$**
Visibility	3.91 %	1.65 %
Total data transmitted (theoretically)	1,303,098 frames	549,969 frames
Received data percentage	15%	35%

Two different GS ranges have been indicated in the table above, specified by an elevation angle. This is due to the fact that a 10° has in fact been designed for in the link budget, but a large final link margin combined with little obstructions on the horizon should have effectively increased the field of view of the Delft GS. Therefore it results that between 15 and 35% of the total amount of packets that could theoretically have been transmitted by Delfi-C^3 *and* received by the Delft GS, has actually been received by the GS.

Of course, due to reasons stated above the actual transmitted amount of data is lower. Also, the received amount is lower due to:

- Tracking, pointing and operation errors. For example, the current rotator connection occasionally causes the antenna system to rotate by 360° mid-pass (see next chapter).

It is interesting to see how the reception of data by the Delft GS varies over time, and the result is surprising. For both cases of 1) data reception per day for a month, and 2) data generation per month for two years, no trend can be found in the results. To illustrate this point only the plot of the daily data reception during the first month of Delfi-C^3 operation is shown below in Figure 14.1. Also added are values representing the maximum theoretical amount of transmitted *and* received frames during these days resulting from daily coverage; an intermediate elevation angle of 5° is selected for the latter.

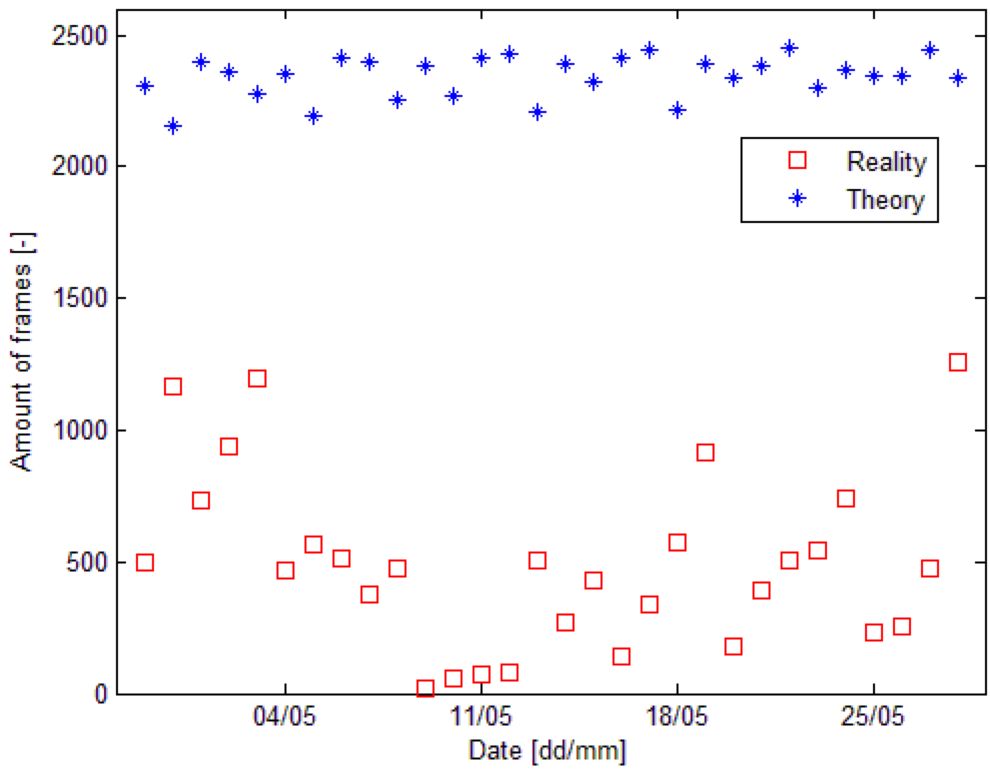


Figure 14.1: Delfi-C^3 theoretically transmitted and actually received data during May 2008

Daily coverage can be seen to be more or less constant. For a 5° elevation angle, STK shows that there are on average 5.9 passes per day, with all amounts being either 5 or 6 as expected. The resulting changes in theoretically transmitted data do in no way explain the variations in data reception shown in Figure 14.1. Delfi-C^3 operational stability is most likely the culprit. A more detailed analysis falls outside of the scope of this document. Also, its relevance to the Delfi-n3Xt COMMS is likely low as it more logically concerns Delfi-C^3 bus stability.

Nevertheless, one interesting aspect shall be shortly reviewed from the graph above. Three points, on the 3rd, 28th and 30th of May can be seen to have very high data reception. What is interesting here is that the database shows they have had 6 or 7 successful passes each. However, according to STK, only 6 passes can be seen within even a 5° elevation angle, or a 5° *artificial horizon*. This therefore is proof that the ground station system not only can make use of all daily passes, *but also passes that in some cases are below 5° elevation angle*. Indeed, STK confirms, 7 passes take place but only within a 1° artificial horizon. This is an important conclusion for assumed Delfi-n3Xt operations.

A further interesting side note is that the amount of passes changes per season, whereas it is more or less constant on a one-year basis. In May the daily average amount of passes is 1 higher for 0°, 5° and 10° elevation angle cases with respect to the yearly average. In December in some cases only 3 passes take place within a 0° artificial horizon! This however has everything to do with the eclipse differences during summer and winter, an aspect that should play no role in case of Delfi-n3Xt if batteries are properly integrated in operations.

A final aspect to be reviewed in slightly more detail is the participation of radio amateurs. The data yield following from the active participation of radio amateurs over two years time is plotted in Figure 14.2 below.

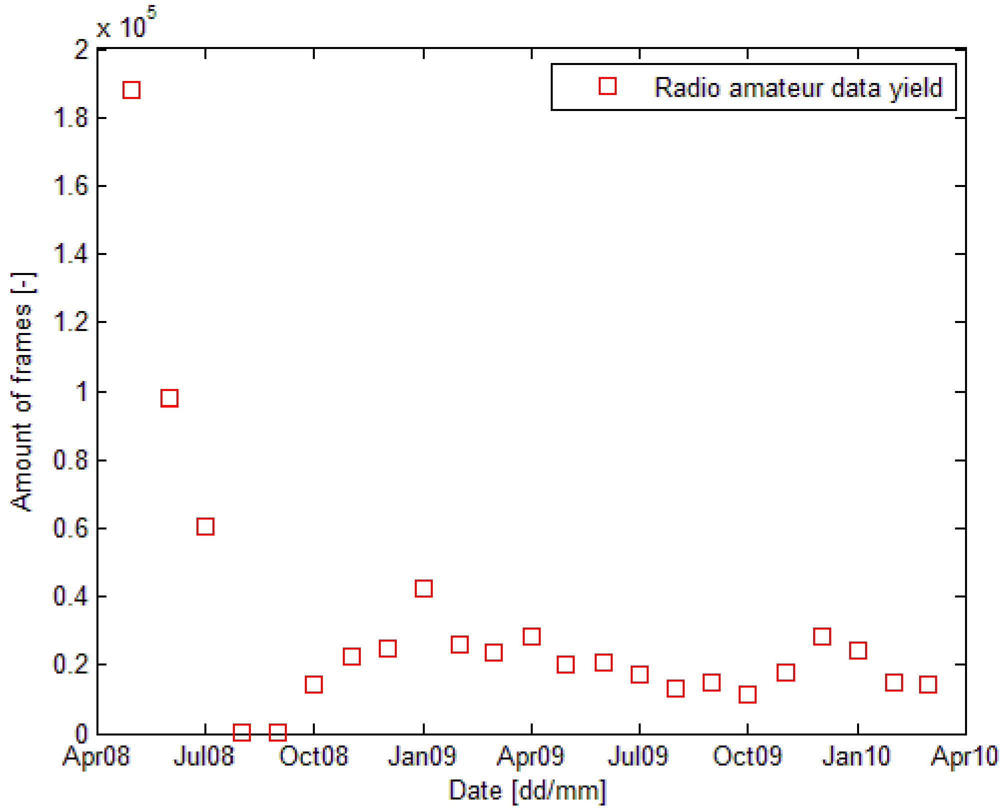


Figure 14.2: Delfi-C^3 data yield due to radio amateur participation over two years time

It can be seen that the radio amateur return appears to be more consistent than the Delft GS return, although admittedly the above plot is on a monthly scale and the values have been compressed towards the bottom more due to the inclusion of high values at the start. Nevertheless, the radio amateur yield does not seem correlated with the monthly values of the Delft GS yield (not shown), and indeed is more constant over time. The more interesting aspect however is the *overall* evolution over time.

At the start of the plot, three very high values can be seen, of respectively 187,738, 97,746 and 60,659 frames. In fact, this first month alone, resulted *in more data forwarded by radio amateurs than data collected by the Delft GS in over two years time*!

Directly after the first three months, two months of silence can be seen; in fact during these months Delfi-C^3 has been in Transponder mode almost constantly, generating nearly no data. This was a deserved return service to these radio amateurs. Finally in the graph, the data yield seems to average out with slight peaks and lows. The most interesting conclusion from this last period is that up until now, a solid handful still tracks Delfi-C^3 on a daily basis, most likely automated. Total daily radio amateur yield is still higher than Delft GS yield alone.

14.2.2 Delfi-n3Xt data yield performance

Delfi-n3Xt primary communication should be similar to that of Delfi-C^3. Data is produced in either of two radio modes, being the default PTRX mode or the secondary ITRX mode. In the other modes no data frames are transmitted, and thereby no telemetry can be received.

The last subsection on Delfi-C^3 performance yields a few valuable conclusions of interest to Delfi-n3Xt:

- **The Delft GS** indeed has shown to be able to make use of all passes within range and a visibility up to a 1° artificial horizon has been shown under certain conditions.
- **A battery** should be able to increase the total data yield by 50% over that achieved by Delfi-C^3, given the time the latter was in eclipse. The Delfi-n3Xt orbit will be one similar to that of Delfi-C^3, however likely with a different local time of the ascending node and orbit altitude; eclipse times will also be different. The inclusion of a battery can also remove the variability in data yield per season due to the varying eclipse times as well as a required boot time after every (pen)umbra. Of course, this assumes that the battery of Delfi-n3Xt will indeed be able to ensure continued operations even in eclipse. *This has not yet been confirmed after the redefinition of the mission.*
- **Radio amateurs** are very valuable, especially in the early mission when data can even be of most interest. Also, the primary mission is designed for three months, during which in case of Delfi-C^3 heavy support was present. A similar encouragement of radio amateurs should therefore certainly be pursued.

Finally, the stability problems of Delfi-C^3 related to the CDHS of the latter, most likely in the form of 'freezes' and consequent required resets, were mentioned; a more stable CDHS in Delfi-n3Xt should remove these problems.

In theory, this would mean that all geometrical passes within range of the Delft ground station should be able to be used for data reception, with the possible exception of the (minimal) time lost due to the tracking mechanism on-ground and the time when other modes have been activated.

Averaged out over a period of two years, this would give the coverage percentages as shown in Table 14.4 below as given for the Delfi-n3Xt range of orbits and for different assumed artificial horizons. The variables are expressed by orbit altitude h and minimum elevation ε respectively.

Table 14.4: Coverage percentages of the Delft GS anticipated for the Delfi-n3Xt mission

	h = 600 km	h = 850 km
ε = 0°	3.93 %	5.72 %
ε = 5°	2.53 %	3.80 %
ε = 10°	1.66 %	2.60 %

Finally, these percentages will be increased by an unknown amount due to the participation of radio amateurs. Using Delfi-C^3 as a measure, some 1-2 times the effective coverage is expected to be additionally supplied by the radio amateur network over the lifetime of Delfi-n3Xt, with a higher peak at the start of the mission.

14.3 STX operations

The operations of the STX will not depend on any satellite attitude pointing capability, just as the UHF and VHF systems do not. If possible, pointing of the ADCS could be used to exploit directivity peaks in the final S-band antenna gain pattern, but for general operations it will not be assumed.

The STX will also not be on continuously. In fact, it is unknown in what periods of time the STX will be activated, and in fact this also depends on its actual performance. As power is limited on Delfi-n3Xt, the STX shall be assumed to be switched on only *if the PTRX transmitter is switched off*. The PTRX TX:

- Is the largest consumer of power
- Provides functionality similar to that of the STX; transmitting

Depending on the performance of the STX, the STX can then be switched on more often. Another possibility is that the ITRX will have the high-efficiency amplifier it aims to have; currently some 60% efficiency is anticipated. This would reduce the power consumption of the ITRX by some 1 W with respect to the PTRX as currently anticipated, which could possibly allow *operating the ITRX and STX at the same time*.

As argued for in chapter 12, it makes sense to, with the STX, aim for the following goals in order:

1) Establish a working link
2) Adjust link budget to actual performance
3) Perform experiments with changing data rates, between and during orbits

Depending on the actual signal power of the received signal, the data rate of the STX can be adjusted between passes or even during pass. The latter would also allow adjusting for the attitude of the satellite during a pass.

In case of operations, the question becomes what operating modes are then required and how these modes are activated. This is discussed in subsections 14.3.1 and 14.3.2 below. Subsection 14.3.3 then introduces the option of selecting the data to be transmitted.

14.3.1 STX operational modes

STX operational modes are independent of other COMMS modes. [SLR 0387] introduces the operational modes in more detail, but the following operational modes can be distinguished:

- **TX OFF mode**

  This is the default mode. The STX is switched on, but its transmission is switched off. In this way it gathers and stores all TM data that is transmitted to PTRX and/or ITRX.
- **TM mode**

  When power is available and likely during a Delft pass with proper satellite attitude, this mode can be activated. Telemetry is transmitted after being read out from the data buffer.

In order to have different achievable data rates, *submodes with different data rate presets can be used*. All transmission submodes are part of the TM mode. Also, the design of the STX as presented in [SLR 0387] includes *two switchable power amplifiers*. In result, it can be decided to switch on just one PA, or both PAs.

In conclusion, within the TM mode the following can be configured:

- **The active PAs** (3 options; either one or both)
- **The data rate of the transmission**

The number of data rate presets should be based on the following:

- **A minimum and maximum data rate**
  The minimum is logically a few kBaud (applying a modulation scheme other than MSK, for minimum performance) and the maximum would be the CC2500 maximum; 500 kBaud.
- **A number of data rate steps in between**
  Basically, any number supported by the CC2500 and the telecommand structure can be used. This value can be as high as 256, specified by an 8-digit number and allowing for 2 kBaud data rate steps. Although this might be excessive, this suggestion illustrates the capabilities. In any case, due to the large insecurities in the link budget and the allowed number of presets, it does not seem useful to base the actual data rates on complex calculations based on pass probabilities and coverage values.

14.3.2 STX activation

The insecurity of the final STX design makes it unlikely that it will in fact allow all gathered mission data to be downloaded. More importantly, this is not the main goal of the technology demonstration. As such, it is more likely that the STX will function in a way similar to PTRX/ITRX; it will send down data using a best-effort approach. Therefore, no 'hand-shakes' or similar procedures are required to establish a proper connection before data transmission; transmission should just commence.

All non-default COMMS modes are activated by means of telecommand. Given the fact that the STX is more experimental, it is logical that also the STX is only switched on by means of telecommand. This leads to two options:

- **Switch on by telecommand**
  A(n) (automated) telecommand is send to the satellite as it is barely in sight to switch on the STX. As long as the STX is then switched on when the elevation angle designed for is approached, operations are nominal. An OBC timer would then most likely switch off the STX after its maximum possible pass time has passed, to ensure a proper switch-off. Of course this approach causes some transmission time to be lost in awaiting the TC reception and STX activation.
- **Schedule by telecommand**
  A more advanced approach would be to schedule the HDM mode to be activated a certain time after a telecommand has been sent. This would maximize the useful transmission time. *Multiple activations can also be scheduled following a number of consequent telecommands*.

The second option is the most preferable in order to maximize downloadable data volume, and should be allowed for by the OBC given the integrated RTC. Also, a simpler timer can be set. The second option is simply a more advanced version of the first.

As such, the option of *activating the STX by means of scheduling by telecommand is preferred*.

A final option, possible only if Delfi-n3Xt capabilities turn out to be (very) potent:

- **Automated switch-on by (simplified) orbit determination**
  Using TLEs and schedule data the STX can be activated during passes over Delft that satisfy specific criteria such as minimum off-track angle. In this case the system can be completely autonomous. This would require orbit determination capabilities on-board of Delfi-n3Xt, possibly as part of the ADCS operations. Also, this requires that the STX performs exactly as designed.

14.3.3 STX data preference

Again, as the STX will not likely be able to downlink all generated data of the mission, it could make sense to select data to have priority over other data. This could imply:

- Data that is generated over the oceans for example, as this information will not likely be successfully downlinked by the PTRX or ITRX.
- Early mission data, as this data can be more valuable.

The data storage size can be increased to more seriously support this functionality.

Preferred data could be selected by means of the frame counters that are included in the TM frames. Of course this would require some form of read-out intelligence for the STX in order to locate data in its storage that meets specific criteria. At the same time its FEC coding approach should be prepared for this. Furthermore extra intelligence in telecommanding and CDHS commanding would be required.

In conclusion, *this functionality shall be of low priority and not be assumed for the STX as long as the STX itself has uncertain performance*. The primary purpose of the STX is to demonstrate successful transmission capabilities, independent of the actual content transmitted.

14.4 S-band downlink performance

The performance of the STX is a hard aspect to discuss given the fact that it has no nominal operational modes. It is to be defined when the STX shall be activated, given the power capacity of Delfi-n3Xt as well as the final capabilities of the STX. What can however be done, is to illustrate the potential added value of the STX given the performance of the PTRX.

The STX is slated to receive the exact same TM data as the PTRX or ITRX. As it is not transmitting, the data is stored. During a pass, the STX can then be switched on. Given information of the pass, a matched data rate is selected.

Consequently, any data rate between a few kBaud and 500 kBaud can possibly be achieved. This can momentarily be **25 times** the *maximum possible* data rate of the PTRX. Any *second* the system might be on and functional, the STX can yield up to *a quarter of a minute* of PTRX transmitted data, when corrected for code rates required. For lower final PTRX data rates than 9.6 kb/s, the STX capability will double or quadruple with respect to the PTRX.

At that, the STX will logically yield data gathered *outside of the pass window*, due to its data storage capability. The overall data yield of the mission can therefore be greatly increased.

Table 14-5 below assesses the data rates required if *all* generated data would be transmitted by the STX, and a *constant data rate is applied throughout the mission*. A single ground station is assumed, that of Delft, and all passes are assumed to be monitored. The data volume has been calculated based on the use of the final DelfiX protocol, assuming maximum information field sizes. Also, it is assumed the PTRX data rate is fully utilized, with higher data rates allowing more data to be generated in the satellite. Frame sizes are assumed to be equal to the bit rate. On the other hand, data protocol overhead required on the S-band transmission link is ignored for the moment.

In Table 14-5, maximum achievable data rates are indicated, based on the optimistic link budget case as discussed in chapter 12. Orbit altitudes h and minimum elevation angles ε_{min} are indicated, which together result in a daily coverage in minutes. The coverage is calculated again using STK, for sun-synchronous orbits over a two-year span.

Table 14-5: STX required bit rates for a complete data dump, using a constant data rate

	Required constant data rate			Maximum achievable data rate
PTRX data rate	1.2 kb/s	2.4 kb/s	9.6 kb/s	-
Daily data volume	90,029 kb	186,278 kb	763,776 kb	-
All passes, ε_{min}=10°,h=850 km → *48 min*	**31 kb/s**	62 kb/s	249 kb/s	*57 kb/s*
All passes, ε_{min}=10°,h=600 km → *31 min*	**49 kb/s**	98 kb/s	391 kb/s	*93 kb/s*
All passes, ε_{min}=40°,h=850 km → *7 min*	223 kb/s	445 kb/s	1.8 Mb/s	*298 kb/s*
All passes, ε_{min}=40°,h=600 km → *4 min*	401 kb/s	802 kb/s	3.2 Mb/s	*576 kb/s*

In the above table, the bold values are lower than the maximum achievable data rate, and would yield symbol rates below 500 kBaud given the 0.4 code rate required. These scenarios would allow all generated data to be transmitted.

If *constant mission data rates are applied,* it can be concluded from the table above that the chances that the STX will be able to transmit *all* mission data are not very large even if the entire pass is monitored, assuming the PTRX data rate and the Delfi-n3Xt data budget will indeed allow for a 2400 kb/s unique data rate, or higher.

As section 7.9 has presented however, a data rate that is changed between passes allows almost **two times** as much data volume to be downloaded for close passes, when compared to the constant data rate case with ε_{min} = 10°. *This would likely allow all data to be downloaded for the case where the PTRX has a data rate of 2.4 kb/s*.

Finally, being able to change the data rate continuously even during a pass, would allow to greatly improve the link budget as the satellite attitude can be taken into account; currently some -5 dB gain is assumed but maximum gain values up to 4 dB are possible. Even more so, optimally changing the data rate continuously could allow downloaded data volume to be increased by almost **five times** with respect to the constant mission data rate case. In this case, the maximum transmit data rate of 500 kBaud would certainly become a limiting factor, but this would likely allow *all generated data of Delfi-n3Xt to be downloaded by the STX*.

This is of course, assuming the optimistic implementation of the STX and its link budget. At the same time this also assumes only one receiving ground station, that of Delft.

15 Ground segment

The ground segment is in fact not part of the COMMS. Better said, the Delfi-n3Xt mission in terms of (CI) components is divided in the *Launch & Orbit Segment*, the *Delfi-n3Xt Satellite or Space Segment*, and the ***Ground Segment***. Part of the Ground Segment deals with satellite communication; it is that part that will be commented on below. The entire Ground Segment is topic of the top-level Ground Segment document, [SLR 0657]. Figure 15.1 below indicates the CIs of importance to COMMS and COMMS functionalities, with proper regard for its context.

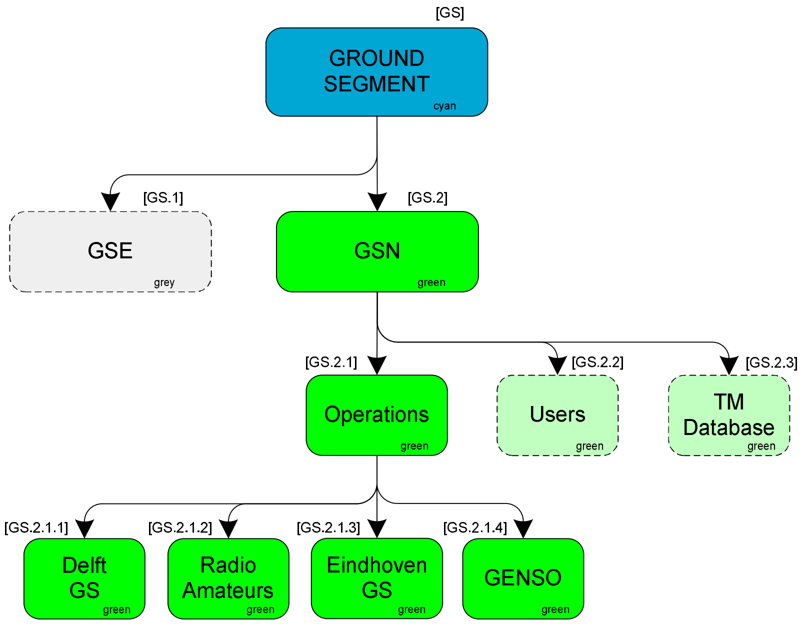


Figure 15.1: Ground segment CIs related to COMMS.

It can be seen that the ground segment consist of two branches, GSE (Ground Support Equipment) and GSN (Ground Station Network). Within the GSN, there is a segment Operations, Users, and TM Database. It is only the first branch that has a (communication) link to the satellite.

The Delft GS is the main ground station. It is to be used for Delfi-n3Xt data reception and for data transmission towards Delfi-n3Xt. More specifically it should receive TM and should be able to transmit TCs. Furthermore, as Delfi-n3Xt follows in the footsteps of Delfi-C^3, radio amateur transmission frequencies are used, and radio amateurs are to be stimulated to receive TM as well. Using a custom designed client, DUDe, this data can be received, interpreted and forwarded to a central TM Database. This database can be seen in the CI tree above as well. The two options above form the main ground segment communication CIs.

Furthermore two other lower level CIs are shown, being the Eindhoven GS and GENSO. The Eindhoven GS has been used in the Delfi-C^3 mission to function as a back-up GS, although never required and GENSO is an ESA initiative that envisions goals similar to using the radio amateur network for increased coverage.

The four operational GSN CIs are discussed in sections 15.1 through 15.4 below.

15.1 Delft Ground Station

The Delft GS is the primary communication point on-ground. Therefore it is the most important part of the ground segment, and its equipment can be used as the baseline for other ground station receivers or possibly transmitters. The data reception software, DUDe, is designed to also function as essential software for radio amateurs.

The Delft GS is fully redundant, which means that all components are present twice. This holds for the antennas plus antenna steering hard- and software, as well as all transceiver hard- and software. Two PC terminals are in place. Also, any failure can be fixed relatively quickly as the GS is on-ground and uses mostly COTS components. Therefore, a back-up GS such as the Eindhoven GS is not strictly required.

The Delft GS is located at the top floor of the faculty of EEMCS (Electrical Engineering, Mathematics and Computer Science) on the campus of Delft University of Technology. Therefore the following specifics can be given (see Table 15.1).

Table 15.1: Delft Command Ground Station specifications.

Position:	51°59’55” N, 4°22’25”
Elevation:	90 m

All requirements of the Ground Segment are logically specified in the top-level document of the latter, but a number of requirements are incurred by the COMMS. Therefore, subsection 15.1.1 presents these requirements. Subsection 15.1.2 specifies the ground station and its main components required for transmission and reception. What is labelled as supporting equipment, being similarly required but less obviously, is presented in subsection 15.1.3. Finally possible upgrades to the current equipment are discussed in subsection 15.1.4.

15.1.1 Requirements

The following Table 15-2 presents all requirements imposed on the Delft GS by the COMMS of Delfi-n3Xt.

Table 15-2: Delft GS requirements imposed by the COMMS

Category	Requirement #	Requirement
FUNCTIONAL	GS.2.2.1-F.01	**The Delft GS shall be able to transmit telecommands over the VHF band** ***Parent:*** *SAT-F.01* ***Rationale:*** *This requirement follows logically from the parent requirement of being able to communicate with the satellite.*
FUNCTIONAL	GS.2.2.1-F.02	**The Delft GS shall be able to receive telemetry over the UHF band** ***Parent:*** *SAT-F.01* ***Rationale:*** *This requirement follows logically from the parent requirement of being able to communicate with the satellite.*
FUNCTIONAL	GS.2.2.1-F.03	**The Delft GS shall be able to receive a Morse Beacon over the VHF band** ***Rationale:*** *This requirement follows from the beacon signal that is transmitted by the PTRX when in Transponder Mode; correct reception of the latter allows verification of the activated mode.*
FUNCTIONAL	GS.2.2.1-F.04	**The Delft GS shall be able to receive data over the S-band** ***Rationale:*** *This requirement is necessary for the STX technology demonstration.*
INTERFACE	GS.2.2.1-F.05	**The Delft GS shall be able to forward received telemetry to the TM Database** ***Rationale:*** *This requirement is actually not imposed by the COMMS, but included for completeness. The Delft GS should forward TM to the TM Database.*
INTERFACE	GS.2.2.1-F.06	**The Delft GS may be able to receive the OLFAR signal over the VHF band** ***Rationale:*** *To test the correct functioning of OLFAR, the signal should be received on-ground. If the data (entire bandwidth) is also to be analyzed, more complex hardware is necessary.*
CONSTRAINT	GS.2.2.1-C.01	**The Delft GS shall be able to receive signals in the VHF range of 144-147 MHz** ***Rationale:*** *These values follow from the transmission frequencies, but are adjusted for Doppler shift and possible oscillator drift.*
CONSTRAINT	GS.2.2.1-C.02	**The Delft GS shall be able to transmit signals in the UHF range of 435-438 MHz** ***Rationale:*** *These values follow from the transmission frequencies, but are adjusted for Doppler shift and possible oscillator drift.*
CONSTRAINT	GS.2.2.1-C.03	**The Delft GS may be able to receive signals in the S-band range of 2400-2450 GHz** ***Rationale:*** *These values are still vague as the exact transmission frequencies within the S-band are not yet specified. As such, the entire ISM S-band is indicated.*

The GS currently does not meet all requirements as it misses:

- UHF/VHF equipment in the case the data rates used for reception or transmission are increased
- The UHF/VHF reception software, as DUDe is being created as an update to RASCAL
- All S-band equipment

The transponder is not actively supported at the Delft GS. The transceiver can currently tune in to the transponder frequencies, and can then receive a transponder signal, but only with a limited bandwidth; not the entire 40 kHz bandwidth but only 2.4 kHz. Depending on the use of the transponder channel, being for example voice transmission or the transmission of different data, different equipment can be required. None of this is required for the Delfi-n3Xt mission so this will not be further commented on. Also, if higher downlink data rates are supported, larger bandwidth reception will be supported.

The transponder beacon can be received as is commented on in the next section.

15.1.2 Delft ground station main equipment

A large amount of equipment is required within the Delft GS in order to facilitate correct communication with Delfi-n3Xt. The main functionalities and equipment required have been introduced in chapters 7, 8 and 12 on transmission techniques and the Delfi-n3Xt communication links. The resulting equipment is introduced in this subsection. As in fact even more equipment is required, the next subsection introduces equipment labelled as *'supporting equipment'*. Currently, no S-band communication is supported so this is covered in section 15.1.4 on ground station upgrades.

Thus, this subsection introduces all components directly involved with transmission and reception of data, each having clear and distinct data processing functionalities. The resulting set-up is presented visually in Figure 15.2 below.

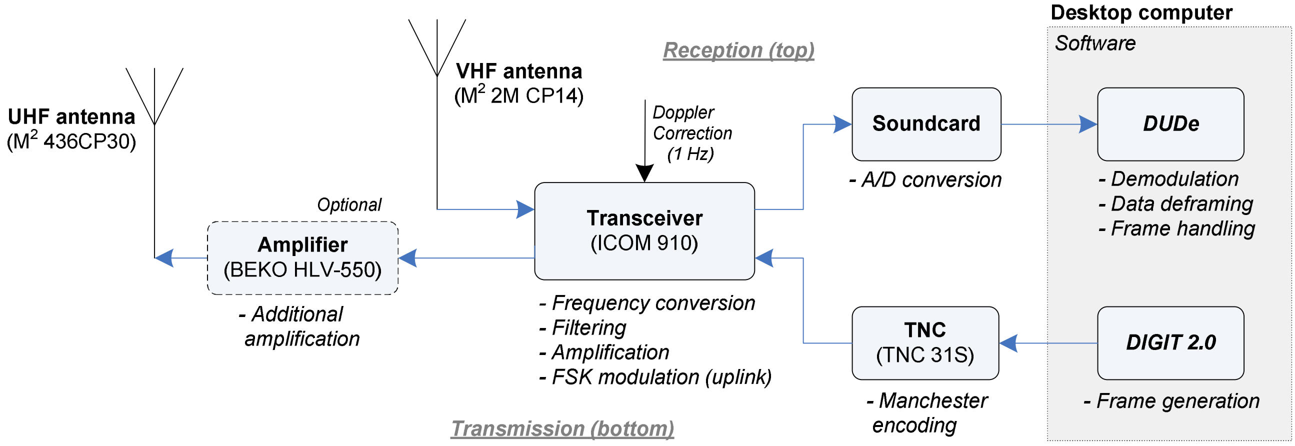


Figure 15.2: The Delft GS main equipment.

The hardware components present in the Delft GS are all indicated, including the functionalities relevant for data transmission and reception. The functionalities required for (de)modulation on the Delft ground station have been commented on in chapter 8 on the UHF/VHF transmission link. All indicated components are shortly commented on below, first introducing the transceiver and then going through the downlink and uplink chain respectively:

- **Transceiver** (ICOM 910)
  This transceiver is used for the uplink and downlink. It:
    - Applies filtering to limit the bandwidth, in downlink and uplink
    - Converts the frequency to IF in case of the downlink (to 1600 Hz)
    - Applies the FSK modulation in case of the uplink using the received 1200 Baud signal
    - Applies a certain amount of amplification, in downlink and uplink (up to 75W)

  The ICOM 910 is designed for use in the UHF and VHF bands, and the exact frequencies can be set on the machine. Also, the transceiver receives a Doppler correction every second, which originates from the orbit prediction software installed (combination of NOVA and an application called Doppler Determination Program). It uses this value to correct the transmission frequencies for the anticipated Doppler shift. This also has an impact on the on-board demodulator, as the frequency shifts abruptly every second. Doppler shift rates are limited to some 100 Hz per second, which translates to a minor effect to say the least; 100 Hz per 145 MHz, with 1200 symbols per second, results in a changed duration per symbol of not even 1/1000th of its original duration.

The transceiver automatically allows for reception of the transponder beacon; it is transmitted over the TM downlink frequency. As the transceiver incorporates a speaker, the Morse signal can simply be heard. A small device (which is present) can also be used to decode the consequent Morse signal.

Downlink

- **VHF antenna** (M^2 2M CP14)
  This antenna is used for signal reception, and is designed for use with the frequencies in the range 143-148 MHz. It is circularly polarized. It can be configured to be either right-handed or left-handed circularly polarized, through a switch incorporated in the antenna. A simple cable can be connected to this switch to change the polarity by means of a short 12V signal.
- **Soundcard**
  The 1600 Hz signal originating from the ICOM 910 is sampled with 38.4 kHz with 32-bit precision. These values are send to software, DUDe in case of Delfi-n3Xt, being a successor to RASCAL.
- **DUDe**
  This program basically fulfils four functions, indicated per *functionality layer*:
  - Physical layer: demodulation of the digitalized waveform
  - Data link layer: the data frames are checked and interpreted
  - Presentation layer: the content of the data frame is interpreted and visualized for the user
  - Network layer: the frames are forwarded to the TM Database

Uplink

- **DIGIT 2.0**
  This program should generate telecommands. DIGIT (1.0) was used for Delfi-C^3, but as telecommands will logically be different for Delfi-n3Xt, this program should be updated.
- **TNC** (TNC 31S)
  The TNC is mainly in charge of applying Manchester coding, and ensuring proper timing of the resulting data stream. As a desktop computer cannot guarantee a constant data stream, due to processing delays and queues, an external device is used. Combining this with the applied Manchester coding, the TNC delivers a perfect 1200 Baud signal. In the current situation, the TNC receives telecommands from DIGIT in the KISS protocol; it then also applies the AX.25 protocol to the TCs. As the DelfiX protocol is preferred, an update is similarly preferred. This is discussed below.
- **Amplifier** (SP7000PA)
  This extra amplifier is optional, and is in principle not required for correct transmission towards the satellite. Nevertheless, in order to be safe it is generally used. It can reach output power values of 500 W, but is limited to 250 W.
- **UHF antenna** (M^2 436CP30)
  This UHF antenna is used for signal transmission, and is designed for use with the frequencies in the range 432-440 MHz. It is circularly polarized, and can be switched between right-handed and left-handed polarization similar to the VHF antenna.

15.1.3 Delft ground station supporting equipment

The supporting equipment can roughly be divided in two functional categories:

- *Signal transmission*
- *Support*

The components dealing with *signal transmission* are not very important from a functional point-of-view in terms of transmission techniques, but they are very important for the link budget for example. *Support* equipment is introduced in order to increase understanding of the communication system, and to illustrate what other components are actually in place on the Delft ground station.

The two categories of supporting equipment are demonstrated in a functional chain shown as Figure 15.3 below. Afterwards again the components are discussed one-by-one, except for the antennas which have been introduced above and are actually not to be seen as supporting equipment but are included for clarity.

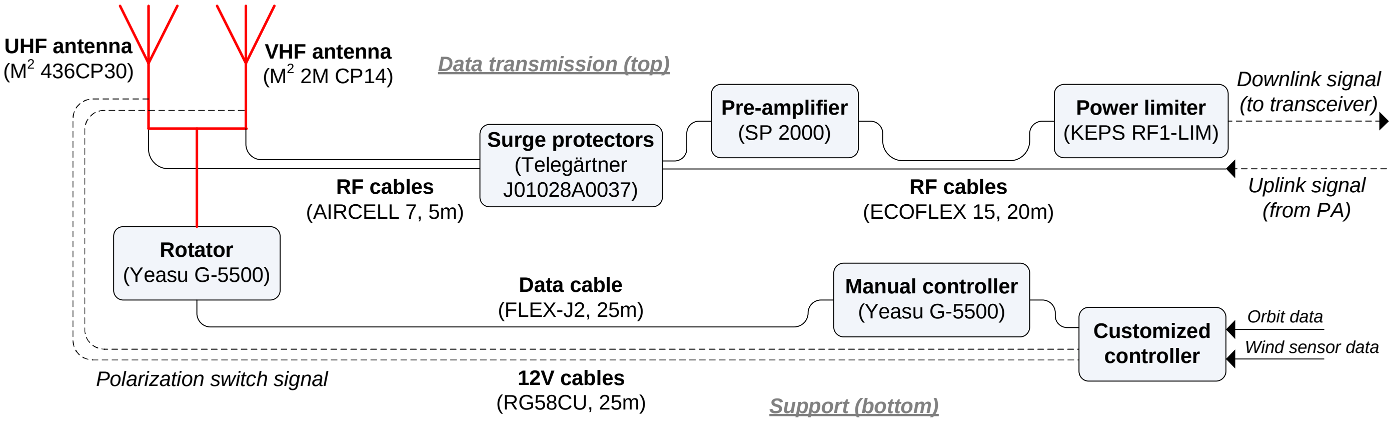


Figure 15.3: The Delft GS supporting equipment

Data transmission equipment

- **RF cables** (AIRCELL 7, 5 m)
  These cables form the connection between the UHF and VHF antennas and the surge protectors. They are relatively short, and are more light-weight and flexible than extremely low-loss cables. These cables should rotate along with the antenna.
- **Surge protectors** (Telegärtner J01028A0037)
  These components are in place for *lightning protection*. These surge protectors in fact function as passband filters. Any alternating voltage of an out-of-band frequency (such as those resulting from a lightning strike) is simply short-circuited, meaning discharged to the ground. In order to do so, these surge protectors are integrated in a box with a ground connection.
- **Pre-amplifier** (SP 2000)
  In case of the downlink only, a pre-amplifier is inserted at this point. The goal of this is to effectively increase the signal strength, before more losses are incurred in the longer distance between the antenna system and the actual GS and receiver. This is taken along in the link budget and decreases overall added noise.
- **RF cables** (ECOFLEX 15, 20 m)
  These cables transport the signals over the significant distance between the roof of EEMCS and the actual GS on the 22nd floor. As losses should be minimized and can otherwise be significant, expensive and heavy, well-protected cables are used with minimum losses.

- **Power limiter** (KEPS RF1-LIM)
  This power limiter serves to protect the downlink receiver equipment from the immense power of the uplink transmitter. While it is not transmitted in the same frequency band, the signal strength scale is vastly different and can 'leak' into the receiver equipment via the antennas.

Support equipment

- **Rotator** (Yaesu G-5500)
  This actuator takes care of the pointing of both antennas towards the satellite. It can rotate through an entire half sphere, with an extra margin in azimuth direction for a total range of 450°.
- **Manual controller** (Yaesu G-5500)
  The Yaesu G-5500 actually comprises of a rotator, and manual controller. The manual controller quite simply allows changing either elevation or azimuth direction. It however has an external control input.
- **Customized controller**
  A customized box includes an automated controller to control the manual controller, based on an ARS-rotor interface card, where ARS stands for antenna rotor system. This card then forms the interface between the computer and its software (ARSWIN) and the manual controller. This software in turn receives orbit data from orbit calculation software and wind sensor data also gathered on the roof. This data (software) is shortly commented on below. The customized controller also has a manual switch to change the antenna polarization from right-handed to left-handed circular polarization and vice versa. This functionality has been commented on above.
- **Data cable** (FLEX-J2, 25 m)
  This cable quite simply relays the controller data to the rotator. Therefore any cable specifics are not relevant.
- **12V cable** (RG58CU, 25 m)
  Again this cable suite simply relays signals, this time from the controller to the antennas, with the purpose of changing the antenna polarization. Cable specifics are again not relevant.

One section of the GS that has not been covered in detail is the equipment and software dealing with orbit data generation, wind sensor data generation, and Doppler shift data calculation. In fact, TLEs collected from the NORAD database are processed in software by means of NOVA/Orbitron, and ARSWIN consequently creates commands that are directly fed to the controller. The first set of programs also yields the Doppler data send to the transceiver by means of a simple data interface. A separate wind sensor on the roof is read out by software called DARCA, and this generates wind sensor data, again forwarded to ARSWIN. All programs are installed on the same PC that houses DUDe, DIGIT, and a sampling soundcard.

These programs and equipment are of less relevance to the COMMS, and therefore not further introduced. The design has been described in [SLR 0393] and it is also shortly addressed in [SLR 0657] along with the top-level design of the ground segment for Delfi-n3Xt.

15.1.4 Upgrades

A number of upgrades are required or possibly required on the Delft GS. These are listed below. First the changes to the current UHF/VHF link are stated, after which the proposed S-band set-up is presented.

For the UHF/VHF link, the first three suggestions are in fact envisioned to be executed to support the design choices made in this document. The possibility of increasing the data rate for the uplink should be assessed, and finally two suggestions are made with respect to the inclusion of error correction on the downlink and the rotator performance.

Required in any case

- **DIGIT 2.0**
  DIGIT should be upgraded for Delfi-n3Xt, resulting in a tentatively called DIGIT 2.0.

Implementing the DelfiX protocol

Currently, DIGIT applies KISS frames and the TNC changes the data protocol to AX.25. It makes sense to move the entire data protocol functionality to DIGIT in PC software. In that case:

- **DIGIT 2.0**
  DIGIT should be configured to support the DelfiX protocol.
- **TNC 31S**
  The TNC software should be updated to only apply Manchester coding.

Implementing higher bit rates on the downlink

- **Transceiver**
  The ICOM 910 transceiver only allows downlink BPSK data rates of up to 1200 b/s. As such, the transceiver should be upgraded, most likely to the solution offered by ISIS. A software based radio is being developed, for which a customer is already found. As such, it is a safe assumption that this transceiver will be available at the launch of Delfi-n3Xt with sufficient margin. The specifics are currently unknown. It will however provide a sampled output signal, thereby removing the need for a soundcard. It will support sampling rates of 4 MHz, widely sufficient for any VHF downlink data rate anticipated or even allowed by radio amateur bands.
- **DUDe**
  The demodulation software should undergo a rather simple update as input frequencies change with changed data rate. As the DUDe user interface is currently already prepared to support higher data rates, the actual changes at a later point are minor.

Implementing higher bit rates on the uplink

The transceiver currently used, the ICOM 910 supports data rates of up 9600 b/s. As such there is only one component that requires an upgrade in case of increased uplink data rates:

- **TNC 31S**
  The TNC integrated modem should be replaced, as it simply only supports 600 b/s. Again ISIS offers an update of the integrated modem, while reusing the same TNC. This update can yield up to 115 kb/s output data rate, again more than allowed by radio amateur bands. This upgrade can be performed relatively easily.

Performing single-bit error correction on the downlink

It has been explained in section 8.5 that (single-bit) error correction can be performed using the CRC used in the DelfiX protocol. The application of error correction would affect only:

- **DUDe**
  An extra software feature should be written that does not discard erroneous frames, but attempts to correct them. Even though this (table-look-up) algorithm can be quite computationally intensive, the time between passes would easily allow for the process to be performed. This would mean erroneous frames would simply be queued for later correction during the pass. The corrected frames can then be labelled (in order to monitor the usefulness and correctness) and send to the TM database. The table-look-up algorithm required has been established and is discussed in several papers (see section 7.7).

Improve pointing performance by correcting rotator orientation

Currently, the combination of software applied for tracking and rotator default orientation causes problems to occur approximately mid-pass for some passes. The reason for this is the (limited) rotation range of the rotator of 450° around its vertical axis. As the software used does not take into account this range before starting pointing, some passes will yield the rotator to reach its maximum value, requiring a 360° 'back-spin'. The software used is closed-source and cannot be easily updated. Therefore the easiest update would be to the:

- **Rotator connection orientation**
  The rotator has basically been oriented randomly when connecting it to the roof. If instead, its point of minimum rotation is placed close to the orientation required for the satellite pass to be followed, a 'back-spin' would never be required.

Receiving an S-band signal

Currently, the ground station has no equipment to receive an S-band signal, with the exception of a small 0.7 m parabolic reflector. ISIS offers a full S-band ground station, which can be adapted to specific needs. Due to the special relation of ISIS they would gladly install the entire system, of course assuming the TU Delft pays for the equipment. Figure 15.4 below shows the global components that should at least be incorporated.

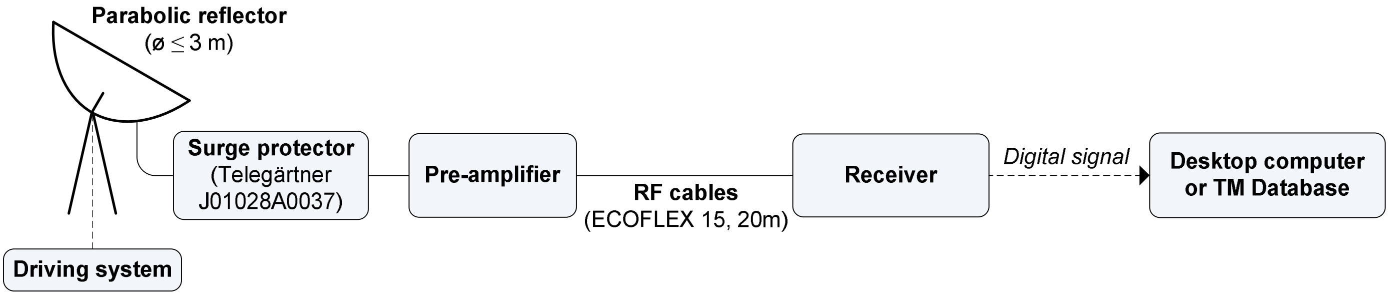


Figure 15.4: The proposed minimum S-band ground station set-up

The minimum required components are:

- **Parabolic reflector**
  This type of antenna is generally used for the S-band due to the high achievable directivity values. ISIS seems to be confident to be able to provide a 3 m dish, which consequently should be able to provide 0.5° pointing accuracy; not taking into account further wind-induced errors and orbit inaccuracies. Of course, the structural impact of such a dish is very large so it should be verified whether this is even achievable. Also, it should be taken into account that heavy winds might cause the antenna to be temporarily out of operation.
- **Surge protector**
  Between the essential pre-amplifier and the parabolic reflector some form of lightning protection is necessary. As the component used for the UHF/VHF bands can also be used for the S-band, it is suggested here.
- **Pre-amplifier**
  As with the VHF downlink, a pre-amplifier is required. On the S-band it is however even more essential, and should have even higher gain to account for the cable losses incurred afterwards. Also, the pre-amplifier should be located more or less directly on the dish.
- **RF cables**
  These cables carry the signal to the actual receiver, and a similar distance and cable type is assumed as for the UHF/VHF connections. It could also be considered to move the receiver to the roof, to get rid of cable losses all together.
- **Receiver**
  The receiver should filter, amplify, frequency convert, demodulate, and decode the signal. In fact this can be done my multiple devices. Soft demodulation is certainly preferred and according to ISIS no problem.
- **Desktop computer or database**
  The digital signal should ultimately be forwarded to the TM Database, but logically is interpreted by a normal desktop computer first.
- **Driving system**
  Finally, as with the UHF/VHF systems, some form of tracking software and hardware is required. This equipment is all grouped under this common header. Due to the proposed size of the dish, a separate system, similar to that as already in place is likely required.

15.2 Worldwide radio amateur network

For the Delfi-C^3 mission the choice has been made to include and motivate radio amateurs all over the world to assist in collecting telemetry. Overall data reception has thereby been largely increased. Therefore, for Delfi-n3Xt again this participation is pursued.

In order to realize this participation, radio amateurs should be able to receive *and* forward telemetry from Delfi-n3Xt. *Data reception* requires specific equipment, in case of the Delfi-C^3 mission limited to an antenna, a transceiver, a soundcard, and software. Also, antenna tracking hard- and software is required. *Data forwarding* is taken care of by software.

The current Delft GS set-up uses much generally used radio amateur equipment. The large participation of radio amateurs has proven that they possess the required hardware. In case of Delfi-n3Xt, increased data rates might change this situation. However, as shown in the previous section, higher data rates on the downlink only affect the transceiver and possibly soundcard. Both the soundcard and transceiver can easily be replaced. At the same time, many (radio amateur) satellites already transmit at 9.6 kb/s so suitable hardware is likely already in place in many cases.

The software, DUDe, can be easily upgraded for higher data rates and will after it has been properly customized to receive Delfi-n3Xt telemetry, be distributed. DUDe will then serve data reception and forward client.

Delfi-n3Xt must, when its launch approaches, be represented at radio amateur meetings in order to get radio amateurs enthusiastic for this mission as well.

S-band reception hard- and software might already be widespread amongst radio amateurs as well, as its support is growing. It will be interesting to see if also radio amateurs can receive S-band signals transmitted from Delfi-n3Xt. Of course, this will depend on the periods of activation of the STX.

If downloaded data yield is to be increased using radio amateur support, a method of forwarding the telemetry will be required. This is not envisioned yet. In any case, the focus of the STX for now lies on proving the technology, so extensive radio amateur support is no requirement given the limited resources.

15.3 Eindhoven Command Ground Station

In case of the Delfi-C^3 mission the Eindhoven GS was used as a backup GS. It was to be used only in case of a failure of the Delft GS. Nevertheless, as mentioned above, a back-up option is not seriously required due to the redundancy amongst Delfi GS components and the accessibility of the latter in case of failures.

Nevertheless, the potential role of the Eindhoven GS can again be discussed in the near-future with the operators. It never hurts to have extra back-up or expanded communication capabilities, especially for the information-critical start of the mission.

15.4 GENSO

A final ground component worth mentioning is that of GENSO. The Global Educational Network for Satellite Operations (GENSO) can theoretically serve as a supplement to the Delft GS.

As a number of universities have already built their own small satellites, each university has its own ground station to operate the satellite. GENSO is an ESA initiative to interconnect all these ground stations to form a large ground station network, so that almost global coverage can be achieved for small university satellites.

GENSO ground stations have the capability to receive data from small satellites. This has the effect that the status of the satellite is also known when it is not in view of the university's own ground station. GENSO can provide support to universities willing to set up a new ground station to expand the community even further. In a later stage, GENSO stations will also have the capability to send telecommands to other satellites, so that satellites can be commanded at any time from anywhere in the world.

All GENSO ground stations operate in the amateur radio bands, most of them in VHF and UHF only. The GENSO specification specifies that the stations should have both uplink and downlink capability in UHF and VHF bands. The Delft GS is in fact fully equipped for this purpose. Since some recent small satellite proposals include a communication system operating in the S-band, it is expected that new stations will have S-band capability as well.

GENSO and its required software have however already been in development for many years, with little actual progress. No importance to the Delfi-n3Xt mission should therefore be assumed. Of course, the GENSO operations fit perfectly with the envisioned radio amateur support so additional ground stations willing to receive and forward Delfi-n3Xt telemetry should be motivated to do so.

For security reasons it is and will be undesired that others have the information required for commanding Delfi-n3Xt. Therefore, for Delfi-n3Xt the only ground station that will in principle be used for commanding Delfi-n3Xt will be the Delft Ground Station.

16 Next steps

From a COMMS point-of-view, next steps can be established for all the components. In general however it holds that the ‘paper phase’ has been finalized; all higher-level (desirable) functionalities have been established and integration and testing should now commence or continue. There are two actions that form exceptions to this statement:

- **Commence frequency filing using the parameters as established in this document**. It should also be decided which transmission and reception frequencies are allocated to the PTRX, and which to the ITRX.
- **The future and support of OLFAR should be determined.** It should be decided whether resources will actively be allocated to OLFAR. This is necessary in order to properly account for its mechanical, electrical and operational integration.

To make a distinction between the primary (essential) communication link, and the experimental S-band link, the two are presented separately below.

UHF/VHF communications link

PTRX:

- As the design of the PTRX is finished on paper, all hardware should be integrated and tested. This in turn allows to test the:
  - *Final data rate achievable (downlink)*. Consequently, the frame size can be confirmed. Also, frequency spectrum meaurements can then be performed to establish the new performance.
  - *End-to-end downlink* using an antenna and the GS equipment to receive and demodulate the signal; this can confirm the use of the DelfiX protocol as well as the higher data rates (requiring an upgraded receiver).

ITRX:

- As the ITRX is a payload, no design is taken care of in-house. Nevertheless:
  - The standard system bus interface should be confirmed pending on the final SSB definition.
  - The DelfiX protocol should be incorporated.

Antenna system:

- Following the definition of the new Delfi-n3Xt solar panel configuration, the UHF/VHF antenna configuration should be confirmed as assessed in [SLR 0036]. Consequently, the system should be integrated and tested, using the phasing circuit and four MABs. Possibly, the performance of the antennas can be measured in a manner similar as done for Delfi-C^3, in cooperation with NLR.

Ground Station:

- DIGIT should be updated following the definition of Delfi-n3Xt TCs.
- DUDe should be finalized.
- The TNC31S software and the ICOM 910 hardware should be replaced to allow for higher downlink bit rates and the DelfiX protocol.

S-band communications link

From a top-level, as well as the link budget it is most important to establish the:

- FEC coding scheme that will be applied
- The size of the ground station antenna that can be implemented

Concrete next steps for the STX are stated in [SLR 0387].
Concrete next steps for the S-band antenna system are stated in [SLR 0036].
Concrete next steps for the STX downlink link budget are stated in [SLR 0106].

Ground Station:

- The Delft GS should be upgraded to properly receive an MSK modulated S-band signal, preferably of varing data rate.

Appendix A Modulation schemes

As introduced in section 7.2, this appendix contains an introduction a large number of modulation schemes. Emphasis is placed on those schemes with most assumed relevance to Delfi-n3Xt. The power performance, bandwidth performance and required complexity are discussed in more detail. Also, different options of complexity which in turn improve on either power or bandwidth efficiency are introduced. Modulator and demodulator schematics are presented.

Subsections A.1 and A.2 discuss the two basic modulation techniques FSK and PSK. A scheme related to both, minimum shift keying (MSK), is assessed in subsection A.3. Subsection A.4 introduces the more general class to which MSK belongs; Continuous Phase Modulation or CPM. Subsection A.5 reviews the third basic modulation technique ASK, with its derivative quadrature amplitude modulation (QAM) discussed in subsection A.6.

The latter three subsections assess techniques that can be used in coherence with (some of) the other modulation schemes; trellis codes modulation (TCM), orthogonal division frequency multiplexing (ODFM) and spread spectrum techniques. TCM is part of the approach in which coding is integrated with modulation as is further explained in subsection A.7. ODFM uses multiple frequencies to directly increase achievable data rates as introduced in subsection A.8. Finally spread spectrum techniques are quickly introduced in section A.9, providing a means to lower the required signal-to-noise ratio during transmission.

A.1 M-ary Frequency Shift Keying (MFSK)

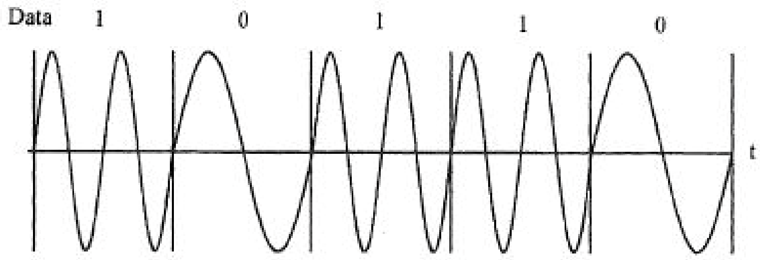
As explained at the start of this chapter, FSK applies different carrier frequencies to represent different bit values. Consequently MFSK indicates the use of *M* frequencies, as such 2FSK or BFSK indicates M=2. When an M higher than 2 is used, the *bit rate* increases with respect to the *symbol rate*. If 4 frequencies are used for example, a demodulated 'symbol' indicates one out of 4 choices. Therefore every choice can represent 2 bits (10, 00, 11, 01).

The difference between the *M* frequencies used for a MFSK signal is usually given by *h*. This value is defined as:

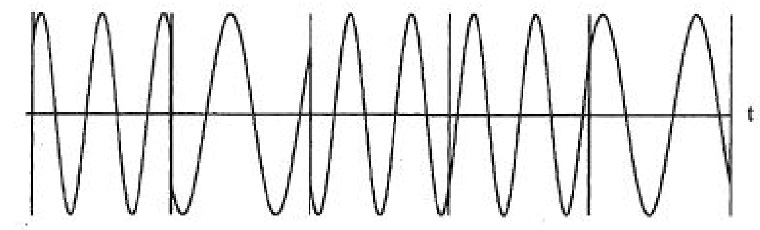
$$h = 2\Delta f T = \frac{2\Delta f}{R_b}, \qquad (6\text{-}9)$$

Δf is defined as the difference between the (apparent) carrier frequency, in other words $2\Delta f$ equals the difference between the two (or more) FSK frequencies. The larger *h* one has, the larger bandwidth is required. Therefore it makes sense to minimize this value.

The following Figure A-1 demonstrates the signals generated for BFSK in case of *h* = 1 and *h* = 3/4. It can be seen that phases are continuous in the case of *h* = 1, also called Sunde's FSK, whereas in the other case they are not. The latter broadens signal bandwidth and is therefore generally not preferred. Sunde's FSK in fact always yields a continuous phase, as long as bit transitions take place at the zero-phase points. It also results in *orthogonality* of the two frequencies, which is important for demodulation.



(a) Sunde's FSK



(b) FSK signal with discontinuous phase

Figure A-1: BFSK modulation with and without continuous phases [SLR 0552]

The minimum *h* for continuous phases and frequency orthogonality is actually *h* = 0.5; this yields Minimum Shift Keying or MSK. In this case however, bits are no longer independent of each other, as phases are not necessarily 0 degrees at bit transitions, but can also equal 90 or 270 degrees. In order to guarantee continuous phase, the following bit frequency should then also start with a 90 or 270 degree phase shift. MSK is further reviewed in section A.3.

An *h* of 1 yields continuous phases and is independent of bit sequence. Other values of *h* in fact also can yield continuous phases (but no bit sequence independence), but this requires more complex modulators and demodulators. This type of modulation is called CPM or *continuous phase modulation*. It is further reviewed in subsection A.4. In fact, pulse shaping, as will be introduced in section 7.4, if applied on FSK schemes and generally used in CPM schemes, also requires more complex (CPM) modulators as they involve slowly changing from one to the other frequency to reduce required bandwidth; this cannot be done by the modulator introduced below.

FSK can be both coherently and noncoherently modulated. Noncoherently modulated FSK modulation uses two signal frequencies with no regard for their initial phases. Coherently modulated FSK modulation applies a slightly smarter approach, with a frequency synthesizer taking care of having equal initial phases, such as in the case of Sunde's FSK. In order to have coherent FSK, both coherent modulation and coherent demodulation are needed. FSK can easily be coherently modulated; it just requires an oscillator, and a frequency synthesizer to extract from this oscillating signal to in-phase frequencies. A multiplexer then decides on which frequency to pass depending on incoming 0's or 1's. This set-up in terms of functional blocks is shown in Figure A-2 below.

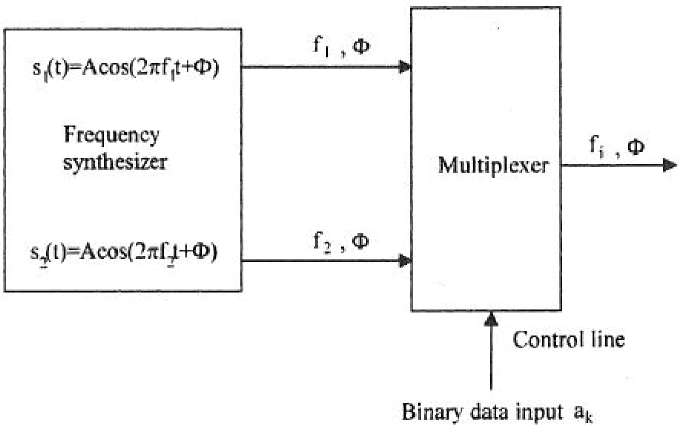


Figure A-2: Coherent BFSK modulator [SLR 0552]

Coherent demodulation is difficult however, due to the need for coherent reference signals (carrier synchronization in terms of phase *and* frequency). Consequently it is not often done, since the resulting performance degradation is little. The difference in performance for BFSK is illustrated by Figure A-4 below. It can be seen that at a *BER* or P_b of 10^{-5} the difference in required E_b/N_0 is less than 1 dB gain; 12.6 dB is required for coherently demodulated BFSK, with 13.3 dB required for noncoherent demodulation. This is similar for higher *M*'s. The more power efficient BPSK is introduced in the next subsection.

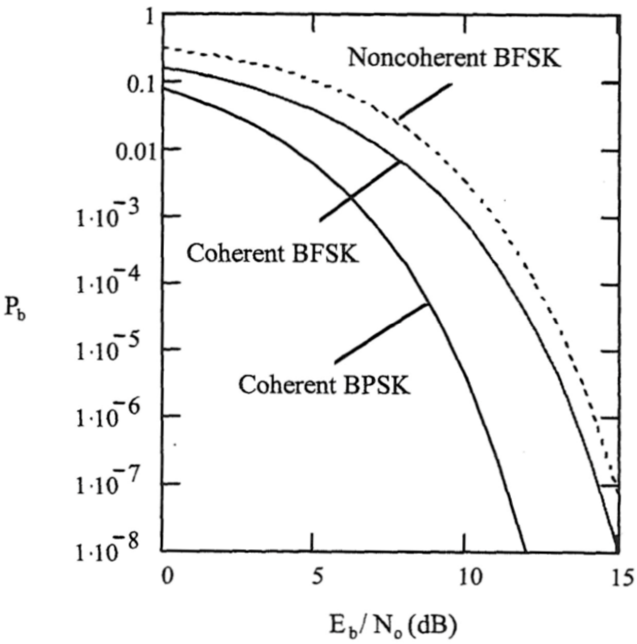


Figure A-3: P_b of BFSK and BPSK versus E_b/N_0 [SLR 0552]

Although noncoherent demodulation is in most cases acceptable, it is important to modulate the signal coherently; coherently modulated FSK manages to have 90% of its power within a bandwidth of 1.23 R_b, and $B_{99\%} \approx 2.12\ R_b$ yielding a practical transmission bandwidth $B_T = 2\ R_b$. In case of noncoherently modulated FSK it is doubled; with B_T practically equaling 4 R_b. These values for coherently modulated FSK are illustrated by Figure A-4 below, which demonstrates properties of Sunde's FSK thus with h = 1. In order to explain energy percentage bandwidth efficiency, some additional matter is required as will be introduced first.

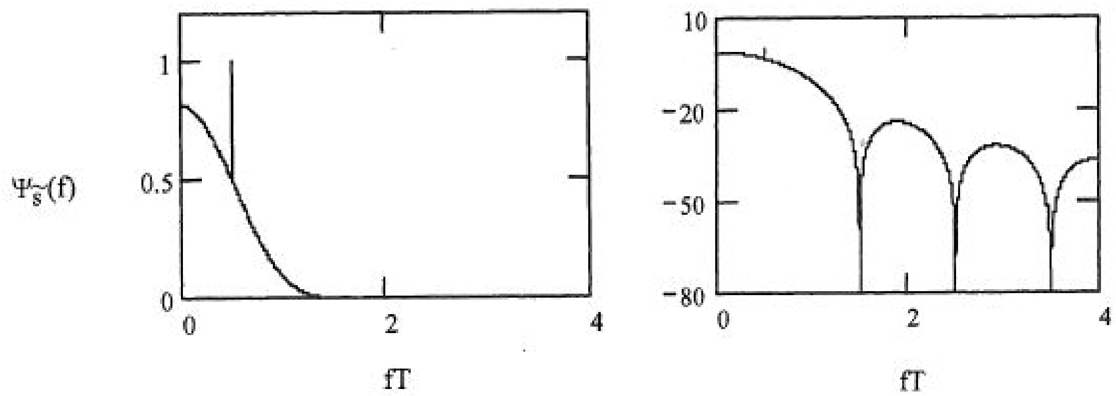


Figure A-4: Power Spectral Density of FSK: linear scale (left), logarithmic scale (right) [SLR 0552]

The mathematics behind these curves are beyond the scope of this text, but an exact derivation can be found in [SLR 0552]. Here the results are used to draw conclusions.

In Figure A-4 above, Ψ_s indicates spectral density, whereas *fT* indicates normalized frequency: *fT* (which stand for frequency times bit length) = f/R_b or frequency divided by bit rate. The amplitude *A* and the period *T* are set to 1 to calculate unity symbol pulse energy. The significance of this normalized frequency is the following. The signal energy, at a frequency away from the transmission frequency (f = 0), contains the indicated amount of spectral power. The point f = 0 in this graphs indicates the *apparent* carrier frequency, which is in between the two actual FSK carriers. Indeed, at f = 0.5 (most clearly in Figure A-4, left), one finds the first FSK frequency with a value of *A* equals 1. As this graph in fact is in fact symmetrical around the zero frequency and also holds for negative frequencies, a value of f = -0.5 gives the other FSK frequency. The distance between these two frequencies is, as expected, 1, the value of *h* introduced before.

Figure A-4 (right) shows the same graph as Figure A-4 (left), but with a logarithmic scale. It can be seen clearly here that modulation at a certain frequency has effects not restricted to that frequency. In case of BFSK, when a 1 Hz data signal is transmitted, it can be said that 8 Hz of frequency bandwidth (2 times 4) away from the transmission frequency, still a residual component of between -35 and -40 dB with respect to the main signal exists. The fact that in fact an infinite frequency spectrum is created by the transmission is called *spectral splatter*. These components are called *spurious components* when they occur outside of the targeted bandwidth. The definition of null-to-null bandwidth is also easily visualized looking at Figure A-4. Finally, the occurrence of in fact an infinite number of lobes is called *spectral regrowth*; some modulation schemes, such as BPSK, suffer from strong spectral regrowth as a relatively large amount of power is contained within these lobes.

Regulations stipulate that signals outside of a designated bandwidth should have a maximum strength. Spectral splatter can be counteracted by pulse shaping, as will be further explained in the next section 7.4. Also, bandpass filtering can be applied after modulation; this simply cuts off part of the spectrum. The result of this however is that part of the original signal strength is lost, therefore requiring more transmission power.

Also suggested by these graphs, is an important relation introduced before: a doubled data rate doubles the frequency spectrum broadness. This is because a doubled bit rate, with a consequently halved bit period, gives a lower *fT* value, which in turn belongs to a higher spectral energy.

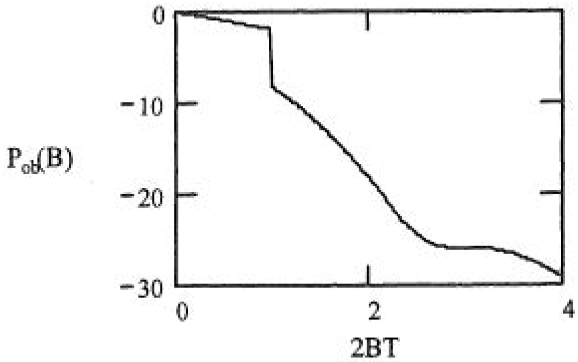


Figure A-5: Power Spectral Density of FSK: linear scale (left), logarithmic scale (right) [SLR 0552]

Using the graphs established above, Figure A-5 above can be established. It is basically the result of calculating the surface underneath the graphs in Figure A-4, between – *fT* and +*fT*, and establishing how much power is present outside of this bandwidth; the out-of-band power. By taking a range of *f* values, a bandwidth *B* is then obtained. And, as also negative frequencies are taken into account, the *BT* value is doubled. The graph then shows the amount of power located within a certain bandwidth, centered on the (apparent) transmission frequency (in dB). It should at this point be said that the calculation of these PSDs assumes equiprobable bit sequences; in other words an equal total number of 0’s and 1’s.

The noncoherent FSK demodulator is not only simple compared to coherent FSK demodulators, but also compared to PSK and QAM demodulators; also for higher *M*. It can be implemented with so-called correlators and squarers, or even simpler through the use of matched filters and envelope detectors. These types of demodulator for BFSK are shown schematically in Figure A-7 and Figure A-6 respectively. Another type of demodulator, using a conventional frequency discriminator can also be used, but as its resulting power performance is some 1 dB less it will not be reviewed here.

The use of filters and envelope detectors is conceptually the easiest, as illustrated by Figure A-6. Two bandpass filters allow for one frequency but filter out the other. Consequently envelope detectors throw away frequency and phase information, but keep the overall amplitude value. Once per bit period, if possible at the middle, both signals are sampled; *k* is an integer equal to the total amount of bits, and *T* is the bit period. The comparator then gives a 0 or 1 depending on which of the two channels contains higher power content. This approach is the simplest as it does not require any carrier synchronization in terms of phase or frequency. Performance however degrades quickly if Doppler shift is seriously present, as the filters can then no longer be very narrow-band. As such, the bandpass filters can be replaced by what are called matched filters, which require carrier frequency synchronization to specify the filter specifics.

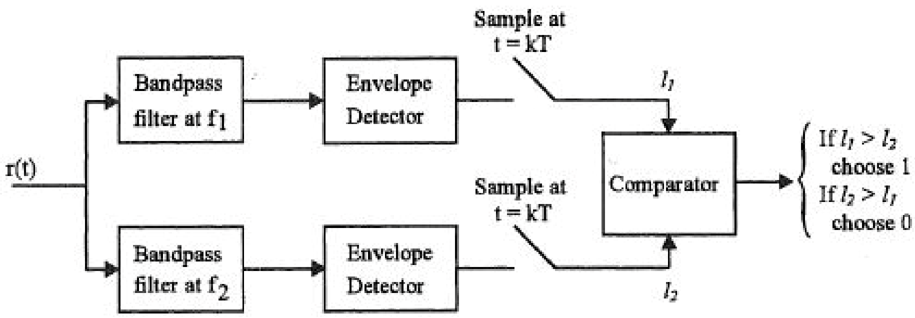


Figure A-6: Noncoherent BFSK demodulator using bandpass filters and envelope detectors [SLR 0552]

A second approach, also requiring carrier frequency but not phase synchronization, is the use of correlators and squarers as illustrated in Figure A-7. A received signal (without noise) can be written as:

$$S(t,\theta) = A\cos(2\pi\, ft + \theta) = A\cos\theta\cos 2\pi\, ft - A\sin\theta\sin 2\pi\, ft\,, \tag{A-1}$$

Where *A* is an amplitude value, *t* is time, *f* is frequency and θ is the unknown phase angle. The received signal can be seen to be partly correlated with $\cos 2\pi\, ft$, and partly with $\sin 2\pi\, ft$. Using the so-called correlators, the frequency products together with the integrators in Figure A-7 below, the signal energy in these two parts are collected. By squaring the results and adding them, the resulting signal becomes independent of the signal phase, as $\cos^2\theta + \sin^2\theta = 1$. Consequently, a comparator indicates which signal frequency was received; the required orthogonality between the FSK frequencies makes sure that when one signal is being detected, the other set of correlators and squarers will only yield a noise component.

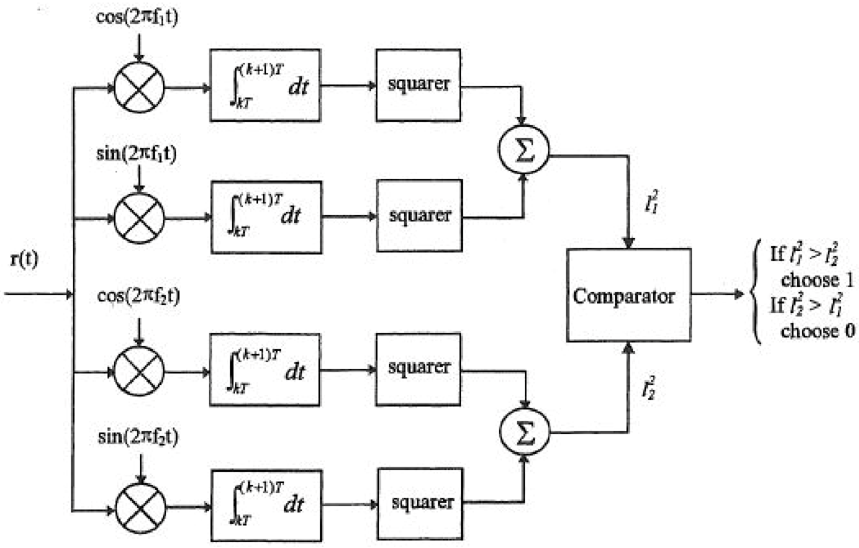


Figure A-7: Noncoherent BFSK demodulator using correlators and squarers [SLR 0552]

The same demodulation simplicity commented on above yields FSK to be used for the telecommand uplink in Delfi-C^3's RAP. However, as Doppler shift is seriously present, the final noncoherent modulator using correlators and squarers was chosen. Also, the modulators and demodulators above can easily be expanded to modulation schemes with higher values of *M*.

The bit error rates of MFSK versus E_b/N_0 can be found in Figure A-8 below. It can be seen that P_b performance goes up with higher M; this comes at the cost of more complexity and less bandwidth efficiency as was illustrated by Figure 7.6.

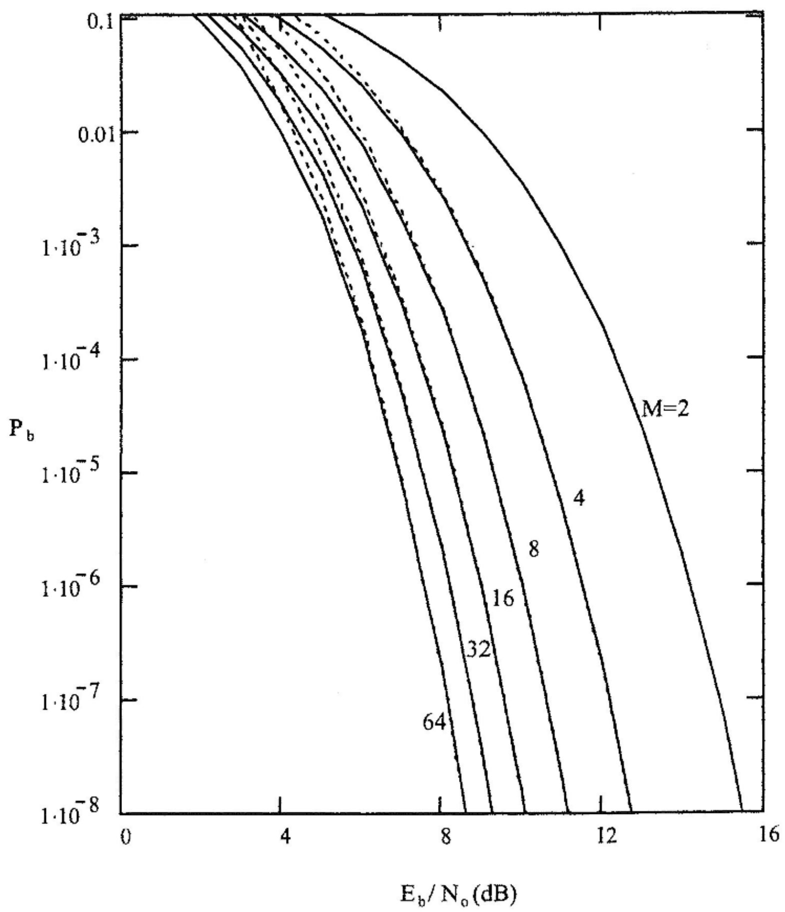


Figure A-8: P_b of noncoherently modulated MFSK versus E_b/N_0 [SLR 0552]

A.2 M-ary Phase Shift Keying (MPSK)

PSK works by shifting the phase of a carrier signal when a bit shift takes place, as illustrated by Figure 7.5. The simplest case with M = 2, shifts only the phase Q of the carrier, the absolute magnitude I remains the same. This is further illustrated in Figure A-9 below.

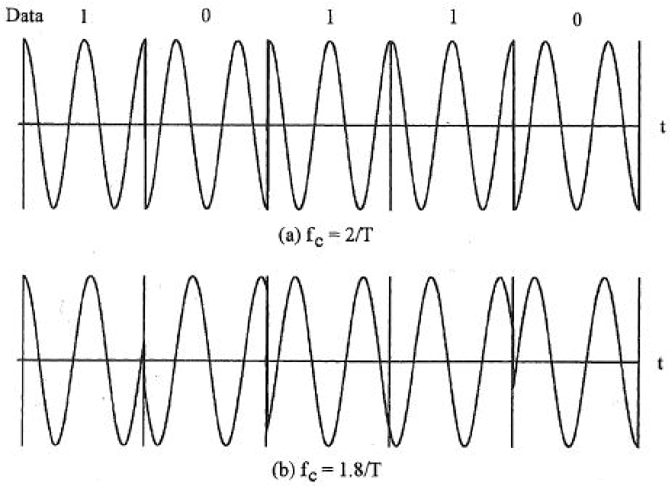


Figure A-9: BPSK modulation for two choices of carrier frequency

As is demonstrated in the top of the figure, case (a), if f_c is an integer multiple of R_b, the phase at bit transitions is either 0 or π . In case (b) however this is not so. In principle, the former condition is required to ensure minimum bit error probability. If however $f_c >> R_b$, this condition is negligible. In case of Delfi-C^3's RAP, the 1200 b/s signal is immediately modulated onto a 10.7 MHz signal; therefore the latter criterion is satisfied.

If M is more than 2, also the phase Q is shifted, hence quadrature modulation is applied. An illustration of QPSK is included as Figure A-10, with the horizontal and vertical axes representing amplitude or intensity I and phase or quadrature Q respectively.

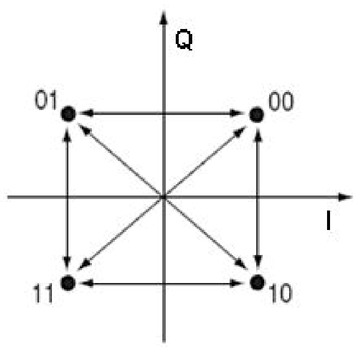


Figure A-10: QPSK state diagram [SLR 0447]

It should be noted that the four states of QPSK can also lie on the axes, in this way in two out of the symbol transitions possible only *I* or *Q* needs to be shifted without crossing the origin; the latter is beneficial for spectral optimization. Also QPSK is equal to 4QAM, as will be reviewed in subsection A.6. The difference between PSK and QAM is that with PSK the states always lie on a circle centered at the origin, thus keeping the signal envelope constant.

BPSK can be modulated very easily, using the functionalities as described in Figure A-11 below. A sequence of bits, usually given by 0 and 1 voltage values, is to be converted to bits indicated by -1 and +1 values. In terms of line codes this transformation can be labeled as *converting unipolar NRZ to (bi)polar NRZ.* Line codes are introduced in section 7.3. Once the bit stream is expressed by positive and negative, but equal amplitude values, they can simply be multiplied by the desired carrier frequency, being an intermediate frequency (IF) or the ultimate radio frequency (RF).

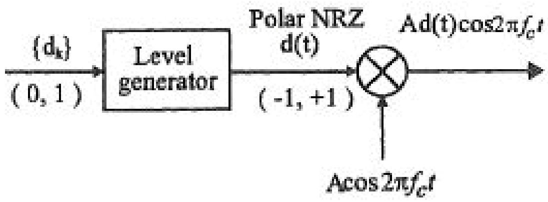


Figure A-11: BPSK modulator [adapted from SLR 0552]

The modulation of QPSK requires more functionality, as is shown in Figure A-12 below. It involves splitting the (bipolar) bit stream in pairs of two bits, and using an oscillator to generate two components being phase-shifted with respect to each other. These components are multiplied with the pairs of bits, and consequently summed.

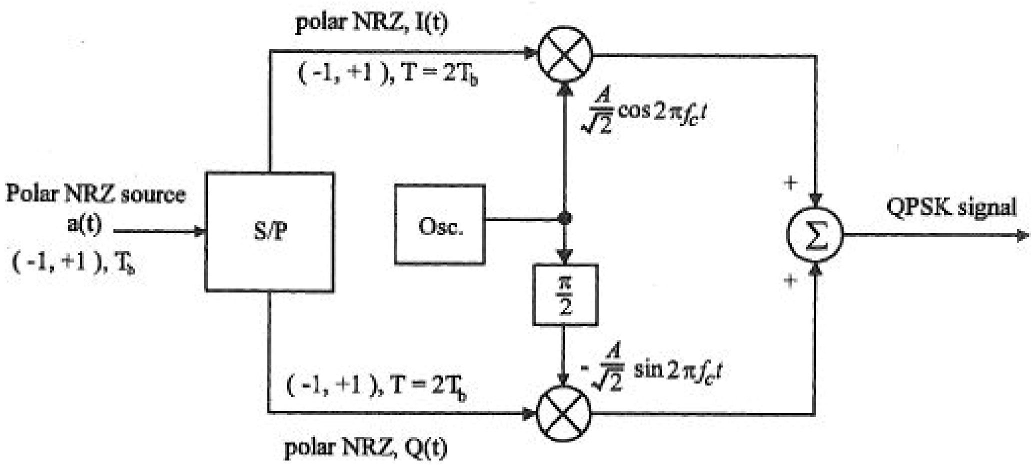


Figure A-12: QPSK modulator [SLR 0552]

The bit error rate performance of (B)PSK is better than that of (B)FSK, as has been shown by Figure A-3. Only an E_b/N_0 of 9.6 dB is required for a *BER* performance of 10^{-5}. It can be seen that BPSK is to be demodulated coherently; simply because data is carried by the phase of the received signal, necessitating both carrier frequency and phase synchronization. Noncoherent demodulation of BPSK can be achieved by differentially encoded BPSK.

Differentially encoded BPSK is called BEPSK. It can be coherently and differentially demodulated. If a differential demodulator is used, the scheme is called DBPSK, or in short DPSK. A differential modulation scheme is actually one that applies a differential line code, and consequently applies the targeted modulation scheme. In case of BPSK, NRZ-I is applied; this encodes bits relatively, thus with respect to the previous bit, instead of absolutely. This means an exclusive OR (XOR) operation is performed between the present bit and previous (from this operation) *resulting* bit; two equal bits give a 0, two different bits a 1 (in effect this means the first resulting bit should be assumed to equal either 0 or 1; this can be shown not to matter in demodulation). The result of this coding however is that the demodulator does not need to know the polarity of the signal; whether what it defines as $+A$ represents a 0 or a 1. The exact characteristics of this code are introduced in the next section, 7.3.

BEPSK adds the feature of polarity indifference at the cost of slight error performance degradation with respect to coherent BPSK, as can be seen in Figure A-15; this is induced because the demodulator is for each bit dependent on the previous bit, exposed to a certain error rate. However, because of the polarity independence in case differential coding is applied, carrier phase synchronization is actually no longer required in the demodulator.

The optimum DBPSK demodulator still applies carrier frequency synchronization to increase *BER* performance. Nevertheless, due to an increase of complexity when Doppler shift plays an important role, matching oscillators in both modulator and demodulator cannot take care of this. Therefore, more complicated carrier recovery is required; mainly achieved by an *M*-th power loop or a Costas loop. The resulting demodulator required is than demonstrated in Figure A-13 below.

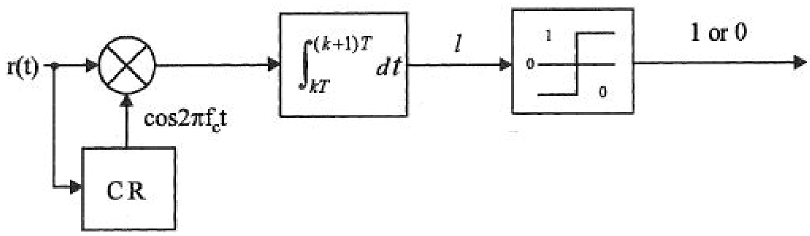


Figure A-13: Coherent BPSK demodulator [SLR 0552]

It can be seen that a CR, or carrier recovery circuit is to be used to return the perfectly synchronized carrier used to modulate the signal in the modulator. The result of the multiplication of the original signal with an identical frequency with unit amplitude is a value of $T/2$ times the bit amplitude, and a remainder which is dependent on the relation between the carrier frequency and data rate. If the carrier frequency is equal to an integer multiple of the data rate, or as mentioned above, the carrier frequency is much more than the data rate, the second term equals or approaches zero. A consequent integration of the bit period duration gives either a positive or negative signal, belonging to a transmitted -1 or +1 value. A final block checks the value l and gives a conclusion as to a 0 or 1 is received.

Carrier recovery however is not required in a third variant, called suboptimum DBPK. Suboptimum DPSK lowers the demodulation complexity and requires only symbol synchronization; in fact a local oscillator is not even required, and its bandwidth filters should only be tuned to allow the entire (possibly Doppler shifted) information signal to pass. The resulting demodulator is shown in Figure A-14.

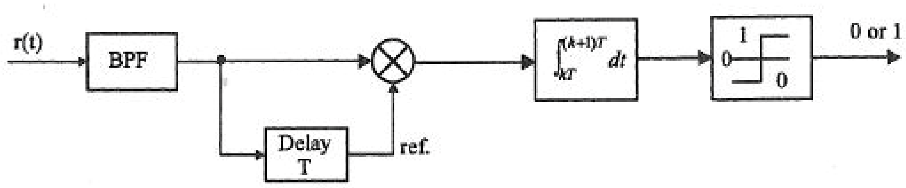


Figure A-14: Noncoherent DBPSK demodulator [SLR 0552]

First, a filter, such as a conceptually easy to visualize bandpass filter, limits the input signal to the targeted frequency bandwidth. Consequently two consecutive bits are mixed together by multiplication, and passed on to an integrator. The signals added together are either equal, or opposite. As such, there are continuously opposite or continuously equal amplitude values added together. Equal values always result in a positive total energy ($A \cdot A$ or $-A \cdot -A$), whereas opposite values result in a negative total energy ($-A \cdot A$). The integrator adds all values during a bit period and yields either a positive or negative output belonging to a 1 or 0. The differential coding is clearly used here, as consecutive bits are added to give a result based on them being equal, or different.

The demodulator schemes for QPSK will not be presented for brevity. Nevertheless, as with the difference between modulator and demodulator for BPSK and QPSK, it involves the same architecture as either Figure A-13 or Figure A-14, but again with two streams between which a phase difference is introduced, after which they are summed together.

The error performances of the DBPSK schemes are shown in Figure A-15 below. In case of optimum DBPSK, a P_b or *BER* of 10^{-5} requires an E_b/N_0 of 10.3 dB and in the case of suboptimum DBPSK some 11.5 dB is required, as opposed to 9.6 dB for coherent BPSK. The suboptimal DBPSK demodulator is the usual-sense demodulator due to its complexity advantages, but only if Doppler shift is limited. In case of Delfi-n3Xt however, as with Delfi-C^3, it is not. As such, also taking advantage of the fact the on-ground demodulation can be more complicated, the coherent DEBSPK scheme was implemented in the Delfi-C^3 mission, with the demodulator as indicated in Figure 7.7 implemented in RASCAL, the demodulator software.

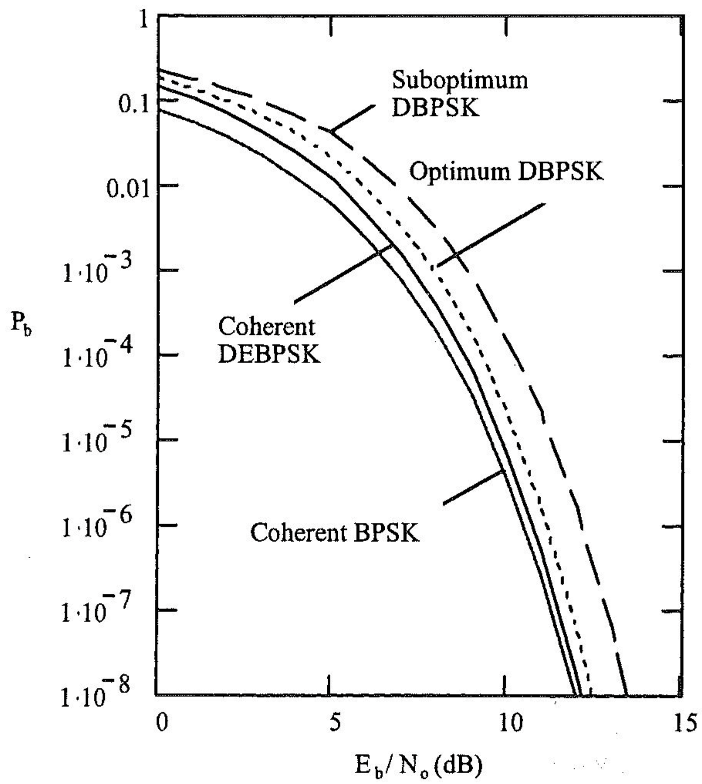


Figure A-15: P_b of BPSK and BFSK versus E_b/N_0 [SLR 0552]

As was shown in Figure 7.6, with an increasing *M*, MPSK schemes increase bandwidth efficiency at the expense of power efficiency. The performance of QPSK compared to BPSK however is a special case, in the sense that QPSK doubles the bit rate achievable, without a decrease in *BER* performance; even more surprisingly without too much added complexity for the modulator and demodulator. Nevertheless, this is the case only for coherently demodulated QPSK.

Optimally differentially encoded MPSK schemes have an increased required E_b/N_0 of 3 dB with respect to their nondifferential counterparts, at $M=\infty$. Of more practical use of course are lower values of *M*; at $M=4$ (QPSK) the increase is 2.3 dB, with a 2.7 dB increase at $M=8$. This means that optimum DQPSK requires an E_b/N_0 of 2.3 more than optimum DBPSK, at the gain of 3 dB in data rate.

A suboptimal differentially decoded MPSK demodulator only exists for DQPSK, next to of course that of DBPSK; higher MPSK schemes can not be demodulated without carrier frequency synchronization. The performance difference between suboptimum DBPSK and DQPSK decreases slightly with respect to optimum versions, to 1.75 dB. As a result, suboptimally and differentially demodulated QPSK still outperforms its BPSK counterpart, as the bitrate can be doubled (+3 dB) requiring only slightly more power (-1.75 dB).

The maximum theoretical efficiency for PSK schemes is high. However, in reality there are some drawbacks; due to the large sudden phase changes belonging to PSK modulation, large spectral regrowth occurs. This can be seen in Figure A-16. Figure A-16 (left) shows that QPSK contains much more energy within 1 Hz bandwidth (*fT* = -0.5 to +0.5). It also reaches first null two times as fast. Figure A-16 (right) clearly shows the spectral lobes. Looking at the trend of the top of the lobes shows that spectral power does not decrease fast anymore, after only small values of *fT*.

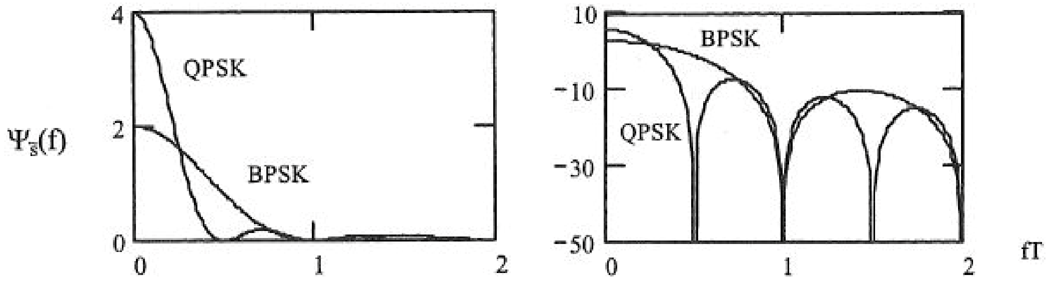


Figure A-16: Power Spectral Density of PSK: linear scale (left), logarithmic scale (right) [SLR 0552]

This spectral inefficiency is even better shown by Figure A-17 below, which indicates the out-of band power. The differences between BPSK and QPSK consistently are 3 dB. But from the graphs in Figure A-17 (left) it can be seen that out-of-band power hardly decreases with increasing bandwidth. As such, in all practical systems, *BPSK and QPSK require filtering and/or pulse shaping*. MSK is also introduced in Figure A-17 and will be discussed in the nest subsection.

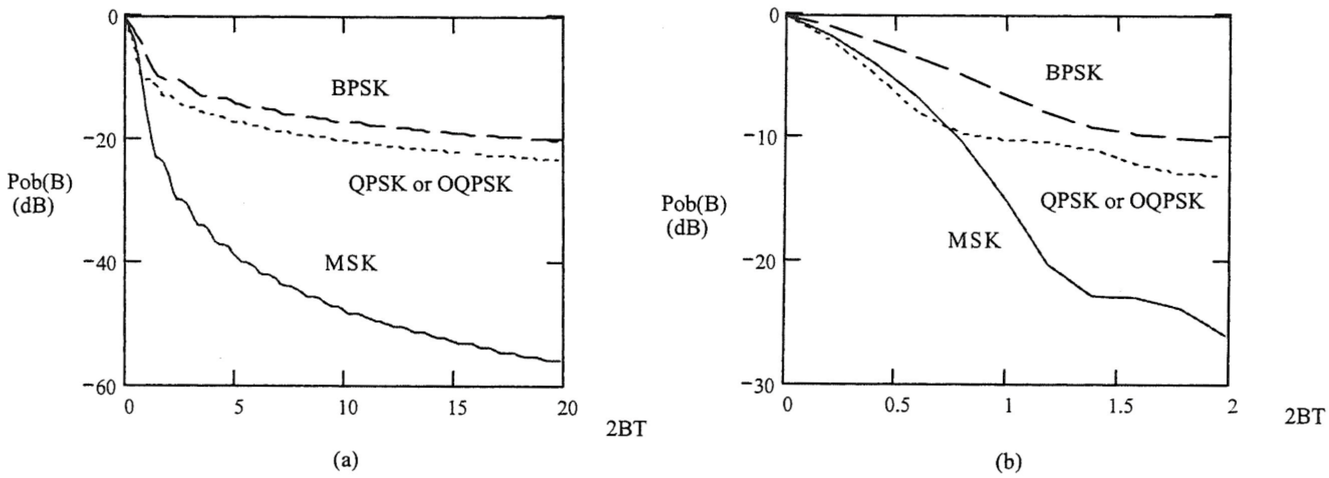


Figure A-17: Fractional out-of-band power of several modulation schemes [SLR 0552]

The following table summarizes the 90% and 99% energy bandwidths of BFSK, BPSK, QPSK and MSK.

Table A-1: Energy percentage bandwidth of popular modulation techniques

Modulation technique	90% energy bandwidth	99% energy bandwidth
BPSK	1.7 R_b	20 R_b
QPSK	0.8 R_b	10 R_b
BFSK	1.23 R_b	2.12 R_b
MSK	0.76 R_b	1.2 R_b

By applying filters before the modulation process, in other words by shaping the bits the energy bandwidth for the PSK schemes can be improved. This is reviewed in [SLR 0447]. Applying a Nyquist filter, usually in the form of a raised cosine, before modulation can largely reduce sideband production, but practically not until the level of energy efficiency of BFSK and MSK. The resulting carrier signal for a shaped BPSK modulation scheme is illustrated in Figure 7.5, under its alias double side band, suppressed carrier or DSB-SC (with shaping). Another name is raised cosine BPSK, or RC-BPSK. The downside of this shaping is next to some added complexity that the modulation scheme is no longer constant envelope, which can be required by non-linear amplifiers if they do not offer some type of envelope elimination and restoration (EER) via an incorporated feedback system.

Finally worth mentioning are the well known related *offset QPSK* or OQPSK and *π/4-QPSK*. Both apply a delay between the *I* and *Q* shifts in the signal modulation, in order to avoid 180° phase shifts. This makes the signal less susceptible to spectral side lobe restoration (large power contents out of the targeted frequency bandwidth, when amplified by a non-linear amplifier after bandpass filtering) compared to QPSK. This is beneficial in satellite communication in case of severe bandwidth restrictions, and non-linear amplification. OQPSK however cannot be differentially demodulated, therefore complicating demodulation. π/4-QPSK however, can. It has little added complexity with respect to suboptimum DQPSK at no extra penalty in error performance.

A.3 Minimum Shift Keying (MSK)

The previous section introduced OQPSK, with its main advantage over QPSK that it exhibits less phase changes at symbol transitions, limiting out-of-band interference. MSK provides even further improvement by making phase transitions completely continuous. MSK can be derived from OQPSK by shaping the pulses to the modulator with half sinusoidal waveforms. Also, and simpler, MSK can be produced as differentially encoded FSK with a value of h of 0.5, which is the minimum frequency separation that still allows for orthogonality between the FSK signals, hence 'minimum' shift keying; orthogonality requires a separation of an integer multiple of $1/2T$. This orthogonality is required for coherent demodulation, if used. Differentially encoded FSK implies applying NRZ-I coding. Since producing MSK signals from OQPSK leaves a 180° phase ambiguity this method requires NRZ-I coding and decoding as well.

It can be seen in Figure 7.6 that the resulting performance of MSK compared to BPSK, BFSK, 4FSK is simply better. The power spectral density of MSK is illustrated in Figure A-18 and shows the lobes that have widths between those of BPSK and QPSK, but the overall power content of the lobes drops off much faster. Thus whereas QPSK has a better null-to-null bandwidth efficiency, its energy percentage bandwidth for large percentages is worse.

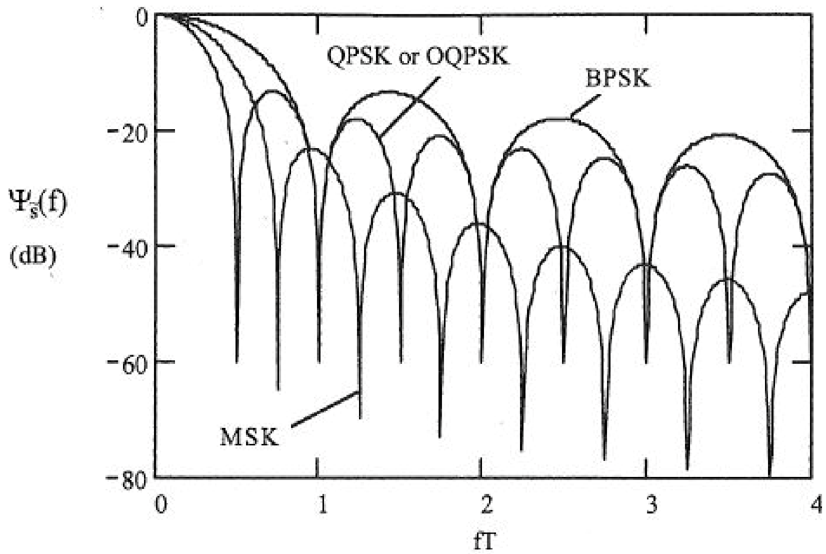


Figure A-18: Power Spectral Density of MSK (logarithmic scale) [SLR 0552]

Together with the bandwidth advantages MSK is easily demodulated noncoherently at the cost of a small loss in E_b/N_0 just as all FSK schemes. However coherent demodulation comes at the cost of little added complexity as phase locked loops (PLLs) can be used to extract the strong spectral components flanking the MSK signal needed for carrier frequency and phase synchronization.

What is interesting about MSK is that in fact it is continuous in phase; but unlike simple FSK schemes with $h = 1$, there are certain bit sequences that cannot exist due to properties of the differential code. An example is the sequence '0101'. As such, knowledge of this yields certain information on the bit sequence to be decoded. The phase continuity and bit sequence information mean a maximum likelihood algorithm such as the Viterbi algorithm can be used to demodulate a MSK-modulated signal. This algorithm is further introduced in section 7.8. Whereas in case of MSK the so-called observation interval usually is $2T$ (to account for the differential coding), it can be increased to $3T$ or $5T$ for an error performance increment of 1 and 1.2 dB respectively. Longer intervals yield minor improvements at complexity increments that do not seem to favor them over simple but efficient MSK.

Gaussian MSK or GMSK applies a Gaussian filter to the data stream before it is modulated. It significantly suppresses spectral splatter further to generally acceptable values at a cost of only up to 1 dB error performance. These performances and the characteristics of Gaussian filtering are further discussed in section 7.4. The application of a Gaussian filter does require a more advanced modulator than generally required, as the frequencies used to indicate 0's and 1's should slowly be varied. GMSK is used in the European Global System for Mobile Communications, or the GSM system, as well as Bluetooth.

Another version of MSK used in high data rate communication is *Serial MSK*, or SMSK, but this scheme is said only to provide benefits at high data rates (1 Mb/s and up), not (yet) applicable to the Delfi programme. Together with MSK these modulation schemes can be grouped under the collective header CPM techniques, or constant phase modulation techniques.

A.4 Continuous Phase Modulation (CPM)

CPM techniques have as the name implies, constant phases. Thereby they improve upon the general bandwidth efficiency of FSK schemes, while keeping good error performance. In fact, due to the notion of memory even error performance can be improved upon. With respect to MSK, which is a simple CPM scheme, either general power efficiency or bandwidth efficiency can be increased. Of course, all comes at the cost of complexity. With respect to QAM schemes, as are introduced below and which have the best bandwidth efficiency, CPM schemes have a constant envelope, thereby allowing for nonlinear amplification.

In comparison to FSK schemes, CPM schemes also have a predetermined M (number of frequencies) and h (distance between frequencies). Furthermore important are specific pulse (bit) shapes (rectangular in the simplest case) and the symbol time L. Pulse shaping is introduced in section 7.4 below, and smoothes the abrupt changes between symbols. MSK then becomes a simple CPM version with $M = 2$, $h = 0.5$, $L = 1$ and a rectangular pulse shape. Due to the introduced of extra variables, guaranteeing a continuous phase, memory is effectively introduced in the phase of the received signal. CPM schemes modulate a signal based on the input bit, but also on the present phase state of the modulated signal.

As well as (generally) introducing a filter, CPM requires added complexity in the demodulator in the form of more complex synchronization, and a maximum likelihood algorithm; the observation interval should be increased for better error performance. In case of the Delfi programme, CPM schemes (other than MSK) will likely not be of much use until the boundaries of either power and bandwidth efficiency are seriously pushed; currently modulation and demodulation complexity is more of a bottleneck.

A final CPM subtype worth mentioned is that of multi-h CPM, or MHPM. As the name suggests, the h can be chosen to vary. The result is that overall error performance can increase, whereas more bandwidth is required, but with less intensity. At the gain of yet more complexity, this can yield even more power efficiency and overall bandwidth efficiency, most useful in channels where the specific channel to be used is not strictly limited, but power efficiency is. In fact its addition can be seen as a form of error correction introduced to the modulation scheme.

A.5 M-ary Amplitude Shift Keying (MASK)

Most of the modulation schemes discussed above have constant envelopes, which can be an essential property if highly efficient yet non-linear amplifiers are used to amplify the RF signal. If a constant envelope is not required, techniques as ASK or QAM can be used, reviewed in this and the next subsection.

The main advantage of MASK is its implementation simplicity. The simplest type of ASK in turn is OOK, as illustrated in Figure 7.5, where the signals shift between M = 2 amplitude levels, of which one equals zero. OOK has the same bandwidth efficiency as BPSK but with a spectral line which can be locked on to by a PLL. The *BER* performance of coherent OOK however is 3 dB below that of coherent BPSK, therefore only if circuit simplicity is of prime concern should OOK be selected for digital modulation. It holds that MASK is simply inferior to MPSK, with as mentioned 3 dB of lesser error performance at $M = 2$, asymptotically increasing to 6 dB at $M = \infty$.

A.6 Quadrature Amplitude Modulation (QAM)

Whereas ASK is generally inferior to PSK schemes, QAM provides a solid alternative, as illustrated by Figure 7.6. QAM is partly similar to PSK as was already explained earlier, but next to phase modulation it also applies amplitude modulation. At low *M*'s, QAM can equal PSK, but at higher *M*'s the amplitude of the signal will be changed over a wide range of amplitudes. The following Figure A-19 illustrates examples of QAM constellations; these constellations can be changed depending on the application. In terms of power and bandwidth efficiency, QAM wins over PSK when *M* is higher than 4 (thus 16 and up). System complexity is also not a particular barrier. The only significant potential downside is its non-constant envelope.

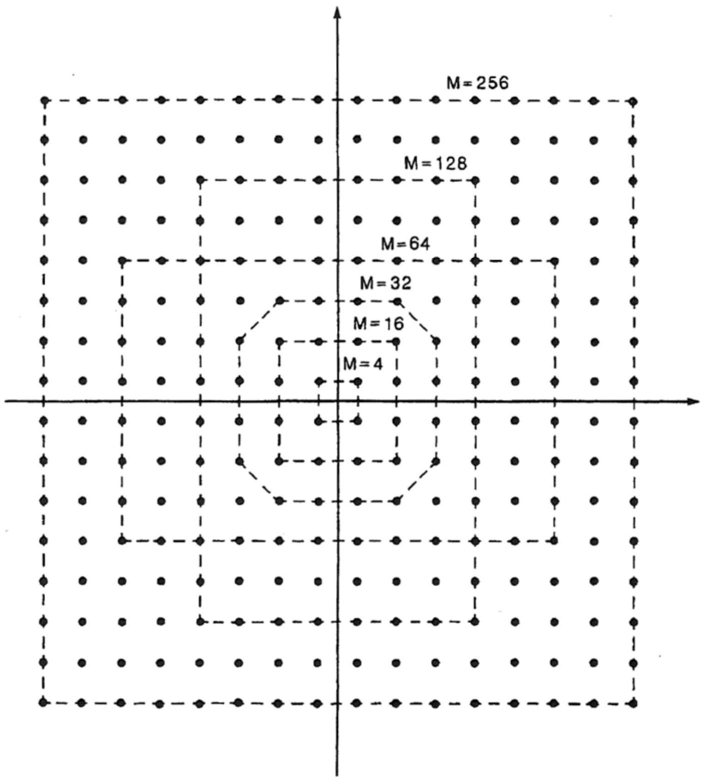


Figure A-19: Possible QAM constellations [SLR 0552]

A.7 Trellis Coded Modulation (TCM)

Trellis coded modulation is an example of a set of combined error control techniques, where channel coding and modulation are carried out jointly. The main driver is the preservation of bandwidth, being a scarce resource in worldwide communication. As further assessed in section 7.8, FEC techniques increase the bandwidth of a signal, thus decrease bandwidth efficiency at the gain of power efficiency. The idea behind TCM is to introduce FEC coding in the modulation process, in order to retain similar bandwidth. In order for this reasoning to apply, TCM can be combined with both AM and PM signals. The idea is simply to increase the modulation scheme from for example 4PSK to 8PSK, and to use the extra bit per symbol for coding. The bandwidth will then stay the same. The signal power requirement per bit will be increased by the introduction of a higher order modulation scheme, but the introduction of coding will decrease the requirement per information bit to a lower value than originally required. This comes of course the cost of more system complexity.

A.8 Orthogonal Frequency Division Multiplexing (ODFM)

A modulation technique worth mentioning due to its widespread application nowadays is that of orthogonal frequency division multiplexing. This technique implies dividing a data stream and modulating multiple carriers, each using a bandwidth efficient modulation technique as QAM or PSK. The total frequency bandwidth used is increased by this logically, but so are the data rates achieved. Nevertheless this technique primarily requires more power too, next to extra complexity and as such is no serious option for a mission such as Delfi-n3Xt, or satellite communication in general. ODFM is a subtype of frequency division multiple access (FDMA) which allows for multiple data streams to concur simultaneously, by using different carrier frequencies.

A.9 Spread spectrum techniques

By using a code to postmodulate data, multiple data streams can be sent over the same bandwidth. These spread spectrum techniques are actually not a type of modulation, but are generally presented as part of modulation. In general the techniques can be applied if the required transmission power for one data transmission in a certain bandwidth is too high, due to regulations for example. Spread signal techniques decrease the power spectral density of a signal, thereby resulting in an increased resistance to interference, a lower probability of detection and the possibility of spectrum sharing. If spread spectrum techniques are used to achieve the latter it is a form of code division multiple access (CDMA). Spread spectrum techniques can also be used to yield accurate range (distance) or range rate (velocity) measurements. A detailed discussion can be found in [SLR 0166].

The basic working is as follows. A signal spreader is introduced after the modulator. When the signal is spread, the energy of the signal remains the same, but the bandwidth of the signal is enlarged. The power spectral density therefore decreases. At reception, the consequent signal-to-noise ratio can be well below 1. As noise is uncorrelated, and the receiver does have information of the received signal, this information can be used to correlate the received signal. When the signal is correlated, despreading occurs, and the original signal is retrieved. The level of this signal should consequently again be above the noise level.

Spread spectrum techniques do not increase the achievable data rate. With the possible exception of the improved capability of bandwidth sharing, the advantages of spread spectrum techniques are consequently not directly applicable to small university satellite missions, but are more of interest for military applications.

Appendix B Block codes

As explained in section 7.3, this appendix introduces the different code classes as demonstrated in Figure 7.7. Attention is paid to the different advantageous and disadvantageous properties, such as bandwidth performance and complexity. Power performance of the basic code classes has already been introduced in section 7.3. The four classes, NRZ codes, RZ codes, Pseudoternary codes and Biphase codes are introduced in sections B.1 through B.4.

Consequently, a line code going by the name of Delay modulation or Miller code, strictly part of the biphase group but with distinct characteristics, is discussed in subsection B.5. Finally more advanced line codes classes are introduced in subsections B.6 and B.7, being the substitution codes and block codes.

B.1 NRZ codes

The (bipolar) NRZ codes, or nonreturn-to-zero codes use two levels, being $+A$ and $-A$. As presented above, the (unipolar) NRZ-L code is the usual code in digital logic as we all know. NRZ-M and NRZ-S are differential forms of NRZ-L; one bit (0 or 1) is defined as a state transition, whereas the other bit causes no transition. In case of NRZ-S, if the bit to be transmitted is a 0, the signal is transitioned from whatever state it is in to the other, whereas in sending a 1 no transition in signal takes place. Another name for differentiated NRZ is NRZ-I, the name which was used to represent this line code in its use for Delfi-C^3. Consequently it depends on convention whether a 1 or 0 creates a transition. Nevertheless, the term NRZ-I is often used.

NRZ codes have no dc component, assuming an equiprobable binary data sequence. If however a larger amount of just 1's or just 0's is transmitted, a dc component will be introduced. This intuitively makes sense as the total averaged signal becomes slightly positive or negative. More importantly, a small stream of consequent 1's or 0's can directly saturate capacitors in this way.

Differential codes as suggested above have one main advantage; the same original bit stream can be recovered, regardless of the polarity of the transmitted encoded signal. See [SLR 0552] for a detailed explanation of this phenomenon, but by symbolizing bits relatively instead of absolutely no knowledge is required on the polarity of the signal. However, as Figure 7.8 suggests, this comes at the cost of reduced *BER* performance. This is caused by the dependence of one bit on previous bits. As also quantatively stated in [SLR 0316], the required E_b/N_0 for a similar *BER* of 10^{-5} is 10.3, or 0.8 dB more. A further disadvantage of differential NRZ coding is a slight increase of complexity, but this is minimal as the differential coding can be integrated in software; basically the bit sequence is just changed.

Finally, NRZ codes do not automatically yield sufficient bit timing or symbol synchronization, as a string of all 0's, all 1's or alternating 1's and 0's yields no transitions, depending on the type of NRZ coding. This can cause synchronization to be lost, which can be prevented by precoding the data to remove these strings. As the AX.25 protocol (see section 7.6) is prepared to use NRZ-S (there simply labeled NRZ-I) coding, it indeed removes long strings of 1's to allow for maintaining bit synchronization through bit stuffing.

The (bipolar) NRZ codes can be made unipolar by changing the lower $-A$ level to 0. This creates a dc component at a level of $A/2$, assuming an equal number of 0's and 1's.

The power spectral densities or PSDs of both (bi)polar and unipolar NRZ are illustrated in Figure B-1 below (left), as well the bandwidth efficiencies (right). The significance of PSDs has been introduced in section 7.2. Per figure a number of aspects can be noted, which are preceded by a small explanation of what the figures exactly demonstrate.

Power spectral densities: normalized to have a maximum at Ψ_s = 1, these graphs indicate how much power is present in a 1-second average pulse at any frequency.

- The zero frequency indicates a constant total power level in one bit or pulse. In the case of bipolar NRZ it is easy to visualize that total signal strength per bit is constant and nonzero, independent of a 0 or 1 is being transmitted. Only the sign changes. This nonzero signal strength, at zero frequency indicates a dc component. Polar NRZ then yields no net dc component if 0 and 1 bits alternate each other, as mentioned above. Unipolar NRZ on the contrary, although slightly hard to see from the figure, has a power impulse of value 0.5 at zero frequency, even if symbols alternate nicely. In other words, the signal always has a net positive power content.
- For both polar and unipolar NRZ it can be seen that the first null is reached at fT = 1; In other words B_{null} = 1 R_b. It can therefore be said that the signal has no spectral energy at the clock frequency.

Fractional out-of-band bandwidths: these graphs demonstrate at any value of BT, how much percent of the power in the signal is out-of-band. As mentioned in the previous section, reading the value at -10 dB on the vertical axis gives the 90% energy bandwidth (log relation). $BT = B/R_b$; an approximate 0.85 at -10 dB indicates $B_{90\%}$ at 0.85 R_b. Furthermore, as can be verified by calculation, $B_{99\%} \approx 10\ R_b$.

What is possibly more interesting about the graphs shown in Figure B-1 (a), is that they are identical to the PSD of BPSK. In fact, as explained before, bipolar NRZ coding is required to modulate BPSK. As the abrupt changes are indeed part of the NRZ code, and not the IF or RF signal mixed with the bit sequence, the code indeed gives the BPSK modulation its frequency characteristics.

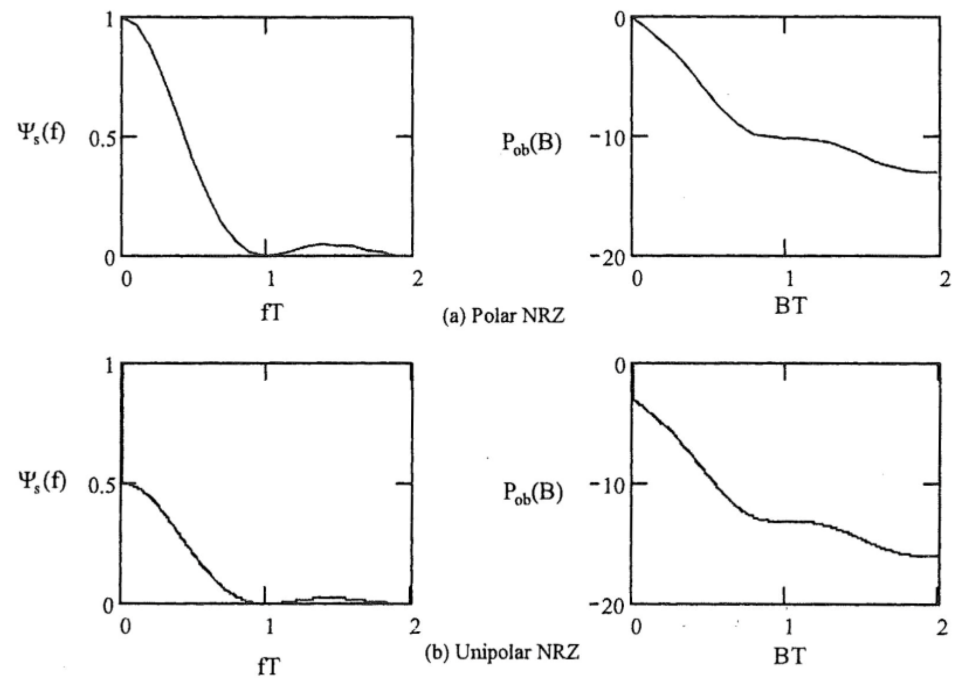


Figure B-1: PSD (left) and out-of-band power (right) of (a) Polar NRZ and (b) Unipolar NRZ [SLR 0552]

B.2 RZ codes

The RZ codes, or return-to-zero codes, introduce extra timing information in the signal, by adding more transitions. The amplitude indicating a bit is simply returned to 0 after half the bit period. Unipolar RZ coding is also possible, where the 0 bits yield no transition with a continuous zero amplitude. The latter has again no bit transitions for long strings of 0's and introduces a certain dc component at the cost of extra complexity, thus is not reviewed further.

The bipolar RZ code guarantees a transition per bit, which comes at no cost in *BER* performance, but only at the cost of increased bandwidth and complexity. A third amplitude level is introduced and the bandwidth is doubled, as can be seen in Figure B-2. In conclusion:

Power spectral density: the shape is the same as that of Polar NRZ, but with half the power and a null bandwidth that is twice as large: $B_{null} = 2\ R_b$. This is due to the fact that the amount of abrupt changes in the bit sequence has doubled. A dc component can arise in a non-perfectly alternating bit sequence.

Fractional out-of-band bandwidth: also more or less twice that of NRZ codes; $B_{90\%} \approx 1.7\ R_b$ and $B_{99\%} \approx 22\ R_b$.

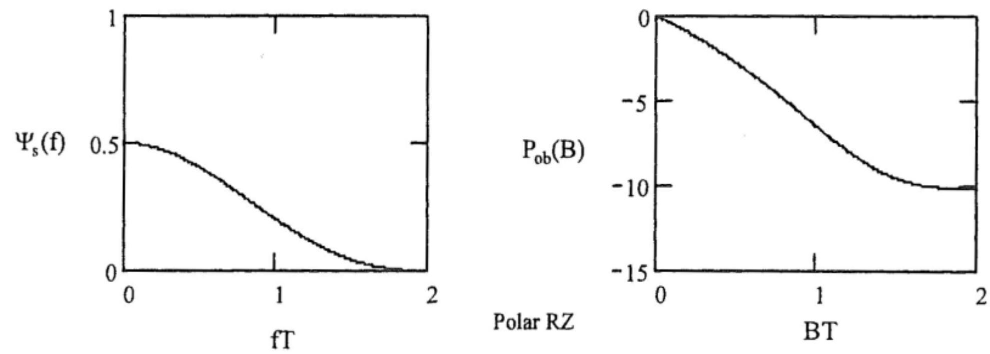


Figure B-2: PSD (left) and out-of-band power (right) of Polar RZ [SLR 0552]

B.3 Pseudoternary codes

Similar to RZ codes, these codes apply three levels, ±*A* and 0. They are also called bipolar codes in the telecommunication industry. AMI codes are part of this group. AMI stands for alternative mark inversion, indicating that consecutive bits, per convention 1's, have opposite signal amplitude. As one can imagine, this can yield RZ and NRZ versions. What this code introduces is a form of error detection, by adding a dependence on previous bits.

These codes also have no dc component. The 0 bits yield no signal strength, thus neither a dc component, and 1 bits are per definition alternated so that an overall dc component is excluded. Nevertheless, little bit transitions might introduce synchronization difficulties. Also, three signal levels introduce additional system complexity and *BER* performance is worse than for simpler line codes.

Next to AMI codes there are the dicodes, which are basically differential AMI codes. This, as expected, comes at the cost of more *BER* performance as illustrated by Figure 7.8. The power spectral densities of the pseudoternary RZ and NRZ codes as well as their out-of-band power are illustrated by Figure B-3 below.

Power spectral densities: an interesting phenomenon can be observed in the graphs on the left; the frequency of maximum power has increased with respect to 0; this indeed illustrates the fact that this signal inherently has no dc component. B_{null} still equals 1 R_b for both the RZ and NRZ types, just as in case of NRZ codes. In case of AMI-NRZ and its differential version, slightly more power is focused on their center frequency; this causes them to have slightly better *BER* performance than their RZ counterparts.

Fractional out-of-band bandwidths: in case of the RZ codes: $B_{90\%} \approx 1.71\ R_b$ and $B_{99\%} \approx 20\ R_b$. For the NRZ codes, as expected due to the higher peak, slightly lower values: $B_{90\%} \approx 1.53\ R_b$ and $B_{99\%} \approx 15\ R_b$.

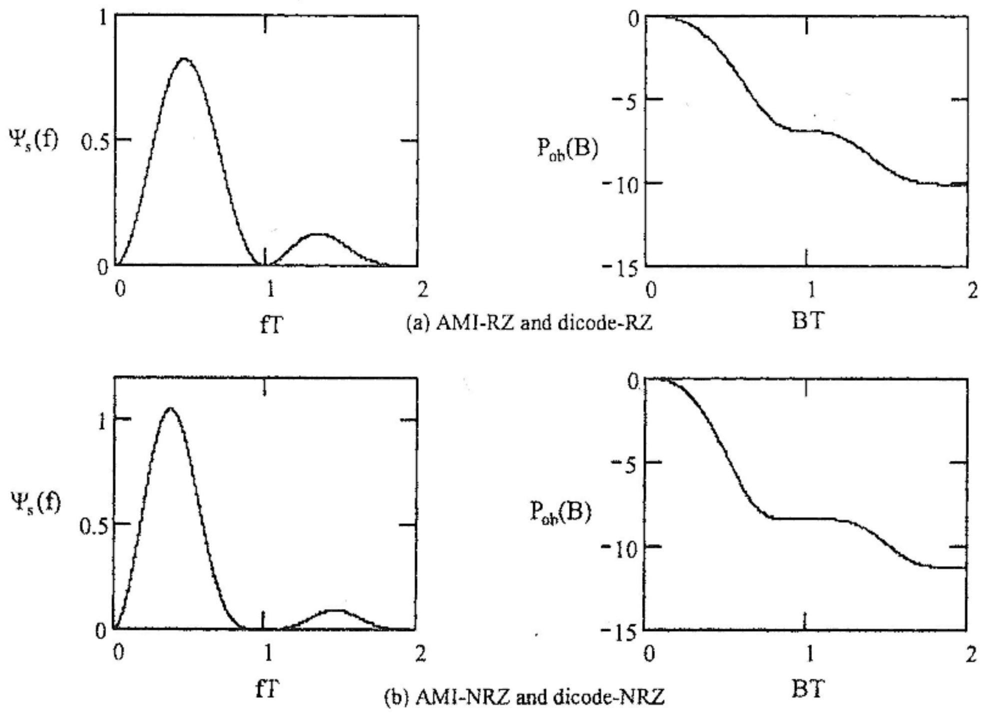


Figure B-3: PSD (left) and out-of-band power (right) of pseudoternary codes [SLR 0552]

B.4 Biphase codes (including Manchester)

Biphase codes apply half-period pulses with possible different phases. The Bi-Φ-L (which stands for bi-phase level) format is better known as *Manchester* code. A 1 is indicated by a higher level first half bit, followed by a lower level second half bit. The 0 is coded vice versa. The pulse shapes can also be exchanged. Other variants such as Bi-Φ-M and Bi-Φ-S ensure a bit transition at the start and ends of a bit, but thereby do not guarantee a bit transition during the bit. This increases coding complexity somewhat but most notably comes at the cost of 3 dB decreased *BER* performance. Finally conditioned Bi-Φ-L is basically differential Manchester coding.

As with RZ codes, used bandwidth is doubled as the two levels per bit are required. This is illustrated by Figure B-4. However, biphase codes only require two different signal levels, to achieve similar *BER* performance. Therefore the impact on system complexity is less. The consequence is that depending on the exact biphase code a transition at the start/end *and* middle of a bit is not guaranteed; offering slightly less symbol synchronization possibilities. Nevertheless, as illustrated by its widespread application this extra synchronization is usually not required. Manchester coding is used in Ethernet, as well as of course the uplink of Delfi-C[3], with success. In the latter case also NRZ-I coding (of type NRZ-S) is applied, effectively applying differentially encoded Manchester; however in the opposite sense as that illustrated in Figure 7.7 as Conditioned Bi-Φ-L. Finally, Manchester coding as well as its differential version has no dc component, which is guaranteed as every bits contains equals parts positive as negative amplitude. As is explained below, Manchester is actually part of a larger block code class, called mBnB codes.

DMI codes or Differential Mode Inversion codes, similarly classified as block codes being further reviewed below, actually have the same spectral performance as biphase codes. The codes add error correction capabilities to the biphase code characteristics; 1's are consequently mapped to a 00 and 11 of each half a bit period, whereas 0's are mapped to 01 and 10. The transition capabilities are consequently lowered with respect to for example Manchester, at the gain of some error detection capability.

Power spectral density: as with the AMI codes, the frequency of maximum power is shifted. The spectrum is however rather broad, so B_{null} = 2 R_b; as with RZ codes the number of transitions per bit have been doubled, forcing the bandwidth occupation to do the same.

Fractional out-of-band bandwidth: these values are quite broad as well, with $B_{90\%} \approx 3.05\ R_b$ and $B_{99\%} \approx 29\ R_b$.

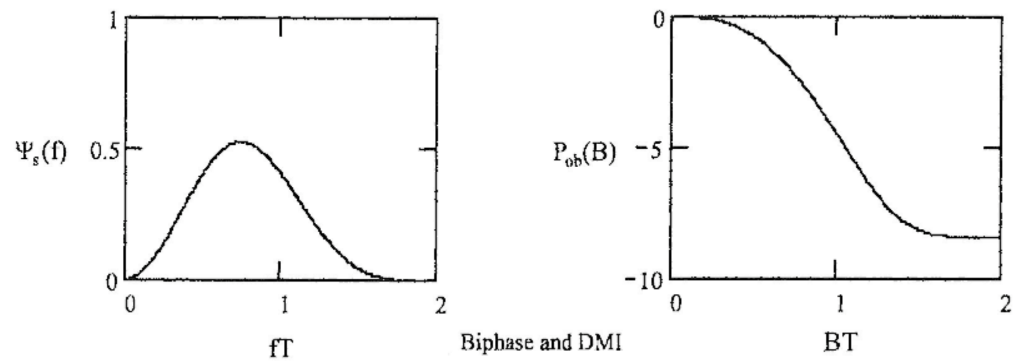


Figure B-4: PSD (left) and out-of-band power (right) of biphase and DMI codes [SLR 0552]

B.5 Delay Modulation

Whereas delay modulation or Miller code can strictly be seen as part of the biphase code group, as it has two phases per bit, if offers some unique behaviour. The representation for a 1 is alternated, being one-half $+A$ and one-half $-A$. A 0 yields no transition, unless it is followed by another 0. Miller codes give little beneficial properties in terms of bit synchronization, but result in a very small bandwidth and a very small dc component. This of course comes at the cost of complexity and non-optimal *BER* performance.

Power spectral density: again a shifted (and high) power peak, as with other biphase codes, but this time yielding a very narrow one. The main lobe bandwidth is only about 0.5 R_b. There is a small dc component introduced, its value depending on the specific momentary bit sequence. This however still renders it suitable for many applications without dc response.

Fractional out-of-band bandwidth: the power spectrum, however having a very narrow peak, converges to zero very slowly. This results in having only 76.4% of the power within 2 R_b and only 83.7% within 250 R_b.

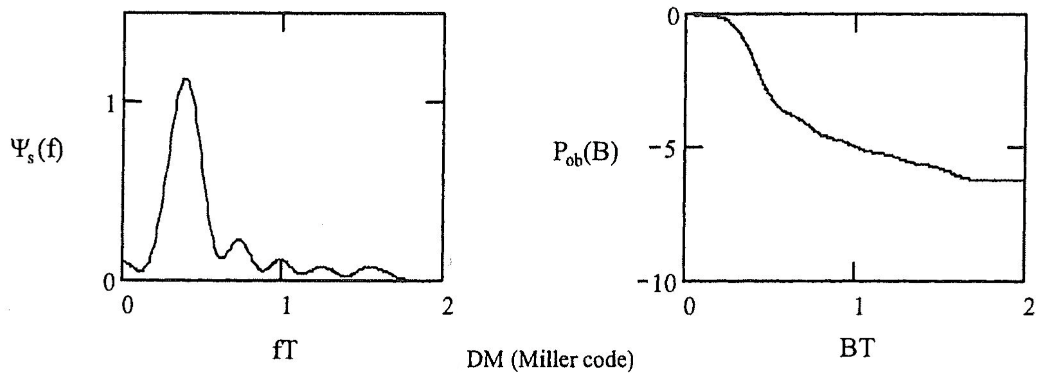


Figure B-5: PSD (left) and out-of-band power (right) of Delay Modulation [SLR 0552]

B.6 Substitution line codes

An improvement of AMI codes in terms of transitions, resulting in better symbol synchronization performance, is given by substitution line codes.

As has been shown above, in a number of line codes a long series of 0's or 1's causes no transitions to take place, complicating symbol synchronization. The idea behind substitution line codes is to replace a certain maximum string of 0's or 1's by a predefined code. The problem of with a predefined code of course is that this code should not occur elsewhere in the bit sequence. Substitution codes solve this problem by using an AMI code, and inserting intentional *bipolar violations*.

AMI codes make use of alternating symbols, indicating 1's. This feature allows for some error detection. If however two consecutive 1's are not represented by an alternating symbol, one speaks of a bipolar variation. This feature can next to error detection be used for more conveyance of information, as done through the use of substitution coding. Of course this lowers the error detection capability slightly, depending on the complexity of the substitution.

B.7 Block codes

Block codes are a general name for more advanced line codes that do not map bits to symbols (1 by 1), but map a number of bits onto a number of symbols. In fact, some of the line codes above can be seen as block codes; all RZ codes map on bit to two symbols, as well as the biphase codes.

Two general approaches can be distinguished:

- Additional symbols can be added; usually to yield better transition characteristics and a virtually zero dc component. This of course comes at the cost of increased bandwidth. The most advanced schemes are used for example optical transmissions, where only on-off signaling is possible (two levels) and bandwidth is relatively unimportant.
- More bits can be mapped to less symbols, usually 4 bits to three ternary symbols; this saves bandwidth. An increase in complexity of the coding alphabet can come at the gain of total *digital sum*, see below. It can however only by used in cases where more than two levels can be used to indicate symbols. These codes are less logically applied in combination with wireless modulation, as in effect they themselves introduce a sort of amplitude modulation.

A code classified as a block code and which adds symbols for increased functionality, as already introduced above is the DMI code. A version called CMI, Coded Mark Inversion, does not alternate the 1 code, and just maps it to 01 or 10. The codes, as well as biphase codes can be classified under the common header mBnB codes, where m bits are mapped to a higher number n.

An interesting code that removes the dc component and adds symbol transitions is the Carter code. It is an mB(m+1)B code; more precisely a 8B9B code. It looks at a block of 8 bits, and then either inverts the sequence, or not; in order to average out the total number of 0's and 1's sent (reducing *disparity* or *digital sum*). At the same time, this introduces transmissions as a sequence of 1's can not be followed by a sequence with a large number of 0's. The ninth bit per sequence is inserted to indicate a conversion.

The Griffiths code also removes a dc component and adds symbol transitions. It improves the characteristic of Carter code that 18 1's or 0's can follow each other. It works by mapping a certain bit pattern onto another pattern and it is usually of the form 3B4B or 5B6B. In fact, its 1B2B form is equal to Manchester. Higher-order codes introduce excessive complexity. The 3B4B Griffiths code uses only zero- and single-disparity words, limiting the maximum number of 1's or 0's to 4 in-a-row. The 5B6B Griffiths code applies only zero- and double-disparity words, limiting the maximum number of 1's or 0's to 6 in-a-row. Manchester, or 1B2B has a maximum of 2 equal symbols in-a-row.

A 2B3B dc-constrained code, as the name implies, does not reduce the dc-component completely but constrains it. The latter approach yields bit redundancy to be used for error correction. It yields a constant dc component of -1/3. It has an average number of changes per level of between 1 and 2, and a maximum number of symbols without change of 7.

Finally mB1C and DmB1M codes add a bit per m bits to incur a transition, by making sure the added bit is different from the last bits of the block. mB1C looks only at the last block bit, whereas DmB1M looks at the entire block and thereby adds a certain error detection capability.

Appendix B

COMMS – STX System Design

SLR 0387

The entire document has been written by the author.

COMMS – STX System Design

Description: Document establishing modules and design options for the STX system

Subsystem(s) involved:	ADCS	CDHS	COMMS	EPS	MechS	STS	TCS	ITRX	MPS	$T^3\mu PS$	SDM	Splash	GSE	GSN	Launch
	X		X										X	X	

Revision Record and Authorization

Issue	Date	Author / Editor	Reviewer checked	PM approved	Affected Section(s)	Description of change
0.1	XX-12-2008	A. Tindemans			All	First issue, functional definition
1.0	10-4-2009	A. Tindemans			All	Entire document (re)written
2.0	16-11-2009	A. Tindemans			All	Total update; mainly operations
3.0	13-06-2010	A. Tindemans			1 2 3 & 4 5-11 12 13 14 15 16	STX goals (re)defined Requirements updated Chapters switched and updated CI chapters updated New chapter Old chapter 12 Component diagram updated and added integration options Old chapter 13 Power consumptions updated Old chapter 14 Next steps added

Table of Contents

1 Introduction

The STX is part of the Delfi-n3Xt communication system, or COMMS. The COMMS is introduced in [SLR 0014] along with its top-level design. The latter document therefore introduces the components that make up the COMMS. They can be said to consist of a primary communication system, consisting of PTRX, ITRX and a respective antenna system, and a secondary communication system, consisting of two *technology demonstrations* and their respective antenna system parts. One of the demonstrations is OLFAR, the other is the S-band transmitter, or shortened: ***STX***.

The STX is an experimental transmitter that is designed to facilitate high data rate one-way communication from the satellite to the Earth, in order to increase the capacity of the Delfi satellite bus. As UHF and VHF frequencies do not generally allow higher data rates than 9.6 kb/s to be achieved on the radio amateur frequency bands, as a minimum performance standard the STX data rate is set to **≥9.6 kb/s**.

Since the conception of the STX during the mission definition of Delfi-n3Xt almost two years ago, two relevant things have changed:

- **Full attitude control is no longer guaranteed**
  Whereas the ADCS is the subsystem where significant advances are to be made, its development is no longer allowed to delay the program and significantly increase the operational risks. A consequence of this choice is also that instantaneously generated power is decreased from 10-18 W to about 3.5 W, although a battery is available.
- **Student resources have turned out to be scarce**
  A long period of little students has been combined with a heavy under-anticipated work-load on the primary radio system, the PTRX. No significant lower-level work on the STX has been performed yet.

The result of these changes has been that the inclusion of the STX on Delfi-n3Xt has been questioned. Nevertheless, during a meeting with staff, students and relevant partners it has been decided that an effort should be made to fly the STX, in order to:

- **Have a proven S-band transmitter for future missions**
- **Increase the novelty and appearance of Delfi-n3Xt with respect to Delfi-C^3**

The matter of requesting a transmission frequency for a system that might in the end not be flown due to limited resources was also discussed, but it was decided the frequency coordination process should be commenced based on the current specifications. The arguments:

- **No regulatory or moral obligations**
  The radio amateur community poses no repercussions for not being able to deliver; project changes due to for example limited resources are acceptable. This has later been confirmed by Graham Shirville, who heads the frequency allocation of the AMSAT community.
- **A substitute can possibly be flown**
  If the STX itself cannot be finished, a substitute S-band transmitter can potentially be bought and accustomed to Delfi-n3Xt, such as the TXS of ISIS B.V.

The above facts lead to the design philosophy of the STX, as presented on the next page.

The design philosophy of the STX can be stated by means of four *design goals*:

- **No pointing shall be required for the technology demonstration**
  Pointing would allow much higher data rates to be achieved. However, if the ADCS is depended on by the STX, the chances of halving a fully operational STX would be halved (simplistically). As such, a more or less omnidirectional antenna pattern shall be implemented, for as far as possible.
- **The STX shall have power consumption no higher than the PTRX (~1.75 W maximum)**
  Power is (very) limited on Delfi-n3Xt; there is little power available on top of nominal functionalities. Although a battery is available on Delfi-n3Xt, the STX would be little flexible if it depends on the latter. At that, it would increase risks for a proper technology demonstration. If the STX consumes less power than the PTRX, the PTRX transmitter can be switched off, to switch on the STX. Ideally, this would replace identical transmission functionalities. Currently, some 1.75 W is allocated to the PTRX. This value might be lowered somewhat by adjusting the gain of the amplifier if the final power budget demands for it.
- **COTS components shall be used wherever possible**
  The use of COTS components logically decreases required (project) resources.
- **All components in the design should be as modular as possible**
  An STX with a *separated processor, data storage, modulator and power amplifier* amongst others can easily be updated if software, power or complexity constraints change (for example in a future mission or later in the Delfi-n3Xt mission).

The actual testability of the STX will depend on the antenna directivity pattern and consequently the attitude of Delfi-n3Xt during a pass. Assuming a (slowly) rotating satellite, STX testing will always be possible.

As partly introduced above, design of systems related to the STX, such as the STX top-level design, has been established in other documents. These documents are:

- [SLR 0014]: COMMS – Top Level Design of the Communication Subsystem
- [SLR 0036}: COMMS – Antenna System Design

The first document introduces not only the STX and the reasons for its inclusion, but also establishes the characteristics of the S-band communication link, taking into account ground segment as well as STX design, in turn introduced in this document. More specifically, the transmission frequencies, modulation scheme, FEC approach, data rate approach and data protocol have been specified in the top level document. Finally, also the anticipated operations of the STX and its activation have been discussed.

The second document establishes the antenna system required for the STX to function. In order to approach an omnidirectional gain pattern, the use of two S-band patch antennas on opposite sides of Delfi-n3Xt is suggested.

Document structure

The objective of the current document is to specify the design of the STX in terms of its low-level components. The required subfunctionalities shall be derived taking as an input the requirements on the STX, its functional description and physical interfaces with Delfi-n3Xt systems. These three aspects are presented in chapters 2, 3 and 4 respectively. Chapter 5 then specifies the subfunctionalities of the STX and from this derives a number of concrete CIs or configuration items.

The lowest level configuration items of the STX consequently form the guiding line for the main body of the report; each of the CI chapters includes a discussion on its requirements, design options, planning and an introduction to required theory. Design options are judged on ease of implementation and thus project impact and risk as well as availability, achievable functionality and added value in terms of innovation and modularity for example. As the Delfi-n3Xt project has already advanced to a stage where attention should be paid to detailed design and as the STX is of low actual mission priority, an important focus lies on the inclusion of available COTS components.

Chapters 6 through 11 then describe these 6 lower level CIs that make up the STX. At the end of each chapter on a CI a concrete list of work yet to be performed is given with some priorities indicated.

Chapter 12 consequently discusses two design options that have been coined in the relation to the STX, being the *variable voltage bus* and a *beacon mode*. In fact, it is explained that these two design options have in fact resulted from the (unfulfilled) desire to have a high efficiency switching amplifier on the STX. This chapter therefore discusses the current position of these two design options.

After all CIs have been discussed, the suggested component lay-out or low-level design of the STX is presented in chapter 13, and it is shown that all design goals as stated above are achieved. This same chapter discusses electrical integration, as well as physical integration, using a PCB or a more exotic choice. Chapter 14 discusses Delfi-n3Xt volume, mass and power budgets.

Finally, all work left to be done in order to have a working Delfi-n3Xt STX and the suggested next steps are presented in chapter 15.

Additionally, Appendix A discusses the possible application of an FPGA within the STX, and argues why this option is not pursued.

2 Requirements

This chapter lists the requirements as specified for the STX experiment. The requirements have been presented in [SLR 0014], introducing also their origins and rationale.

The STX requirements are composed of a set top-level requirements, applicable to all lower level systems, and a number of specific STX requirements. These two sets of requirements are shown in Table 2-1 and Table 2-2 respectively.

Table 2-1: Generally applicable STX requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT-F.03	The satellite shall be able to switch off any transmitter on the satellite
CONSTRAINT	SAT-C.01	All satellite systems shall comply with the mass budget, as given in [SLR 0018]
CONSTRAINT	SAT-C.02	All satellite systems shall comply with the volume budget, as given in [SLR 0303]
CONSTRAINT	SAT-C.03	All satellite systems shall comply with the power budget, as given in [SLR 0017]
CONSTRAINT	SAT-C.04	All satellite systems shall comply with the data budget, as given in [SLR 0282]
CONSTRAINT	SAT-C.05	All satellite systems shall comply with power and data bus interfaces, as specified in [SLR 0263]
CONSTRAINT	SAT-C.06	All satellite systems shall be able to withstand the launch environment
CONSTRAINT	SAT-C.07	All satellite systems shall be able to withstand the space environment
FUNCTIONAL	SAT.2-F.02	All satellite bus systems shall generate housekeeping data when of interest to satellite operation
CONSTRAINT	SAT.2-C.01	All satellite bus systems shall adhere to reliability standards, as specified in [SLR 0263]
PERFORMANCE	SAT.2.3-P.01	All transmitters shall provide a maximum permitted power level for spurious emissions in dBc according to -43 - 10 log (P), where P is the transmitted power
PERFORMANCE	SAT.2.3-P.02	The bit error rate of all digital data transmission links shall be designed to be at most 10^-5
CONSTRAINT	SAT.2.3-C.02	Any data transmitted to and from the satellite other than telecommands shall not be encrypted in any way
CONSTRAINT	SAT.2.3-C.05	All UHF and VHF antenna connections and transmission lines on the satellite will have an impedance of 50 ohm

Table 2-2: STX-specific requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.1-F.01	The STX shall be able to transmit telemetry (payload and housekeeping data)
FUNCTIONAL	SAT.2.3.1-F.02	The STX shall have a buffer to store data that must be transmitted
FUNCTIONAL	SAT.2.3.1-F.03	The STX shall have an interface with the Antenna System to forward an analogue data signal
FUNCTIONAL	SAT.2.3.1-F.04	The STX shall allow for transmission of data with a data rate of at least 9600 bits/s
FUNCTIONAL	SAT.2.3.1-F.05	The STX Data Storage must have a size that represents at least 24 hours of all unique payload data and housekeeping data
INTERFACE	SAT.2.3.1-I.01	The STX is assumed to transmit the exact same raw data as the PTRX, with raw data indicating data excluding transmission protocol specific data
CONSTRAINT	SAT.2.3.1-C.01	The STX transmission signal shall be within the S-band frequency band between 2400 and 2450 MHz

In chapter 5 the Configuration Items (CIs) are established; these visualize the different hardware components within the STX system. Per CI a set of (low-level) requirements will be specified. A number of requirements mentioned above are generally applicable to all lower level CIs; these are indicated in Table 2-3 below.

Table 2-3: Higher level requirements applicable to all STX CIs

Level	Requirement(s)
Satellite	SAT-C.01, SAT-C.02, SAT-C.03, SAT-C.04, SAT-C.05, SAT-C.06, SAT-C.07
Satellite Bus	SAT.2-C.01
COMMS	SAT.2.3-P.02
STX	SAT.2.3.2-P.01

Requirement SAT.2.3-P.02, stating the maximum allowable *BER*, strictly does not apply to all components, but most. It should be taken along in all decisions related to transmission, being modulation, filtering and amplifying.

3 Functional description and modes

According to the general and STX-specific requirements, the STX should be able to fulfil the following functionalities:

- Transmit telemetry (TM)
- Generate housekeeping data
- Store data that is to be transmitted

Whenever the STX board is active, housekeeping data should be generated. Data should be stored constantly, as Delfi-n3Xt has no other form of data storage and thus generated TM frames should be forwarded to the STX continuously. The high power requirements of telemetry transmitting however prohibit there to be a single operational mode; transmission cannot take place continuously.

As such, two operational modes should exist:

- **TX OFF mode**

  This is the default mode. The STX is switched on, but its transmission is switched off. In this way it gathers all TM data that is transmitted to PTRX and/or ITRX. Also, this mode provides the required functionality according to regulations (requirement SAT-F.03), stating that signals transmissions should be able to be switched off.
- **TM mode**

  When power is available, and likely during a Delft pass with proper satellite attitude, this mode can be activated. Telemetry is transmitted, read out from the data buffer. As is further explained in [SLR 0014], *submodes* should be defined with different data rates and possibly different data content. The suggested data rates that would belong to these submodes are also introduced in [SLR 0014].

As is discussed in chapter 12, a *beacon mode* had previously been defined for the STX as well. However, as the variable voltage bus is no longer assumed to be a useful feature as far as the STX is concerned (as is discussed in the same chapter), there is no functional use for a separate beacon mode anymore.

4 Interfaces

This chapter quickly assesses all the interfaces the STX system has with other subsystems. Next to the payloads there can be said to be seven subsystems in the space segment of the Delfi mission. These are the STS, MechS, TCS, EPS, CDHS, ADCS and COMMS. For more information on these subsystems one is referred to corresponding technical notes. The STX is of course part of the COMMS.

While the COMMS consists of multiple radios, there are no active connections betweens them. The STS of course provides structure, while the TCS provides thermal control. The CDHS and EPS have definite and active interfaces with the STX as demonstrated in Figure 4.1.

In fact, the illustrated interface diagram shown below is outdated. The final design of the Standard System Bus (including power and data interfaces) is still to be determined.

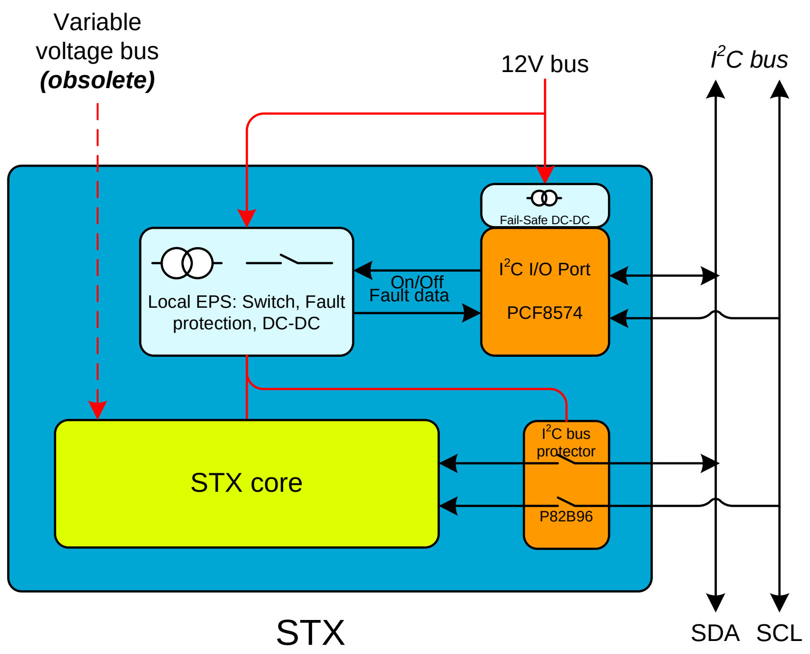


Figure 4.1: STX interface with the CDHS and EPS [adapted from SLR 0029]

The figure above shows the connections of the *Standard System Bus*, being the bus that takes care of power and data interfaces. It characteristics are defined in [SLR 0263], as stipulated by requirement SAT-C.05. SDA and SCL stand for serial data line and serial clock line respectively. All components shown are to be integrated on the STX PCB. Next to the STX core there is what is defined as the *local EPS* and *local CDHS*; indicated by light blue and orange boxes respectively.

The basic working of the structure shown in the figure is as follows. The I/O Port is commanded by the OBC to switch the STX system on or off and therefore communicates with the OCB. The I/O Port consequently controls the power switch (including fault protection and DC/DC conversion) and can receive fault data from the local EPS system. When power is switched on the I^2C bus protector opens up communication between the STX core and the OBC. The power interface connects to the regular 12V bus, which supplies power to all STX support components and the STX core.

In the original design a connection to the variable voltage bus was envisioned, but this will no longer be pursued as is further explained in chapter 12.

The STX core currently has four physical links as can be seen in Figure 4.1, two of which are the different voltage buses and the other two are to the OBC. The clock line just provides the clock signal, whereas the STX is required to communicate to the OBC over the data line. This means the STX core must follow the I^2C protocols in terms of commanding and addressing in its communication with the OBC. Also the STX core is supposed to put housekeeping data such as amplifier temperature and used current on the bus (see also chapter 8 on the housekeeping sensors).

A top-level requirement has been stipulated to shut down any transmitter in case necessary (SAT-F.03). As with other radios and subsystems, the proposed implementation of local CDHS/EPS means that the STX transmitter can be shut off by shutting down the entire STX board. This can be done by the I^2C I/O port, commanded by the OBC. However, this would also shut down the *data storage* capability. Therefore the TX OFF mode has been introduced in the previous chapter.

5 Defining the CIs

In order to develop a hardware and consequently work package architecture this chapter first establishes the subfunctionalities of the STX required to perform its main functionalities presented in chapter 3. This is covered in section 5.1. Consequently these subfunctionalities are related to discrete hardware packages and attention is paid to the interfaces between multiple packages in section 5.2. Finally, the CI tree for the STX is given in section 5.3.

5.1 Subfunctionalities of the STX system

In order to transmit TM in case of the STX, a number of subfunctionalities are required. General transmission techniques and functionalities have been introduced in [SLR 0014], including also the superheterodyne principle again applied here. In case of the STX, data storage is also required for correct operation. The resulting functional flow diagram required for STX data transmission is shown in Figure 5.1. This chain is more or less applicable to any transmitter.

The functionality of gathering housekeeping data is not reflected by the subfunctionalities shown below, as it is not part of the transmission flow. This functionality is again taken into account when defining the CIs in section 5.3.

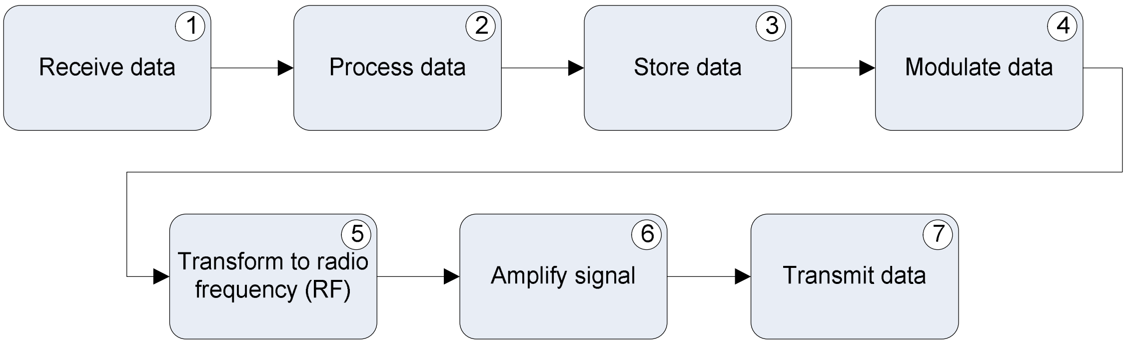


Figure 5.1: STX functional flow diagram

The functions as described in Figure 5.1 are further explained below:

1) ***Receive data***: Data is sent by the OBC over the I^2C bus, and as such the STX should be able to communicate over this bus correctly. This has implications on both communication protocols and data rate.
2) ***Process data***: The data received should be digitally prepared to be transmitted at this stage. This includes putting the data in correct data frames, adding error correction and possibly applying data compression.
3) ***Store data***: Data storage as argued is required due to the possible high transmission speeds and the limited TM data generation speeds within Delfi-n3Xt. The application of data framing, error correction and data compression needs some form of memory too; large data storage allows data to be pre-processed and stored before actual transmission.
4) ***Modulate data***: Modulation is required to impress data on a signal used for wireless transmission. The modulation schemes optional for the STX have been commented on in [SLR 0014].
5) ***Transform to RF***: Due to physical constraints regarding signal processing in terms of signal power, technological complexity and thus also cost it is better to perform work on a baseband signal, then to perform signal processing at transmission frequency. Therefore the frequency of the signal is

usually up-converted to the radio frequency (RF) after digital data processing and modulation, according to the superheterodyne principle.

6) ***Amplify signal***: The higher the signal power level, the more losses are incurred during signal processing. Therefore it is better to work with voltages or currents at low power for as long as possible, and to have serious power dissipation in the antenna and power amplifier only. The signal of low power level is then amplified for transmission at this stage.
7) ***Transmit signal***: After amplification the signal is ready for transmission, as exemplified by the last required functionality. The element used for transmission is per definition an antenna. In case multiple antennas or even multiple feeds per antenna are to be used proper phase difference should be introduced. Phasing is important to achieve a certain antenna gain pattern and signal polarization.

5.2 Establishing hardware packages

Since the functionalities introduced in the previous section are different in nature, different pieces of hardware will take care of these functions. Therefore this section will sketch the STX in terms of hardware packages and will link the functionalities to hardware.

Figure 5.2 demonstrates the hardware components required to fulfil the functionalities as shown in Figure 5.1. The functionalities in Figure 5.1 are numbered, and as such the components or combinations of components in Figure 5.2 are indicated by numbers to represent the respective functionalities they can fulfil.

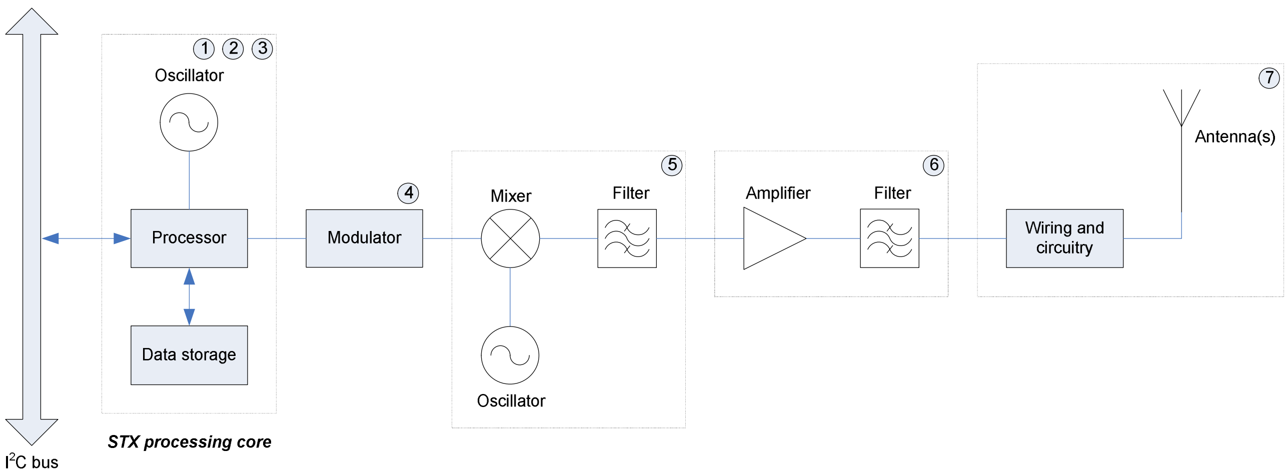


Figure 5.2: STX hardware diagram

A step-wise description:

- **Processing core**: Once the STX processing core (as indicated in the figure) is powered and thus switched on as discussed in chapter 4 on interfaces, data can be exchanged between the STX core and the OBC via the I^2C bus. Some form of data processor will be necessary for this function. In other words the processor will take care of receiving the data, but will also be in charge of processing the data. In order to perform the data processing subfunctionalities (1, 2 and 3) as described above and in more detail in chapter 6 some form of memory is necessary, and an oscillator to generate a clock signal required to perform calculations.
- **Modulator**: The modulator will logically take care of the modulation functionality.
- **Frequency up-converter**: A combination of an oscillator, a mixer and a filter can take care of transforming the signal frequency. A filter is necessary after mixing in any case to remove frequency images introduced in the mixing process (commonly called an *image-rejection filter*). The mixer can be replaced by a more complicated phase locked loop (PLL) scheme. Nevertheless, the general approach is shown here.

- **Power amplifier**: The amplifier logically amplifies the signal. It is accompanied by a filter to limit the final transmission bandwidth, which might or might not be required. As multiple amplifiers can exist along the chain, but only one will actually provide the required transmission power, this one is called the *power amplifier*.
- **Antenna system**: After the signal is amplified it is to be passed on to the antenna system. In the figure the components to be used for transmission are labelled 'Wiring and circuitry' and 'Antenna(s)'. The signal will leave the STX configuration item (CI) and will be forwarded to the S-band Antenna System. As will be shown in this document, the additional 'circuitry' is in fact not required.

The components in Figure 5.2 are described as usual discrete components. The oscillator can in fact also be an integrated part of the processor, as can the data storage. The former is generally not done due to the need for highly stable (crystal) oscillators in space. The latter can be done, although a relatively large amount of memory is required in the STX. A larger amount of components could even be integrated into a field programmable gate array (FPGA). At the moment this does not seem feasible for Delfi-n3Xt; this topic is covered in Appendix A.

5.3 Determination of configuration items (CIs)

In the Delfi-n3Xt project use is made of a configuration item tree, or CI tree based on hardware [SLR 0325]. The highest-level CI is the Delfi-n3Xt Mission, with the Delfi-n3Xt Satellite at one level lower, with the subsystems (satellite bus) and payloads at the next. The STX itself is a CI of the COMMS branch.

The CI tree for the STX can be derived using the hardware diagram as pictured in Figure 5.2.

The CI tree is inserted as Figure 5.3; the entire CI tree can be found in [SLR 0325].

For clarity in function a division has been made between the *transmission and support parts of the STX*,

- **STX TX**: This branch takes a digital data stream and outputs an amplified analogue signal. Subfunctionalities 4, 5 and 6 from Figure 5.1 are covered by the Modulator, Frequency Up-converter and Power Amplifier respectively. Filtering is included in the CIs of both the Frequency Converter and Amplifier, since the nature of the process (possibly) requires filtering afterwards. The Frequency Up-Converter also requires the oscillator as shown in Figure 5.2.

- **STX Support**: This branch is involved with the satellite and its operations through communicating and providing housekeeping data. It delivers the digital transmission signal to the STX TX. The required STX processing core delivers the *STX Processor* and the *STX Data Storage*. The processor also incorporates an oscillator. The additional *STX Housekeeping Sensors* were as explained before not part of the hardware breakdown presented, but are logically added to the STX Support.

The cables and circuitry required for signal transmission are part of another branch of the complete CI tree, being part of the Antenna System (SAT 2.3.3).

No requirements will be specified on the level of the STX TX and STX Support, as these are included for reasons of functional distinction; if a person performs work at this level he/she can work with all lowest-level requirements.

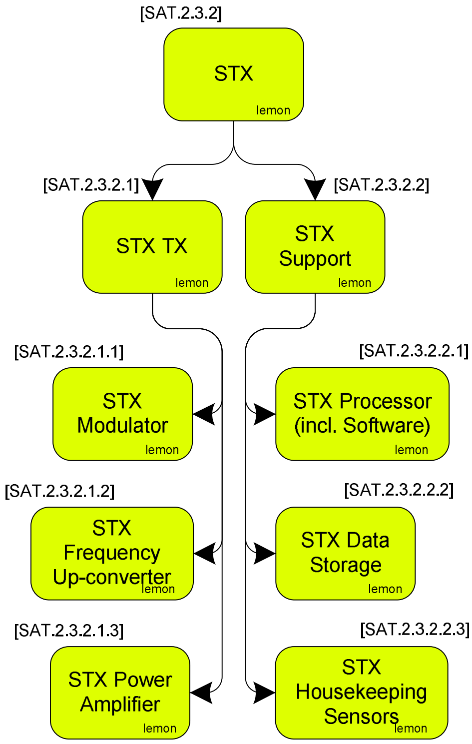


Figure 5.3: STX Configuration Item tree [SLR 0325]

The only aspect not covered directly by the CI tree is that involved with the electrical interfaces between the components, and consequently the PCB design. This can only take place at a later stage when all components have been specified. Integrating PCB designs in the CI tree is further complicated by the fact that components represented by CIs from different branches can be combined on PCBs.

Using the configuration items as a guideline, the *following six chapters discuss the six lowest-level CIs* in more detail; by describing functionalities more specifically, providing operational background information, mentioning design considerations and where possible delivering hardware choices. This is done in light of project constraints, as well as availability and technological feasibility. The requirements are established per CI, and as such the requirements plus the description given in the respective chapters should provide sufficient context on the CI to be able to continue work at a lower level; including the electrical design of the component.

Chapter 6 through 8 deal with the three STX Support CIs, and chapters 9 through 11 consequently deal with the three CIs of the STX TX branch respectively. Afterwards the integration and interfacing of the different CIs is discussed in chapter 13.

6 STX Processor (incl. Software)

The STX Processor lies at the core of the STX Support, since this component takes care of the communication with the OBC over the I^2C bus, and will deliver data to the STX TX. Since the STX should house data storage and provide housekeeping data, both the STX Data Storage and the STX Housekeeping Sensors are required; these components in turn need the STX processor to read data from and/or write data to them.

The interfaces as described above, as well as the main functionalities and data flows of the STX Processor are the topic of the first part of this chapter. Section 6.1 establishes the functions, interfaces and data flows of the processor. Sections 6.2 and 6.3 then investigate the interfaces and data flows in more detail respectively. Section 6.4 will use this knowledge to discuss the hardware implications and options. Finally section 6.5 states where the priorities lie within the STX Processor CI.

The STX Processor hardware and software are not split amongst two CIs for the reason that they are naturally intertwined in their operation and requirements often hold for both. This chapter will present information relevant to selecting the processor as well as writing the software. The STX Data Storage on the other hand is a separate CI, while enough memory can possibly be integrated in the processor. This is further assessed in section 6.4 of this chapter and in chapter 7.

6.1 Requirements definition

In order to present a set of requirements it should be clear what the exact functions and interfaces of the STX Processor are. The following Figure 6.1 demonstrates these.

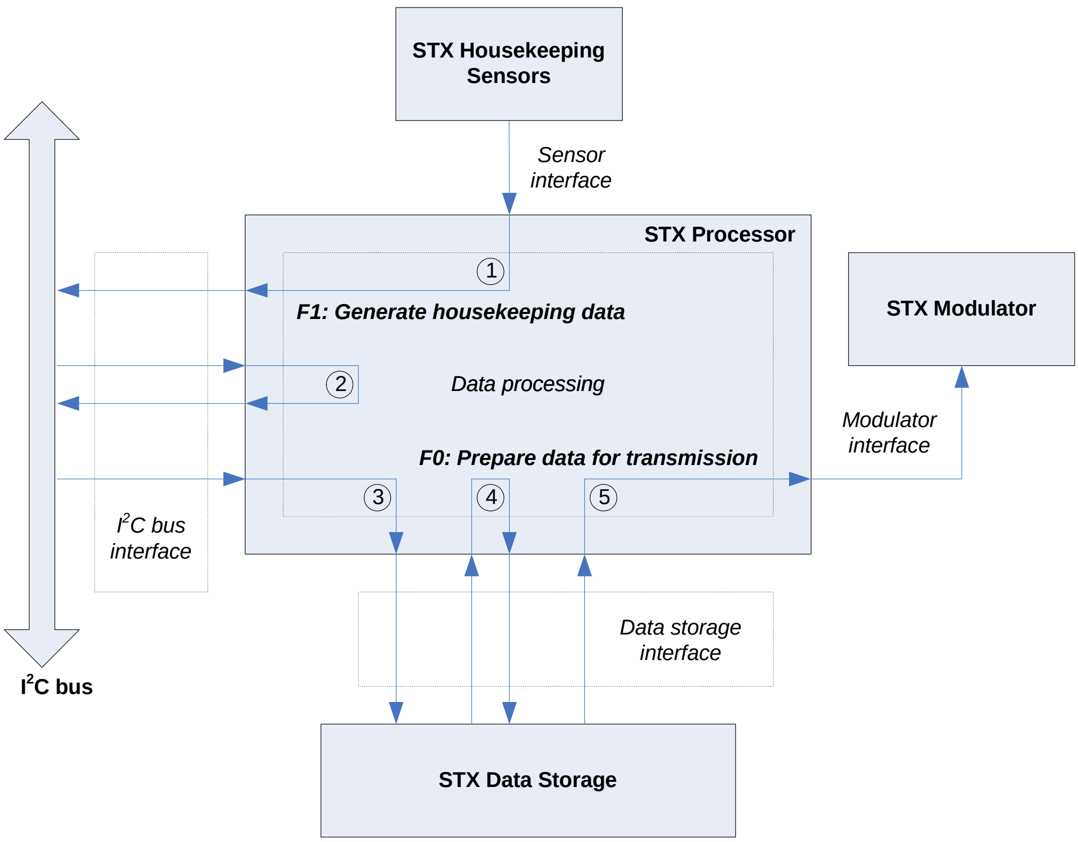


Figure 6.1: STX Processor data flow

In Figure 6.1 a number of things can be seen regarding:

- ***Functions***: There are two main functions to be executed by the STX Processor, labelled F0 and F1 in the figure. The STX Processor should prepare digital telemetry data for transmission, and the STX core should produce housekeeping data which shall be read out, processed and delivered to the OBC by the STX Processor. All interfaces and data flows are in place to provide these two functionalities.

- ***Interfaces***: Four different interfaces have been indicated in *italic*. In order to fulfil the two main functions these four interfaces are necessary. These interfaces are the topic of section 6.2.

- ***Data flows***: Five different data flows take place in the STX Processor. These data flows are numbered 1 through 5. Each of these data flows requires specific interfaces and provides different subfunctionalities in order to be able to as a whole fulfil the two main functionalities. These data flows are listed below, and are the topic of section 6.3:
  1) Housekeeping data flow
  2) Command data flow
  3) Transmission data storage flow
  4) Transmission data process flow
  5) Transmission data output flow

While in Figure 6.1 the STX Data Storage is illustrated as a separate piece of hardware, it can physically be integrated with the processor. The reason is that the functionalities do not change in this case, only the physical architecture. The processing core of the processor should still interface with the, in that case, internal memory.

The functions and interfaces introduced above yield functional and interface requirements. A number of high-level requirements that are generally applicable to all CIs and thus also the STX Processor was already presented in chapter 2. One other high-level requirement applies only to the STX Processor and not the other lower level CIs, and is presented in Table 6-1 below. All specific STX Processor requirements are given in Table 6-2 shown afterwards. In the second table an eventual parent is stated in case the requirement follows from a higher level requirement, and a rationale for the requirement is given if necessary.

Table 6-1: Higher level requirements applicable to the STX Processor

Category	Req. #	Requirement
CONSTRAINT	SAT.2.3-C.02	Any data transmitted to and from the satellite other than telecommands shall not be encrypted in any way

SAT.2.3-C0.2 is to be taken into account in the software, and simply means no encryption shall be included.

Table 6-2: STX Processor requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.2.1-F.01	**The STX Processor shall prepare the digital data to be transmitted by the STX** ***Rationale:*** *As explained above.*
FUNCTIONAL	SAT.2.3.2.2.1-F.02	**The STX Processor shall generate housekeeping data of interest to satellite operation** ***Parent:*** *SAT.2-F.0.2* ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.2.1-I.01	**The STX Processor shall have an interface with the OBC to exchange digital data** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.2.1-I.02	**The STX Processor shall have an interface with the STX Data Storage to exchange digital data** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.2.1-I.03	**The STX Processor shall have an interface with the STX Housekeeping Sensors to gather digital data** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.2.1-I.04	**The STX Processor shall have an interface with the STX Modulator to provide digital data** ***Rationale:*** *As explained above.*
OPERATIONAL	SAT.2.3.2.2.1-O.01	**The STX Processor shall be able to switch off the STX transmitter** ***Parent:*** *SAT-F.0.3* ***Rationale:*** *As a shut-off of the entire STX also shuts off data storage, it makes sense to have an 'off' mode directed by the STX Processor.*

6.2 Interfaces

As was indicated in Figure 6.1 there are four interfaces between the STX Processor and other components, as well as five data flows. Whereas the data flows pass through the processor and consequently require processing; interfaces can be seen as languages. In order to receive data to be processed and deliver processed data the processor should be able to communicate with different external components. Each of these interfaces has different characteristics such as its data type (digital or analogue), data rate and the protocols required for communication. These four interfaces are shortly reviewed below in subsections 6.2.1 through 6.2.4.

6.2.1 Sensor interface (analogue to digital)

The sensor interface is different from the other interfaces, as sensing is anticipated to imply reading analogue data, whereas the processor works with digital data. The data provided by the sensors is further discussed in chapter 8, but measurements are anticipated to deliver (analogue) voltage or current signals. For example the reflected power amplifier current can be directly measured as a current, but a temperature sensor, if analogue, will also provide either a measurable current or voltage. As such there is a likely need for an A/D converter; either separate or integrated in the actual processor. This is further discussed in section 6.5.

6.2.2 I^{2}C interface (digital)

A further definition of this interface is given in [SLR 0263], as it is linked to requirement SAT-C.05 which stipulates that all (sub)systems should adhere to standard system bus specification. This bus consists of the power bus and the data bus; the latter operates via the I^{2}C protocol.

In order to communicate with the I^{2}C bus a number of subrequirements follow. These are:
- There should be a physical connection to the serial data line (SDL)
- There should be a physical connection to the serial clock line (SCL)
- The STX Processer should be able to communicate at 100 kb/s
- An I^{2}C address should be appointed to the STX processor core
- The STX Processer should function as a slave and should listen to the SDL for data addressed to the STX processing unit, possibly via broadcast
- The STX Processer should be able to reply using the proper protocols (i.e. adding addresses)

As was mentioned by the requirement SAT.2.3.1-I.01, all data that is to be delivered to the STX is assumed to be the exact same as that provided to the PTRX and ITRX, to minimize impact. This then suggests that a broadcast can be used by the OBC to provide PTRX, ITRX and STX with raw data at the same time.

6.2.3 Data storage interface (digital)

The exact physical interface between the STX Processor and the STX Data Storage will depend on the exact processor chosen and the type of memory. The memory, being either the integrated processor memory or a combination of this and external memory should be digitally accessed by different processes.

6.2.4 Modulator interface (digital)

The modulator interface is a very simple interface; the STX processor shall output a prepared data stream to the STX Modulator. The STX Modulator is in charge of performing digital to analogue conversion so a digital bit stream is expected. The only thing that is important to be specified is the data rate of the digital bit stream. As explained before, the minimum would be 9.6 kb/s, but a (much) higher data rate should be allowed by the STX Processor. Also, the output data rate can be varied depending on pass and link budget specifics.

6.3 Data flows

As explained before the data flows require explicit processing in the STX Processor. A data flow architecture has been incorporated in Figure 6.1 with five distinct data flows. This division is based on functionality and could be slightly different in integration as is also explained below. Nevertheless in the current division the functionalities of each flow are clearly distinct, and as such every data flow can be handled by different routines. Every data flow also has its own data rate, mostly determined by its interfaces.

6.3.1 Housekeeping data flow

As was explained for the sensor interface, every sensor shall in principle provide a continuous measurable (likely analogue) signal. As such the processor itself can determine when the signal is read out and consequently when digital (housekeeping) data is produced. Normal operations will have the OBC send a 'gather housekeeping data' command, followed by a 'send housekeeping data' to each subsystem, spaced by a certain amount of time for each subsystem processor to gather and temporarily store the data in internal processor. The processor will consequently send the housekeeping data over the I^2C bus, complying with the rules of the I^2C bus.

6.3.2 Command data flow

The command data flow is envisioned to deliver the executable commands to the STX Processer, but also deliver responses to the OBC. At the moment possible commands envisioned are related to:

- Mode selection – see chapter 3
- Data rate determination – see [SLR 0014]
- Housekeeping data requests – as discussed in subsection 6.3.1

6.3.3 Transmission data storage flow

Since the only anticipated data connection between the OBC and the STX Data Storage is via the data line of the I^2C bus, the STX Processor will have to pass on the data to the STX Data Storage. The transmission data storage flow is exactly this, and nothing more. The data rate of this flow is directed by the OBC and the I^2C bus. In reality the boundaries between this and the next data flow can be artificial.

6.3.4 Transmission data process flow

The telemetry and housekeeping data provided by the OBC should be processed before it is send on to the STX TX. The following actions (can) take place:

- Addition of FEC
- Addition of data protocol
- Application of data compression (possibly but unlikely)

These processes have been introduced in detail in [SLR 0014] for the S-band communication link.

Looking at Figure 6.1 there are three different moments where this data processing can take place; during the *transmission data storage, process or output flow*. It does not make sense to prepare data for transmission just before sending it out, due to high and possibly unknown processing delay. Therefore processing directly on reception or more precisely after a specific amount of data is received (taking into account operations as data framing or FEC coding, which need minimum chunks of data) is most logical. Therefore, a separate data flow results which exchanges data between processor and data storage. After data is processed and prepared for transmission, it is stored again awaiting actual transmission.

6.3.5 Transmission data output flow

When the STX is transmitting this data flow will simply read out data from the memory storage at a certain data rate, and deliver this to the STX Modulator.

6.4 STX Processor

At the core of all the functionalities, interfaces and data flows lies of course the actual STX Processor. The CI that is the topic of this chapter is labelled "STX Processor (incl. Software)", thereby emphasizing the fact that in order for the CI to perform its functions there is the need for an *actual processor* and *operating software*.

As is covered in [SLR 0300], the default processor chosen to be used in every subsystem in the Delfi-n3Xt mission is the MSP430, more specifically the *MSP430F16X family*, since this family can fulfil all default requirements. The MSP430 processor is also the *only microcontroller for which (positive) data on radiation testing is available*.

All members within the family have equal connections and peripherals, so they can be interchanged rather seamlessly; in terms of interfaces and software. Within the family however the preferred processor is the MSP430F1612, having the largest integrated program memory. Since interfacing and Delfi-n3Xt software design will be most simple if similar or even identical processors are used, *this processor family is also the default choice for the STX system*. This increases modularity and decreases integration complexity.

In order to accept this default choice it should be verified whether:

A. The default processor can meet all requirements as imposed on the processor
B. The are no other processors meet that A. and provide extra benefits

To answer to point B. there are processors that incorporate modulation functionalities. While this can be an added benefit, it also removes the option to replace the modulator without changing the processor. For a test bed such as the STX this might be useful. Also, the modulator that is introduced in chapter 9 is in fact designed to work seamlessly with the MSP430. Furthermore, no other processors have the proven space performance that the MSP430 has.

Most importantly, modularity as stated is *one of the design goals of the STX*. Using the standard Delfi-n3Xt processor will lead to minimal integration difficulties of the finalized STX design. This is important given the limited available resources.

As such, *the MSP430 processor is indeed selected as the default processor on the STX.*

It remains to verify that the processor can indeed meet the requirements as posed on the STX Processor. The functions and interfaces are related to processing power, memory, physical connections and software. Some requirements do not differ substantially from those of the other processors on Delfi-n3Xt, such as those related to housekeeping sensors and the I^2C interface and commanding. Then the processing tasks are exactly those listed in subsection 6.3.4 to be those related to *transmission data processing*, together with the overhead created by the communication with the data storage.

The reading from and writing to a form of data storage is not assumed to pose a severe load, as is the addition of a data protocol. Data compression, as argued in [SLR 0014], is not likely to be applied. FEC coding will have the most significant impact. But, as with any FEC coding scheme the most significant impact will be on-ground. Nevertheless, in case high bit rates will be used, the application of the FEC coding will become the limiting factor. Functional testing will provide the final answer. Final transmission speeds are in any case limited to by the current modulator to 500 kBaud as introduced in chapter 9.

Subsections 6.4.1 and 6.4.2 investigate the possibilities of A/D and D/A converting within the MSP430F16X family, and the need for external clocks respectively. The available integrated memory within the MSP430F16X family is commented on in subsection 6.4.3.

6.4.1 A/D and D/A converting

In subsection 6.2.1 the constraint was anticipated that the STX Processor CI should incorporate an A/D converter. The MSP430F16X subfamily [SLR 0454] incorporates this converter. In Delfi-C^3 also use was made of internal A/D converters of the PICs to read out housekeeping sensors [0245], which offered 10-bit precision. As such it would seem the internal A/D converter functionality in this case could fulfill the same goal. The MSP430F16X family provides 12-bit A/D conversions, but as argued in [SLR 0014], 8-bit is most likely enough.

The MSP430F16X also has an incorporated D/A converter, so optionally this could also be used in relation with the STX Modulator. However, assuming the use of the CC2500 modulator as assessed in chapter 9, no D/A conversion is expected to be required.

6.4.2 Clock signals

In order to process data and to communicate over its interfaces a processor needs a clock signal. The MSP430F16X has an integrated clock signal tunable to be at maximum about 8 MHz, but the responsible oscillator is not stable under varying temperature conditions. The I^{2}C bus will provide a clock signal, but this will be at 100 kHz. For the processor's operations this frequency is certainly uselessly low and potentially also for the modulator interface, and even more so standard system bus requirements specify an internal clock of at least 1 MHz. Therefore a higher frequency oscillator should be externally connected to the processor. This oscillator could produce a clock signal close to the original maximum 8 MHz, to allow for the maximum processing speed for the MSP430F16X. Consequently this signal can be changed by an internal frequency manager to produce clock signals suitable for all processor operations. A suitable clock will most likely be selected for general use with the MSP430F16X's within the Delfi-n3Xt mission. Consequently the speeds of the different processes should be determined in close coherence to software design.

In case of the STX the clock signal could also originate from the oscillator required to up-convert the frequency to radio frequency. The STX Frequency Up-converter is further reviewed in chapter 10. The MSP430F16X as mentioned has an internal frequency manager, which can produce multiple clock signals by division from up to two external oscillators. Therefore it needs an external clock signal other than that of the I^{2}C bus, which is at minimum equal to the required processor speed. Also, the integrated frequency manager cannot handle a clock signal above 8 MHz, and as such an external high frequency oscillator should be connected to a separate controller to decrease the clock signal to a usable one.

The CC2500, the 2.4 GHz transceiver chip suggested as the STX Modulator (see chapter 9) provides exactly this option. This is further reviewed in chapter 10.

6.4.3 Integrated memory

It has been mentioned that the MSP430F16**12** is the preferred processor within the MSP430F16X family, due to its largest program memory. Nevertheless the MSP430F16**11** actually offers the most process memory, in the form of volatile SRAM; Table 6-3 illustrates the differences. In all other aspects the two processors are identical.

Table 6-3: Memory difference of the MSP430F1611/12

Processor	***Program memory (kb)***	***Program memory (kB)***	***Process memory (Mb)***	***Process memory (MB)***
MSP430F1611	384	48	81.920	10.240
MSP430F1612	440	55	40.960	5.120

Since the MSP430F1612 has been primarily selected to be able to fulfill all OBC functionalities [SLR 0300], the small difference in program memory in case of the STX Processor is not likely to make much difference. The increase in process memory on the other hand can have larger implications. If the process memory proves to be enough for the processor to perform its operations and at the same time to temporarily store all required telemetry data to be transmitted in a downlink event, there would be no need for an external piece of data storage. In other words the STX Data Storage could be physically integrated in the STX Processor.

As the next chapter reviews however, the maximum integrated process memory would be on the light side for the envisioned operation. But in any case, since the STX Processor will be much more involved with actual data processing than processors on other subsystems, there could be a good reason to maximize process memory. Working with internal memory should allow for quicker processing than making use of external memory. This would favor the *MSP430F1611 processor*.

The next chapter addresses the requirements on the data storage and as such also addresses the process memory of the MSP430F16X further.

6.5 Work to be done for the STX Processor (incl. Software)

Work to be done:
- The processor (MSP430F16X) should be selected, depending on the choice of data storage
- The software should be written, offering correct interfaces and data flow functionality

Highest priority:
- None

Clarification:
- The main priority within the STX CI lies with creating a working transmitter. This can be done using any MSP430F16X processor. Test software can then be written at the same time. The final software including all operational aspects and coding etc. can be written at a later stage.

The priority of the STX Processor with respect to the other CIs is discussed in chapter 15

7 STX Data Storage

The second CI in the branch of the STX labelled STX Support, as can be seen in Figure 5.3 is the STX Data Storage. The previous chapter has already elaborated extensively on the link between the STX Data Storage and the STX Processor, including their possible integration in terms of components.

This chapter presents the requirements on the STX Data Storage, and consequently derives what this means for the hardware. The requirements are first defined in section 7.1. Afterwards sections 7.2 and 7.3 discuss memory size and memory type respectively. Section 7.4 assesses some operational aspects of the STX Data Storage, and Section 7.5 concludes with an assessment of where the priorities of the STX Data Storage CI lie.

7.1 Requirements definition

The STX Data Storage is a relatively simple component, since it is an entirely passive component. The STX Processor is in charge of reading and writing to it. Therefore the data storage has only one interface. It also has no higher level requirements applying only to itself. The result short list of requirements on the STX Data Storage is presented in Table 7-1, as before with a rationale and parent included if applicable.

Table 7-1: STX Data Storage requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.2.2-F.01	**The STX Data Storage shall provide storage for all data to be transmitted by the STX** ***Parent:*** *SAT.2.3.2-F.02* ***Rationale:*** *direct derivative of parent.*
INTERFACE	SAT.2.3.2.2.2-I.01	**The STX Data Storage shall have an interface with the OBC to exchange digital data** ***Rationale:*** *As explained above.*
CONSTRAINT	SAT.2.3.2.2.2-C.02	**The STX Data Storage must have a size that represents at least 24 hours of all unique payload data and housekeeping data** ***Parent:*** *SAT.2.3.2-P.02* ***Rationale:*** *direct derivative of parent.*

It can be concluded that the main driver for the STX Data Storage is its memory size. The constraint requirement was based on the desire to use the STX for a high speed down link once per 24 hours. While this desire is no strict requirement, it forms a good guide line; in any case no smaller size would be wise.

7.2 Required memory size

As stated by requirement SAT.2.3.2-P.03, the STX shall receive the exact same data as the PTRX or ITRX. This means that the STX will receive a data density directly dependent on this in turn. As argued in [SLR 0014], a data rate of 9600 b/s is likely achievable, but it might not be functionally used in terms of data generation. A minimum data rate is 1200 b/s. The amount of payload and housekeeping data is less due to overhead required in data transmission. The maximum amount of telemetry data will be produced if the final DelfiX protocol [SLR 0014] will be used. Excluding the bits of the Frame Type ID field, 72 bits are used per second for the protocol overhead, and 86 bit stuffing bits per 1200 bits. In case of a bit rate of 1200 b/s this would then yield 1042 bits per second to be delivered to the STX.

The following Table 7-2 presents the resulting daily data production amounts for the cases that a data rate of 1200 b/s, 2400 b/s or 9600 b/s will be designed for.

Table 7-2: Raw data storage requirement

PTRX bit rate	*DelfiX overhead*	*Raw data rate*	*Daily data production (b)*	*Daily data production (Mb)*	*Daily data production (MB)*
1200 b/s	*DelfiX fields*: 72 bits *Bit stuffing*: 86 bits	1042 b/s	90,028,800	90.029	11.254
2400 b/s	*DelfiX fields*: 72 bits *Bit stuffing*: 172 bits	2156 b/s	186,278,400	186.278	23.285
9600 b/s	*DelfiX fields*: 72 bits *Bit stuffing*: 688 bits	8840 b/s	763,776,000	763.776	95.472

It can be seen that the daily data rate production at minimum is a little over 10 MB. *In fact, this is already more than the MSP430F16X family can offer*.

Some effects that would increase the required storage size even further:

- **FEC coding** will have big effects on required data storage, as this is not likely applied only on-the-fly due to a heavy processing load. An advanced FEC scheme, such as the AO-40 FEC coding scheme increases the data rate or symbol rate to *2.5 times its original value*, due to a code rate of 0.4 (1 out of 2.5 bits is an information bit).
- **The data protocol** adds bits in the form of fields to the message; this might be stored before transmission. These fields however cause little physical (number of bits) overhead.
- **Bit error prevention**: data storage is in theory sensitive to bit errors due to the space environment, most notably radiation. Depending on the actual type of storage and its reliability, an approach could therefore be preferred in which data is stored redundantly, possible in triple. A comparison between all instances of the same data could then yield the most likely original data. An approach with less overhead would be to calculate check sums over stored data, so that errors can at least be detected. This all comes at the cost of increased processor load and complexity, and of course storage capacity, and is hopefully not required for sufficient operations.

And in case internal memory would be used:

- **Processing memory** is also required, next to the amount required for direct storage. A certain amount should be identified and reserved for data processing.

However, two arguments for a smaller required storage size are the following:

- **Lower actual download performance**: there is a good chance that not all produced data can be downloaded by the STX, not even in the best possible case, as reviewed in [SLR 0014]. In this case the choice could be made to store not all daily data, but just a certain amount.
- **Data compression** would if feasible decrease the storage size for the same time period, as data can be compressed before being stored.

In conclusion, *it is not likely that the MSP430F1611 internal data storage capacity is sufficient*, as even with a minimum data rate and without FEC coding it is too small. Reducing the required storage size due to the data compression is not likely as the latter functionality is not anticipated, and limiting download capability by minimizing the storage size by choosing a smaller storage window would be unappealing as it lowers overall potential.

Then, holding on to the requirement of having data storage size equal to the data production of one day, an estimated 2.5 times these values should be assumed, taking into account only additional FEC coding. This yields a *minimum storage size of almost 30 MB, and a maximum of 240 MB*. Larger amounts would in turn allow for much more data storage to take place, which can be of interest to request missing data.

For example, early mission data can be stored for the first days or even months, after which this can be downlinked bit by bit at request. Sticking to the assumption that not more data can be received by the STX than the PTRX or ITRX will receive, an anticipated 3 month mission lifetime would yield between 1 and 8 GB based on the above numbers. Two years would then yield 8 to 64 GB of data. Having several GBs of storage is by no means off limits thinking of current (small) COTS storage memory card applications, and according to [SLR 0300] 4 or 8 GB should be interfaceable to an MSP430F16X without problems.

Only if integrated processor data storage turns out to be absolutely preferred, possibly due to a lower design complexity or due to more predictable radiation behaviour of the MSP430 processor, the storage capacity might be reduced.

7.3 Memory type

When talking about memory an important consideration next to size is its type. The main subdivision in memory types is between volatile and non-volatile memory. [SLR 0105] gives a good introduction to the two types and its subtypes. Also Appendix A reviews some memory types in FPGAs.

The difference between volatile and non-volatile memory is the ability to retain information after power is removed; the latter types are able to do so while the former do not. Also, volatile memory types are much faster in their operation. In case of the STX Data Storage it is actually not really important to retain information when no power is available, since in nominal operations the STX Data Storage should always be powered. *However, the capability of the battery to ensure continuous operations during all eclipses has not been confirmed yet.*

If emphasis on serious data storage in the STX in increased, for periods longer than 24 hours, non-volatile memory is preferred. At the same time, the added speed offered by volatile memory is not directly required (in both cases still in the order of 10 nanoseconds).

Due to higher power requirements and a higher probability of single event upsets (SEUs – due to space radiation) SRAM seems to win over DRAM, also since physical size is not too important at these scales. PROM is not an option since it cannot be rewritten, and EPROM is mentioned to be outdated compared to EEPROM and FLASH, which are similar. Therefore the choice seems to be between SRAM, and EEPROM/FLASH, being volatile and non-volatile respectively. [SLR 0300] makes a case for FLASH however, due to its lower general power consumption and its application on a nano-satellite as Swisscube. FLASH adds the advantage of non-volatility and availability, as it is the type of memory storage used in USB drives and memory cards (e.g. SD cards). These chips can easily be taken out of their package and interfaced with an MSP430F16X.

As is discussed in [SLR 0104] read/write cycles play a role for non-volatile memory storage, but the endurance is said to be between 10,000 and 100,000 write cycles per sector. This is confirmed by [0311]. These numbers would largely suffice for the 730 daily cycles over a 2 year mission, including sufficient margin for multiple read/write sessions per day related to FEC coding, data framing and compression as well as more general processor operations. Also, a larger memory size would allow for less effective read/write cycles.

[SLR 0300] reviews a number of memory units on the basis of power consumption, reliability, radiation resistance and storage size, amongst others. Approaching the actual graduation of Armin Noroozi the final version of the latter document will hopefully allow to straight-forwardly select a specific data storage chip to be applied on the STX.

To be further investigated

- What specific data storage device should be used as the STX Data Storage?
- Is a type of redundant data storage necessary?

7.4 Operation

The operation of the STX Data Storage was assessed in detail in chapter 6, since all activity of the storage device is directly dependent on the STX Processor. Different processes might have to be able to access the storage device more or less simultaneously; not at all uncommon to general data storage device operations.

An aspect that has not been covered in the previous chapter is that of its buffer functionality; new data is constantly added to the data storage, and since the requirement is that all data from the last 24 hours should be stored, data could also be erased or overwritten outside of the interval of the last 24 hours. Data can also be erased after a read out during data transmission, but this would assume all data can be send only exactly once.

It is most simple to use the data storage as a buffer regardless of whether data is actually being read from it; this means that the oldest data should simply be overwritten when new data is added, depending on storage size. A maximum amount of available data storage should be defined for storage purposes (in case the processor will use the data storage also for value storage) and the memory should function as a *push-out buffer*. The processor can use a number of sectors in an identical sequence and start from the beginning when the last sector has been written to.

7.5 Work to be done for the STX Data Storage

Work to be done:

- The data storage device to be used should be selected
    - Is consequently a type of redundant data storage necessary?

Highest priority:

- The selection of the data storage device

Clarification:
Continuing from work to be presented by [SLR 0300], a specific data storage device should be selectable relatively straight-forwardly. Some research should be done with respect to radiation resistance at the same time, and the interfaces with the MSP430F16X can be tested afterwards.

The priority of the STX Data Storage with respect to the other CIs is discussed in chapter 15.

8 STX Housekeeping Sensors

The final CI as part of the STX Support branch is that of the STX Housekeeping Sensors. The goal of the housekeeping sensors is to provide data to be able to spot irregularities that indicate system errors or failures, or vice versa to be able to pinpoint where problems originate after an error or failure occurs. Section 8.1 first presents the current requirements on the sensors. Section 8.2 establishes the variables that are to be measured, after which section 8.3 introduces the components required and some concrete options. Again, the final section 8.4 concludes on where the priorities lie within the STX Housekeeping Sensors CI.

8.1 Requirements definition

The functionality of the STX Housekeeping Sensors has been described in short above. All housekeeping sensors have one interface to the STX Processor. Momentarily this leads to only two requirements on the STX Housekeeping Sensors, as can be seen in Table 8-1 below.

Table 8-1: STX Housekeeping Sensors requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.2.3-F.01	**The STX Housekeeping Sensors shall provide housekeeping data of interest to satellite operation** ***Parent:*** *SAT.2-F.0.2* ***Rationale:*** *The main and only functionality of the housekeeping sensors; therefore they are solely required due to the parent requirement.*
INTERFACE	SAT.2.3.2.2.3-I.01	**The STX Housekeeping Sensors shall have an interface with the STX Processor to provide its housekeeping data** ***Rationale:*** *The housekeeping data will ultimately be presented to the OBC by the STX Processor.*

8.2 Housekeeping data specification

As was explained in subsection 6.3.1 on communication with the I^2C bus, the OBC will regularly announce housekeeping data collection, after which the responsible subsystem processor will gather the information and store it in its internal memory. After a second command the data is then put on the bus addressed to the OBC. The PTRX housekeeping data has been determined in [SLR 0014], with a number of five variables at least being measured by the PTRX microcontrollers itself. As the STX is only a transmitter, this number is logically less. *However*, as argued in chapter 11, it is suggested to use *two* power amplifiers to supply two different patch antennas with a signal. As such, a number of housekeeping values can be gathered twice.

Following the same argumentation as presented in [SLR 0014], the following values are determined to be of value. The first value is to be measured by the Local EPS on the STX PCB.

Current consumption values:

- ***STX current (8 bits)***: In order to get an idea of whether the receiver is operating without serious electrical problems, the current consumption can be monitored to check for nominal values.

Signal strength indicators:

- ***2 x Forwarded power (8 bits)***: The forwarded power, together with the reflected power, indicates the total power forwarded to the antenna system from the transmitter. In case of an antenna failure, these two values allow to notice and draw conclusions on the failure that has occurred. This value can be measured in an analogue fashion using a power detector and an ADC port of the MSP430, similar to how the PTRX does.
- ***2 x Reflected power (8 bits)***: See forwarded power.

Temperature values:

- ***2 x Power amplifier (PA) (8 bits)***: The power amplifier (PA) is the only component that significantly heats up during nominal operations, if no high-efficiency PA is used. Thus, it is essential to monitor this heating. A digital temperature sensor identical to the PTRX PA temperature sensor can be used that yields a digital value. The (8-bit) precision of the value can be increased to produce a better thermal model.

Also for the STX, it might be desired to have more temperature measurements to improve upon an overall thermal model. Also, as the STX is experimental this might be even more preferred. Optional:

- ***Processor (8-12 bits)***: MSP430 processors have an integrated temperature sensor. As such, these temperature values can easily be communicated to the OBC and transmitted to the ground if the data budget allows for it. As MSP430 processors are to be implemented everywhere on the satellite, a general approach should be chosen. 12 bits is the precision of the value provided by the MSP430 processors, but 8 bits might be enough.
- ***Modulator (8-12 bits)***: In the next chapter the CC2500 modulator chip is introduced. This chip also has an integrated temperature sensor, and as such could if deemed useful also have its sensor output connected to the processor to provide it with temperature data. It provides an analogue value, which can be converted to an 8-12 bit value by the MSP430 based on what seems necessary.
- ***PCB (8-12 bits)***: One or multiple PCB temperature values can be measured to get a good idea of the temperature distribution on the PCB and its components. This would however require dedicated temperature sensors at strategic locations. This is a choice that should be made on the satellite level, as the same argumentation holds for all PCBs, albeit with more or less activity. Again 8-12 bits can be used.

Of course, as the STX design is not finished, extra variables could be available or required to provide extra insight into the system during operations. Keeping this in mind the future design of the STX should be reviewed for additional options. Nevertheless, for now the subset of housekeeping variables based on what is measured on the PTRX and RAP shall be assumed, being the *forwarded power*, *reflected power* and *PA temperature*. All these variable are however *to be gathered twice* for each PA. The seventh value, the *STX current*, is as said before not to be measured by the STX itself.

To be further investigated

- What exact variables should be measured?

8.3 Sensor specification

Using the variables to be measured, the sensors and the hardware required can be specified. The architecture of the PTRX allows for easy reimplementation. The sensors are required twice each in case of having two PAs.

- ***Forwarded power and reflected power***: In case of the PTRX, a power detector called the *AD8361* is used. It is to be connected to a bidirectional coupler (the ADC-10-1R), which in turn is placed right after the PA, and gives (analogue) data on the forwarded and reflected power. Similar to with the PTRX, an MSP430 processor can read out these values with integrated A/D convertors. According to the datasheets of both components ([SLR 0110) and [SLR 0719]), they correctly function up until signal frequencies of 2.5 GHz and 2.6 GHz respectively, which is perfect. It should however be mentioned that the inserted bidirectional coupler yields an insertion loss of 1.2 dB. Therefore at a later stage its required functionality should be verified.
- ***PA temperature***: Again there seems to be no reason to assume why the same sensor as used on the PTRX cannot also be used on the STX, being the SMT16030 [SLR 0692]. It yields a digital duty cycle output, which means the average time its output is kept 'high' or 'low' indicates the temperature value. Pre-written software for the PTRX Frame Generator (microcontroller) can be used. Some PAs also have integrated temperature sensors.

The processor temperature is automatically measured, as well as the CC2500 temperature if indeed the CC2500 is to be used. In the latter case the temperature should be converted to a digital signal by the MSP430. Other integrated circuit (IC) solutions such as amplifiers can have similar embedded temperature sensors which similarly would require a link to the STX Processor. PCB temperature sensors would require dedicated sensors, which can be of the same type as that used for measuring the PA temperature.

8.4 Work to be done for the Housekeeping Sensors

Work to be done:
- The exact required variables to be measured should be confirmed and/or determined
- Depending on the variables, the hardware to take care of these measurements should be selected

Highest priority:
- None

Clarification:
The complexity of these sensors is assumed to be quite low, and its implementation is closely linked to electrical integration to be executed near the final stages. Also near the end the actual required measurables can be better established.

The priority of the STX Housekeeping Sensors with respect to the other CIs is discussed in chapter 15.

9 STX Modulator

The second CI branch of the STX is called the STX TX, in other words the part of the STX that takes care of preparing a digital data stream from the STX Processor for transmission. The STX TX modulates a data signal onto a carrier, transforms the carrier frequency to radio frequency (RF) and amplifies the signal strength for transmission. These three functionalities are covered by the three CIs STX Modulator, STX Frequency Up-converter and STX Power Amplifier. This chapter deals with the STX Modulator, whereas the next two chapters deal with the other two CIs.

The selection of the modulation method for the S-band communication link has been performed in [SLR 0014], and the result has been that *the preferred modulation scheme for the STX is MSK*. No line coding is deemed necessary, and at the same time no additional pulse shaping seems necessary. If the latter would change, Gaussian filtering can be applied to reduce the produced bandwidth to 0.75-1.1*R [Hz], where R is the data rate.

Section 9.1 first defines the requirements for the STX Modulator. Section 9.2 consequently assesses the hardware that can be used in the STX to perform the preferred modulation. Finally section 9.3 concludes on the STX Modulator near future priorities and planning.

9.1 Requirements definition

The STX Modulator is simple in its interfaces, in that it receives a digital signal from the STX Processor and is supposed to deliver an analogue modulated signal to the STX Frequency Up-converter. The requirements have flowed down from higher level requirements as presented before and are shown in Table 9-1 below.

Table 9-1: STX Modulator requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.1.1-F.01	**The STX Modulator shall modulate a received digital signal onto a carrier signal** ***Rationale:*** *The logical function of the modulator.*
INTERFACE	SAT.2.3.2.1.1-I.01	**The STX Modulator shall have an interface with the STX Processor to receive a digital signal** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.1-I.02	**The STX Modulator shall have an interface with the STX Frequency Up-converter to deliver an analogue signal** ***Rationale:*** *As explained above.*

Next to the requirements stated above, chapter 11 on the STX Power Amplifier presents the following preference:

It is preferred that the STX Modulator applies a constant envelope scheme, with the possible exception of on-off keying (OOK), to allow for future switching mode amplifiers that might be non-linear.

9.2 STX Modulator hardware

Whereas it would be an option to design an STX Modulator in-house, there are two prime reasons against it. First of all, if acceptable solutions already exist, there is no reason to do so, provided that a) performance improvements that can otherwise be made are minor and b) technical information (in relation to testing and adaptation) is available. Second of all the STX is not a mission priority; complete electrical design seems not very feasible in terms of present and future resources. Therefore a critical eye should be turned to available COTS products that can provide S-band transmission.

Robin van Eijk has looked into COTS Modulators [SLR 0643], and after making a selection has purchased a certain promising COTS S-band transceiver, which is the CC2500 from Texas Instruments, for its datasheet see [SLR 0556]. [SLR 0643] also presents the results of playing around with some required settings. At the same time it includes an introduction to the chip as well as some conclusions, in turn being a summary of the work below. The transceiver is introduced in subsection 9.2.1, after which its modulation schemes, packet handler functionality and electrical integration are dealt with in more detail in relation to Delfi-n3Xt in subsections 9.2.2 to 9.2.4.

9.2.1 The CC2500 2.4 GHz transceiver

In terms of CIs the CC2500 could be the STX Modulator. It needs a separate processor to provide it with data, and it also needs an external oscillator and power amplifier. It incorporates a frequency synthesizer and has processing capabilities and as such can possibly draw somewhat into the STX Processor and STX Frequency Up-converter CIs. This will be reviewed in the sections below and the next chapter on the STX Frequency Up-converter.

The CC2500 is designed to be used together with an MSP430 processor, which is the default Delfi-n3Xt processor. This also allows making use of the proven radiation resistance of the latter.

The CC2500 from Texas Instruments is as mentioned actually a transceiver, but it can be set in transceiver, transmitter or receiver mode; due to the system design of the STX no use will be made of the receiver mode. This would require significant additional circuitry in the antenna system, as well as proper signal amplification and extra complexity in the processor functionality and consequently the OBC. Next to this there is simply no mission need for the receiver, since two receivers are already in place and a high data rate receiver provides little added value since only telecommands are uplinked. In a future mission the functionality might be reconsidered however.

Due to its availability and its promising functionalities the CC2500 is shortly reviewed here with a focus on its possible incorporation in the STX. The most relevant specifics of the CC2500 can be found in Table 9-2 below.

It can be seen that the frequency range, data rate range, and temperature ranges should easily fit the requirements. The data rate range is further discussed in the next subsection. Regarding the frequency range it can be said that there is an additional requirement due to the nature of the frequency up-conversion taking place in the CC2500; RF frequencies at $n/2$ times the crystal frequency should be avoided due to strong spurious signal creation, with n being any positive integer. This is further assessed in the next chapter on the STX Frequency Up-converter, in relation to the choice of crystal frequency and expected STX transmission frequency.

The supply voltage range is also very suitable, as it is identical to that of the MSP430; the latter is run at 3.3 V. 3.6 V is preferred in case of the CC2500 for maximum output power. The output power should be taken along in the discussion on the STX Power Amplifier. The current consumption is also sufficiently low.

The energy percentage bandwidth is an indication for the required bandwidth to send and properly receive and demodulate the signal. The bandwidth is to be taken along in the precise frequency coordination, which is pending. In fact at the edges of this allocated bandwidth more than 99% signal reduction is required, so also the digital channel filter bandwidth is indicated which gives another boundary to the bandwidth. The exact required bandwidth is commented on in [SLR 0014] and should be determined by means of functional tests later on.

Furthermore it can be seen that a temperature sensor is integrated which can be connected to the STX Processor and deliver housekeeping data. A frequency synthesizer is integrated as well which means there is most likely no need for a separate mixer circuit for the STX Frequency Up-converter, as well as a separate signal filter. The packet handler functionality is further discussed in subsection 9.2.3, as well as the use for the available modulation schemes in subsection 9.2.2.

Table 9-2: CC2500 relevant specifications

Modulation schemes	• BFSK -> *data rates 1.2-500 kBaud* • MFSK -> *data rates 1.2-250 kBaud* • OOK -> *data rates 1.2-250 kBaud* • MSK -> *data rates 26-500 kBaud*
Programmable data rates	1.2-500 *kBaud*
Frequency range	2400 – 2483.5 MHz
TX current consumption	~21.5 mA
Operating supply voltage	1.8 – 3.6 V
Storage temperature range	-50 – +150 °C
Operating temperature range	-40 – +85 °C
Max. output power	1 dBm / 1.3 mW
99% energy percentage bandwidth	117 kHz @ 10 kBaud, BFSK 296 kHz @ 250 kBaud, MSK 489 kHz @ 500 kBaud, MSK
Digital channel filter bandwidth	232 kHz @ 10 kBaud, BFSK 540 kHz @ 250 kBaud, MSK 812 kHz @ 500 kBaud, MSK
Noteworthy extra features:	• Integrated temperature sensor • Complete on-chip frequency synthesizer, no need for extra filters • Incorporated packet handler functionality

Finally it can be said that the CC2500 could be used as well in the ground station to demodulate received data, at least for early functional testing. A development kit [SLR 0559] is being sold which features the complete electrical integration circuit as further assessed in subsection 9.2.4 including the CC2500 and an MSP430 processor as well as an USB connection, and operating software is available. This same kit can be used to configure the CC2500 to be used in the STX. However, the CC2500 does not provide soft-decision demodulation; so for high performance a separate demodulator on-ground is required.

In the last chapter it was noted that the MSP430F16X incorporates a D/A converter, but since indeed the CC2500 expects a digital data flow no use shall be made of that in the anticipated configuration.

9.2.2 CC2500 modulation schemes

The CC2500 provides a number of simple modulation schemes, which have undoubtedly been selected for their technological simplicity, widespread application and possibly bandwidth efficiency. It has been stated at the start of this chapter that MSK is the preferred modulation scheme for the STX.

The CC2500 can be selected to use solely MSK modulation, however in order to do so it should be with a data rate of at least 26 kBaud. If this data rate is not reached, the achievable data rate is immediately halved to 13 kBaud approximately due to the 3 dB approximate power efficiency difference compared to the other modulation schemes.

As FEC coding is added to the signal, the data rate in Baud will go up, the bit rate in b/s will not. The following convention shall be used in the context of the STX; telemetry data, along with data framing bits will solely define a certain bit rate, whereas added FEC coding will only increase the Baud rate. Whenever the term data rate is used, it usually implies the bit rate unless it is specifically stated.

In HDM a data rate of 26 kBaud should be achievable, even though it is not strictly required from a mission point of view. Nevertheless as was assessed in chapter 7, in order to transmit all accumulated data at the STX a much higher data rate would be necessary, and a maximized data rate is therefore desirable. And, if FEC coding will be used with coding rates of possibly more than ½, the data rate in Baud shall be possibly more than double the bit rate.

9.2.3 Packet handler functionality

The CC2500 is not just a modulator, as was already addressed by demonstrating some added functionalities. As such it also provides packet handler functionality. This means that it can be set to receive only a raw data stream and it can package the data itself, to provide for necessary information. This is done as the CC2500 is designed to also be able to receive data it itself has sent (using 2 identical CC2500 chips). As such it should be able to add preamble and possibly synchronization words to the data. The purpose of preamble and other data protocol fields has been addressed in [SLR 0014].

As the CC2500 is not designed to have or be connected to data storage, all packet handling will have to be performed 'on-the-fly', or in real time. Therefore logically only simple FEC or data protocol schemes can be applied.

All possible additions that are part of the CC2500 packet handler functionality can be seen in Figure 9.1.

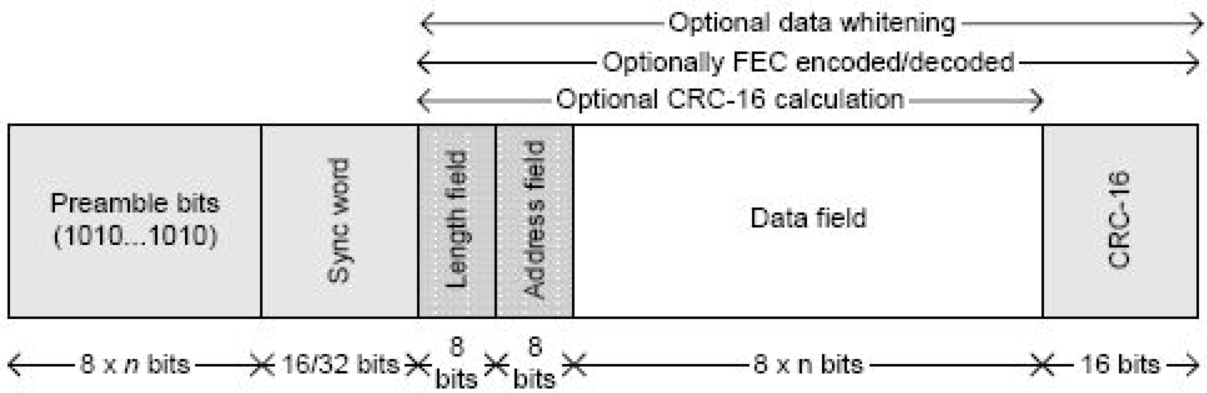


Figure 9.1: Options of the C2500 Packet handler functionality [SLR 0556]

Next to the preamble and synchronization words a number of other fields can be added if required:

- ***Length field***; if the total package length is under 256 bytes
- ***Address field***
- ***CRC-16***: 16 bit cyclic redundancy check as a form of error detection

Also, over a number of fields *data whitening* as well as *FEC coding* can be applied:

- ***Data whitening*** is a functionality that can 'randomize' the data to be sent; this can be done in order to avoid large sequences of 0's or 1's that can complicate signal synchronization.
- ***The FEC coding scheme*** that the CC2500 provides is relatively simple. It provides a convolutional coder with a memory size of 4 bits, having a rate of ½; thus producing 2 bits per received bit. In order to deal with burst errors an interleaver can be applied as part of the FEC coding scheme, which is one with a buffer size of 16 bits (matrix size of 4 by 4).

For more information on general FEC coding, see [SLR 0014]. This document also discusses the possible relevance of the offered FEC coding scheme to Delfi-n3Xt. It is suggested in that at least the addition preamble can be done by the CC2500. The FEC coding scheme, possibly in combination with data whitening will only be applied if no other FEC coding scheme shall be feasible. It can in any case be preset as an alternative.

9.2.4 Electrical integration

As the CC2500 should be integrated, likely on a PCB, as is further discussed in chapter 13, any COTS solution should provide clear details on its electrical interfaces, mainly through a complete pin lay-out. The CC2500 data sheet [SLR 0556] provides this, as well as an exact suggested electrical diagram. The resulting circuit integration diagram is inserted here as Figure 9.2.

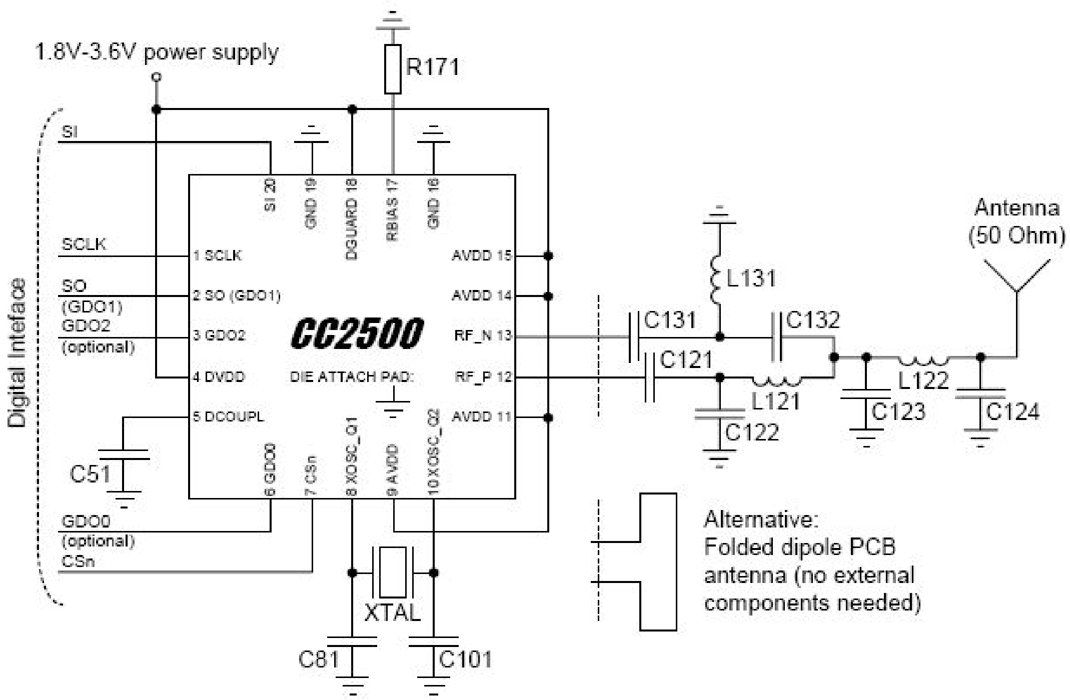


Figure 9.2: CC2500 circuit integration [SLR 0556]

The detailed explanation for this figure can be found in [SLR 0556]. The digital interface to the processor can be seen on the left; which also includes the digital output from the temperature sensor. The power supply is shown on the top, and the integration of the piezoelectric crystal resonator is illustrated, indicated by XTAL. Together with the shown electrical connections and capacitors this forms the oscillator circuit. An antenna connection with a preferred impedance value of 50 Ω can be seen on the right side of the figure. The development kit as mentioned before incorporates the simpler folded dipole antenna as illustrated. Next to these interfaces a number of electrical ground connections can be seen as well as standardized electrical components.

The crystal oscillator is a critical component in the sense that it should be stable under widely varying temperature conditions as in space. This is further assessed in the next chapter as part of the STX Frequency Up-converter.

9.3 Work to be done for the STX Modulator

Work to be done:

- The use of the CC2500 as the STX modulator should be confirmed after a short up-to-date review of the alternatives, or possible updates.
    - The chip should be tested with simple programming
- The CC2500 software should be written

Highest priority:

- The confirmation of the CC2500 usage
- The integration of the CC2500 is a working transmitter section

Clarification:
As the CC2500 is confirmed, it should be tested with the MSP430F16X to basically test simple STX operations. These tests should be operational, but also environmental (temperature for example).

The priority of the STX Modulator with respect to the other CIs is discussed in chapter 15.

10 STX Frequency Up-converter

According to the superheterodyne principle, the frequency is only up-converted to *radio frequency* (RF) after all relevant signal processing has been performed. Therefore one or more practical *intermediate frequencies* (IFs) are selected for data processing and preparation, after which the signal frequency is up-converted. Due to the nature of this up-converting a filter should be used after the up-conversion, as signal images and harmonics are created. The components necessary for the actual up-conversion and the filtering afterwards make up the CI STX Frequency Up-converter.

A usual approach to up-convert a signal frequency is to select an oscillator to equal the desired RF frequency when summed with the IF, and to use mixer circuitry to combine the signals. An ideal linear mixer will create two resulting frequency signals; $f_1 = f_{oscillator} + f_{IF}$ and $f_2 = f_{oscillator} - f_{IF}$. As was done in the RAP and as generally done by radio amateurs, an *image-rejection filter* is placed after the mixer. This filter then passes only the desired f_1 or f_2. This approach is explained in [SLR 0014].

As in reality mixers are not ideal linear devices, signals with frequencies which are combinations of the two mixed frequencies can be created depending on the specific mixing technology used. These should also be filtered out.

The downside of the above method is that *it requires an oscillator frequency near RF*. Crystals that can provide these frequencies *do not exist for S-band frequencies*. Therefore a *frequency multiplier* can be used.

The simplest implementation of this idea is shown in Figure 10.1 below. The nonlinear device should be tuned to create an apparition of the original bandpass signal at the *n*th harmonic of the supplied carrier signal frequency. The bandpass filter consequently filters out the frequency components outside of the required bandwidth and leave only frequencies in the vicinity of $n \cdot f_{carrier}$.

The downside of this simple approach is that it is very power inefficient. All combinations of signal and crystal frequency will be present in the signal after it has passed through the nonlinear device. The signal strength of the multiples (high values of *n*) will become decreasingly less with *n*.

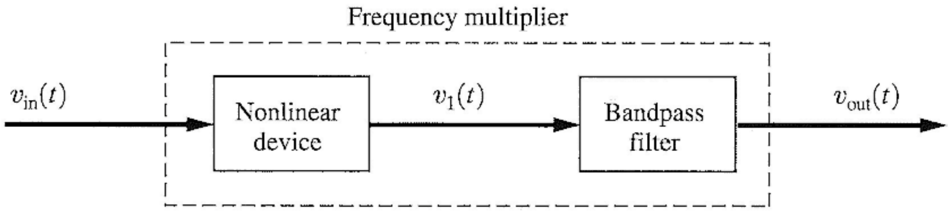


Figure 10.1: Block diagram of a frequency multiplier [SLR 0443]

The figure nevertheless illustrates the approach of how to generate high-frequency signals; a low-frequency crystal can be used of which the frequency can then be multiplied. Actual implementations of the above system will in fact use phase-locked loops and other control hardware to generate *only the frequency multiplication of interest*.

The suggested frequency multiplication circuitry is in fact included in the CC2500 package.

As done in previous chapters, the requirements for the STX Frequency Up-converter are specified in section 10.1, after which sections 10.2 and 10.3 deal with the oscillator circuit and the bandpass filter respectively. In these sections their functional details are reviewed more closely, and hardware choices are elaborated on. The priorities of the STX Frequency Up-converter are established in section 10.4.

10.1 Requirements definition

The function of the STX Frequency Up-converter has been assessed above, and its interfaces are also straight-forward. A constraint describes the RF-range that should be used. These requirements are summarized in Table 10-1 below.

Table 10-1: STX Frequency Up-converter requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.1.2-F.01	**The STX Frequency Up-converter shall convert a modulated signal to RF** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.2-I.01	**The STX Frequency Up-converter shall have an interface with the STX Modulator to receive a signal** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.2-I.02	**The STX Frequency Up-converter shall have an interface with the STX Power Amplifier deliver an RF signal** ***Rationale:*** *As explained above.*
CONSTRAINT	SAT.2.3.2.1.2-C.01	**The STX transmission signal shall be within the S-band frequency band between 2400 and 2450 MHz** ***Rationale:*** *direct derivative of parent* ***Parent:*** *SAT.2.3.2-C.02*

The frequency range within which S-band communications should take place is the *ISM band*, and its selection for Delfi-n3Xt S-band communications has been argued for in [SLR 0014]. Precise frequency coordination is underway in cooperation with the Amateur Satellite Community; the preliminary center frequency is in consultation established at a little above 2400 MHz.

10.2 Oscillator circuit

As was explained above a frequency multiplier with a customized control loop can be used for frequency up-conversion, with a far lower frequency oscillator than otherwise required. This is exactly what is required by the CC2500 if it is to be used, with a required crystal frequency of 26 to 27 MHz. Since the integration of the CC2500 is the primary design choice, these sections will cover frequency up-conversion in light of this scenario.

The suggested oscillator circuit for use with the CC2500 is illustrated in the CC2500 data sheet [SLR 0556] and is inserted as Figure 10.2. The components that make up the circuit are a piezoelectric crystal resonator or crystal and two capacitors. The capacitances of the latter components are dependent on the load capacitance of the crystal, as is further explained with some example values in [SLR 0556].

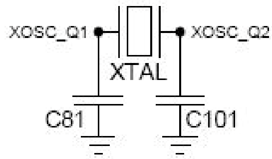


Figure 10.2: CC2500 suggested oscillator circuit [SLR 0556]

The development kit [SLR 0559] has been mentioned before, and integrates the CC2500, a crystal as well as an MSP430 processor. It uses one crystal to generate a clock signal for the processor, but at the same time to support the CC2500.

As was discussed in subsection 6.4.2, the MSP430F16X does not accept an external clock signal higher than 8 MHz. It does contain a frequency synthesizer to decrease the received clock signal for internal operations and it can have different processes operating at different speeds depending on one external clock.

The CC2500 however can *output a decreased frequency clock to the MSP430 processor*. This technique can ideally be used for the STX as well, reducing the need from two crystals to just one.

Special attention should be paid to the temperature stability of the crystal. Temperature stability is of large importance in space flight, as also indicated by the experiences with Delfi-C^3. The temperature stability is of even more importance since the crystal frequency is multiplied by a large *n*, of about 90 to 92 to be at an approximate frequency of 2425 MHz. A number of other required characteristics are mentioned in the CC2500 [SLR 0556], such as equivalent series resistance (ESR) and maximum start-up time. The operation temperature dependency is expressed to the crystal tolerance, which is set at ±40 parts per million (ppm), in other words ±40 µHz/Hz. This tolerance should also account for aging, crystal loading variations and initial crystal tolerance.

Another thing that should be taken into account in the crystal selection is the center frequency of STX operations. The CC2500 data sheet mentions the *n*/2 crystal frequencies should be avoided due to spurious signal creation, where *n* is any positive integer. In the case of a 26 MHz crystal 2405, 2418, 2431 and 2444 MHZ within the ISM S-band should be avoided. A frequency near 2500 MHz would then pose no problems.

10.3 Bandpass filter

As was indicated by Figure 10.1 at the start of this chapter, depending on the method of frequency multiplication a bandpass filter is required to filter out the frequency components outside of the range of interest. A bandpass filter can be created using a number of simple electrical components such as a capacitor and inductor in parallel, of which the impedance value should be tuned to the required frequency.

Also, a bandpass filter should be incorporated after the information signal is up-converted, as non-ideal up-conversion techniques will create images. Also, modulation with finite bit signals will create an infinite frequency spectrum, so the bandwidth should be limited if too broad.

In case the CC2500 will be used, all required filters are incorporated in the device and as such there is no likely need for extra filters. Digital bandpass filter bandwidths of the CC2500 have been presented in chapter 9. As assessed in [SLR 0014], this will cause produced spectrum to be reduced to a sufficient extent.

10.4 Work to be done for the STX Frequency Up-converter

Work to be done:

- The frequency multiplication crystal should be selected, when the transmission frequency is established
- If a modulator other than the CC2500 is to be used, it should be assessed whether no bandpass filter is required

Highest priority:

- The selection of a preliminary crystal

Clarification:
A preliminary crystal can be selected, assuming a near-2400 MHz transmission frequency (avoiding *n*/2 crystal frequencies). This crystal can then be used to try and create a working transmitter sections.

The priority of the STX Frequency Up-converter with respect to the other CIs is discussed in chapter 15.

11 STX Power Amplifier

After the signal has been converted to RF it should be amplified to a power level sufficient for transmission. As amplification will also amplify spurious frequency components (possibly non-linearly), there is the possible need for a bandpass filter after the amplification, which limits the bandwidth of the signal to a required range. Since this is the last step before transmission, the bandwidth restrictions that regulations specify should be met, possibly by the use of an (extra) filter.

Due to the project time constraints, combined with the simple fact that nobody at the moment is looking into a detailed design for a high efficiency and possibly variable supply voltage amplifier to be used at S-band frequencies, the short term vision should be focused on COTS products that are generally less ambitious. As has been explained before the STX is labelled an experiment and as such available resources do not allow for extensive research into the topic.

Section 11.1 defines the requirements for the STX Power Amplifier, after which section 11.2 moves into the topic of the actual amplifier. Section 11.3 discusses the possibly required bandwidth filter after the power amplifier. Finally, the priorities of the STX Power Amplifier are again assessed in section 11.4.

The next chapter, chapter 12 discusses the option of connecting the STX to a variable voltage bus, also in light of the COTS amplifiers that are presented in this chapter.

11.1 Requirements definition

An amplifier is a straight-forward device in its functionality and interfaces. The only performance requirement stipulated at the moment is one flowed down from a higher level, and is mentioned as a requirement here simply because after this stage it should be met. The modulator itself however can already provide a sufficiently narrow bandwidth.

The aforementioned requirement is listed in Table 11-1 below, with other and specific STX Power Amplifier requirements stated in Table 11-2.

Table 11-1: Higher level requirement applicable to the STX Power Amplifier

Category	Req. #	Requirement
PERFORMANCE	SAT.2.3-P.01	All transmitters shall provide a maximum permitted power level for spurious emissions in dBc according to 43 + 10 log (P), where P is the transmitted power

Table 11-2: STX Frequency Up-converter requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.2.1.3-F.01	**The STX Power Amplifier shall increase the power of the RF signal to a level sufficient for transmission** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.3-I.01	**The STX Power Amplifier shall have an interface with the STX Frequency Up-converter to receive a signal** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.3-I.02	**The STX Power Amplifier shall have an interface with the S-band Antenna System to deliver a signal** ***Rationale:*** *As explained above.*
INTERFACE	SAT.2.3.2.1.3-I.03	**If feasible, the STX Power Amplifier shall be connected to the variable voltage bus to receive power** ***Rationale:*** *Both the STX and the variable voltage bus are experiments, and no guaranteed solution for the connection exists. If feasible however this application should be pursued.*

11.2 Power amplifier

As was explained above a focus on a COTS power amplifier makes most sense in the current project schedule. This section provides both guidelines and a starting point to a more detailed amplifier system design, by presenting a non-exhaustive overview of a number of COTS amplifiers following some preliminary research and concluding on their possible implications on the STX system design. Some 20 theoretical options were found after which 4 most-useful options were extracted.

The ISM band in the S-band frequency range is widely being used by most notably WLAN, Bluetooth, and cordless (home) phones, next to industrial applications such as microwaves. In other words it would make sense to conclude that plenty of time has been invested in the development of high efficiency S-band amplifiers, and this is also true. All of the aforementioned applications require signal amplification, with or without a direct need for high efficiency. In order to supplement the discussion on amplifiers, subsection 11.2.1 discusses general amplifier classes and their efficiencies, and pays extra attention to switching amplifiers. Afterwards subsection 11.2.2 presents the aforementioned overview of available COTS amplifiers. Subsection 11.2.3 discusses the possible application of these amplifiers in the STX, after which subsection 11.2.4 reassesses the desired application of a switching amplifier and its consequence.

11.2.1 Switching amplifiers

Amplifiers are often classified using classes. An introduction to classes can be found in [SLR 0556]. Linear amplifiers such as class A amplifiers can only offer a low efficiency, whereas higher class amplifiers have more difficulty in maintaining linearity. Classes A to C function using a conventional transistors approach. Due to bias currents and dissipating elements the efficiencies of class A and B amplifiers are theoretically limited to 78%, with class C amplifiers possibly having efficiencies over 90% at the cost of large nonlinearities. A class A amplifier with efficiency below 20% has been used in the Delfi-C^3 RAP.

Class D amplifiers and up, or *switching mode amplifiers* can theoretically yield 100% efficiency by switching on and off all active devices at a very high frequency. Class D amplifiers work by switching the power supply on and off at a frequency typically ten or more times the highest frequency of interest in the input signal. Through a process named *pulse width modulation* the original waveform is recreated but with a higher power content. Due to this high switching frequency the application of class D amplifiers to S-band frequencies is difficult (yielding operating frequencies up to 25 GHz).

Class E amplifiers do not have this same required high switching frequency of an order higher than the application frequency, but in the process they discard the amplitude information of the original signal, while maintaining phase and frequency information. This can only be circumvented by integrating more complicated circuitry, varying the supply power in coherence with the input power such as through envelope elimination and restoration (EER). Therefore, these high efficiency amplifiers can only readily be used to amplify FM or PM signals, as opposed to AM signals. The design effort to design a high frequency switching amplifier (s-band) is sizeable, even more so to prepare it for AM signal amplification.

Mainly the technological and consequently the design complexity of these switching amplifiers are higher than lower class amplifiers, however once the design has been made switching amplifiers provide a [SLR 0014]:

- Reduction in size and mass of the amplifier
- Reduced power waste as heat dissipation, and hence smaller (or no) heat sinks
- Reduction in cost due to its smaller heat sink and compact circuitry
- Very high power conversion efficiency, usually around 90%
- Possibility of having a variable supply voltage, at least theoretically

Due to the nature of their design, class E amplifiers amplify their input to match the amplitude of the supply power. The consequence is that at least theoretically the supply voltage can be variable. In practice the design, due to the integrated electrical components, will have a certain operating range. Nevertheless, a switching amplifier could allow practical use to be made of a variable voltage bus. Operation with the latter therefore was a design target for the STX, but as will be addressed in the next chapter no longer is.

The following subsection presents a preliminary research into available COTS solutions, related to their applications.

11.2.2 Applicable COTS amplifiers

S-band amplifiers are mainly available for WLAN applications (802.11b/g and other versions), Bluetooth applications, cordless phones, and some other products such as wireless gaming controllers and model plane controllers. For all these applications, similar restrictions in terms of maximum transmitted power hold, since they all work in the ISM band and are terrestrial services. 802.11 uses a slightly larger bandwidth but still covers also the ISM bandwidth at S-band frequencies. For example class 1 Bluetooth, which is the class with the largest range of up to 100 meters, has a maximum allowed power transmission of 100 mW, or 20 dBm. The unit dBm is usually used to denote the power in the context of transmission and amplifiers. This value can slightly differ per location but the point is that amplifiers are mainly designed to tailor these applications.

The same holds for linearity; 802.11g at its highest data rates applies ODFM with 16-QAM and 64-QAM, which require linear amplification as explained in [SLR 0014]. Together with the fact that switching amplifier design complexity is high at S-band, and possibly that efficiency is not too much of a driver, this yields the conclusion that most COTS amplifiers are designed to amplify a signal linearly and to a power value of about 23 dBm, with power added efficiency (PAE) of at best 50%. Nevertheless there are some exceptions which deliver higher efficiencies, however none of the found COTS components currently available uses switching amplifying. More applications in the (higher part of the) S-band are radar and direct-to-home satellite television which together with similar high power applications leaves the realm of low power and integrated circuit (IC) technology and offers comparably huge amplifiers delivering many watts of power.

As was mentioned at the start of this chapter a short review is presented of noteworthy available amplifiers within the size and power restrictions. A larger compilation of amplifiers has been made but only the most relevant ones on the basis of a few simple criteria have been selected, being input power, output power and efficiency. A more detailed research could take into account more parameters and should focus on testing actual behaviour in our envisioned system. All amplifiers are IC amplifiers and consequently very small; not more than a few mm thick and not over 1 cm^2 in surface area. The resulting list is implemented as Table 11-3, sorted on output power and consequently inversely sorted on efficiency. A number of notes on the amplifiers are included directly beneath the table. Table 11-4 provides some conversion factors from dBm to Watts.

Table 11-3: Potential STX Power Amplifiers, sorted on delivered output power

Amplifier manufacturer	*Amplifier name*	*SLR # datasheet*	*Suggested input power (dBm)*	*Expected output power (dBm)*	*Power added efficiency (%)*
TNO	S-Band Class E PA	0437	15	35.5	25 (->50?)
Tyco M/A-Com	AM42-0055 (2 stages)	0576	5-12	33	29
Tyco M/A-Com	AM42-0055 (1 stage)	0576	5-8	30	29
TriQuint Semiconducter	CGB240	0578	3	25	48
SST	SST12LP00	0579	-5	25	58

The mentioned values for output power and efficiency are optimal data sheet values, and depend on supply power and input power as well as other operating conditions. The real values might consequently differ, as is partly assessed per amplifier below. Each of the amplifiers except for that of TNO accepts an input power of 1 dBm, which is what the CC2500 as discussed in chapter 9 delivers. A value differing from the suggested input power however implies slightly differing expected output power values and PAEs, as is also assessed below. A pre-amplifier can be used to increase the input power before final amplification. Pre-amplifiers are not uncommon components and with no direct implications for high efficiency amplifying since only small amounts of supply power are required.

Table 11-4: dBm to watt conversion for some useful values

Power (dBm)	*≈ Power (mW)*
35.5	3,550
33	2,000
30	1,000
28	630
25	315
23	200
15	31
1	1.3

The TNO S-band Class E amplifier is a switching amplifier. Nevertheless it is still under development and its specifications are preliminary, but produced by testing on actual models. The PAE of 50% is in current specifications only at 2.75 GHZ and up; for 2,4 GHz the PAE drops to 25%. As such operation at a lower frequency might also lead to unwanted amplification of spurious signals and harmonics. Discussion with TNO has taken place to discuss the amplifier's application to Delfi-n3Xt, but unfortunately no serious adaption is foreseen. Nevertheless its type will be taken along in the analysis in this chapter, to represent a possible switching mode amplifier. This amplifier is also the only amplifier of the four ones mentioned that absolutely requires pre-amplification since its required input power is considerable.

The Tyco M/A-Com AM42-0055 is optimized for either a supply voltage of 5 V or 8 V, yielding a 30 or 33 dBm output power approximately, or 1 or 2 W. With only 1 dBm input power these values will drop to approximately 28 and 31 dBm. A 5 dBm input power is designed for, and the resulting values are shown in the table. At 1 W mode and an input power of 8 dBm, output power increases by only 1 dBm to 31 dBm; higher values are off the chart shown in the data sheet. Pre-amplification is in any case preferred.

The TriQuint Semiconducter CGB240 can reach a maximum PAE of about 52%, however at an optimum input current of 3 dBm and a supply voltage of 3.4 V. If this supply voltage is changed to approach 5 V a higher output power can be reached with a slight decrease in efficiency. The input power can be decreased to 1 dBm, at a maximum power output decrease of 1 dBm. This amplifier has four cascaded amplifiers stages internally, as such lower output power values of only 7, 12 and 17 dBm typically can be delivered, by applying a certain control voltage to a single control voltage pin. Requesting these output power values however comes at the cost of low PAEs of only 10, 20 and 32% respectively.

The SST SST12LP00 is optimized for an input power of some -5 dBm, applying amplification up to some 25 dBm at a very high efficiency. It however has an indicated maximum power input of 10 dBm, but its graphs only demonstrate the effects until about -4 dBm. The expected output power and efficiency are seen to be rising right until the point of cut-off, so possible even higher output power values are reached at the cost of some efficiency.

It should be mentioned that all amplifiers have a permitted *range of supply voltages*, but the PAE quickly drops with a voltage different from the ideal supply voltage. Therefore these amplifiers will not automatically function in a practical way when connected to a variable voltage supply bus.

From the available amplifiers it can be concluded that either at efficiencies of up to and over 50% radiated power values of a little over 300 mW can be obtained using a single power amplifier, as well as values of up to 2 W with efficiencies of not more that 30% most likely, not taking into account the still uncertain TNO amplifier. High PAEs are thus intuitively inversely related to high output power values.

11.2.3 STX options

The amplifiers shown above can in theory all be used to amplify the transmission signal of the STX to a level sufficient for transmission.

As sun-pointing is no longer guaranteed on Delfi-n3Xt, the system becomes heavily power limited. In the introduction of this document it has been stated that a design goal of the STX is that:

- **The STX shall have power consumption no higher than the PTRX**

Currently, 1750 mW or 1.75 W is allocated to the PTRX in the power budget. As some amount of power will be required to power other STX components, *roughly 1.6 W can maximally be used to power the PA*.

However, it is also an option to reduce the total power consumption of the PTRX by reducing the gain of its power amplifier, given the power constraints of Delfi-n3Xt and the available link margin. Whether this shall be done is unsure; if so, the total reduction of power consumption will likely depend on the final power budget. To give an extreme case however, a power reduction of 3 dB power output will save an estimated 600 mW. In this case, the STX will have some 1 W available.

The amplifier options for the STX are concretely discussed below, starting with the two high-power amplifiers and finishing with the two low-power amplifiers.

Low-power amplifiers

The two low-power amplifiers, the TriQuint Semiconducter CGB240 and the SST SST12LP00 can only yield a power output of approximately **315 mW** singly. This comes at the cost of respectively *~650 and ~550 mW*.

This would easily fit the power requirement stated above. This can be done achieving high efficiencies. As the SST12LP00 quite simple has a better efficiency than the CGB240, the former is preferred. Unfortunately, no higher power value is delivered by these amplifiers. Another 1 W of extra amplification power is available when compared directly to the PTRX power consumption.

However, another design option exists. In theory, two identical amplifiers can be used in parallel to amplify an identical signal, after which their power can be combined. In that configuration *two identical amplifiers* can supply over 600 mW final transmission power, at the cost of about 1300 mW. The difficulty in this approach however is that the signals should be *perfectly in-phase* when re-combined. If not, losses are incurred.

The design of the antenna system however gives an ideal customization. It is suggested in [SLR 0036] that the use of two patch antennas on opposite sides of Delfi-n3Xt is preferred. The gain pattern of these patches *will not interfere* with each other. As such, *two different PAs can be used to amplify the signals sent to both patches*. At the same time, this application will allow to switch on and off the different PAs, effectively reducing the gain pattern from an omnidirectional pattern to a hemispherical gain pattern, at the gain of less power consumption. This latter option could also be applied when the PTRX will ultimately consume less power and less power is available to operate the STX.

High-power amplifiers

Both the TNO amplifier and the Tyco M/A-Com AM42-0055 can single-handedly account for a sizeable output power. In both case however, they require a power input that is *not supported by the current power budget*.

The two high power amplifiers would yield the fixed power requirements as shown in Table 11-5.

Table 11-5: HDM power requirements using COTS amplifiers

Amplifier manufacturer	***Amplifier name***	***Required power (W)***	***Output power (W)***
TNO	S-Band Class E PA	7	3.5
Tyco M/A-Com	AM42-0055 (2 stages)	6.6	2
Tyco M/A-Com	AM42-0055 (1 stage)	3.3	1

The last option presented in the table above, using only one stage of the Tyco M/A-Com AM42-0055, has the least power impacts. It allows for a reasonable 1 W output power, similar to other developed CubeSat S-band transmitters. Unfortunately, it generates this output power at the cost of rather low efficiency.

Using both stages of the Tyco M/A-Com AM42-0055 can deliver 2 W, but already requiring a very significant 6.6 W.

The TNO amplifier will be a switching amplifier. Unfortunately, as it should in case of the STX be used out of its designed band, its efficiency drops to 25%, losing a prime benefit of switching amplifiers. It could however support a variable supply voltage, but as is further argued in the next chapter this feature will no longer be pursued.

Both mentioned amplifiers would require pre-amplification, to obtain the preferred input power values of about 15 and 5 dBm. In both cases the TriQuint Semiconducter CGB240 can be used, as it has 4 internal amplifier stages. Another dedicated amplifier can be searched for as well.

Conclusions

A distinction has been made in the discussion above between the low-power amplifiers and the high-power ones. The low-power ones deliver only some 315 mW singly, but *are the only ones that are acceptable given the power budget*. Of these, the SST12LP00 dominates the other in turn of its specifications.

The suggestion has consequently been made to use two of these amplifiers in parallel. The antenna system design allows for optimal implementation of this suggestion, at the gain of switching capability.

In result, *the current suggested power amplifier for the STX is the SST SST12LP00*.

The other (high-power) amplifiers, most notably the Tyco M/A-Com AM42-0055, might become useful, for example when higher power consumption would be allowed on a future mission. They do however have very low efficiency.

The SST12LP00 should be acquired and tested for its exact functionality and performance. Other (future) amplifiers should be looked out for as well, especially if electrical integrating and/or testing will not be performed soon. The SST SST12LP00 should however be able to yield a working STX that is acceptable given the current mission constraints. As [SLR 0014] illustrates, acceptable data rates can most likely be achieved with this amplifier.

To be further investigated

- What are the exact characteristics of the SST12LP00 when tested?
- Are there alternatives to the mentioned amplifiers?

11.2.4 Switching amplification in the STX

Switching amplifiers have been introduced above, and the main conclusion is that project constraints do not momentarily allow for an in-house S-band switching amplifier design and COTS amplifiers do not readily offer switching capability. The only candidate at the moment is the TNO amplifier, but as discussed is not envisioned as a serious candidate due to its low efficiencies at 2400 MHz and its high minimum supply power.

Due to the possible higher efficiencies, switching amplifiers however remain preferred if new options turn up. If one turns up, it is likely that it will not allow for AM signals to be amplified, due to the need for increased complexity in the form a feedback loop. Therefore, the following can be stated:

It is preferred that the STX Modulator applies a constant envelope scheme, with the possible exception of on-off keying (OOK), to allow for future switching mode amplifiers that might be non-linear.

While OOK is strictly spoken a version of AM, the signal can be seen to be either off or on as opposed to of multiple amplitude levels, so that it can be amplified by nonlinear amplifiers.

The statement above is not in the form of a *requirement* as this would be too strong; no switching mode amplifier is currently envisioned. This preference will have been taken into account by an implementation of the CC2500 as explained in chapter 10.

11.3 Bandwidth filter

Requirement SAT.2.3-P.01 states that the STX Power Amplifier should deliver a maximum signal power outside of a prescribed and allowed bandwidth. An illustration of the allowed bandwidth and a value of the required reduction are presented in [SLR 0014]. In fact, based on the 0.315 W transmission output, a reduction to -38 dBc should be realized at the boundary of the allotted bandwidth. A value in dBc indicates the power difference in dB relative to the maximum transmitted signal power frequency component.

As the amplifier amplifies also spurious signals there is the possible need for a bandpass filter afterwards. And since the power amplifier is the last component in the STX where signal power content is significantly changed the requirement from the ITU on the spurious signal power levels is to be met by this CI. In this case, companies as Filtronetics and muRata deliver filters in all types of frequency ranges; use of these brands will also be made on the PTRX most likely. Filters are typically large compared to other components; nevertheless a quick search into Filtronetics catalogue of S-band RF filters delivers sizes of not more than 5-6 mm typically.

However, [SLR 0014] takes into account the bandwidth produced by the CC2500 and argues that this bandwidth is likely acceptable over the radio amateur S-band. Also, the required bandwidth can be reduced by adjusting the digital channel filter bandwidth in the CC2500 internally. As such, it can currently be assumed that no additional bandwidth filter is required.

11.4 Work to be done for the STX Power Amplifier

Work to be done:

- An up-to-date review should be done of available amplifiers
- The SST SST12LP00 and possibly the CGB240 should be bought and tested for their actual characteristics, such as efficiency and output power, with varying input power and possibly varying supply power
    - The amplifier(s) should be selected
- Depending on frequency bandwidth allocation, the need for an extra bandpass bandwidth filter should be reassessed

Highest priority:

- An up-to-date review of available amplifiers
- The acquisition of most notably the SST SST12LP00 and its consequent testing in a transmitter section

Clarification:
In order to create a working STX transmitter section, a working amplifier is essential. Therefore an amplifier, most notably the SST SST12LP00 should be bought and analyzed in order to yield this transmitter section. A short up-to-date review and consequent/simultaneous purchase of a number of other amplifier options can give the concrete alternatives.

The priority of the STX Power Amplifier with respect to the other CIs is discussed in chapter 15.

12 The variable voltage bus and beacon mode

When the idea of the STX was conceived, it was influenced by the fact that:

- Current (UHF/VHF) transmitters on CubeSats do not exceed 9.6 kb/s data rates
- Delfi-n3Xt should allow for pointing, allowing even higher data rates to be achieved

As this newly envisioned transmitter was brought into life exactly to outperform its older brothers on the UHF/VHF band, a number of ideas evolved from each other:

- A *high-efficiency switching mode amplifier* could be integrated
- As power variability depending on the pointing capabilities could become high, and as this type of amplifier can support a variable supply voltage, the use of a *variable voltage bus* was suggested to amplify the transmitted STX signal
- As the STX would not be continuously fully operational, but the power variability would be continuous, an additional operational mode named the *beacon mode* was envisioned.

Of these suggestions, the switching mode amplifier has already been discussed in the previous chapter. Unfortunately, no COTS component seems to be available and ready for implementation on the STX. As in fact the ideas evolved from each other, this could logically lead to exclusion of the other options.

This chapter explains indeed why the other two options are no longer supported in the current design of the STX. The variable voltage bus is commented on in section 12.1 followed by a discussion on a beacon mode in section 12.2.

12.1 The variable voltage bus

A variable voltage bus would allow to get rid of excess power. As generated power on a satellite is not constant, it is desirable to use this excess power for a functional means as critical operations will be designed for minimum available power. The STX and its possible switching mode amplifier would theoretically present this means, allowing the excess power to be used to quite simply amplify the transmitted signal. A number of significant arguments can however be stated against connecting the STX to the variable voltage bus.

These arguments are brought forward as subsections 12.1.1 through 12.1.3 below. Section 12.1.4 will afterwards conclude on the variable voltage bus in light of the STX.

12.1.1 The need for an excess power handling system on Delfi-n3Xt other than the STX

The design of the excess power dumping system should be such that it always works, and also is single-point-of-failure-free. Even if the STX dumps all excess power through its transmission, it can never be the only power dumping system, unless it is fully redundant. This will not be the case. Also this would mean the STX could not be switched off; by current protection systems or by the required option to shut off the transmitter due to regulations.

As such, while the STX could optionally yield a functional usage of excess power, it can never be the only option of dealing with excess power. <u>As such, some other method of dealing with excess power on Delfi-n3Xt will be required in any case.</u>

Two logical options would exist:

- A power shunt on a strategic place on the Delfi-n3Xt satellite could radiate excessive power in the form of heat. The thermal consequences as well as of course its location of having a shunt will have to be assessed carefully by the responsible TCS systems engineer. In this case the variable voltage bus can still be included on Delfi-n3Xt as excess power will have to be delivered to the shunts, and possibly to the STX if feasible.
- Power can be left on the solar panels. The effect of a similar use of the solar panels should be investigated, but it would pose a simple solution if indeed power can simply be left without serious consequences. In this case a variable voltage bus is not necessarily required anymore, other than for possible probing functions, if no power is delivered over it to the STX.

Nevertheless the argument that the excess power is more desirable to be used for a functional means, such as increasing transmission power still holds. Therefore an STX power use is preferred, but can only be implemented as an *extra feature*, requiring extra resources. The same holds for the variable voltage bus itself, if power can simply be left on the solar panels. In this case the variable voltage bus itself is not required.

12.1.2 The limited power generation of the redefined Delfi-n3Xt

Based on pointed Delfi-n3Xt operations, the power left to be carried over the variable voltage bus would be anywhere between 0 and 8 W. This was due to up to 7 W of contingencies and left-over power in satellite modes [SLR 0017], but also due to a low assumed pointing accuracy of 25°. Without pointing however, power generation on Delfi-n3Xt falls back to some 2-4 W, also drastically decreasing the possible variability in available power, as well as its possible use. In this case, it seems power can easily be left on the solar panels as addressed in the last section.

12.1.3 The limited functionality of available COTS amplifiers

Since project constraints lead to the choice of COTS amplifiers, no smart variable supply voltage amplifier will be designed. A number of COTS amplifiers have been introduced in the last chapter, and will be used here to draw conclusions on the usability of a variable voltage bus.

First of all, in directly available COTS components, there is no smart variable supply voltage amplifier. Next to possible technical and functional complexities, a reason for this is that this function is simply not desired in for example WLAN or Bluetooth applications; the main 2.4 GHz amplifier design drivers. The TNO amplifier is the only amplifier found, but it is still in development. However, TNO has produced several working models and the preliminary specifications are based on them. Efficiency is likely to be low (25% as opposed to 50% for 2.4 GHz signals), but the amplifier should allow for the variable voltage bus to be effectively used. It does have very sizeable power requirements, being designed for a minimum power of 7 W. Given the fact that power variability on Delfi-n3Xt cannot provide this 7 W, the battery should be used, thereby *removing any benefit of using the excess power for transmission as it could just be used to power the battery*.

Second of all, a smart solution using the COTS amplifiers can theoretically be found. Such a solution could make use of the fact that COTS amplifiers can have a large accepted input voltage range. Usually this accepted input voltage range is 2.7 to 3.6 V, but some amplifiers such as the TriQuint Semiconducter CGB240 offer a range of 2.0 to 5.5 V. It was mentioned above however that these amplifiers do not deliver a serious change in output power related to a changing power supply. If the supply voltage is raised above the optimal supply voltage, efficiency quickly drops and the extra added power is thus simply turned into heat, having the amplifier function as a power shunt. In other words deliberate use can be made of the accepted supply voltage range to use up left-over power, but this power will never seriously go into a stronger transmission. Using this functionality it could be possible to functionally use a variable voltage bus, by cascading multiple amplifiers, *but only by effectively displacing the power shunt capability*. Nevertheless, if multiple amplifying steps can be used, multiple power outputs can be realized.

In order to illustrate this point, the possibility of the Tyco M/A-Com AM42-0055 as well as the TriQuint Semiconducter CGB240 to functionally use a variable voltage bus is described below.

The Tyco M/A-Com AM42-0055 its supply voltage range superficially matches the old possible variable supply voltage of 0-8 W the best. Too little about the amplifier design is known to give concrete conclusions, but it is certain that the amplifier can accept a supply voltage of 5 V as well as a supply voltage of 8 V, over two different pins. The maximum accepted supply voltage is 10 V. Therefore possibly the amplifier can accept almost the entire supply voltage range, divided by a switching system over 1 W or 2 W operation. This use of the variable voltage bus could be perfectly in line with normal STX operations, as no other components would be required. However, the new Delfi-n3Xt definition will no longer yield up to 8 watts of power over the variable voltage bus, only very low values too low for the AM42-0055 to function**.**

The TriQuint Semiconducter CGB240 has a supply voltage range of 2-5.5 V, and four separate stages of amplification which should provide some flexibility. An analog control voltage selects the number of stages activated. Therefore the amplifier yields the power requirements as shown in Table 12-1.

Table 12-1: TriQuint Semiconducter CGB240 power requirements

Stages	***Required power (mW)***	***Output power (mW)***
1	50	5
1,2	80	16
1,2,3	155	50
1,2,3,4	400	200

Due to the high supply voltage range and its consequent effect on efficiency, the CGB240 can theoretically handle all power supply inputs varying from effectively 0 to 400 mW. The datasheet of the CGB240 [SLR 0578] gives however even more information with respect to the maximum supply power, as the CGB240 also accepts an higher supply voltage at the cost of efficiency. Table 12-2 shows the power supply range for the case in which all four stages are used, calculated using given PAEs and power output values for a supply voltage of 2.0 and 5.0 V. No data is given for when the supply voltage is over 5.0 V.

Table 12-2: TriQuint Semiconducter CGB240 supply power range, all four stages

Supply voltage	***PAE (%)***	***Output power (dBm)***	***Output power (mW)***	***Required power (mW)***
2.0	38	18.5	71	187
5.0	48	25	316	658

Consequently it can be seen that using only the CGB240 a variable power supply of 0 to about 650 mW, probably slightly higher for a supply voltage up until 5.5 V, can be reached. The delivered output power is above 300 mW, which should be able to yield a minimum data transfer. If at this the possibility exists that multiple of these amplifiers can be connected in parallel, a larger amount of power can possibly be accepted by the system.

In any case this usage of the CGB240's variability would require some regulation to provide a proper control voltage to select the stages used. However it is most likely also possible and more practical to just use all four stages, and have the amplifier(s) be inactive if supply voltage is below some 180 mW. Control voltages can possibly be taken care of by the MSP430F16X. The CGB240 could be the same physical amplifier that provides pre-amplification in normal STX operations anyhow, nevertheless this would probably increase complexity unnecessarily as amplifiers sizes and mass are of little relevance.

As a result, the implementation of the CGB240's variability would require a separate electronic system to be designed, different from the normal STX transmission path used for TM mode. As a consequence, this application would again add an extra feature to an already experimental STX.

12.1.4 Conclusions

In the above three subsections the main arguments against having a functional variable voltage bus in combination with the STX have been presented. In summary:

- The implementation of a variable voltage bus connected to the STX will not allow for the elegant implementation that was once envisioned, as another system to take care of excess power will still be required.
- The new definition of Delfi-n3Xt has much less use for a variable voltage bus, as much less power will be available on this bus.
- COTS solutions do not offer the required characteristics, as no real functional use can be made of the delivered range of power. Even though some smart solutions can be constructed using the available COTS amplifiers, this would allow for very limited actual use at the cost of increased required resources, while at the same time displacing the power shunt capability to the STX.

As a result, it is suggested that *no connection shall exist between the STX and the variable voltage bus, if the latter will even be incorporated on Delfi-n3Xt*.

12.2 Beacon mode

As a variable voltage bus is no longer assumed to be useful for STX operations, if makes sense to assume the beacon mode or BM is no longer useful as well. Indeed, this is the case. But where a BM was supposed to use the variable voltage bus and thus the changing amounts of excess power, it also represented the option of having a continuously transmitting STX during satellite operations, allowing radio amateurs to receive a signal over the S-band, at other moments than during Delft passes.

As no excess power will be simple available over a separate bus, a CDHS decision will have to be made to switch on any STX mode. This will no longer automatically create an STX beacon when power happens to be available, but power should be monitored and the STX can be activated. It then becomes a choice of having the STX switched off and *having it switched to normal TM mode.*

If then it is desirable to have a relatively high power signal received on-ground, for example to cater radio amateurs with less powerful equipment, a low data rate can be used according to one of the data rate presets of the TM mode. And, of course, the STX can be activated using this low data rate even when not above Delft.

For a beacon mode it was suggested that it might be nice to transmit data other than TM data in case of this BM, as a low-power and continuous signal would favour other options. As such the transmission of a real and very low-power actual beacon was suggested, or names of radio amateurs which have participated in TM forwarding. However, due to limited power architecture of the STX (in terms of flexibility) and the sporadic nature of the TM mode activation when not above Delft, it is suggested that TM data is simply transmitted, as in normal TM mode operations. Again, this is in line with normal TM mode operations.

13 Integrating the components

The six configuration items as reviewed in detail in the six chapters 6 through 11 together make up the STX as defined in the configuration item tree [SLR 0325]. Per CI the interfaces to other CI's as well as to different subsystems have been reviewed. This chapter discusses their integration to one working system, also by paying attention to other components required.

Section 13.1 concludes from all chapters until now the most likely hardware lay-out of the STX. It presents a hardware diagram required to fulfil all top-level functionalities. Section 13.2 comments on the electrical interfacing furthermore required.

Sections 13.3 and 13.4 will consequently discuss the physical integration of the STX. Since the STX is (per default) to be integrated on one PCB, this PCB and its exact contents are discussed in Section 13.3. A quick consideration of more exotic integrations of the STX is given in section 13.4.

Finally, as another method of integrating components and functionalities, the option to use an FPGA in the STX is assessed in Appendix A, including the reasoning as to why use of an FPGA is not feasible in the Delfi-n3Xt STX. The use of an FPGA is also linked to the possible design of a modular baseband system.

13.1 The STX component lay-out

In chapter 5 the hardware components of the STX were derived from its (sub)functionalities, and consequently the CIs were derived from these components. As the CI's have been discussed in more detail this subsection derives again the important hardware components independently from CI structure to focus on their interconnection.

The resulting component diagram represents the most likely or appealing configuration, see Figure 13.1. A name or concrete description of the most likely component per component type is given. In other words, *Figure 13.1 presents the currently suggested STX component lay-out*.

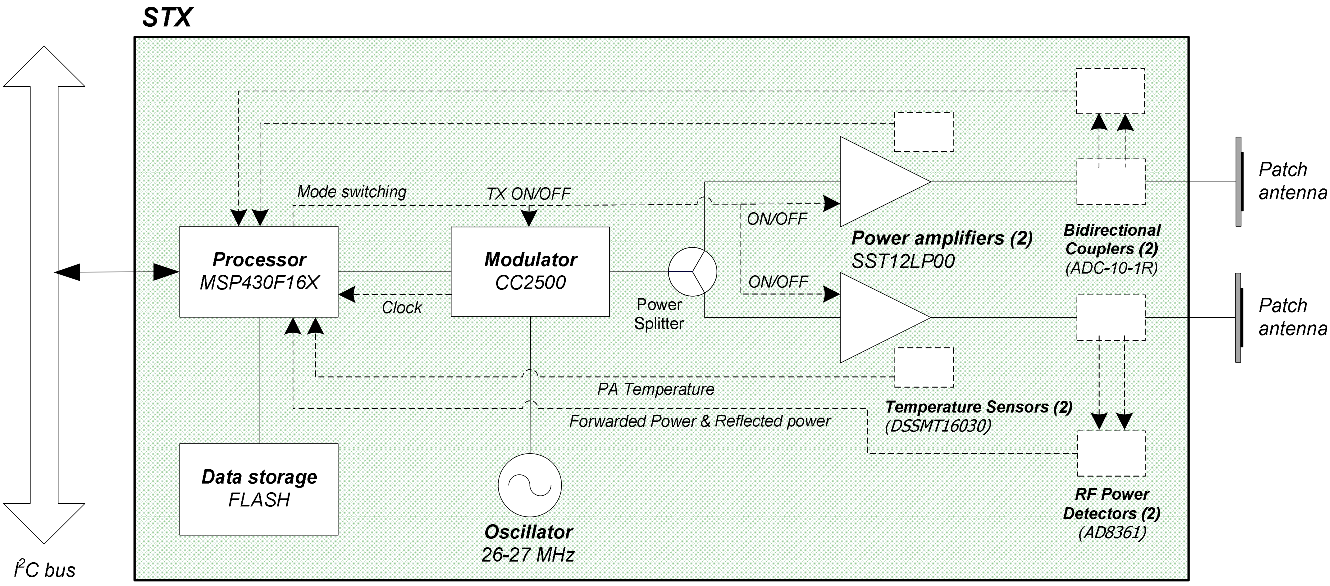


Figure 13.1: STX hardware components and their interfaces

The figure above is of course subject to change, as more research into each of the components is done.

From what is known about all of the components, this arrangement would at least yield a working STX, making both a *TX OFF mode* and *TM mode* possible as the CC2500 transmission can be switched on and off as indicated in the figure. At the same time, the power amplifiers can be switched on and off by the processor. According to the datasheet of the suggested PA [SLR 0579], it has an integrated on/off pin. The fact that two PAs are used gives flexibility as to what PA is activated, if not both. This allows lowering the overall power consumption, possibly without a reduction of radiation performance as each patch antenna radiates only to one side of the satellite [SLR 0036].

Also indicated in the figure are the signals delivered to the processor, being a clock signal from the CC2500 as well as various housekeeping sensor data signals from the various sensors.

Concluding from the above figure it can be said *that all STX design goals have been met*. These were stated in the introduction of this document:

- **No pointing shall be required for the technology demonstration**
- **The STX shall have power consumption no higher than the PTRX**
- **COTS components shall be used wherever possible**
- **All components in the design should be as modular as possible**

First of all, [SLR 0014] shows that moderate to high data rates can be achieved using the configuration demonstrated above, without relying on pointing.

Second of all, using the two PAs as suggested in the figure will indeed ensure that the STX will use less power than the PTRX, as currently anticipated. The next chapter shows that there is still sufficient margin. At the same time, it is possible to switch on only one PA; reducing the power consumption of the STX by some 45%.

Third of all, it can be seen that all components used are indeed COTS components. It will therefore likely require minimum effort to integrate these suggested components.

Finally, modular components have been used. *Almost every CI has been integrated as a single component*. This means that all components can in principle be changed without affecting the others:

- **Processor**
  If available processor power turns out to be limited, or a different default processor would be chosen in a current or future mission, this component can be replaced
- **Data storage**
  If a different data storage type or size is required, this component can be replaced
- **Modulator**
  If the modulation scheme should be changed, or the maximum data rate turns out to be a limiting factor, this component can be replaced
- **Power amplifier**
  If a future power budget or a future mission allows for more power consumption, or if a more advantageous power amplifier becomes available, this component can be replaced

13.2 The STX electrical interfaces

In order to implement all of the components as reviewed in the previous section supporting electrical components are necessary to glue everything together. A suggested electrical interface for the CC2500 has been presented in subsection 9.2.4, and pin configurations for all of the suggested components are available. First essential electrical integration should take place to yield a fully functional STX. Afterwards, final circuit board design can take place.

13.3 The STX PCB

In Delfi-n3Xt payloads and subsystems are in general allocated some volume, usually surrounding a PCB. The printed circuit boards (PCBs) will be custom designed per subsystem or payload, and will connect all subsystem components electrically as well as structurally. The PCBs are stacked in the structure of the satellite to form Delfi-n3Xt. [SLR 0169] deals with the STS or Structural Subsystem in detail.

The STX has one PCB allocated to it. Its volume as well as its mass are assessed in chapter 14. Figure 4.1 shows the STX actually encompasses more than just the 6 CIs as discussed in detail so far, representing the STX. In order to function in the satellite a number of components should be incorporated on the STX PCB, as was discussed in chapter 4. These components in terms of CIs are the *Local EPS @ STX* and the *Local CDHS @ STX*. The first CI falls under the responsibility of the EPS systems engineer, whereas the last falls under the management of the CDHS systems engineer. Together, these components belong to the *standard system bus*.

Within the COMMS system a number of other components are essential to the functionality of the STX; the antenna system components. The Antenna System is discussed in [SLR 0036]. As argued in this document and suggested by Figure 13.1 above, the antenna system should in fact only require a cable connection between the signal resulting from the Pas (or sensors) and the patch antennas. As such, the STX PCB requires *Antenna Wire Connectors* to connect the wires to.

In conclusion the CIs that are incorporated to form the STX PCB are shown in Figure 13.2 below. The colours indicate subsystem responsibility, as is further explained in [SLR 0325].

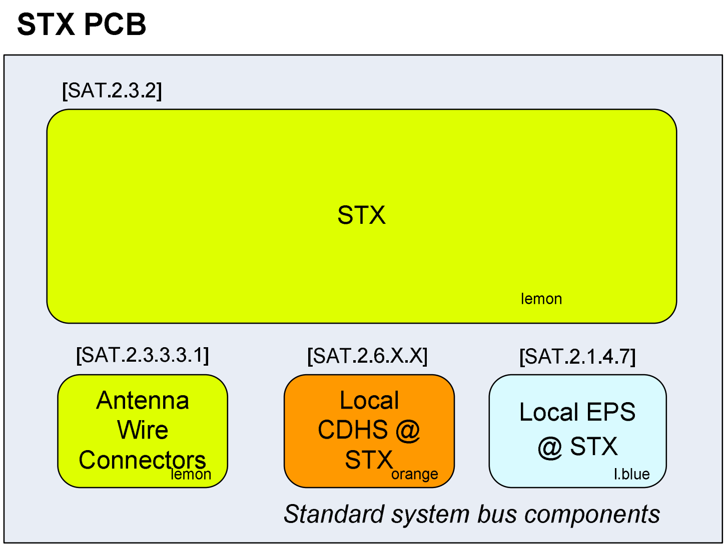


Figure 13.2: CI integration on the STX PCB

A certain area of the actual PCB cannot be used to allow for structural connections and the integration of bus connectors, as is the case for any PCB.

Possibly and likely a cooling fin shall also be added to the STX PCB as the power amplifiers will generate sizeable heat. The cooling fin will fall under the responsibility of the thermal control subsystem, or TCS. No lower level CIs have been specified for the TCS as of yet.

13.4 STX integration without a PCB

A more exotic solution for the STX would be to integrate the entire system on a circuit board other than a PCB. As the STX involves a relatively low number of components as has been shown above, this might be possible. It does not make sense to integrate the STX on a PCB other than its own, for example to save space. Delfi-n3Xt currently does not have any mass and volume issues. Also, this would make the combined systems less modular.

Generally it does present some advantages to integrate the STX somewhere close to where its antennas might be. This would reduce any additional losses that might be incurred due to long cable lengths; currently some -0.3 dB is included in the link budget to be prepared for a worst case cable length. As the antenna patches require a ground plane, the components can be attached *directly to this ground plane*.

If integration near its own antenna is achievable, instead of having one STX (with two amplifiers) and multiple antennas, it can be chosen *to have multiple STXs with their own respective antenna systems*. This would create a nice 'all-in-one' package.

Of course, two aspects become more difficult for these propositions:

- Mechanical connection
- Connection to the standard system bus

In all cases it would require a customized connection. As this would likely come at the cost of additional required resources with respect to the PCB integration case, it does not seem preferable. Also, for the proper technology demonstration of the STX and its essential components, integration on a PCB would be without problems. Also, PCB integration can be seen to be in fact very modular given the CubeSat standard. Finally, Delfi-n3Xt is not volume and mass limited.

However, to increase the compactness or flexibility of the STX (with respect to future missions or even other satellites), a more exotic integration can be reinvestigated in the future.

14 The STX in the Delfi-n3Xt budgets

The STX is forced to comply with a number of budgets. With regard to communication only, the operations of the STX should comply with the link budget, so that a proper connection can be established between Delfi-n3Xt and the ground station(s). All systems on Delfi-n3Xt however should comply with the volume, mass and power budgets. As such these three budgets are elaborated on in light of the STX system, in sections 14.1 through 14.3 below. All of these budgets focus on the use of an STX PCB, including the components as assessed in the previous chapter.

14.1 Volume budget

The volume budget yields some requirements to which the STX should comply, which can be seen as areal requirements and height requirements:

- All STX components and electrical connections should fit on the usable area of the allocated PCB.
- The height of all STX components should allow for them to be integrated on the allocated PCB.

The outer edges of each PCB side form an area of 90 by 90 mm. Nevertheless to allow for structural connections and bus connectors the exact usable area is smaller, and defined in [SLR 0169]. A qualitative discussion suffices here.

A PCB in principle has two usable sides. Whereas large components will almost always require a through-board connection, electrical connections can run on both sides of the PCB simultaneously. Nevertheless the STX shall most likely be designed to fit on one side, simply because the required area to house all components and connections together is not assumed to be a serious restriction. In comparison the Delfi-C^3 RAP used two sides, one for the transmitter or TX side and one for the receiver or RX side. As the STX incorporates just a TX system, one side could logically be enough. All components selected so far do not seem to suggest differently. Detailed PCB design of course has to take place at a later stage.

The height of the STX system is by the volume budget [SLR 0303] defined to be 14 mm; see Table 14-1.

Table 14-1: Height restriction of STX components

Allocated STX height	14.0 mm
PCB thickness	*-1.6 mm*
Soldering connections on the PCB side not in use	*-3.0 mm*
Maximum STX component height	***9.4 mm***

In the case of the RAP and also the PTRX as currently under investigation, the largest components are the filters and the crystals. The COAX connectors in the case of the RAP were vertical connectors, and as such took up unnecessary height. These connectors will most likely be replaced by horizontal connectors or smaller (micro) connectors; requiring anywhere between a few mm to about 8-9 mm. As was mentioned in chapter 11, a quick search for RF bandpass filters in the required frequency range yields components not higher than 5-6 mm, and these are likely not even required. Due to the use of the CC2500 no other IF filters should be required. A similar quick search for crystals of Filtronetics yields 26-27 MHz crystal packages not higher than 4-5 mm. Finally cooling fins might be required, which can be put at the side of the PCB design (along with the Pas) in order to 'stick out', or if the height restrictions allow for it they could be located anywhere else.

It can be concluded that the allotted height restriction likely should not pose problems for the STX design and at the same time leave sufficient margin. Also, since the volume budget was conceived Delfi-n3Xt is no longer volume limited so the volume available for the STX could be increased somewhat, if deemed necessary.

14.2 Mass budget

The current STX PCB budget as incorporated in the mass budget has a maximum mass of 123 grams. This value equals the RAP final mass. As the RAP incorporated a larger number of components since it was a transceiver, the mass of the STX should logically be less. The only concrete component type not on the RAP compared to the STX is data storage, as opposed to all RX and transponder related components not necessary on the STX. Some antenna circuitry might also be required on the STX PCB.

The mass budget consequently could therefore incorporate a better estimate, or at least one that is better arguable. The number of uncertainties nonetheless still would be very large at the moment. Also there are, as for the volume budget, no hard arguments to think that certain components will likely cause to violate current estimates, and mass on this scale is not very mission critical. As more knowledge on components is known detailed estimates can and should be made.

14.3 Power budget

As opposed to the volume and mass cases, up-to-date power estimates are of continuous importance in the Delfi-n3Xt, due to its power limitations and different switchable systems. As such, all mode scheduling takes into account all individual subsystem and payload power consumptions.

The active components of the STX during its operations can be said to be the preliminary components as specified in Figure 13.1. The additional components of the standard system bus are taken into account separately in the power budget [SLR 0017].

As there are two operational modes, there are also two different STX power consumption totals. However, the option to switch on just one PA leads to a third. The following Table 14-2 gives for each of the probable or possible STX components an estimation of power consumption, as well as the modes in which they are active; *1 indicates TX OFF mode and 2 indicates TM mode*. All components are logically active in TM mode, whereas in TX OFF mode no amplifiers have to be active and can be switched off. In the latter mode the CC2500 can just pass on the clock signal to the processor, yielding a minimal CC2500 power consumption. The sensors are logically always active as they continuously provide a check of correct operations, they have low power consumption and finally no on/off interface with the processor is then required.

Table 14-2: STX PCB components power estimation; 1 TX OFF mode, 2 = TM mode

Component type	***Component***	***Mode(s)***	***Power consumption***	***Argumentation***
Processor	MSP430F16X	1,2	~2.0 mW	Maximum current consumption at supply voltage of 3.3 V [SLR 0581]
Data storage	*TBD*	1,2	~60 mW	Assuming FLASH memory, worst case (write power). Reading only would be less, and neither would cost no power [SLR 0311]
Modulator	CC2500	1,2	(1) 22 mW (2) 65 mW	(2) Only frequency synthesizer running (1) Maximum transmitter power at maximum power supply [SLR 0556]
Oscillator	*TBD*	1,2	100 μW	Estimate from the Filtronetics website
Power amplifier (2)	SST12LP00	2	543 mW	Assuming an -1 dBm input power and 58% efficiency [SLR 0579]
PA temperature sensor (2)	SSMT16030	1,2	1 mW	Consumption at supply voltage of 5 V (indication) [SLR 0692]
RF Power Detector (2)	AD8361	1,2	3.3 mW	Consumption at supply voltage of 3 V (indication) [SLR 0110]

[SLR 0643] provides the results of a measurement of the consumed power of the ensemble of the CC2500 and the MSP430 processor; perfectly in line with the above numbers (65 + 2 mW) it gives a value of 65.4 mW.

The above power consumption values can be added in coherence with the modes in which they are active to yield power consumption totals per mode. As contingencies in the actual power budget already take into account insecurities of the values, no estimations will be made for additional electronics that are potentially necessary. Some DC/DC conversion losses will however be added, assuming a conversion efficiency of 85%.

Consequently the estimated power consumption per STX mode is given in Table 14-3.

Table 14-3: Approximate STX power consumption per mode

Mode	***Power consumption***
TX OFF mode	110 mW
TM mode (1 PA active)	799 mW
TM mode (2 PAs active)	1447 mW

As mentioned above, the actual power consumption required to operate the STX will be slightly higher due to the additional PCB components which are required to operate the system related to the standard system bus.

15 Next steps

This document has so far presented the functionalities and requirements of the STX, and combined this with required theory and available hardware options to derive design options. As such it can be said that a preliminary STX design has been established. This chapter provides the concrete next steps to be done to ultimately end up with a complete and working STX system on Delfi-n3Xt, based on the work to be done per CI presented in the end of each CI chapter.

On a level different of the STX itself, the S-band Antenna System [SLR 0036], the Delft S-band ground segment [SLR 0014] and the confirmation of the S-band transmission frequencies and bandwidth are also required to ultimately having a working S-band communication link between Delfi-n3Xt and the Delft GS.

All STX CIs are incorporated in Figure 15.1, with an indication of their priority level. These three priority levels also correspond to the three steps that are required to have working STX on Delfi-n3Xt.

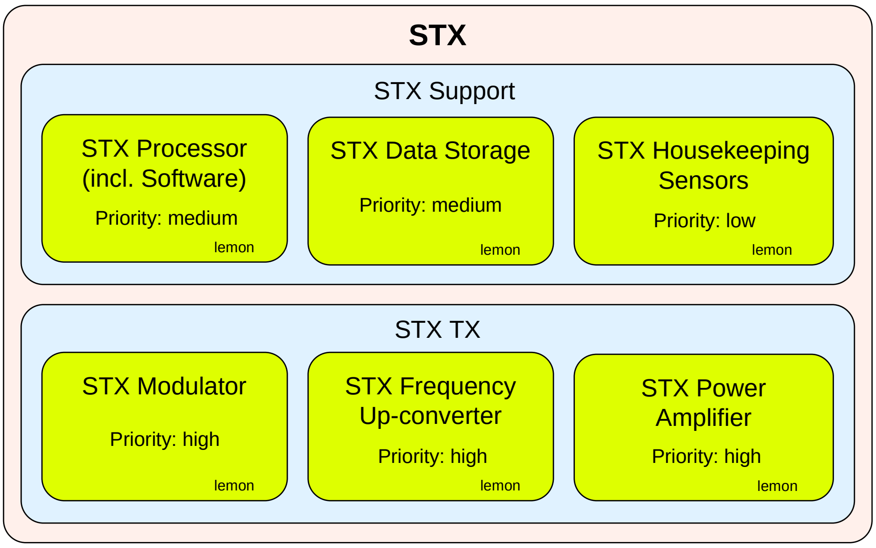


Figure 15.1: STX priorities

The three steps towards the Delfi-n3Xt STX are commented on below. In all cases electrical and functional integration and testing should be performed, in order to verify the correct operation.

Step 1: STX TX – High priority

The number one priority within the STX lies with creating a working transmitter, or STX TX. In order to do so, the hardware diagram as presented as Figure 13.1 in chapter 13 can thereby be used and followed.

The STX Modulator and STX Frequency Up-converter are integrated, as explained, when using a CC2500. In that case just a crystal and some simple circuitry are required. In testing their use, the MSP430F16X can be used.

STX Modulator - most important:
- The confirmation of the CC2500 usage
- The integration of the CC2500 is a working transmitter section

STX Frequency Up-converter – most important:
- The selection of a preliminary crystal

STX Power Amplifier – most important:
- An up-to-date review of available amplifiers
- The acquisition of most notably the SSMT16030, and its consequent testing in a transmitter section

Suggested approach
The components most important to a working transmitter are the ones stated above. A quick up-to-date reassessment of the CC2500 should confirm its usage, or give a suitable replacement. A significant change of component is not advised, given the work performed on analyzing and testing the CC2500 already. An updated model could however be used.

A similar quick review of up-to-date alternatives to the PA should be performed and the resulting PA should be acquired. A preliminary crystal, which is not required to be a highly stable or space-proven one, can be acquired to provide the required oscillator signal. Then, any MSP430F16X processor can be used and the resulting set of components should be interfaced to result in an ultimate working transmitter section.

In this manner, data storage is not yet required, neither is the confirmation of the exact MSP430F16X processor or the writing of its complete software. Logically, this forms the next step.

Step 2: STX Processor and Data Storage – Medium priority

After a working transmitter section has been established, the focus can shift to creating the brains of the system, being the processor along with its data storage capability. This in turn will basically create a working STX. The STX should in this phase however not yet be customized for Delfi-n3Xt, in terms of operational software and final PCB integration. The focus at first shall remain on creating a working STX.

STX Data Storage:
- The selection of the data storage device

STX Processor:
- The selection of the processor (MSP430F16X), depending on the choice of data storage
- Writing the basic operational software

Suggested approach
Continuing from work to be presented by [SLR 0300], a specific data storage device should be selectable relatively straight-forwardly. Some research should be done with respect to radiation resistance at the same time, and the interfaces with the MSP430F16X can be tested afterwards. Having selected the type of data storage and having a working transmitter section, gives all the constraints required to set the processor type.

Step 3: Finalization of the Delfi-n3Xt STX – Low priority

The final step towards having a working Delfi-n3Xt STX is the customization effort required to prepare the STX for integration on Delfi-n3Xt, assure the proper operation on Delfi-n3Xt and assuring bandwidth and transmission frequency requirements are met by means of having proper filters and a stable oscillator. The required actions:

- The exact crystal type should be confirmed or selected and implemented
- The exact required housekeeping variables to be measured should be confirmed and/or determined
- Depending on the variables, the hardware to take care of these measurements should be selected
- It should be determined whether additional filters are required to meet bandwidth specifications
- All components should be integrated on a PCB adhering to all system specified requirements
- The CC2500 software should be written, with the specified submode data rate presets included
- The operational software of the STX Processor should be written
- STX integrations tests on Delfi-n3Xt should be performed

Appendix A Implementing an FPGA

A design option has been suggested to implement a field programmable gate array, or FPGA. FPGAs provide certain benefits over discrete component designs, most notably that of design and possible operational flexibility. This appendix gives an introduction to FPGAs, and provides a comparison to application specific integrated circuits (ASICs). The knowledge on FPGAs in this section is derived from [SLR 0384], unless stated otherwise. This section provides by no means a detailed research into FPGAs; only noteworthy specifications are discussed in relation to a possible implementation in the STX.

The conclusion however is that FPGAs shall not be used in the STX system.

Subsection A.1 introduces the concept of an FPGA, after which subsection A.2 presents the different FPGA types. Subsection A.3 draws a comparison to ASICs, and finally subsection A.4 points the topic to the STX and draws conclusions on the usability of FPGAs. The design option to produce a modular baseband module is also assessed in light of FPGAs in the last subsection.

A.1 Field programmable gate arrays (FPGAs)

As the name indicates, FPGAs are field programmable, in other words they can be configured and reconfigured ‘in-the-field’, after production. FPGAs consist of a very sizeable number of gates, or transistors. Using a hierarchy of reconfigurable interconnects the gates can be wired to perform complex processor functionalities as well as represent simple logic. Software can be written in a language such as Verilog or VHDL, after which this software can be used to configure or reconfigure an FPGA. Consequently the biggest benefit of an FPGA is the design freedom it offers.

Due to their flexibility FPGAs are nowadays widely used in prototyping systems, and consequently also in final system designs. The FPGA configuration can also be changed during operations, as a means to change the functionality.

Implementing a program for an FPGA can range from relatively straight-forward to being very complicated. Programs can be written by oneself, but also preconfigured modules or intellectual property (IP) can be bought or obtained and be implemented. An FPGA configuration to implement for example processor functionality can be obtained. Also FPGAs usually have these common functionalities already implemented. An optimal program is written especially for the FPGA type and brand at hand; developing a complete configuration by oneself is very complicated and time consuming.

A.2 Types of FPGA

Three types of FPGA are used nowadays. The three archetypes are explained in [SLR 0384], and the following summarizing Table A-1 is adapted from the same source.

Table A-1: FPGA programming technologies [adapted from SLR 0384]

Feature:	**SRAM**	**Antifuse**	**E2PROM/FLASH**
Technology node	State-of-the-art	One or more generations behind	One or more generations behind
Reprogrammable	Yes (in system)	No	Yes (in system or offline)
Reprogramming speed (incl. erasing)	Fast	-	3x slower than SRAM
Volatile (must be programmed on power-up)	Yes	No	No (but can be if required)
Requires external configuration file	Yes	No	No
Good for prototyping	Yes (very good)	No	Yes (reasonable)
Instant-on	No	Yes	Yes
IP security	Acceptable	Very good	Very good
Size of configuration cell	Large (six transistors)	Very small	Medium-small (two transistors)
Power consumption	Medium	Low	Medium
Radiation hard	No	Yes	Not really

The radiation hardness of only the antifuse FPGAs is said to be good, which is an important requirement for space application. This avoids the use of complicated and/or bulky shielding. While SRAM FPGAs are currently state-of-the-art they also require reconfiguration every time after power up. And while E2PROM/FLASH FPGAs have a similar technological back log as Antifuse FPGAs they are reconfigurable. Only antifuse FPGAs however have low power consumption and are thus radiation hard. Also it mentioned in [SLR 0384] that antifuse FPGAs can achieve somewhat higher speeds, but this can be counteracted by the technological back log. And antifuse FPGAs lack the reconfigurability of the other types. Size and consequently mass differences are less easily pronounced.

Very intelligent design of FPGA software, taking into account the exact hardware of the specific FPGA to be configured can lead to maximum processing speeds, minimum power consumption and minimum size. Nevertheless, for the current discussion aimed at relatively low-complexity designs this argument shall be discarded. The resulting conclusion is that FPGA software can be written in a language such as Verilog or VHDL, after which any FPGA type can be configured. Therefore a reconfigurable FPGA can be used in the system design, after which an antifuse FPGA can be used for actual spaceflight.

Nevertheless, there is one big downside to antifuse FPGAs, and that is their cost. The AtmoCube team is implementing an FPGA for their particle measurement tests; due to the clock frequencies at which they will be working an FPGA apparently yields the best power consumption. Following the conclusions on radiation-hard antifuse FPGAs they wanted to include an Actel antifuse FPGA, but its cost was around $3000. In conclusion they have decided to go for a non-antifuse FPGA with a triple redundancy check on most functions. These FPGAs have prices that are of little impact.

The main vendors for FPGAs are Xilinx, Actel, Altera and a number of others. Xilinx has already released a radiation tolerant, non-antifuse FPGA, the Virtex-4QV, whereas Actel is planning on releasing a similar radiation tolerant PROASIC3 FPGA. These FPGAs have a lower cost, with the added benefit that they retain reconfigurability.

A.3 Comparison between FPGAs and ASIC designs w.r.t. STX

The largest difference between FPGAs and application specific integrated circuits (ASICs) is the ability of adopting changes. Whereas ASICs consist of dedicated circuits, FPGAs are reconfigurable through software. This involves different design processes and costs. Usually FPGAs are more cost and time efficient if they are used during actual development of a complex system. However, as a design is finalized and to be used for mass production, losing the need for reconfigurability, ASICs show cost, size and/or power benefits. Consequently ASICs are used in phones, remote controls, radios, and an almost infinite number of other applications.

It is hard to give an exact estimate of the time efficiency or cost efficiency of using either approach but in light of the application on satellites some relevant specifications can be addressed. On satellites important aspects are mass, size, power requirements and environmental resistance, next to general important aspects such as cost and speed. The environmental resistance is expressed especially through radiation resistance, as has been assessed above for multiple FPGA types.

It should be mentioned, that regular microprocessors as well as data storage devices do not automatically provide radiation tolerance either. Nevertheless, as not all logic and signal processing is performed by *changeable* sets of transistors, radiation tolerance is of lesser impact as it targets mainly the processing memory. With FPGAs, everything can be seen as memory. Microprocessors such as the MSP430 family are and have been flown in space, and whether they are actually radiation tolerant or not they seem to be performing well.

The conclusion on the topic of radiation tolerance is ambiguous. The most accurate conclusion is probably that radiation tolerance should not be the main driver in the selection of FPGA usage or not. Other aspects as mentioned are the size, mass, power consumption, cost and speed achievable. The environmental resistance of FPGAs other than its radiation resistance, thus varying operability under temperature changes, is not assumed to differ much from discrete components for FPGAs.

FPGAs can be very large, to implement many functions. Depending on the application at hand, a suitable size can be selected. Consequently, both size and mass are not serious consideration factors for the application of FPGAs on satellites.

Power consumption is a more complicated factor. In general however, for an optimized discrete component design with a minimal number of components the power consumption will be the least. This is especially true if use is made of an exceptionally low power consuming microprocessor such as the MSP430 family [SLR 0581]. The AtmoCube anecdote however demonstrated that at highly demanding speeds FPGA power consumption can be less; this will not likely be the case for the STX.

In terms of speeds achievable, there is no real minimal requirement that would arise for use in the STX that cannot be met both by FPGA usage and use of discrete components. There could be one interesting application of FPGAs, if their internal processing speed could be at the S-band transmission frequency; this however is not the case for current FPGAs. Therefore speed is not much of a factor.

Finally there is the aspect of cost. There is monetary cost, and timely cost. The monetary cost has been addressed quickly above, with the conclusion that antifuse FPGAs can be expensive, but more common FPGAs have acceptable price ranges, similar to discrete components. The timely cost, or development time cost however poses large constraints on the possible usage of FPGAs in the STX, as a plethora of COTS components are available that can be used to develop a working STX system within a minimum time, whereas implementation of an FPGA is far less transparent. The argument of development time and that of reconfigurability will be addressed in light of the STX in the next subsection.

A.4 Conclusion with regard to the STX

As FPGAs do not seem to be able to achieve processing speeds of up to 2.4 GHz, an FPGA in the STX could integrate the components as illustrated in Figure A-1. An FPGA needs an external clock signal, especially as a highly accurate crystal would be required for space operation under varying environmental conditions. FPGAs do incorporate dedicated clock managers so that one oscillator signal should be enough for all functionalities.

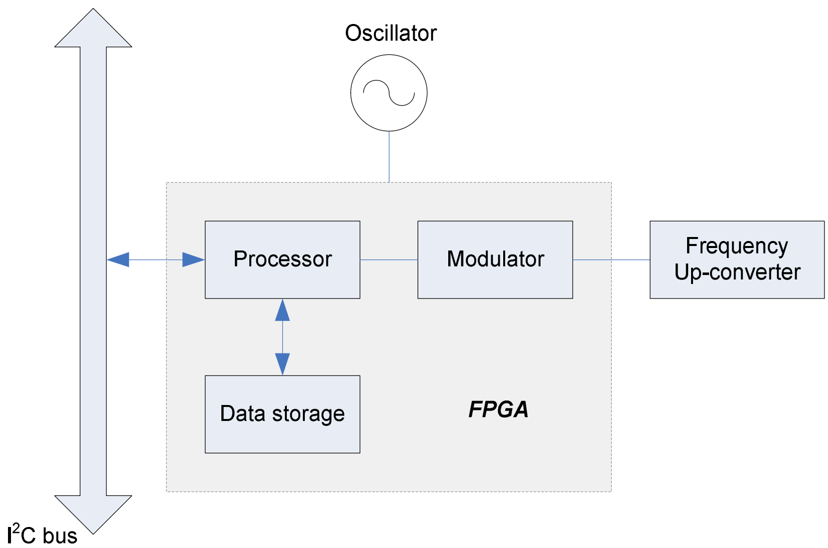


Figure A-1: Possible FPGA functionalities in the STX

As in the STX, due to time constraints as well as priority establishment as has been assessed before, the choice has been made for maximal usage of COTS components as opposed to in-house design, the motivation for an implementation of an FPGA largely disappears. The design benefit that FPGAs offer, also during testing and tweaking, largely dissolves as the FPGA 'functionality blocks' would first have to be designed, or at least be bought or obtained and interfaced. The three required functionality blocks however, as shown in Figure A-1, are available as highly standardized COTS components as is shown in this document.

Another possibility for reconfiguration is during the actual mission operation. If no antifuse FPGA would be used, the FPGA could be reconfigured to perform different functionalities in space. Nevertheless this kind of setting changes can also be implemented by implementing software modes, as also the CC2500 allows for adapting its modulation scheme and data rate. Nevertheless any changing configuration is simply not part of the mission requirements, and as it would come at the cost of increased development time it is for the project at hand not viable.

It has been suggested to create a modular baseband module; to basically have a circuit that can deliver a signal to be up-converted to any desired frequency in order to supplement radios in different frequency ranges. An FPGA could ideally deliver this, as is demonstrated in Figure A-1. It requires however, that the integrated processing software and modulator are prepared for a number of modulation schemes and data rates in order to be truly modular. Especially the variable data rate is relevant as an important reason for selecting transmission frequencies is the required data rate. The mission and operational aspects have significant influence on the achievable data rates, by means of orbital altitude, encoding schemes, transmission power and receiver/transmitter antenna directivity patterns, amongst others. The resulting modular baseband module is by no means impossible to create, but it would certainly require a dedicated development. As such, it is by no means a possible time investment within the current bounds of the Delfi programme.

Also, one would have to ask oneself what exactly would be the value of having a modular baseband transmitter module. Applying an FPGA or not, this module would in this case just incorporate a processor, an oscillator and a modulator (and possibly data storage). A processor is in principle modular as it requires software to specify its functionalities. Would it then be useful to have a modulator that can support a range of modulation schemes as well as data rates, just to be able to call the module modular? Possibly for a transceiver on the other hand a true modular IF system might be useful and worthwhile to develop, as more variables can be supported. In any case again for the STX transmitter on Delfi-n3Xt a modular baseband module is not feasible within the project constraints.

Appendix C

COMMS – Antenna System Design

SLR 0036

The entire document has been written by the author.

COMMS - Antenna System Design

Description: This document describes the preliminary design of the antenna system.

Subsystem(s) involved:

ADCS	CDHS	COMMS	EPS	MechS	STS	TCS	ITRX	MPS	$T^3\mu PS$	SDM	Splash	GSE	GSN	Launch
		x			x									

Revision Record and Authorization

Issue	Date	Author / Editor	Reviewer checked	PM approved	Affected Section(s)	Description of change
1	05/09/2008	Martijn de Milliano	JG	JB	All	First issue
1.1	05/11/2008	Martijn de Milliano	JG	JB	All	Outdated changes
2.0	29/01/2009	Martijn de Milliano	AT	JB	All	Outdated changes
3.0	18/06/2010	Arthur Tindemans	AT		***All:*** 1-6 Old chapters	***Complete update***: Completely new chapters Moved to appendices

Table of contents

No appendices are included in the thesis version of this document.

1 Introduction

The Antenna System is part of the Delfi-n3Xt communication system, or COMMS. The COMMS is introduced in [SLR 0014] along with its top-level design. The Antenna System and its design however are not discussed in detail; this document serves that purpose.

Two primary transceivers are to be flown on Delfi-n3Xt, being the PTRX and ITRX. Both radios should be able to transmit and receive signals. More precisely, the VHF-band is to be used for transmission, and the UHF band for reception. Furthermore, a technology demonstration experiment called the STX is aimed to be flown, which is a high-speed transmitter using the S-band.

All three radios are integrated on PCB's, and will have conduits for radio-frequency signals. In order to actually transmit or receive signals, a supporting Antenna System is required. As such, the resulting Antenna System can be sketched as Figure 1.1 below.

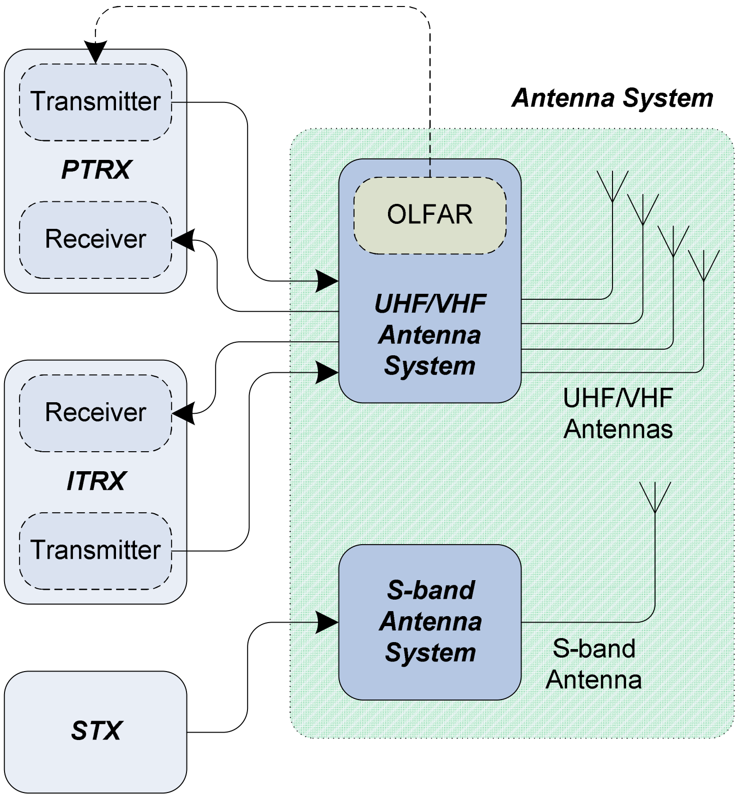


Figure 1.1: Components of the Antenna System and their interfaces

The Antenna System can be seen to be composed of two main and separate parts; the UHF/VHF Antenna System and the S-band Antenna System. Integrated on, but not actually part of the UHF/VHF Antenna System is the OLFAR experiment, if it will be flown.

The exact content of the Antenna System is first introduced in chapter 2, after a presentation of the requirements. The resulting antenna system components, the UHF/VHF Antenna System, the S-band Antenna System and the required Antenna Wiring are the topics of respectively chapter 3, 4 and 5. Next steps are discussed in chapter 6.

The design changes of Delfi-n3Xt have had significant effect on the Antenna System. In fact a detailed analysis of the antenna system has been made (by the previous author, Martijn de Milliano) based on the old solar panel configuration of Delfi-n3Xt, that of having four solar panels *in the same plane*, on one side of the satellite. This solar panel configuration was of course based on full pointing capability of Delfi-n3Xt.

The decision has been made to no longer rely on any pointing capability of Delfi-n3Xt. As a result:

- The solar panel configuration will have to be changed
- The S-band antenna system will no longer aim to be directional, but omnidirectional instead

Also, time has made clear that little resources are and will likely be available for the realization of the STX and its antenna system, so that the focus on COTS components has become even more important. This is important for the design of the S-band antennas for example.

In conclusion, all old chapters and sections of this document have been become severely outdated. In fact drastically so, so that all material in the chapters of this document is new. The old chapters *have been moved to appendices*. References are made to the old appendices on several accounts. The content of the appendices is as follows.

An introduction to antenna modeling software is given in Appendix A. Consequently, the influence the satellite geometry has on the antenna gain patterns and the different antenna systems have on each other is discussed in Appendix B. Appendix C presents the determination of the old UHF/VHF antenna configuration. Appendix D and E discuss the S-band antenna system design in terms of its antenna and antenna configuration respectively. Afterwards, a description of a patch antenna design performed by using antenna modeling software named FEKO is given in Appendix F, and the effect of different mesh sizes in both FEKO and 4nec2 is discussed in Appendix G.

Finally, an updated derivation of the preferred S-band antenna system, based on the constrained pointing model of having continuous sun-pointing *while trying to point with the STX* has been added by the author. This text has however also been written before the Delfi-n3Xt design changes and is similarly outdated. It is included as Appendix H.

Of interest to the latest version of this document; Appendix B in fact demonstrates that the S-band antenna system and the UHF/VHF antenna systems can be modeled separately, and have negligible influence on each other.

In the version of this document as appended to the thesis paper, the appendices have been removed as they are not relevant for the evaluation of the author's work and would come at the cost of a large increase of paper weight.

2 The components of the Antenna System

In order to determine the exact elements and thus Configuration Items or CI' that compose the Antenna System, this chapter states the requirements imposed on the Antenna System and looks at the functionalities to be performed by it. As such, section 2.1 presents the requirements, followed by the specification of the CIs in section 2.2.

2.1 Antenna System requirements

In [SLR 0014] generally applicable COMMS requirements as well as Antenna System requirements have been presented. These requirements are stated below in Table 2-1 and Table 2-2 below. Underneath each table some comments are made.

Table 2-1: Generally applicable Antenna System requirements

Category	Req. #	Requirement
CONSTRAINT	SAT-C.01	All satellite systems shall comply with the mass budget, as given in [SLR 0018]
CONSTRAINT	SAT-C.02	All satellite systems shall comply with the volume budget, as given in [SLR 0303]
CONSTRAINT	SAT-C.03	All satellite systems shall comply with the power budget, as given in [SLR 0017]
CONSTRAINT	SAT-C.04	All satellite systems shall comply with the data budget, as given in [SLR 0282]
CONSTRAINT	SAT-C.05	All satellite systems shall comply with power and data bus interfaces, as specified in [SLR 0263]
CONSTRAINT	SAT-C.06	All satellite systems shall be able to withstand the launch environment
CONSTRAINT	SAT-C.07	All satellite systems shall be able to withstand the space environment
CONSTRAINT	SAT.2-C.01	All satellite bus systems shall adhere to reliability standards, as specified in [SLR 0263]
CONSTRAINT	SAT.2.3-C.05	All UHF and VHF antenna connections and transmission lines on the satellite will have an impedance of 50 ohm

Most requirements in the table above are self-explanatory. The requirement with respect to the power and data bus interfaces is important when using deployables, which are to be activated by means of the system bus.

Table 2-2: Antenna-System-specific requirements

Category	Req. #	Requirement
FUNCTIONAL	SAT.2.3.3-F.01	The Antenna System shall be able to receive the uplink signal over the UHF frequencies
FUNCTIONAL	SAT.2.3.3-F.02	The Antenna System shall be able to radiate the downlink signal over the VHF frequencies
FUNCTIONAL	SAT.2.3.3-F.03	The Antenna System shall be able to radiate the downlink signal over the S-band frequencies
INTERFACE	SAT.2.3.3-I.01	The Antenna System shall have an interface with the PTRX to forward and receive an analogue data signal
INTERFACE	SAT.2.3.3-I.02	The Antenna System shall have an interface with the ITRX to forward and receive an analogue data signal
INTERFACE	SAT.2.3.3-I.03	The Antenna System shall have an interface with the STX to forward an analogue data signal
CONSTRAINT	SAT.2.3.3-C.01	The polarization of the Antenna System shall be circular
CONSTRAINT	SAT.2.3.3-C.02	A single antenna deployment failure should not cause loss of the link

The table above demonstrates the main functionalities of the Antenna System, as well as the three resulting interfaces. The requirement of circular polarization is necessary due to the fact that the satellite moves with respect to the ground station and it is the only way to guarantee a working link. Finally the requirement of having no antenna deployment being mission-critical is related to the general requirement of adhering to reliability standards, SAT.2-C.01.

2.2 Antenna System CI's

From the requirements in the previous section it can be seen that the Antenna System has three main functionalities, being to receive UHF signals, transmit VHF signals, and transmit S-band signals if the STX will be flown. Therefore, the primary components required are:

- **Antennas:** Antennas take care of the physical transmission of signals; dissipating energy in the form of radiation.

As antennas cannot be integrated on standard PCB's, due to size constraints and the fact that they should preferably be on the exterior of the satellite, some kind of connection should be designed. Furthermore, as the length of the antenna will be at least $1/4^{th}$ of the wavelength used, UHF and VHF antennas require relatively long antennas. As such, required are:

- **Antenna connections and deployment mechanisms**

The use of multiple antennas requires some form of power splitters/combiners, and, when phasing is to be introduced, phase shifters. Also, the impedance of the antenna should be matched to that of the cables so that impedance matching circuitry is required. This leads to require:

- **Phasing circuit(s) and impedance matching circuitry**

Finally, as antennas are not located at the same place on the satellite as the radios, the need for wires is introduced, capable to transport the high-frequency RF signals. Thus, also required:

- **Antenna wires and antenna wire connectors**

The UHF and VHF antennas are combined into one set of antennas, as further explained in the next chapter. This therefore creates a UHF/VHF Antenna System, necessitating the above components. A separate system however is required for the STX, using frequencies of a different order and thus different antennas. Therefore all components described above are effectively necessary for a UHF/VHF Antenna System *and* an S-band Antenna System. The resulting CI tree can be seen in Figure 2.1 below.

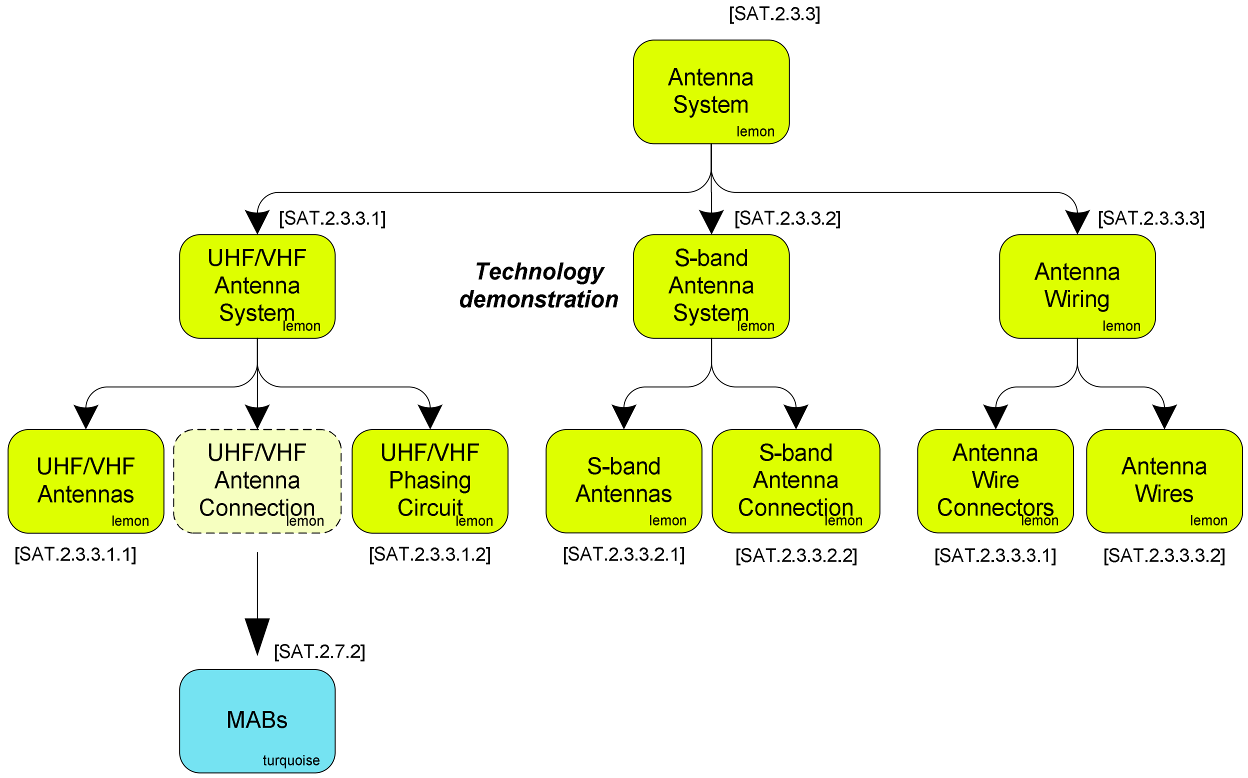


Figure 2.1: Context of the S-band Antenna System; the three lemon-colored CI's

Figure 2.1 above shows three branches, these are shortly commented on below:

- **UHF/VHF Antenna System**: This system is primary and mission-critical; it should provide a reliable link to the ground station. It consists of a number of antennas and a phasing circuit to properly phase the signals fed to or received from the different antennas. Also, as the UHF/VHF antennas are combined, this phasing circuit should take care of creating a UHF and VHF path that are effectively independent of each other. Finally, a UHF/VHF Antenna Connection CI can be seen, but in effect this CI is replaced by the MABs CI. As the connection and deployment of the UHF/VHF antennas is slated to be provided by the same Modular Antenna Boxes or MABs used in the Delfi-C^3 mission, these are listed as a separate CI under *Mechanisms* (SAT 2.7). Nevertheless they are introduced in this document as well.

- **S-band Antenna System:** The S-band Antenna System is as indicated focused on *technology demonstration*. It will only be flown if a working S-band transmitter is flown. The S-band antenna system comprises of its antennas and the connection of the antennas. The latter CI is related to mechanical aspects. No deployment is necessary for the S-band antenna system as is further explained in this document. Similarly, no additional (phasing) circuitry is required within the S-band antenna system other than what is included on the STX.

- **Antenna Wiring:** As mentioned before, antenna wiring connects radios to the two antenna systems. In both cases connectors and actual wires are required. The connectors and wires required for the UHF band, VHF and S-band are joined as their required properties are similar.

The three main CIs that make up the Antenna System are the topics of the next three chapters.

3 The UHF/VHF Antenna System

The UHF/VHF antenna system is the primary antenna system. It is supposed to receive signals over the UHF band and transmit signals over the VHF band. This functionality is identical to that of the antennas of Delfi-C^3. As such, it makes sense to make use of *proven heritage* of Delfi-C^3, for a number of its components. The antennas and their supportive systems have been tested and perfected for Delfi-C^3, and there are no indications of unexpected performance reductions.

The UHF/VHF antenna itself is introduced in section 3.1. For Delfi-n3Xt, it is suggested to combine the UHF/VHF antennas, as explained in section 3.2. The resulting antenna configuration is presented in section 3.3. The required antenna connection and phasing circuit are introduced in sections 3.4 and 3.5 respectively. Finally, section 3.6 discusses the Deployment and Antenna Board (DAB) on which the antenna components will be integrated.

3.1 The UHF/VHF antenna

The UHF/VHF antenna has in fact proven to be very simple of design. In the design of the Delfi-C^3 antenna system, it was found that regular metal measuring tape has proper electromagnetic characteristics. When cut to the proper length, this measuring tape could perfectly form an antenna for either UHF or VHF frequencies.

Antenna theory states that significant power transfer can be achieved using either a half-wavelength dipole antenna fed in the middle, or a quarter-wavelength unipole antenna fed at one end. If the feed-point is near a *conductive surface*, the latter antenna will then behave as a half-wavelength dipole antenna. The latter approach has been chosen for Delfi-C^3, as it logically minimizes required antenna length while maximizing performance. Antennas could similarly be chosen to have three-quarter wavelength, or one-and-a-quarter wavelength, etc. for equal power performance.

To illustrate, the final transmission frequencies of the Delfi-C^3 downlink were 145.870 and 145.930 MHz. Dividing the speed of light by the frequencies yields the wavelength; ~2.056 and 2.054 m respectively. A quarter-wavelength yields an approximate 51.4 cm. The final length of the measuring tape as measured on the spare model in the clean room is ~51.2 cm. The latter is of course subject to slight measuring errors as the antennas are partly enclosed by the MABs introduced below, and at the same time electrical connections to the antenna affect the apparent size. In conclusion however, the quarter-wavelength is near spot-on. The same holds for the uplink frequencies.

For Delfi-n3Xt, [SLR 0014] argues that the same transmission frequencies as used during the Delfi-C^3 mission can be used *for both the up- and downlink.* In other words, the exact same antennas as used on Delfi-C^3 can be used. A suggestion that has been made however is to *combine the UHF and VHF antennas*.

3.2 Combination of the antennas

For Delfi-C^3, it was decided to have a total of eight deployable antennas, of which four were used for the VHF frequencies and four for the UHF frequencies. The antennas belonging to the respective transmission bands had lengths matching the optimum quarter-wavelength. The *antenna configurations* of the two separate antenna systems however, *were identical*. The antenna configuration refers to the relative positioning of the four antennas. One side (top or bottom) of Delfi-C^3 housed the UHF antennas, and the other side housed the VHF antennas. In both cases the antenna pointed away from the satellite; the configurations were in fact *mirrored* with respect to each other. This antenna configuration (as introduced in section 3.3) together with the phasing introduced to the signals (as introduced in section 3.5), in turn determines the *antenna gain pattern*.

The antenna gain pattern is designed to be omnidirectional. In this case, the relative attitude of the satellite is unimportant. Given the fact that both the UHF and VHF antenna configurations were designed to approximate an omnidirectional pattern, the relative placement of the antennas on the satellite is unimportant. From a location-point-of-view then, the antennas can then be combined.

For an electrical point-of-view, it is required that the antennas are suitable to receive both signals. As explained above, a single-quarter-wavelength antenna is similar in performance to a three-quarter-wavelength antenna. As it turns out, VHF transmission frequencies (~145 MHz) are almost perfectly three-quarters of UHF transmission frequencies (~435 MHz). Theoretically, the VHF antennas can then be used for reception of the UHF signals.

The advantages of combining the antennas are:

- **Only four antennas and deployment mechanisms are required**; this allows a volume saving of about 10% of the satellite given the height of the MABs of about 3 cm, next to a reduction of mass and complexity.
- **All components can be located on one side of the satellite**; this additionally reduces the amount of components required and complexity involved. If the radios are placed near the antennas a minimal decrease of transmission line loss is also realized.

The effect of combining the antennas has two effects which should be prepared for:

1) The length of the antennas of at least one of the two frequency ranges becomes slightly non-optimal, as the ratio between the transmission frequencies is not exactly 3.
2) The signal strength received by the antennas on different frequencies is changed due to the direct connection.

Firstly, the effect of a changing antenna length with respect to the transmission frequency is a *change of impedance*. If this is not corrected for, an *impedance mismatch loss* will result. The impedance mismatch loss can however be prevented (for reasonable amounts of required correction) by introducing circuitry to adjust the phasing loss.

Secondly, the received signals should not be allowed to disturb the desired signals. The actual transmission frequencies are logically spaced far apart, but *harmonics* of the frequencies are important. Ironically, the same ratio of three that allows the antennas to function as both UHF and VHF antennas causes the *third harmonic* of the VHF signal to be close to the UHF frequencies that are to be received. Also, the downlink signal of the transmitter is many orders larger than the signal to be received over the uplink.

The harmonics are in fact caused by the power amplifier or PA. All other harmonics incurred by for example the modulator are cut off by means of a 3.3 MHz filter in the transmission circuitry. After amplification however, filtering would cause large signal loss.

A quantitative assessment of this second effect is given below.

The important values:

- **Difference between uplink and downlink signal strength**
  According to the link budget [SLR 0106], the difference between the received signals is about 116 dB (!) assuming a true omnidirectional gain pattern and a maximum satellite altitude.
- **Difference between the uplink and downlink transmission frequencies**
  Three times the (minimum) downlink frequency results in a centre frequency of 437.631 MHz. The uplink frequency however is between 435 and 436 MHz (not stated directly for reasons of security). As such, the third harmonic of the downlink signal is spaced apart from the uplink frequency by some 2 MHz.
- **Signal strength and bandwidth of the third harmonic of the VHF transmission frequency**
  These values are illustrated by the measurement figures below.

The signal strength and bandwidth of the third harmonic of the transmission signal *of the Delfi-C³ spare model* are measured using a spectrum analyzer. The measured signal strength is illustrated by Figure 3.1 and Figure 3.2 below.

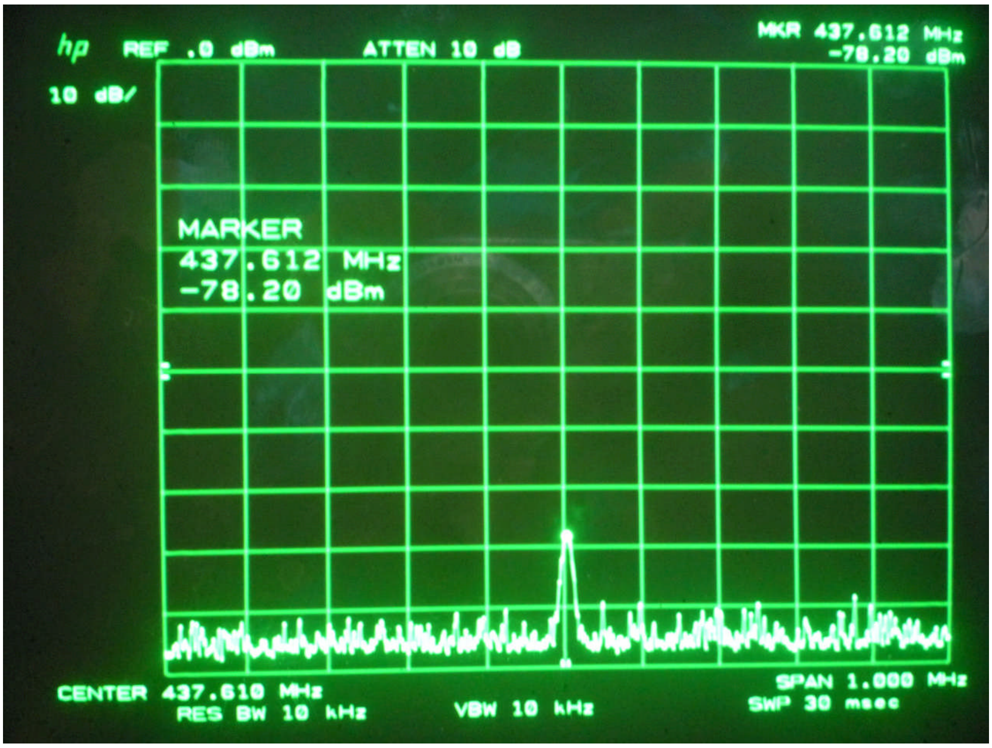


Figure 3.1: Signal strength (in the test set-up) of the third harmonic of the VHF downlink

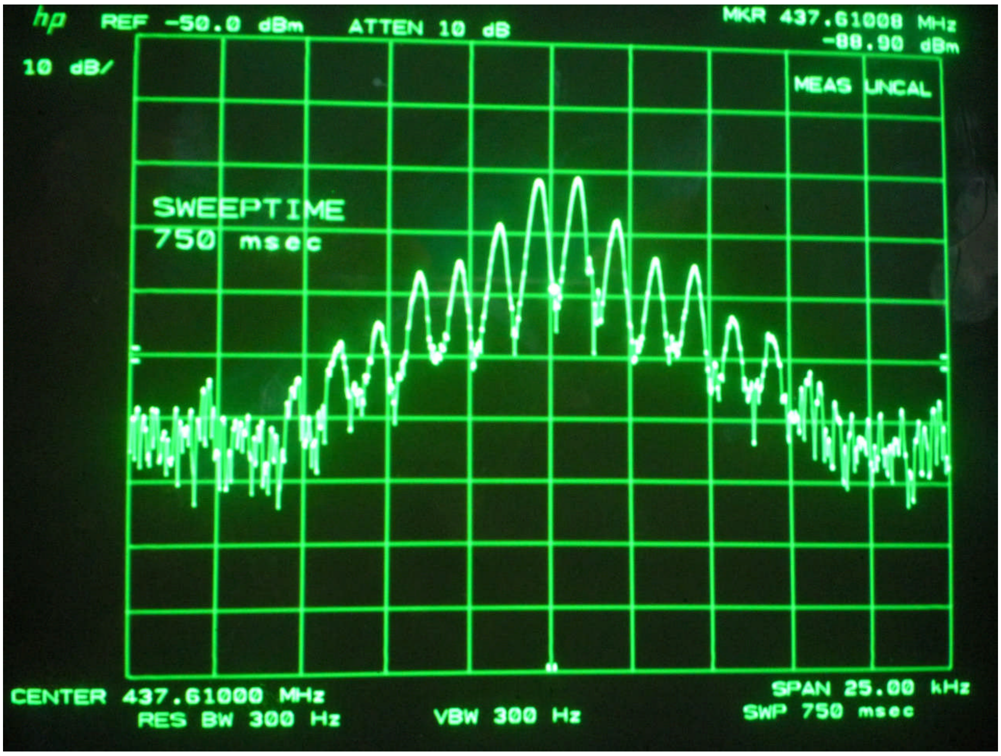


Figure 3.2: Frequency spectrum of the third harmonic of the VHF downlink measured over a 25 kHz span

The signal strength is illustrated by Figure 3.1. In fact, it is a figure that is taken on the same scale as figure 7-17 in [SLR 0014]. In the latter figure the peak has a value of -20.20 dBm. In Figure 3.1, the peak can be seen to have a -78.20 dBm power value (indicated by the marker). As a result, the third harmonic is measured to have signal strength of -58.20 dBc.

Figure 3.2 zooms in on the third harmonic, to conclude on the bandwidth. It shows that within about 7 vertical squares, or 17.5 kHz, the signal strength changes another -40 dB. For lower values of signal strength, the noise floor of the measurement equipment and set-up is an obstruction. In the set-up applied the Delfi-C^3 spare model transmission is measured over a 1 m wireless channel and consequent reception by a simple monopole antenna.

In conclusion, it can be seen that the overall signal strength of the third harmonic is -58.20 dBc, and it drops off to -98.20 dB within about 17.5 kHz. A total signal strength reduction of -116 dB within a 2 MHz bandwidth is required, to be able to resist the worst case. Indeed, it would seem this should be no problem.

A few notes:

- Signal strength drop-off in fact decreases with frequencies away from the carrier frequency; in other words a doubled bandwidth will not result in halved signal strength. The additional -20 dB required for a worst case will therefore still require a sizeable bandwidth.
- A larger data rate will result in larger occupied bandwidth of the main carrier as well as its harmonics.

Still, *it seems very unlikely that the third harmonic of the downlink signal will disturb the uplink signal in case of a data rate of up to 9.6 kb/s; for (much) lower data rates the effects will surely be negligible.* Of course, final integration tests are required to be absolutely sure; also to see if there are no other non-linear effects causing signal strengths at reception frequencies.

Finally, the first effect of combining the antennas, as mentioned above is the resulting requirement of additional *impedance matching circuitry*. There are limits to what can be corrected for with circuitry, but given the fact that the above calculated 2 MHz is only ~1.4 % of the transmission frequency on the UHF band, circuitry is assumed to be able to correct for this. In fact, *this has been done*. Wolter van der Kant has presented in [SLR 0715] that a phasing circuit can be designed that performs similarly (in terms of power losses) to that of Delfi-C^3, but designed to support the combined antennas. This phasing circuit is further introduced in section 3.5 below.

3.3 Antenna configuration

It has been argued above that the antennas can in fact be combined, at least assuming the same antenna configuration (and transmission frequencies) as those used for Delfi-C^3. It then remains to confirm the reuse of the antenna configuration.

A very extensive trade-off of possible antenna configurations has been performed by Martijn de Milliano, but unfortunately this trade-off has been based on the *old solar panel configuration of Delfi-n3Xt* (this trade-off can be found in Appendix C). The result of the trade-off however has been to:

- Reuse the antenna configuration applied on Delfi-C^3 for the VHF antennas
- Integrate the UHF antennas on the solar panels

The latter was chosen because the antennas that would be on the same side as the solar panels would be negatively influenced by the panels. Antennas oriented along the panels are minimally affected. As the UHF antennas are in fact small enough to be placed alongside of the panels, the UHF antennas were chosen to be integrated with the panels as opposed to the VHF antennas. For the VHF antennas, it had then been decided to opt for the use of heritage, giving good performance. The old solar panel configuration and the respective antenna configurations results are shown in Figure 3.3 below.

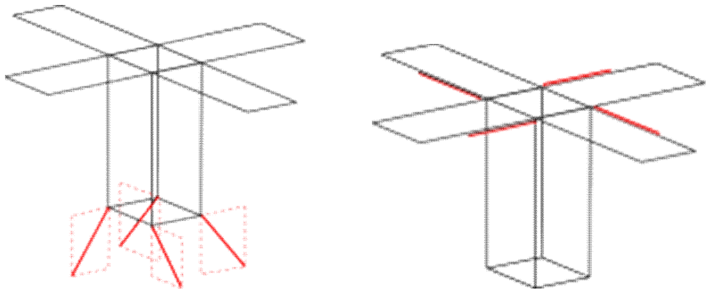


Figure 3.3: (Old) Delfi-n3Xt antenna configuration for the VHF antennas (left) and UHF antennas (right)

For the old solar panel configuration, the choice of having combined antennas would logically lead to having all antennas located on the bottom of satellite, following the suggestion of the VHF antennas above.

As suggested above, the redefinition of Delfi-n3Xt has resulted in a change of the solar panel configuration. Unfortunately, the new solar panel configuration has not been determined yet. Currently, the investigation into the reconfiguration is being performed by Johannes Bürkle. Two conclusions:

- The old Delfi-n3Xt configuration is no option, due to its non-omnidirectional performance
- The Delfi-C^3 configuration is an option, but is undesirable. This due to:
    - A relatively low power generation (compared to more ambitious configurations), at the same time being not enough for desired nominal operations
    - The fact that it covers all long sides, which is undesirable for the $T^3\mu PS$ payload

As said, the investigation for new panels is currently ongoing. This research includes power impacts as well as impacts related to mechanical aspects and complexity. Two concrete suggestions are however shown in Figure 3.4 below. All indicated solar panels are in fact *double-sided*.

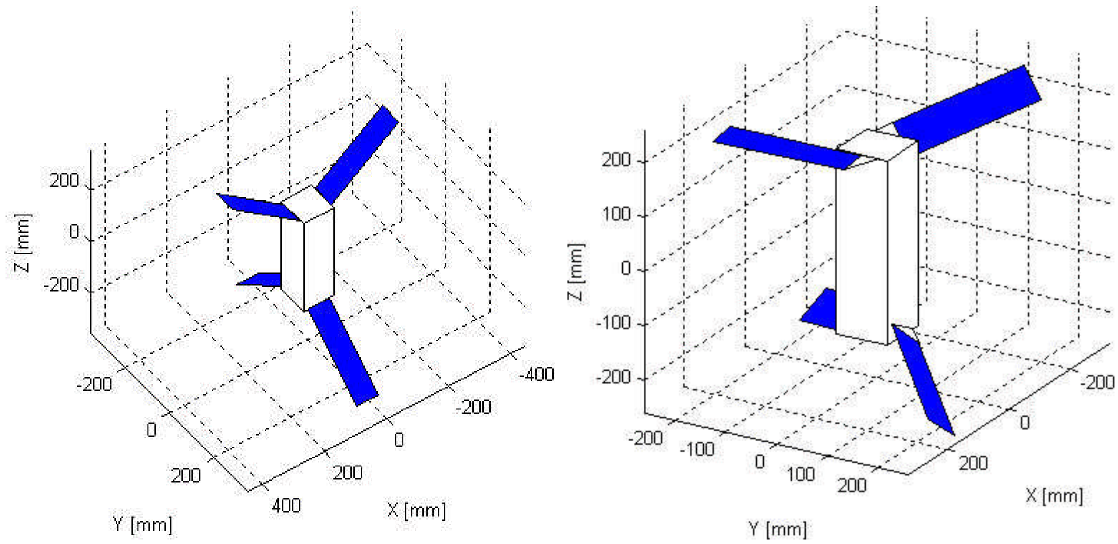


Figure 3.4: Two suggestions for Delfi-n3Xt solar panel configurations

The exact fold-out angles in the above figure will not be given here. The suggested configurations serve to show that larger fold angles will likely be used in order to ensure no blockage of the sides, and the use of double-sided solar panels allows the generation of more power.

Given the above configurations, it is in fact rather hard to give a desired antenna configuration without further analysis. Nevertheless, some concrete conclusions for the UHF/VHF antenna system can be drawn.

At the start of this section, it has been shown that with all solar panels on one side, the Delfi-C^3 antenna configuration is preferred. As such it is arguable that with the above (new) solar panel configurations, the panels located on the far side from the antennas will not change this conclusion. This then leaves the two antennas on the same side as two panels. As Appendix C shows, a set of two opposite antennas actually performs (slightly) better than a set of four. *The reason to include four is to avoid single-point-of-failures*. It can therefore be assumed that, any solar panel that affects an antenna, will minimally, or *differently* affect its redundant copy that is *oriented at 90° with respect to the long axis*. This is a rough assumption but it does sound feasible given the rotational symmetric properties of the antenna configuration.

As said however, the above solar configurations are still very tentative. Currently, *more options will be reviewed with all solar panel hinges located on one side*. In that case however, the antenna disturbances will arguably always be less than those of the old Delfi-n3Xt orientation, given the fact that the projected area of the panels will be less (as they will have to be inclined to approach an omnidirectional power generation pattern). Then again, the Delfi-C^3 antenna configuration but with combined antennas is preferred.

In conclusion, *it is assumed that Delfi-n3Xt will reuse the Delfi-C^3 antenna configuration*, but with only four antennas and located on one side (top or bottom). If all solar panel hinges are located on one side, the antennas would logically be located on the other side. The resulting orientation of the antennas is shown in Figure 3.5 below. It also shows the connection of the antennas to a PCB, using the Modular Antenna Boxes or MABs, as introduced in the next section. As assessed above, requirement SAT.2.3.3-C.02 is met, as a single deployment failure will not cause a loss of the link; only a slightly deteriorated performance as the gain distribution becomes unequal.

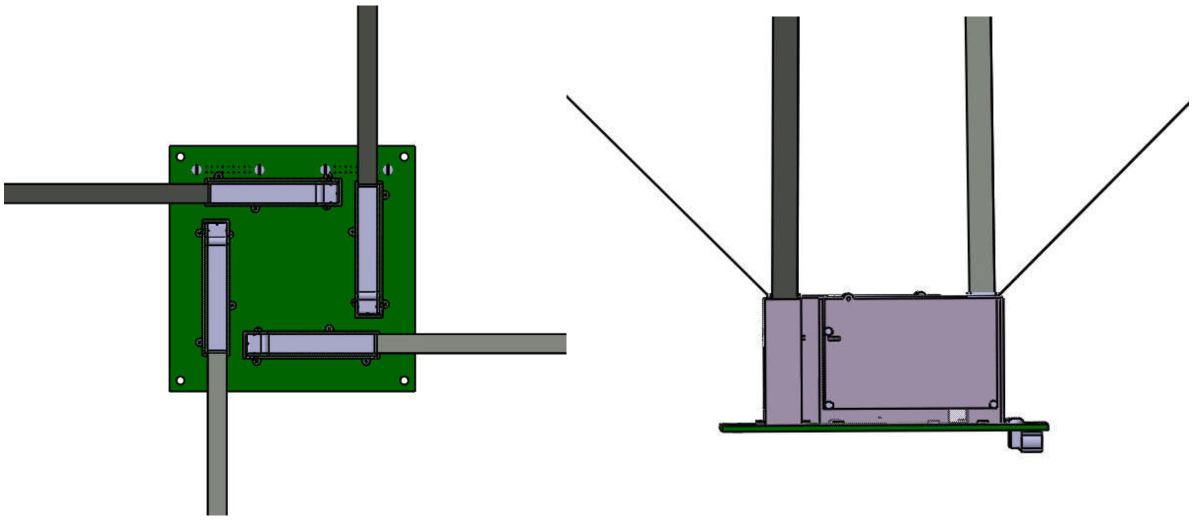


Figure 3.5: DAB including MABs and antennas; top-view (left), side-view (right)

The final performance of the UHF/VHF antenna system will have to be modelled, or better yet measured. For now, any estimates will be based on the old models of the Delfi-C^3 configuration, logically applying a Delfi-C^3 solar panel configuration. A real-life test has been performed using a mock-up of Delfi-C^3 and its antennas. During the tests the antenna lengths have been optimized. Unfortunately, no (well-documented) overall gain pattern has resulted. As such, the digitally modelled Delfi-C^3 antenna pattern is shown below as a performance indication. Figure 3.6 shows both the right- and left-handed circularly polarized (RHCP respectively LHCP) gain patterns of the VHF antenna configuration, as presented in [SLR 0248]. This assumes the application of proper *phasing*, as introduced below.

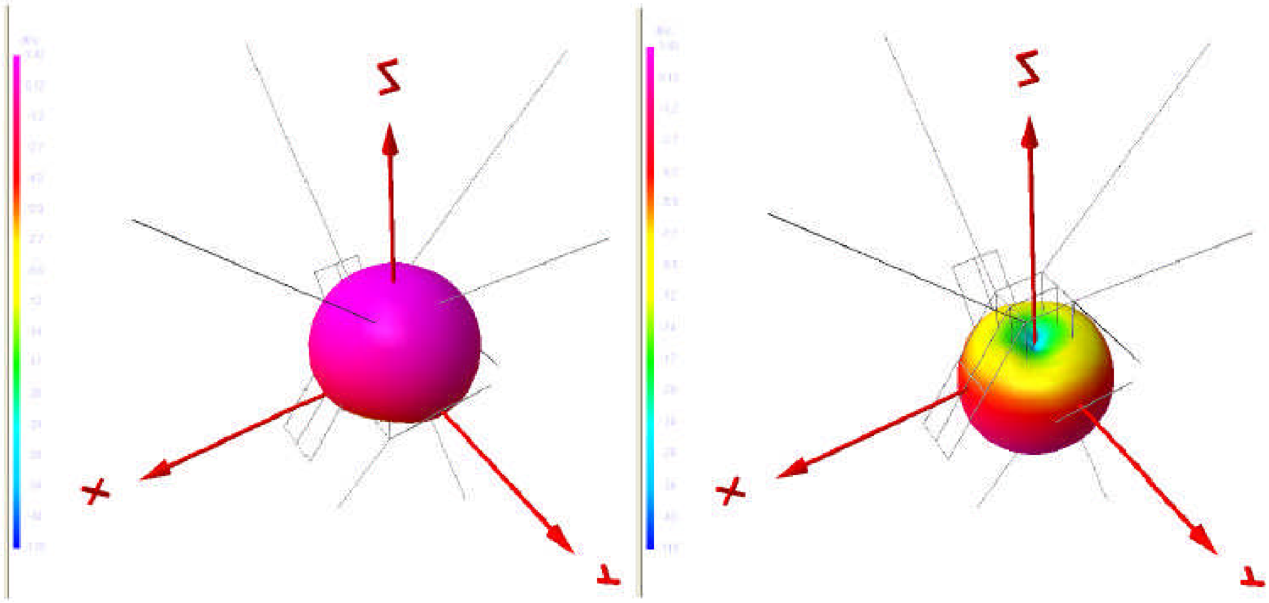


Figure 3.6: Theoretical RHCP (left) and LHCP (right) gain pattern of the VHF antennas on Delfi-C[3]

In the above figure the indicated values are rather small and unclear. The maximum indicated however is 1.42 dB, with the minimum value on the scale being -11.7. The RHCP and LHCP gain pattern are *exact mirror images*. As the ground station is capable of receiving both RHCP and LHCP signals, the minimum gain value of the superimposed gain patterns is important. In other words; that at an angle of 90° or -90° with respect to the Z-axis. This value (approximately -5 dB) is used in the link budget as the minimum gain value.

For the UHF antennas, the resulting gain patterns are shown in Figure 3.7 below.

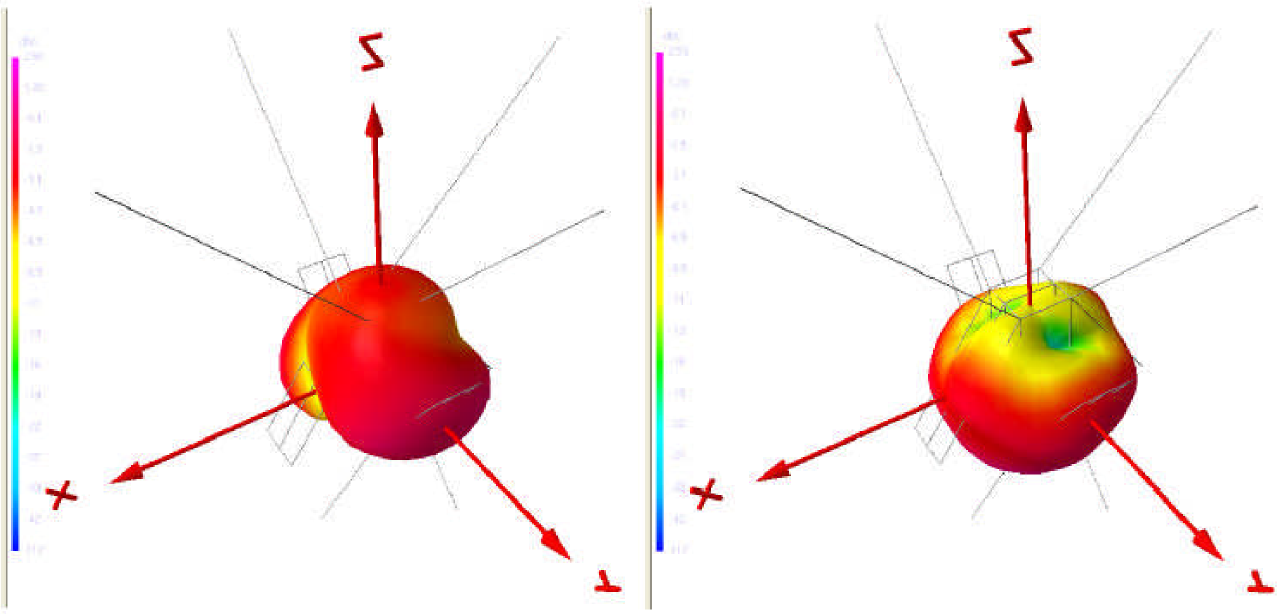


Figure 3.7: Theoretical RHCP (left) and LHCP (right) gain pattern of the UHF antennas on Delfi-C[3]

In the above figure, the indicated maximum and minimum values are respectively 2.58 dB and -11.7 dB. Performance can be seen to be worse than for the VHF antennas. A value of -10 dB at an angle of 90° or -90° with respect to the Z-axis is taken as the minimum gain value.

3.4 Antenna connection

All Delfi-C^3 antennas were connected to the PCB using modular antenna boxes (MABs). These same MABs were used to *deploy* the antennas. These MABs have proven to work perfectly for Delfi-C^3. Logically, these are then reused for Delfi-n3Xt.

A modular antenna box works as follows. The (measurement tape) antenna is initially rolled up in the antenna box. The lid is held in place by a Dyneema wire. During deployment, the wire is melted by means of heating a resistor. Due to the force exerted by the metal strip the lid is opened. Due to the properties of the measurement tape, the antenna surges out of the box and stretches to a linear shape. As there is no gravity in space, it will remain in this shape to form a perfect unipole antenna. An illustration of the MAB is shown in Figure 3.8.

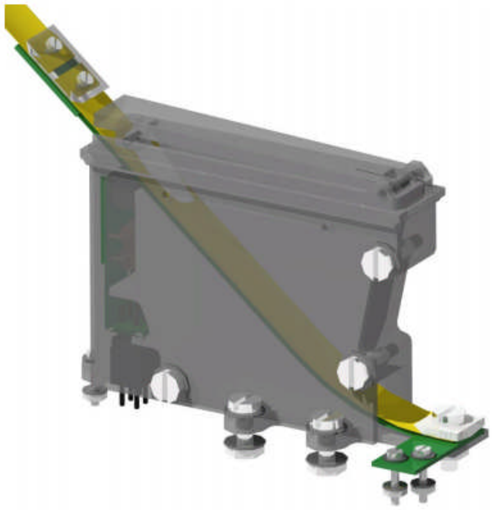


Figure 3.8: Drawing of the modular antenna box

3.5 Phasing circuit

The Delfi-n3Xt phasing circuit has three purposes:

- Introduce proper phase shift between the antennas for the resulting gain pattern to be omnidirectional and *circularly polarized*
- Isolate the UHF frequencies from the VHF frequencies
- Correct for the non-optimal length of the UHF antennas when combined with the VHF antennas

In case of Delfi-C^3, the first two functionalities were also required. However, as the antennas were separated, the signal strengths of the UHF and VHF frequency bands were less strong on respectively the VHF and UHF antennas.

The required phasing circuit has actually been designed and tested (as a separated component) and is described in [SLR 0715]. Some losses are incurred in the phasing circuit, but these are comparable to those on Delfi-C^3. They have been taken into account in the link budget.

3.6 Deployment and Antenna Board

The Deployment and Antenna Board (DAB) is the PCB that houses all components as introduced above. In summary:

- Four MABs including quarter-length VHF antennas (~51.2 cm)
- The UHF/VHF Phasing Circuit

The DAB and the integration of the four MABs have in fact already been shown in Figure 3.5 above. In the middle it can be seen sizeable area is available for the (electrical) integration of the phasing circuit.

As [SLR 0014] explains, a third component can in fact be integrated on the DAB:

- OLFAR

If OLFAR is to be flown, the DAB is the only logical location for it. As it involves only very little components, it should likely pose no problem to integrate it on the DAB, not taking into account the increased complexity. With respect to the Delfi-C^3 PCB similar to the DAB, the Delfi-n3Xt DAB has less required components if the solar panel deployments systems are not integrated on the DAB.

4 S-band Antenna System

The S-band antenna system is a secondary system; not required for essential operations. It serves to perform technology demonstration in combination with the STX on-board of Delfi-n3Xt. It is completely separated from the UHF/VHF antenna system.

It has been established in [SLR 0387] that one of the design goals of the STX is that it should assume the use of an omnidirectional antenna configuration. In other words, no pointing is assumed. Therefore, the antenna gain pattern should provide this omnidirectional pattern for as far as possible.

S-band antenna types are introduced first in section 4.1. Also, the preferred antenna type, some representative versions and their performance are presented. Section 4.2 then discusses the possibly antenna configurations that can be made using the S-band antenna versions, and introduces the preferred configuration. Section 4.3 concludes with an assessment of the integration options of the S-band antenna system on Delfi-n3Xt.

4.1 Antenna type

There are several options for antennas in the S-band which are generally applied. It should be realized that the Delfi-n3Xt S-band frequencies are around 2400 MHz, so that the wavelength is only ~12.5 cm. The quarter-wavelength is then some 3 cm. Different S-band antenna types are the:

- Helical antenna
- Wire antenna
- Parabolic dish
- Patch antenna

The helical antenna and parabolic antenna can be used to obtain a (highly) directive antenna gain pattern. They however require sizeable volume. Wire antennas and patch antennas are more low-profile. Patches have some very favorable characteristics. For example, they are mechanically robust and they are very flexible in terms of resonant frequency, polarization, pattern and impedance. Compared to wire antennas, they can be easily body-mounted, as they are relatively flat, resulting in gain pattern perpendicular to the patch *and* the body. A wire antenna would require a *deployment mechanism*. Patches are generally used on the S-band whenever size is available; although they are flat they require some area, patches with total areas of up to 80x80 mm are introduced below.

A final argument for the use of S-band patches is delivered by a quick search for existing CubeSat applications of S-band transmitters. In fact, all (three) successful CubeSats with S-band transmission capability have applied *patch antennas*. A number of larger microsatellites similarly did so. They in fact also suggest possible antenna configurations and are thus introduced in more detail below.

In conclusion, a short and simple trade-off leads to conclude that for Delfi-n3Xt, *the use of S-band patches is preferred*. A more detailed trade-off could be performed to verify this, but is not deemed necessary.

As an omnidirectional gain pattern is preferred, the gain pattern resulting from a patch should also preferably be *as broad as possible*. In fact, Martijn de Milliano has designed (in software) a patch antenna, but this was with the goal to get one as directional as possible. This design (process) is described in Appendix D. Its results are taken along belong, but it should not be forgotten that these are *theoretical* values.

An important focus within the S-band communication link design has come to lay with COTS products. S-band patches are generally available, but in little cases do websites offer detailed gain patterns over all angles. Also generally, single patch antennas, or single-chip solutions (possibly including multiple internal patches) are sold, under the header of being 'omnidirectional'. This then either holds for one axis, or actually means a very low gain value (up to or below -10 dB) is accepted in some directions.

Two options have however been found:

- **SSTL patch**: Of all the online shops selling CubeSat parts, one of them sells an S-band patch antenna; that of Surrey Satellite Technology Ltd. (SSTL). It is rather large and (relatively) heavy, but achieves high-directivity while retaining sizeable gain other directions. It is space-proven. At the same time, this might make it (very) expensive.
- **SPECEMC patch (PA28-2450-120S)**: A website was found, ran by a company called Spectrum Control, Inc. that delivers custom made and standard patches. The respective website is www.specemc.com. In a conversation with one of the design engineers, the measured performance data of the most relevant patch was obtained. It is optimized to have a gain pattern as broad as possible. It is not space proven, but much smaller than that of SSTL.

Even though these patches might not be those used in the end, *the general performance of these patches will be used as a guideline*. In order to asses the performance of the patch antennas, the following Figure 4.1 demonstrates the applied coordinate system. In the following, the patch antenna will be placed in the *xy*-plane.

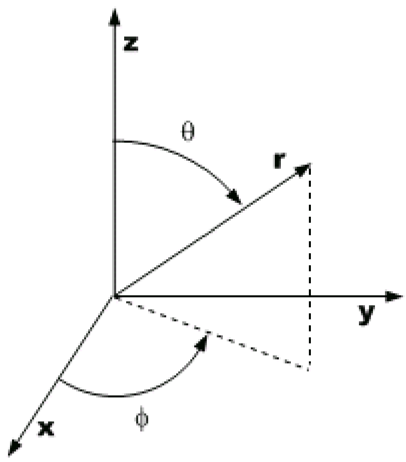


Figure 4.1: Spherical coordinate system used below

θ is labeled the *angle off-boresight*. For different values of θ, the gain of the patches has been measured by the manufacturer. At the same time, the gain values *do not vary with* ϕ. Of course this latter will never perfectly hold but is assumed to hold with sufficient accuracy.

For both the SSTL and SPECEMC patches, the gain values are plotted for $-90° \leq \theta \leq 90°$ in Figure 4.2 below. As the patch will be placed on a surface, the exact effect of the gain in reverse directions is unknown; potentially it could reflect to increase the transmitted signal slightly. Nevertheless, the gain values for $90° < \theta < 270°$ are negligibly low. Gain values are stated in dBic, which is the gain with respect to a hypothetical isotropic circularly polarized antenna.

The values for the SSTL patch have been read from the plot given in [SLR 0472], whereas those for the SPECEMC plot are given in [SLR 0743]. It should however be noted that only the results on the *first page* of this latter document are actual measurement results, the other results in the file are theoretical (and wrong).

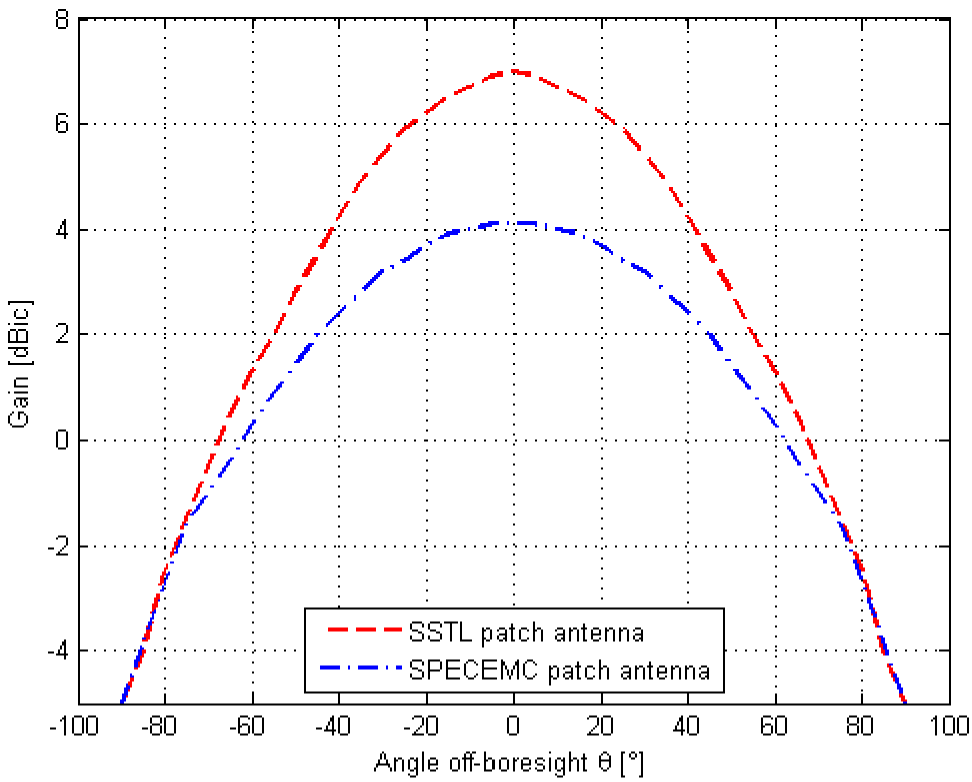


Figure 4.2: Gain values in dBic for different angles off-boresight of the SSTL and SPECEMC patches

In summary, the resulting gain values are stated in Table 4-1 below, for different values of θ. The gain values of the theoretical patch antenna design are also given for comparison, these are indicated by italics. The angle that yields a gain of -2 dB for the two existing patches is also included, as it will be referred to in the next section.

Table 4-1: dBm to watt conversion for some useful values

Angle off-boresight θ	***Gain value [dBic]***		
	SSTL patch*	SPECEMC patch	*Theoretical patch*
0°	7.0	4.1	*5.6*
15°	6.5	3.9	*5.0*
30°	5.5	3.2	*3.7*
45°	3.5	2.0	*1.9*
55°	2.0	0.9	*0.3*
60°	1.5	0.3	*-0.6*
75°	-1.5	-1.6	*-2.7*
78°	-2.0	-2.0	*-2.8*
90°	-5.0	-5.0	*-3.0*

*The values specified are read from a plot and are therefore of slightly less accuracy than those of the SPECEMC patch.

Some characteristics of the different patches are summarized in Table 4-2 below. All antennas are prepared for a 50 Ω antenna connection; somewhere at the back there is a pin connection to which a coax wire can be connected. These characteristics will again be taken along in section 4.3 in a discussion of the integration options.

Table 4-2: dBm to watt conversion for some useful values

	SSTL patch [SLR 0742]	**SPECEMC patch*** [SLR 0744]	***Theoretical patch*** *[Appendix D]*
Area	82x82 mm^2	45x45 mm^2	*40x40 mm^2*
Thickness	20 mm	6.4 mm + ground plane	*< 3 mm*
Mass	< 80 g	< 50 g (estimated)	*< 40 g (estimated)*
Polarization	RHCP or LHCP	RHCP or LHCP	*RHCP*
Bandwidth	Tunable for 2.4-2.5 GHz	2450 MHz ± 50 MHz	*2410 ± 24 MHz*
Comments:	Space proven	Non-space proven	*Theoretical*

To give some general conclusions:

- The SSTL patch is the only space-proven one. It is directly available and can straight-forwardly be applied. It is however very big, both in terms of area and thickness, and it might be (very) expensive. The SSTL patch has comparable performance to the SPECEMC patch, however with a higher peak value of almost 3 dB more. Although not interesting for an assumed omnidirectional pattern, this would yield extra functionality in case pointing will be achieved for Delfi-n3Xt.
- The SPECEMC patch is smaller than the SSTL patch, and quite drastically so. In turn it yields lower maximum gain, but similar gain at high angles of θ when compared to the SSTL patch. It is not space-proven, so proper attention should be paid to the choice of materials. It is available as well, and a sample version has been ordered.
- The theoretical patch seems to perform worse for low angles off-boresight, but better at higher angles off-boresight. It has been designed for minimal size, in order to fit on multiple locations of Delfi-n3Xt. Nevertheless, it is completely theoretical and its actual values can deviate significantly. Also, significant effort will have to be made to actually produce and test the antenna, which is not desirable for the S-band communication link.

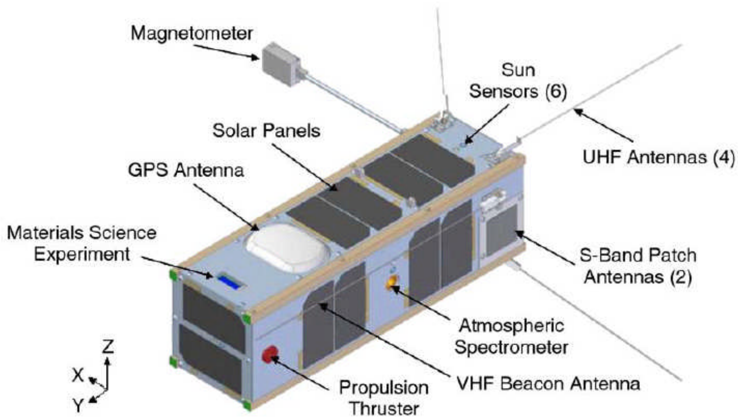
4.2 Antenna configuration

The patch antennas presented above can be used to draw some quick conclusions on possible configurations. The discussion below focuses on determining the options for getting a proper S-band antenna system to support the main goal of demonstrating STX transmission capability, using COTS components and proven design choices.

Three successful CubeSats have delivered working S-band transmitters to space to the best of the author's knowledge, being GeneSat, PharmaSat and CanX-2. The first two are developed by the National Aeronautics and Space Administration (NASA), and PharmaSat is the successor of the first one. The CanX-2 satellite is a result of development by both the University of Toronto Institute for Aerospace Studies (UTIAS) and the Space Flight Laboratory (SFL). Its (S-band) technology follows from the MOST microsatellite. The S-band antenna approaches of the two different groups are introduced below.

GeneSat and PharmaSat: Both GeneSat and PharmaSat applied a single S-band patch [SLR 0745] (see Figure 4.3 (left)). By using passive three-axis attitude control, the rotation of the satellite was dampened and the orientation constrained. In result, the one patch would yield sporadic communication (more or less 50 % of the time) depending on the satellite orientation.

CanX-2: As said, the S-band design approach of CanX-2 is in fact based on and identical to the larger and heavier MOST (Microvariability and Oscillations of Stars) microsatellite (65x65x30 cm). Both satellites apply two patches on opposite sides to approach an omnidirectional pattern [SLR 0746].



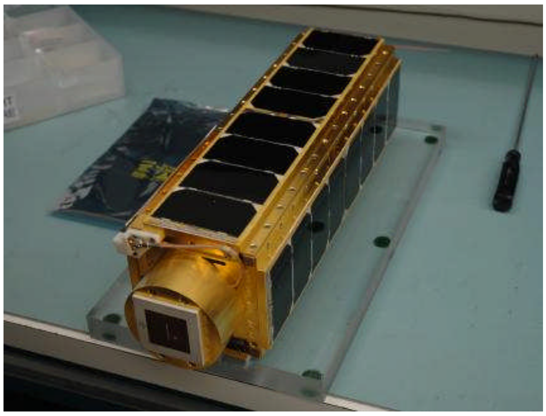


Figure 4.3: (S-band patch antenna) lay-out of CanX-2 (left) and PharmaSat/GeneSat (right)

In line with the MOST and CanX-2 satellites, a cube shaped (20 cm per side) microsatellite named TugSat also applies two patches on opposite sides of the antenna. Finally, the data sheet of the SSTL patch [SLR 0742] in fact *suggests* using two patches on opposite sides of the antenna.

In case of Delfi-n3Xt, it has been decided to aim for an antenna system that is independent of any pointing capabilities that Delfi-n3Xt might have [SLR 0387]. In other words, the most desirable pattern would become the perfect omnidirectional one. Although this pattern would give no guaranteed positive directivity in any direction, it assures communication for all possible attitudes. As illustrated by the UHF/VHF antenna system however, a true omnidirectional pattern is impossible to obtain. In fact, in case of the VHF antenna pattern, it means accepting a -5 dB minimum gain value, and for the UHF antenna pattern even a -10 dB gain value.

In the discussion below, the possibility of attaining an omnidirectional pattern is assessed using the performance of the existing patches introduced above.

The satellite has six sides. Patches have been shown to only have a (relevant) gain pattern on the front side of the patch. In other words, two options become possible to obtain an omnidirectional pattern:

- **Two patches on opposite sides with an hemispherical gain pattern**
- **Six patches on all sides with sufficient gain up until angles off-boresight of ~55°**

Intuitively, one could think that for the last case only a gain pattern until an angle off-boresight of 45° would be necessary, as four times 90° closes to 360°. However, if six of these rotationally symmetric patches are used to cover a three-dimensional spherical area, it can be seen that six cones would need to have higher angles; if not, the 'diagonal' cones would not touch. The relation $\theta = \tan^{-1}\left(\sqrt{2}\right) \approx 55^\circ$ then gives the proper angle.

The gain values at the relevant angles have been presented for the different patches in the previous section. By applying and simultaneously operating a number of patches, an overall power loss per patch is incurred, resulting in a gain loss. The resulting minimum gain values of the gain patterns that can optimally be created using the patches presented above are shown in Table 4-3 below.

Table 4-3: Minimum gain values of the resulting gain pattern

# of patches	**Gain loss per patch**	θ	**SSTL patch**	**SPECEMC patch**	***Theoretical patch***
2	-3 dB	90°	-8 dB	-8 dB	*-6 dB*
6	-7.8 dB	55°	-5.8 dB	-6.9 dB	*-7.5 dB*

It can be seen that for the already existing antennas, the patterns consisting of six patches perform slightly better. In reality however, this performance will be reduced due to the need for signal splitting and the need to feed the antennas exactly equally. In any case, complexity is largely increased. More importantly however, *six patches cannot be integrated on Delfi-n3Xt*:

- Except for the case were solar panels are exactly at 45° with respect to their connecting surfaces, negative reflections will occur due to the solar panel configuration
- It seems unlikely six sides will have space available for patches; this holds for area required as well as volume due to the patch thickness
- Six patches will have significant, volume and complexity effects

In conclusion, an omnidirectional pattern is only logically obtained using two patches, at opposite sides. Indeed, this has been the configuration suggested by CanX-2 amongst others introduced above.

The resulting minimum gain value is then some -8 dB for both existing patches. The theoretical pattern outperforms the other ones, but shall not be taken along due to its insecurity. **The resulting -8 dB is the minimum predicted gain pattern for an omnidirectional S-band antenna system**.

This knowledge can be combined with knowledge of the STX design as discussed in [SLR 0387]. It is shown in this document that *the use of two patches can optimally be combined with the use of two separate power amplifiers* (PAs), so that:

- The power output of the STX is effectively doubled
- The two different amplified signals do not have to be *exactly* in-phase as the signals will feed two patches on opposite sides of the satellite.

Using this implementation, three options become possible, *switchable by telecommand*:

- **Omnidirectional pattern, two PAs activated**
  This scenario assumes -8 dB gain but the use of two PAs effectively wins back 3 dB.
- **Near-omnidirectional pattern, two PAs activated**
  It can be chosen to *accept a certain range in which the link budget will not close*. In this case, a higher antenna gain value can be used. This option if further assessed below.
- **Hemispherical pattern, one PA activated**
  Using the integrated on/off ports on the PA, one PA and thus one respective patch can also be used at a time, effectively allowing switching between the two patches. This *lowers overall power consumption*, but will not generate a higher gain. This option can ideally be used for satellite pointing, if the final Delfi-n3Xt ADCS will allow for it.

The latter option effectively would approach the approach of GeneSat and PharmaSat. In effect, *the proposed solution combines the options of all proven CubeSat S-band antenna systems*. It also allows for some redundancy as the S-band communication link can still be tested (at least sporadically) if one patch or PA fails.

The only downside of choosing this set-up is that an omnidirectional pattern can only be achieved when both PAs are activated. This has implications on the power budget. Nevertheless, the resulting power consumption of the STX as presented in [SLR 0387] of 1.4 W should be acceptable, as it is less than the current PTRX power consumption. This has been stated as one of the design goals of the STX.

A small explanation is given below with respect to the second option introduced above; of having a *near-omnidirectional gain pattern*. Its reasoning is based on the following. For the UHF/VHF antenna system, communication should be guaranteed for all angles. For the S-band antenna system, *this criterion might be stretched a little*; if a better performance comes at the cost of a small percentage of time where no communication is possible, this can be deemed acceptable.

For the presented patch gain patterns, gain drops off quickly for high values of θ. Although many different values can be chosen, the angle that given a minimum gain value of -2 dB is selected here; $\theta \approx 78^\circ$. The resulting gain pattern when two patches are activated simultaneously is then -5 dB.

Assuming the satellite rotates without preference of direction, it can be calculated which percentage of time the antenna system is then assumed to yield no communication. In fact this can be calculated by means of the *solid angle*. It is given by:

$$\Omega = 2\pi\left(1-\cos\theta\right) \tag{4-1}$$

For $0^\circ \leq \theta \leq 90^\circ$ the percentage of time can then be calculated by dividing the solid angle by the maximum solid angle of a hemisphere:

$$\text{Percentage of time the link is closed} = \frac{\Omega}{2\pi} = \left(1-\cos\theta\right) \tag{4-1}$$

For a suggested value of $\theta = 78^\circ$, this percentage of time becomes ~79%.

This option has been taken into account in the link budget as a second option; it offers a 3 dB gain at the cost of losing only ~20% of the communication time. In result, **the second option delivers a minimum gain value of -5 dB for almost 80% of the time.**

In fact, the S-band communication link will be designed to support changing data rates. Therefore optimally, the data rate can be optimized for the satellite attitude during the monitored pass. At the same time, this can mean that the activation of only one patch at the same time is required. This is to be assessed at a later point, when more about the final performance of the STX is known following its integration. The antenna configuration suggested above gives maximum flexibility, and satisfies all requirements. The mechanical aspects and the respective required space on Delfi-n3Xt are assessed in the next section.

4.3 Integration options

It has been concluded that the preferred S-band antenna system consists of two S-band patches, applied on two opposite sides of Delfi-n3Xt. It remains to assess whether this is possible, and at which sides. Basically two options exist:

- **One patch at the TOP, one patch at the BOP** (top panel and bottom panel respectively; these are the short sides of Delfi-n3Xt)
- **Two patches on opposite long sides**

There are three aspects important for the possible integrations:

- Location of the solar panels
- Area available and area required
- Space available and thickness required

The first aspect has in fact been addressed in the last chapter for the UHF/VHF antenna system as well. Two preliminary solar panel configurations have been shown, but are still very tentative. It has also been noted that it is preferred to have a configuration with at least one of the panels not blocking the sides of Delfi-n3Xt, due to the inclusion of the $T^3\mu PS$ payload. Similarly, a solar panel with all mechanisms (hinges) on one side would be preferred. In fact, both preferences will then likely lead to having two opposite sides free due to reasons of symmetry. A general conclusion that shall be drawn is:

- If all Delfi-n3Xt sides are covered by the solar panels (angles with respect to the surface of less than 90°), the top and bottom panels are likely free. If however some sides are free, the top and bottom panels are likely constrained.

The conclusion then is that the solar panel configuration will likely allow only one of the two possible integration options. Which one that is remains to be confirmed. However, given the preference of the $T^3\mu PS$ payload, the option of having two long sides free is more desirable. Nevertheless, both options are further evaluated below taking into account the two other aspects of area and space/thickness.

4.3.1 Two patches on TOP and BOP

In terms of **area**, both the TOP and BOP are constrained. They have a physical size of 10x10 cm.

- On the BOP, four MABs are placed. In the center of these four MABs, an area of 44x44 mm remains. In other words, a patch (ground plane) of some 40x40 mm could theoretically be squeezed in. However, the MABs could be placed somewhat further apart; *possibly*, 80x80 mm can be available given the width of 9 mm of the MABs.
- On the TOP, the (old) Delfi-n3Xt design is pretty full. Of this area, about half (~45x90 mm) is taken up by the SDM experiment, and the other half (~45x90 mm) is required for the sun sensor. In fact, both occupations are currently unsure; the sun sensor might no longer be placed and the SDM can possibly be moved to a side. The SDM however has a preference similar to the S-band patches; not being covered by a solar panel. Nevertheless, half the top panel might still be available.

In terms of **space**, the TOP and BOP are rather flexible:

- For the TOP, there is no reserved space underneath it. Given the large available volume on Delfi-n3Xt after the removal of the MPS payload, a patch could be lowered into the TOP.
- For the BOP, the maximum space is defined by the distance from the top of the MABs to the components on the DAB. MABs are some 3 cm high. *Possibly*, 2 cm is available.

In the above, it has been shown that it seems possible that a patch such as the SPECEMC patch can be integrated on the TOP and BOP. The main uncertainties are the placement of both the SDM and a sun sensor, and the solar panel configuration. On the BOP, the ground plane can possibly be decreased in size (from 45x45 mm to say 42x42 mm) to fit between the MABs in the current set-up.

The SSTL patch can *possibly* be integrated, but requires the maximum volume theoretically available on both the TOP and BOP. This might not be very desirable if even possible.

4.3.2 Two patches on opposite long sides

Area is simply no problem on the sides, at least as long as the specific location is not constrained.

Space is a more difficult variable, as the patches will be placed effectively against the side of the PCB stack. The following distances are relevant:

- 3-4 mm available between the solar panel when folded and the exterior of Delfi-n3Xt
- 1 mm thickness of the Delfi-n3Xt outer shell
- 4 mm between the outer shell and the PCB stack; only one side will house the bus from top to bottom.

In result, the sides would allow for limited thickness, but provide abundant area. Assuming the ground plane of the SPECEMC patch to be some 1 mm, the sides would allow for it but would require the patches to be lowered into the inner body of Delfi-n3Xt. They would not touch the PCBs however. Of course some type of connection would also be required to constrain the patch with respect to the outer shell and inner structure.

A final note however; the theoretical patch has in fact been designed to fit within the 3 mm between the outer shell and the solar panel when folded in. In theory then, a smaller patch could be used in the S-band antenna system. Unfortunately, such a patch is not (yet) found to be available commercially-off-the-shelf.

4.3.3 Conclusion

In conclusion, both options presented above of having two patches on the TOP/BOP or the long sides are possible under certain circumstances. The SSTL patch is most likely problematic, at least for the long sides and probably for the 'short' sides. A patch similar to the SPECEMC side would likely fit at different places, but would require lowering it into the outer shell of Delfi-n3Xt.

The mechanical and integration aspects can be seen to be incorporated as the CI *S-band Antenna Connection*, which together with the S-band Antenna(s) make up the S-band Antenna System.

A test model of the SPECEMC patch has been requested and has been sent from the Germany office to the Netherlands.

Integrating the patches on the long sides of Delfi-n3Xt would give one other advantage; the STX can be placed on the same height as the patches, requiring minimum cable length between the STX and the patches. On the S-band, signal attenuation per meter is sizeable. This is further assessed in the next chapter.

5 Antenna Wiring

The final CI that makes up the total Delfi-n3Xt Antenna System next to the UHF/VHF and S-band Antenna Systems is that of the Antenna Wiring. Its two components are quickly addressed in the next two sections 5.1 and 5.2. Nevertheless, near the final integration phases the mechanical and integration aspects should be assessed in more detail. For a communication point-of-view, losses are most important and are discussed below along with several cable and connector options.

5.1 Wire connectors

In case of Delfi-C^3, two types of connectors were used at different places. The indicated diameter values are the outer diameters of the plug.

- **SMA connectors** (Ø ≈ 6.4 mm): These connectors were used to connect the coax cables between the RAP and the antenna boards.
- **MMCX connectors** (Ø ≈ 2.3 mm): These smaller connectors were used to connect the MABs to the antenna boards.

The main difference between the two types is the size and their method of connecting; SMA connectors have screw thread, and the MMCX connectors are 'clicked'. Similarly, the MMCX connectors require a certain force to loosen them; logically this force is rather high to withstand launch loads for example. The disadvantage of this is that this force can damage the solder connections when integrating and testing, as has happened on some occasions with Delfi-C^3. Nevertheless, the size difference is sizeable, and proper handling (experience) lowers the negative impacts.

Smaller connectors generally have the disadvantage of less area to handle them, and slightly increased (theoretical) losses. These losses are however still relatively low and a general assumption is made in the link budget, due to the lack of detailed data, which covers all connector types.

Another significant impact of the chosen SMA connectors related to size is that of their orientation; they were oriented vertically. In other words, the cable was inserted at the top of the connector. The effect of this was that the volume required for a connector (jack) and cable (plug) was already almost 3 cm. All other required components on the radio PCBs can fit within a maximum of 2 cm.

For Delfi-C^3, size was not an issue as volume was largely available. For Delfi-n3Xt this is less so, although volume is not critical (anymore). Nevertheless, it seems desirable to reduce required volume, at least to create a more compact modular product. Some concrete suggestions for the large connectors used for the connection between the PTRX/ITRX and DAB are given below.

- ***Choose horizontally oriented SMA connectors as opposed to vertical ones***. In this manner, some extra space is required horizontally, but the overall height can be limited to some 1 cm. The downsides of the increased horizontal size required can if necessary be compensated for by creating a small cavity in the PCB.

- ***Choose MMCX connectors or similar SMP connectors***. The SMP connectors have not been introduced above but are very similar in size and configuration to MMCX connectors. The downside as mentioned above is the required force when assembling and disassembling. This can however partially be compensated for by orienting these connectors vertically as well; in this manner the force is still applied to the connection but also in the direction of the PCB as opposed to perpendicularly to it. A possible SMP integration is shown in Figure 5.1 below.

- ***Choose a Harwin connector with integrated coax connectors***. The standard coax connectors have a diameter of 2 mm and are clickable as well. Harwin connectors are used for all bus interfaces. Although it does not seem preferably to integrate the coax connections with the bus connector for reasons of modularity (testability), it might be an option to use connectors similarly sized as the Harwin bus connectors.

An illustration of a possible vertical SMP connector integration and that of the Delfi-C^3 horizontal SMA connector is shown as Figure 5.1.

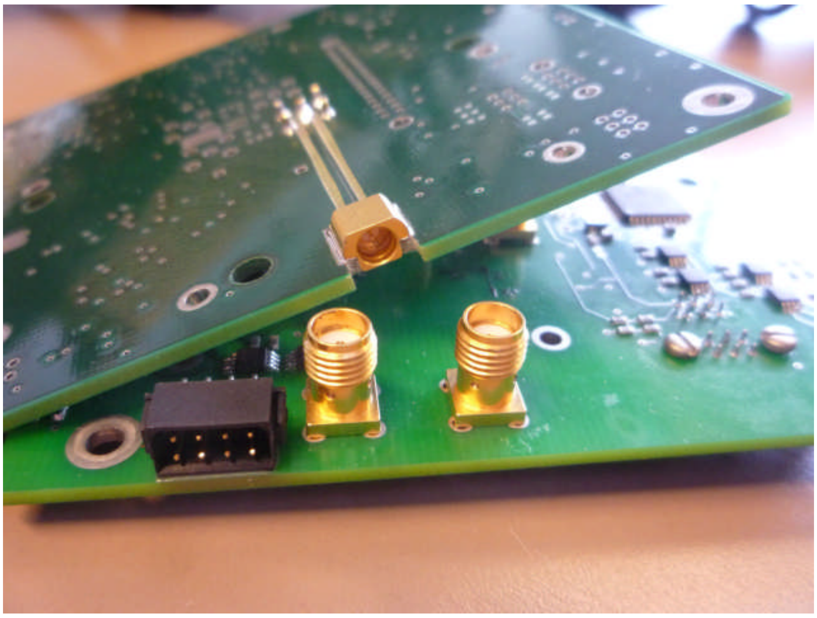


Figure 5.1: Relative difference between a vertical SMA jack (bottom) and a horizontal SMP jack

5.2 Wires

For Delfi-C^3 a specific type of wire was used, with type number: *RG178BU*. It has an impedance value of 50 Ω as required, and it has an outer diameter of almost 2 mm [SLR 0273]. It is a very flexible wire and has reasonably low attenuation values.

- For **UHF and VHF frequencies** the cable loss is less than -0.1 dB for both the up- and downlink, based on an assumed cable length of 10 cm.
- For **S-band frequencies** the attenuation depends on the final cable length, but higher attenuation values per m are realized; some -2.3 dB per m. The datasheet however gives only attenuation values for 1 GHz and 5 GHz (between -1.5 and -3.4 dB), so an assumption is made here. Depending on the final integration approach of the S-band patches, required cable length might or might not be significant; between ~5 and ~20 cm. Given the above attenuation per m, the incurred losses could amount to -0.5 dB.

The datasheet of the SPECEMC patch in fact suggests a standard cable to be connected to the patch, with type number: *RG-316*. It similarly has an impedance of 50 Ω, and has a slightly increased diameter of 2.5 mm. It can be assumed to be slightly less flexible than the RG178BU cable.

- For **UHF and VHF frequencies** the datasheet [SLR 0747] gives slightly lower attenuation values than those of the RG178BU. Given the short cable length however only gains of less than 0.05 dB can be realized with respect to the cable type suggested above.
- For **S-band frequencies** the RG-316 gives attenuation values of only 1.36 dB/m for S-band frequencies. This cable is preferred over the RG178BU as it gives up to 0.2 dB less loss than the old cable over 20 cm.

The following Table 5-1 summarizes the results.

Table 5-1: Minimum gain values of the resulting gain pattern [SLR 0273 & 0747]

Cable type	**Attenuation/m for VHF frequencies**	**Attenuation/m for UHF frequencies**	**Attenuation/m for S-band frequencies**
RG178BU	0.49 dB/m	0.82 dB/m	2.30 dB/m
RG-316	0.30 dB/m	0.57 dB/m	1.36 dB/m

Of course thicker cables could also be applied with even lower loss values. An example is the RG142B/U cable, which applies *double shielding* (two braided shields around the central conductor) and therefore has an increased diameter of 5 mm. it attains losses of only 0.44 dB/m for 1 GHz signal frequencies (the highest value in the data budget) compared to 0.86 dB/m of the RG-316. For low frequencies its loss values are also more or less halved. Therefore, this type of cable could be used to achieve even lower attenuation losses, but of course at the cost of even lower flexibility.

In fact, the data sheet of the *RG-316* states that the minimum bend circle is some 15 mm. The bend radius of the thicker RG142B/U cable however can be found to be about 1 inch, or 25.4 mm. This latter value is sizeable and is assumed to be undesirable if the cables are relayed along the PCBs.

It is therefore concluded that *the preferred cable for Delfi-n3Xt RF signal transmission is the RG-316*. A non-exhaustive search has been executed based on a suggested option for the S-band. Its flexibility is deemed sufficient and it is only 20% thicker than the RG178BU applied on Delfi-C^3 applied cable, but improves the attenuation performance by ~40% at S-band frequencies.

For the UHF/VHF frequencies, it can be concluded to reuse the *RG178BU*, given the lower absolute attenuation at lower frequencies. It has the benefit of having slightly increased flexibility, and of course has proven heritage although this argument is less important for a simple cable. In final integration it can be decided which of the two types will be used for the UHF/VHF signal transmission.

6 Next Steps

The Antenna System of Delfi-n3Xt has been shown to be directly divisible into three components or CIs: The UHF/VHF antenna system, the S-band antenna system and support equipment in the form of wiring and connectors. The next steps are then logically assessed again for each of these three CIs separately.

For **the UHF/VHF antenna system**, it has been shown that the UHF and VHF antennas can be combined, and that most likely the reuse of the Delfi-C^3 configuration and hardware will provide sufficient or even best performance.

The main insecurity is:

- The new Delfi-n3Xt solar panel configuration

The next steps:

- When the new solar panel configuration is known, it should be confirmed whether the Delfi-C^3 antenna configuration is still preferred. Most likely this can be done by applying common sense. However, a more detailed digital analysis can also be done.
- Final integration tests should be performed, to confirm that:
    - The combination of the antennas and the new phasing circuit do not lead to unwanted effects
    - The gain pattern of the final antennas is acceptable for all angles

For the **S-band antenna system**, it has been shown that the use of two patch antennas applied to two opposite sides of Delfi-n3Xt is preferred.

The insecurities are:

- The new Delfi-n3Xt solar panel configuration
- The specific patch that will be used and its performance

The next steps:

- Depending on the new solar panel configuration, it should be decided whether the antenna patches are applied to the TOP/BOP or to two opposite long sides of Delfi-n3Xt.
- The patch that will be used should be determined. Possibly, it can be either of the two patches presented in chapter 4. One SPECEMC sample patch has been ordered and is on its way; it can consequently be tested. If however, following a more in-depth research of available antenna patches, it is desired to choose another patch, it should most likely have a gain pattern similar to that of the SPECEMC and SSTL patches. Given the omnidirectional application, it is most important that the gain value at off-boresight angles of 90° is sufficient.

For the **antenna wiring**, concrete suggestions for the wires and connectors have been made. Whenever testing and integration take place, the suggestions can be taken along and using arguments directly relevant to the integration the final decisions can be made.

Appendix D

Delfi-n3Xt Link Budget

SLR 0106

Note: the link budget Excel sheets are appended to the end of this document, in the following order:

1. PTRX downlink, orbit altitude = 600 km
2. PTRX downlink, orbit altitude = 850 km
3. PTRX uplink, orbit altitude = 600 km
4. PTRX uplink, orbit altitude = 850 km
5. STX downlink, *optimistic scenario* as introduced in [SLR 0014]

Within this document, the following sections are based on texts from the earlier versions of the project documentation. These earlier versions have not been written by the author. Per section a short line explains the extent of the adoption. In all cases have the texts been updated and edited for correctness and relevance.

- **Section 2.3 Channel losses** (~3 pages)
  The content of the subsections on path loss, atmospheric and Ionospheric losses and polarization mismatch loss (2.3.2-4) is largely based on the old texts.
- **Section 2.5: Received E_b/N_0** (~1/2 page)
  The content of this section is based on the old content.
- **Link budget sheets**
  The lay-out and content of the appended link budget sheets is based on the pre-existing sheets.

Delfi-n3Xt Link budget

Description: Companion document to the actual link budget, offering explanations and background.

Subsystem(s) involved:

ADCS	CDHS	COMMS	EPS	MechS	STS	TCS	ITRX	MPS	$T^3\mu PS$	SDM	Splash	GSE	GSN	Launch
		X											X	

Revision Record and Authorization

Issue	Date	Author / Editor	Reviewer checked	PM approved	Affected Section(s)	Description of change
1.0	22/01/2009	Martijn de Milliano	AT			First issue
2.0	13/04/2010	Arthur Tindemans			1 2	Completely rewritten Multiple changes: -Better introduction & structure -List of variables completed -Expanded on all variables and added actual values -Irrelevant data deleted
2.1	13/06/2010	Arthur Tindemans			2 3 4	Values updated Chapter added Next steps updated

List of Used References

SLR code	Version	Data/Variable
0106	3.0	Companion document - link budget

Table of Contents

1 Introduction

A large number of budgets are kept on the satellite level for Delfi-n3Xt, that is to say the volume budget, mass budget, power budget and data budget. The link budget is a special case, as in fact it is not kept on the satellite level, but it describes the interface between the communication subsystem of Delfi-n3Xt and the ground station. It can be used to:

- Determine the final link margin
- Establish achievable data rates

The link budget is actually kept in a separate file with the same SLR code [SLR 0106], that and this file are *companion files*. This document introduces which sheets can be found in the other (spreadsheet) file, and more importantly serves to explain the values that are integrated into the link budget.

Three transmission links should physically exist in case of the Delfi-n3Xt mission, being the VHF downlink, the UHF uplink and the S-band downlink. As currently the orbit altitude has not yet been determined, multiple sheets are kept to quickly demonstrate the differences. The data rate of the PTRX downlink and uplink which can be maximized can be changed on the sheets. In case of the STX, one sheet is kept as there are still many variables that should be considered. Finally, the link budget of the ITRX is not in the link budget file; this link budget is created and maintained not by the Delfi-n3Xt team but by the provider of this payload, ISIS B.V.

As a result, currently 5 actual link budget sheets are kept:

- PTRX downlink
    1) 600 km
    2) 850 km
- PTRX uplink
    3) 600 km
    4) 850 km
- STX downlink
    5) Flexible link budget

The structure of this document is simple. Chapter 2 introduces and presents all link budget parameters. Chapter 3 then investigates the correctness of the VHF downlink link budget using data from Delfi-C^3. Finally chapter 4 concludes by presenting the next steps required.

The results of the link budget calculations are not discussed in this document, but in [SLR 0014].

2 Link budget parameters

The link budget is basically a sum of all losses and gains along a link. The starting point is the signal power produced by the power amplifier (PA) of the transmitter and the finishing point is the amplifier of the receiver. Theoretically, using a perfect medium, any signal, however low in power content, can be received and amplified to have useful strength again. This however no longer applies due to *noise* present in all transmission systems. Space, the main transmission channel, adds more less perfect Gaussian noise to the signal, implying equal noise at all frequencies. However, transmitter and receiver add noise as well, due to having nonzero temperatures for example. In the end, it is then the ratio of signal-to-noise that is of importance, as amplification will take along the noise present. In case of a digital link, the signal-to-noise ratio is expressed as the bit-energy-to-noise spectral density ratio E_b/N_0, being the energy contained within one bit (duration) over the noise. The link budget is specified using decibel (dB) values.

The following Figure 2.1 illustrates all gains and losses incurred or possibly incurred in the Delfi-n3Xt-to-GS transmission link. This figure shall be used as a guideline to present the parameters incorporated in the link budget.

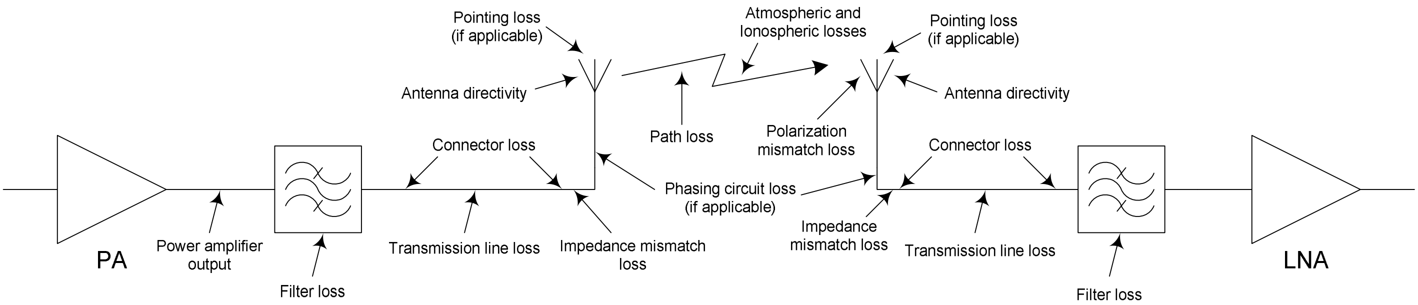


Figure 2.1: Losses incurred in a transmission link.

It can be seen that transmitter and receiver side are nearly identical, the only exception being the *power amplifier output* on the transmitter side. Also not shown are *coupler loss*, *pre-amplifier loss* and *surge protector loss*, incurred in some links of the specific Delfi-n3Xt implementation case. The end-result of the transmitter section, after the addition of the *antenna directivity*, gives the *EIRP* or *effective isotropic radiated power* which is a value often referred to.

Next to the transmitter/receiver sections, a channel section can be seen, where *path loss* and *atmospheric and Ionospheric losses* are incurred. Also *pointing loss* and *polarization mismatch loss* will be included in this category; these losses can be said to be incurred in and due to the addition of the channel, but more importantly this allows to respect the usual convention of having the *EIRP* as the end-result of the transmitter section.

Using the power amplifier output values and the end-to-end gains and losses, the effectively received E_b/N_0 can be calculated using the *system incurred noise* and *data rate*. Comparing this value to the required E_b/N_0 for correct data interpretation gives the final remaining link margin. Usually, a minimum link margin is also taken into account.

The following sections will introduce all parameters of interest in calculating the link budget. At the same time, the values applicable to Delfi-n3Xt scenarios are presented. The input, being the power amplifier output, is commented on in section 2.1. Section 2.2 discusses all transmitter/receiver gains and losses, including the three Delfi-n3Xt specific losses at the end, and section 2.3 consequently describes channel losses. Section 2.4 introduces signal noise. Section 2.5 then introduces the relation and parameters required to determine the received E_b/N_0 after which section 2.6 presents the required E_b/N_0. Knowledge of the latter two ratios logically results in a final link margin, and thus a confirmation of whether the link closes or not. Section 2.7 finally explains the inclusion of a minimum link margin.

2.1 Power amplifier output

The power amplifier output values have resulted from design considerations within the different systems (see [SLR 0014] and [SLR 0387]). Every transmitter has a power output value, thus there are three values in case of Delfi-n3Xt link budgets, one for each link.

The resulting values are:

- VHF downlink transmitter PA: 0.2 W or -6.99 dBW
- UHF uplink transmitter PA: 250 W or 23.98 dBW
- STX downlink transmitter PA: 0.315 W or -5.01 dBW

In case of the STX PA, the use of two PAs connected to two patches as assessed in [SLR 0387] will not be indicated by an increase in PA output, but by an increase in antenna gain. In this manner, incurred losses after the PA do only have to be counted once.

In case of the UHF uplink transmitter, present in the ground station this value assumes the use of the post-amplifier which in practice is used. The ICOM 910, which is the transceiver, can provide 75 W output power (according to its data sheets). The BEKO HLV-550 can yield 550 W of transmission power. The allowed transmission power is however limited to 400 W on radio amateur bands. Also, as high values cause reflections in the antenna which turn off the amplifier as a build-in safety measure, the actual transmission power used for Delfi-C^3 commanding is about 250 W.

2.2 Transmitter/receiver gains and losses

As introduced above, transmitter and receiver incur mainly the same losses. These losses are explained in this section. The consequent subsections 2.2.1 to 2.2.6 discuss all losses portrayed in Figure 2.1 from left to right on the transmitter side up until antenna directivity. Subsections 2.2.7 through 2.2.9 discuss the extra losses of relevance to the Delfi-n3Xt implementation.

2.2.1 Filter loss

A filter, possibly integrated in a larger transceiver, can be in place in the transmitter and receiver. In both cases it logically serves to limit the bandwidth, but its insertion causes an *insertion loss*. In the transmitter a filter is required if the bandwidth produced does not meet certain spurious signal reduction requirements. Possibly, filtering has already taken place before the PA and it might not be required. In the receiver a filter limits the bandwidth to be amplified and processed afterwards.

In case of all ground station receivers and transmitters, no separate filter is in place. The filters are instead integrated in the transceiver equipment. In case of a ground station transmitter, the ultimate power amplifier output is known after which no filtering takes place anymore, thus no (additional) filtering loss occurs. The ground station receiver applies no separate filtering and its final loss before amplification is the pre-amplification loss, see subsection 2.2.8. The S-band receiver should apply similar pre-amplification to have no additional filter loss.

In case of the PTRX transmitter no filter is in place after amplification. The PTRX receiver does filter using a SAW filter; in case of Delfi-C^3 mismatched 100 Ω filters were used as matching versions were not available. However, as these are available now, a suggested update is a Filtronics SF434040A [SLR 0213] leading to a typical 1.5 dB insertion loss, and a maximum of 3.5 dB. Therefore, 2 dB is assumed. For the STX no downlink filtering is likely necessary as the CC2500 has internal (digital) bandwidth filters.

In result, the only nonzero loss:

- UHF uplink receiver: -2 dB

2.2.2 Connector loss

As generally the physical transmitter and receiver are separate from the antenna, a connection between the two is necessary by means of a disconnectable wire. The ends of the wire then require *connectors*. As however any connection is slightly imperfect a connector loss is incurred.

On Delfi-n3Xt, the connector types are still to be confirmed. Delfi-C^3 however applied SMA connectors between the RAP and the antenna boards, and smaller MMCX connectors between the antenna board and the MABs. Specific data for the connector losses is not available, but typical losses are in the range of a few tenths of a dB depending on the RF. For UHF/VHF frequencies, -0.20 dB is then assumed for all three connectors, and -0.20 dB for S-band frequencies for a single connector and the patch antenna connection.

On the ground station, gold-plated N connectors are used. These have very little losses, with little exact specification, but a maximum of -0.15 dB at 10 GHz stated at various sources. As UHF and VHF transmission frequencies are only 0.145 and 0.435 GHz, a conservative constant value of -0.10 dB for all connectors will be used, even though this impact drops below the actual accuracy of the link budget. For the S-band, -0.20 dB will be assumed for two connectors in total.

Therefore the following values result:

- VHF downlink transmitter: -0.2 dB
- VHF downlink receiver: -0.1 dB
- UHF uplink transmitter: -0.1 dB
- UHF uplink receiver: -0.2 dB
- STX downlink transmitter: -0.2 dB
- STX downlink receiver: -0.2 dB

2.2.3 Transmission line loss

Next to the connectors, the wire or *transmission line* in between also causes losses to occur, which can be sizeable for high-frequency RF signals. The loss depends on the length of the lines and the attenuation per unit length. The attenuation depends on the type of cable used, and increases with frequency. In case of all Delfi-n3Xt transmission lines, a thin, low-weight coax cable is used [SLR 0036]. On-ground, two cable types [SLR 0014] are used, separated by surge protectors. However, in case of the downlink a pre-amplifier effectively takes the second type out of the equation, as is explained in section 2.2.8 below.

The total transmission lengths per cable type are given below. The actual loss is then calculated in Excel, using also the attenuation per meter for the transmission frequency at hand. For the UHF and VHF satellite cables, 0.10 m is taken per connection, assuming all radio are housed relatively close to the antennas. The S-band cables are assumed to be 0.20 m. This assumes the maximum required length where the STX is located in the middle of the satellite and cables run towards either the BOP, TOP or both. In result:

- VHF downlink transmitter: RG178BU (0.1 m)
- VHF downlink receiver: AIRCELL 7 (5 m)
- UHF uplink transmitter: ECOFLEX 15 (20 m) & AIRCELL 7 (5 m)
- UHF uplink receiver: RG178BU (0.1 m)
- STX downlink transmitter: RG178BU (0.2 m)

For the STX ground station set-up, the loss incurred by the few metres of high-quality cable between the antenna and pre-amplifier would already be very significant, given losses of some -0.40 dB/m. Then, it becomes important to have the pre-amplifier directly connected to the antenna. As however also a surge protector shall be required, some cabling might be necessary. As such, the transmission line loss for the S-band downlink receiver is set to a conservative 0.5 dB.

- STX downlink receiver: -0.5 dB

2.2.4 Impedance mismatch loss

A loss in power can occur due to a mismatch in impedance basically at all connections, but usually the interface between the transmission line and the antenna is the most important and used to indicate the possible impedance mismatch loss. The transmission wire can change the effective impedance of the signal and should be matched to the antenna, possibly by introducing additional circuitry. This can be done for example by adding a tuning circuit consisting of passive components such as capacitors or inductors, in which case there is no (theoretical) loss and all power is put onto the antenna. For a patch antenna, matching can be done by properly positioning the feed on the patch surface.

In case of both the VHF and UHF antenna systems, the impedance mismatch losses are incurred on the phasing circuit. As also other power losses are incurred on this phasing circuit, the losses are indicated as *phasing circuit losses*. These losses are explained in the next subsection. The UHF and VHF impedance mismatch are then zero. On the S-band antenna system, a value of -0.2 dB shall be assumed, assuming a Voltage Standing Wave Ratio of 1.5:1, as also used in a link budget of ISIS for their S-band link. On-ground 50Ω matched COTS products are used and no extra losses are assumed other than already incurred connector and transmission line losses.

The only resulting values:
- S-band downlink transmitter: -0.2 dB

2.2.5 Phasing circuit loss

In both transmitter and receiver on-board of Delfi-n3Xt (at least in case of the UHF/VHF link) a phasing circuit is used to properly phase the signals between the four antennas. This circuitry can also be seen as impedance matching circuitry; all losses can be allocated to the phasing circuit. The phasing circuit causes sizeable losses to be incurred and should be taken along.

The phasing circuit losses for the UHF/VHF phasing circuit are stated in [SLR 0715]. For the STX a separate phasing circuit does not seem to be necessary. In result:
- VHF downlink transmitter: -1.7 dB
- UHF downlink receiver: -2.15 dB

2.2.6 Antenna directivity

An antenna never is perfectly isotropic, or omnidirectional. As such, there is a direction of minimum directivity, and similarly a direction of maximum directivity. Some antennas aim to approach an omnidirectional pattern, and then apply no active pointing, such as the Delfi-n3Xt UHF/VHF antenna system. It that case, the minimum directivity should be used in the link budget, if permanent coverage is to be guaranteed. In other cases, active pointing is performed and the maximum directivity can be used, but a pointing loss is introduced as introduced in the next section.

The UHF/VHF antenna patterns are commented on in [SLR 0036] and figures are presented from which minima can be read. The UHF/VHF antenna gains on-ground can be read from the antenna specifications of both the VHF antenna *M^2 2M CP14* and UHF antenna *M^2 436CP30* [SLR 0734 and 0736]. However, these values state the gain in dBdc, being the gain referenced to a half-wave dipole antenna. In order to convert this to the required gain with respect to an isotropic (omnidirectional) antenna, 2.15 dB should simply be added.

For the S-band antenna system on-board of the satellite, an omnidirectional directivity pattern shall also be approximated, given the uncertain characteristics of the final ADCS. Two scenarios are described in [SLR 0036]; that of having a -5 dB loss and a -2 dB loss (with a closed link in ~80% of the time).

On-ground, S-band signals are to be received by means of a *parabolic reflector*. ISIS will likely be responsible for the installation of the S-band ground segment, and they say that they are able to provide a 3 m dish. A similar dish is used for the reception of the MOST (Microvariability and Oscillation of Stars) satellite [SLR 0738], and is said to have 35 dB maximum directivity. Similarly, [SLR 0219] mentions 3 m dishes to have 35 dB maximum directivity. A second, back-up option is the dish already in possession at the faculty of EEMCS (although without operating mechanics). This dish has a diameter of 0.7 m, leading to a maximum directivity of 22.3 dB. As the difference is sizeable the first option is definitely preferred.

The resulting values:

- VHF downlink transmitter (minimum): -5 dB
- VHF downlink receiver (maximum): 12.35 dB
- UHF uplink transmitter (minimum): -6 dB
- UHF uplink receiver (maximum): 16.40 dB
- STX downlink transmitter:
    - Fully omnidirectional: -5 dB
    - Achievable in ~80% of time: -2 dB
- STX downlink receiver (maximum)
    - 3 m parabolic reflector: 35 dB
    - 0.7 m parabolic reflector: 22.3 dB

2.2.7 Coupler loss

In case of the PTRX transmitter and assumably the STX as well, a *bidirectional coupler* is used to extract data on the signal strength right before transmission [SLR 0014]. In combination with a *power detector* this allows to determine the power forwarded to and reflected by the antenna system; in turn yielding valuable housekeeping data on operation and antenna deployment. In case of the PTRX this coupler also serves to provide the transponder AGC with the final signal strength. Unfortunately, the addition of this component after final amplification comes at the cost of an additional insertion loss. According to the data sheet of the integrated ADC-10-1R [SLR 0719], this loss is between 0.8 and 1.2 dB, thus assumed to be 1 dB, for VHF frequencies. In case of S-band frequencies this loss rises to 0.9-1.5 dB, or 1.2 dB averaged.

In result:

- VHF downlink transmitter: -1.0 dB
- STX downlink transmitter: -1.2 dB

2.2.8 Pre-amplifier loss

In case of receiver systems were a significant distance exists between the actual receiver and the antenna system, it is common to use a *masthead pre-amplifier*. It then makes sense to also see this amplifier as the end of the chain, even though some losses are still incurred and noise is still introduced behind the amplifier. But, as the later losses are incurred with respect to the *new* signal strength it makes sense to neglect the later-incurred losses of the original signal as the orders of magnitude become very different.

However, this added pre-amplifier comes at the cost of an insertion loss. In case of the UHF downlink it is equal to 0.2 dB according to its specifications to be found on the internet (SP-2000 amplifier). In case of the S-band downlink a similar masthead pre-amplifier has not yet been specified. However, the pre-amplifier loss resembles the connector losses and surge protector losses in value, at least for the VHF downlink. Therefore, a value of -0.3 dB shall be assumed.

In result:

- VHF downlink receiver: -0.2 dB
- STX downlink receiver: -0.3 dB

2.2.9 Surge protector loss

Finally, all ground station receivers as well as transmitters with antennas at elevated positions need lightning protection systems. Therefore simple bandpass filters designed to ground lightning-incurred frequencies and voltages, or *surge protectors* are inserted directly next to the antennas. Again, these cause an insertion loss to be incurred. In case of the VHF downlink and UHF uplink, the Telegärtner J01028A0037 is used [SLR 726], which is said to have an insertion loss of 0.2 dB for N-type connectors, for frequencies up to 2.5 GHz. The same surge protector can then be used for the S-band downlink.

Then:

- VHF downlink receiver: -0.2 dB
- UHF uplink receiver: -0.2 dB
- STX downlink receiver: -0.2 dB

2.3 Channel losses

The losses that have been defined as channel losses are the pointing loss, path loss, atmospheric and Ionospheric loss and finally the polarization mismatch loss. These losses are discussed in subsections 2.3.1 through 2.3.4 below.

2.3.1 Pointing loss

The pointing loss factor is the loss due to an error in pointing of both the spacecraft as well as the ground station, if applicable. The error in pointing introduced is then the difference between the actual line-of-sight vector and the direction of the maximum of the antenna major lobe. The effect of a certain pointing error increases when the directivity of the antenna is higher, thus is most important on-ground. The UHF/VHF antenna system on Delfi-n3Xt will in any case not apply pointing, and similarly the S-band satellite antenna system will no longer be assumed to. The ground station will apply pointing for all transmission links.

The pointing accuracy of the ground station antenna is limited by the TLE accuracy, antenna rotor mechanism and data processing delay as well as environmental conditions (e.g. oscillations due to strong winds). Consequently, a pointing error incurs a non-zero angle with respect to the point of maximum gain, and as such the actual pointing loss is dependent on the exact antenna gain pattern.

In case of the UHF and VHF antennas, general assumptions are made, confirmed by Wouter Weggelaar (who does now manage the GS) to be representative for the Delft GS; these are -0.5 dB and -1 dB for respectively the VHF and UHF systems.

In case of the STX, the on-ground pointing capability can be estimated. Along with the promise of providing a 3 m dish, ISIS seems to be confident to achieve 0.5° pointing accuracy with the tracking mechanism. To this possible orbit errors and weather effects should be added. [SLR 0738] mentions that with the MOST mission, 0.5° *overall* pointing accuracy was even achieved, while shutting off during unfavourable wind conditions. For now, an overall 1.5° pointing error shall be assumed. Heavy winds might cause the antenna system to be out of operation.

[SLR 0219] can be used to calculate the pointing loss given the pointing accuracy. For a 3 m parabolic reflector, an 0.5° pointing accuracy would yield 0.3 dB loss, whereas a more realistic 1.5° causes 3.2 dB pointing loss. Similarly 1.5° pointing accuracy would yield 0.2 dB loss for a 0.7 m parabolic reflector.

Resulting values:
- VHF downlink receiver: -0.5 dB
- UHF uplink transmitter: -1.0 dB
- STX downlink receiver
  - 3 m parabolic reflector: -3.2 dB
  - 0.7 m parabolic reflector: -0.2 dB

2.3.2 Path loss

The path loss L_p is the loss in signal power resulting from an increasing distance between transmitter and receiver. It is dependent on the wavelength of the signal, and is given by the following relation [SLR 0316]:

$$L_p = \left(\frac{\lambda}{4\pi \cdot S} \right)^2 , \tag{2-1}$$

Where

λ	wavelength of the signal
S	distance between transmitting and receiving antenna (also called slant range)

The slant range S depends on the elevation of the satellite above the horizon [SLR 0316]:

$$S = R\left(\sqrt{\frac{(R+h)^2}{R^2} - \cos^2 \varepsilon} - \sin \varepsilon \right), \tag{2-2}$$

Where

R	radius of the Earth (6378 km)
h	orbit altitude in km
ε	satellite elevation above the local horizon in radians

The variation of the slant range over a range of elevation angles is shown in Figure 2.2 below for the Delfi-n3Xt range of orbit altitudes.

The resulting path loss in decibels versus elevation angle is plotted in Figure 2.3; a lower (negative) decibel value is actually worse, so these are plotted highest on the Y-axis. Losses become (significantly) lower with elevation angle, due to the similarly large decrease of slant range.

The distance typically used in calculating the path loss is the maximum slant range for which a communication link with the satellite is desired, in other words by selecting a minimum elevation angle at which the link budget is supposed to close. This then assumes a constant data rate through the pass, or even the mission. Theory on selecting this minimum elevation angle by pursuing a maximum downloadable data volume is presented in [SLR 0014].

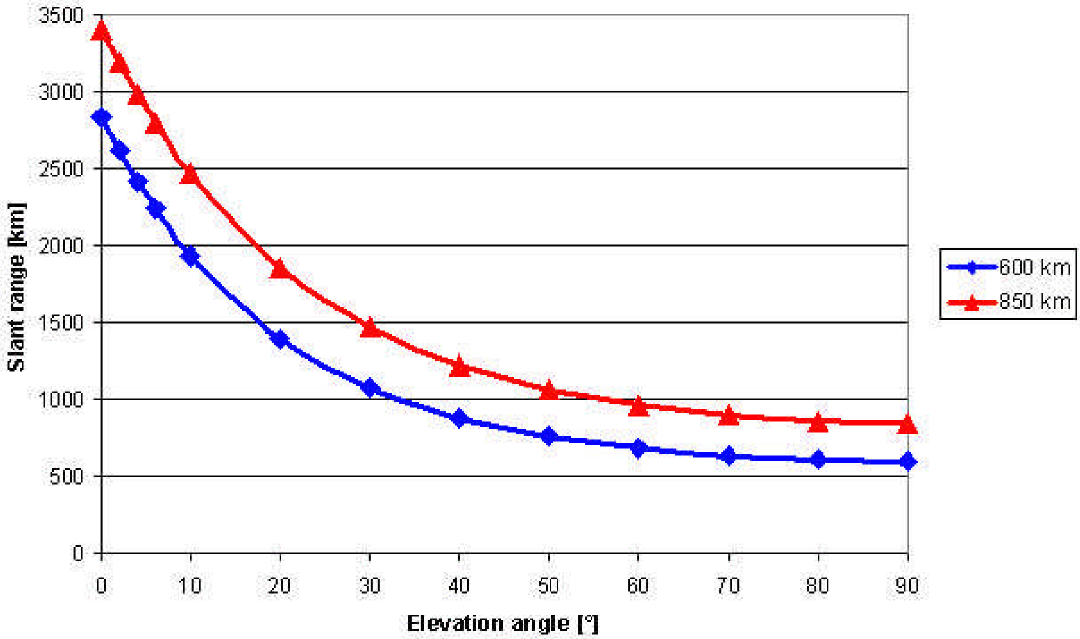


Figure 2.2: Slant range versus elevation angle for the Delfi-n3Xt range of orbit altitudes

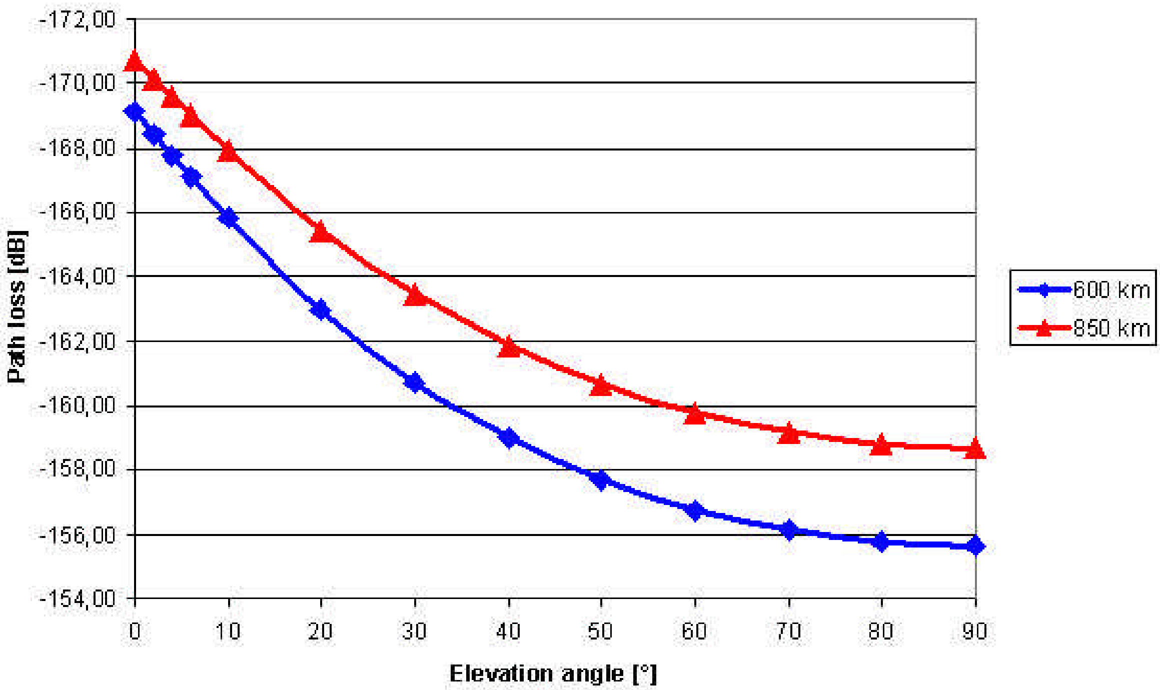


Figure 2.3: Path loss versus elevation angle for the Delfi-n3Xt range of orbit altitudes

In case of the primary UHF/VHF link the link budget will de designed for maximum contact time; in other words assuming a worst case path loss. The consequence is that lower data rates can be achieved. This choice has been discussed in [SLR 0014], but results in maximum data yield when all (off-zenith) passes are utilized. Due to increasing atmosphere losses, antenna noise and obstructions at low elevation angles a minimum elevation angle of 10° is selected for all VHF and UHF communications.

For the STX, the data rate is supposed to be changeable. Then, the path loss value does not need to be based on a worst case value. As data transfer is maximum for a zenith pass with a data rate based on a minimum elevation angle of about 40°, a worst case value and an optimum path loss value are given.

Also, in all cases the path loss depends on orbit altitude and Delfi-n3Xt orbit altitude is currently scheduled to be between 600 and 850 km.

The resulting path losses in dB:

- VHF downlink, 600 km, 10° elevation angle: -141.40 dB
- VHF downlink 850 km, 10° elevation angle: -143.53 dB
- UHF uplink, 600 km, 10° elevation angle: -150.94 dB
- UHF uplink 850 km, 10° elevation angle: -153.07 dB
- S-band downlink, 600 km, 10° elevation angle: -165.77 dB
- S-band downlink, 850 km, 10° elevation angle: -167.90 dB
- S-band downlink, 600 km, 40° elevation angle: -158.97 dB
- S-band downlink, 850 km, 40° elevation angle: -161.83 dB

2.3.3 Atmospheric and Ionospheric losses

When the signal traverses space towards a massive body, two effects cause degradation of the signal:

- Absorption by atmospheric gasses
- Phase delay introduced and reflections caused by the ionosphere

The absorption of atmospheric gasses decreases with increasing elevation (the signal has to pass through less atmosphere), and losses are approximately independent of transmission frequency up until 2 GHz transmission frequencies. This includes the UHF- and VHF-bands. For the S-band, a small rain attenuation correction (additional loss) shall be given below. Documented values are listed in Table 2.1.

Table 2.1: Losses due to the presence of atmospheric gasses [SLR 0219]

Elevation angle [°]	Loss [dB]
0	10.2
2.5	4.6
5	2.1
10	1.1
30	0.4
45	0.3
90	0.0

Ionospheric effects can be researched in large detail, but is not essential for UHF-, VHF- and S-band frequencies. These frequencies are far less influenced than lower transmission frequencies. As such, usable Ionospheric attenuation values are given in Table 2.2.

Table 2.2: Ionospheric losses [SLR 0219]

Frequency [MHz]	Loss [dB]
146	0.7
438	0.4
2410	0.1

In case of high frequency signals, rain fall becomes very relevant. At S-band frequencies, this is only marginally so. The atmospheric losses take into account the presence of water in the atmosphere. However, one can imagine that the presence of clouds and rain distorts the figures. [SLR 0730] gives the losses per km in which there is a rainfall of anywhere between 0.25 mm/h and 150 mm/h. In the Netherlands, with an average rainfall of not even one meter per year, 10 mm/h would already be very exceptional. For possible STX passes above other regions, the value might be higher. The resulting rain loss as presented in [SLR 0730] for a 10 mm/h value is already off the chart; below 0.01 db/km. Assuming clouds to be not above 10 km altitude, but taking into account low elevation angles, a conservative loss of 0.1 dB is taken into account in the link budget.

The total atmospheric and Ionospheric losses relevant for the Delfi-n3Xt link budget can be obtained by adding the values presented above. In result:

- VHF downlink, 10° elevation angle: -1.8 dB
- UHF uplink, 10° elevation angle: -1.5 dB
- S-band downlink, 10° elevation angle: -1.5 dB
- S-band downlink, 40° elevation angle: -0.5 dB

2.3.4 Polarization mismatch loss

The electric field of an electromagnetic wave can be seen as a vector in a plane perpendicular to the direction of travel. The vector can be decomposed into two orthogonal components, for example an E_x component and an E_y component, where the x and y direction are perpendicular. The relative amplitude of and the relative phase between these components determines the polarization of the wave. There are three types of polarization [SLR 0261]:

- ***Linear polarization***: the relative phase between the components is 0° or 180°
- ***Circular polarization***: the amplitude of both components is equal and the relative phase between the components is 90° (left-hand polarized) or 270° (right-hand polarized)
- ***Elliptical polarization***: when the wave is neither linearly nor circularly polarized

The polarization of the receiver antenna must be matched to the polarization of the incoming wave. If the wave is linearly polarized, then the antenna must be linearly polarized for maximum energy transfer. If the polarization directions of both antennas are perpendicular, no signal is theoretically received. Also, linearly polarized signals will be elliptically polarized by the channel due to an incurred Faraday rotation of the Earth.

As the attitude of the satellite is not fixed with respect to the ground station, and due to the circularization of the Faraday rotation, circularly polarized signals will be used for communication between Delfi-n3Xt and the ground. For the UHF and VHF systems, polarization mismatch shall be set to 0 as the minimum directivity values of the satellite have been taken at which circularly polarized transmission is achieved. On-ground, one can switch between left-handed and right-handed circular polarization (LHCP and RHCP respectively), depending on satellite attitude. The S-band patch directivity values are similarly based on the RHCP directivity pattern.

In result:

- All links: 0 dB

2.4 System noise

As is introduces in the next section, the combination of the Boltzmann's constant and system noise temperature gives the noise spectral density N_0. System noise temperature values are not self-evident, thus are given more attention and explained here.

System noise can be expressed with respect to the LNA using the following equation [SLR 0219]:

$$T_S = \alpha T_a + (1-\alpha)T_0 + T_{LNA} + \frac{T_{2ndStage}}{G_{LNA} \cdot \beta} \tag{2-3}$$

Where:

α	Signal transmission efficiency between antenna and LNA (1 - total loss)
T_a	Antenna or sky temperature (not a physical temperature)
T_0	System temperature and reference temperature (physical temperature)
T_{LNA}	LNA noise temperature (not a physical temperature)
$T_{2ndStage}$	Noise temperature of the second amplifier (not a physical temperature)
G_{LNA}	Gain of the LNA
β	Signal transmission efficiency between LNA and second amplifier (1 - total loss)

In the relation above a number of temperature values are incorporated. However, in case of active devices, these are not actually physical temperature values, rather just representative values. For passive devices however actual temperature values are used.

Also, it is shown that all noise is assumed to be incurred in the receiver, but taking into account sky temperature. Due to the large signal strength at the transmitter, noise incurred there is not relevant to total system noise. Furthermore, a 2nd stage is shown, but no third stage; generally there are only two stages assuming a pre-amplifier and the actual receiver. Nevertheless, a large number of mixers and IF amplifiers can increase the amount of stages. As however the extra incurred noise temperature for an additional stage is to be divided by the gains of all preceding amplifiers, third and later stages can safely be ignored with little loss of precision.

It should also be noted that no decibel values should be entered in the above equation, rather just linear values. Decibel values can be recalculated to 'normal' linear values using the following relation:

$$X(linear) = 10^{\frac{X(\log)}{10}} \tag{2-4}$$

Also, in order to calculate noise temperatures T_X using a given noise figure F of devices such as amplifiers, the following relation can be used [SLR 0316]:

$$T_X = (F-1)T_0 \tag{2-5}$$

As opposed to the procedure followed for signal losses and gains occurred in the system, the system noise parameters will be discussed per case, being the VHF downlink, UHF uplink and S-band downlink. This is done in sections 2.4.1 through 2.4.3 below. For a good idea of the chain of components and their functionalities one is referred to [SLR 0014], where extensive diagrams are presented of GS, PTRX and STX receiver architectures.

2.4.1 VHF downlink

In case of the VHF downlink, the important section is the GS VHF receiver section. The first stage, up until the first amplifier, is formed by cables (plus connectors) between the antennas and amplifier, as well as an inserted surge protector. The second stage is formed by all cables and circuitry (power limiter and connectors) and the ultimate receiver, the ICOM 910. The resulting values to be used in equation 2-3 are shown and explained below:

α = 0.81 — This value is a sum of all losses up until the pre-amplifier, as indicated in the link budget. This includes connector, transmission line, surge protector and pre-amplifier insertion loss, summing up to -0.9 dB or 0.81 (using relation 2-4).

T_a = 150 K — Two sources, being SMAD [SLR 316] and TRASH (the radio amateur satellite handbook, the latter being referred to via different websites) indicate a value of 150 K. In the first source, 150 K is indicated as a usual system noise temperature at 0.2 GHz, which is close to the 0.146 GHz of the downlink. The other source indicates the same value for VHF frequencies, but mentions it to be worst case value for direct pointing at the horizon at minimum elevation. Furthermore, the Delft GS is located very high up with little ground interference. As such, 150 K seems to be a good, possibly slightly conservative value.

T_0 = 290 K — The physical temperature of the cables is set to be 290 K.

T_{LNA} = 58 K — The LNA has an indicated noise figure of 0.8 dB; using T_0 given above and relations 2-4 and 2-5 this can be calculated to yield 58 K

$T_{2ndStage}$ = 473 K — The noise temperature of the ICOM 910 can be calculated using the sensitivity of the transceiver, indicated in [SLR 0737] to be "0.11 µV for 10 dB signal-to-noise in 2.4 kHz bandwidth". [SLR 0248] gives the rather lengthy calculation, and arrives at a noise figure of 4.2 dB. Using again relations 2-4 and 2-5 and a T_0 of 290 K gives a noise temperature of 473 K.

G_{LNA} = 20 — [SLR 0725] indicates that a gain of 12-14 dB should be aimed for (which can be adjusted on the amplifier). An intermediate 13 dB equals a gain of 20.

β = 0.81 — Summing all losses after the pre-amplifier, being additional connector losses (estimated at -0.1 dB), transmission line losses (-0.7 dB based on cable attenuation data given in the link budget spreadsheets) and finally the power limiter loss (-0.1 dB according to [SLR 721]) gives -1 dB, or 0.81.

The resulting T_S or system noise temperature is ***264 K***.

2.4.2 UHF uplink

In case of the UHF uplink, the section of relevance is that of the PTRX receiver. Its architecture can be seen in [SLR 0014], and clearly has a large number of stages. Nevertheless as explained above, two stages are enough to give an estimate of the noise temperature. A small margin will be implemented at the end to account for additional noise. The values of relevance:

α = 0.23	This value is a sum of all losses up until the LNA, as indicated in the link budget. This includes phasing circuit, impedance mismatch, connector, transmission line and filter loss, summing up to -6.41 dB or 0.23.
T_a = 290 K	SMAD [SLR 316] indicates a value of 290 K for all frequencies in between 0.2 and 20 GHz, as well as 40 GHz. In other words, the noise temperature is more or less constant and depends mainly on the Earth albedo. In fact, this value is based on a pointed antenna directed at the Earth, as opposed to an omnidirectional antenna as used in the UHF/VHF links. Therefore, the value should logically be less. Nevertheless, as a conservative value it is kept. Also, due to the high losses (α) the noise temperature has only limited impact on overall noise temperature.
T_0 = 313 K	The physical temperature of the satellite systems generally should be between -40° C and +80° C for correct operation. A rather high but reasonable temperature of 40° C will be assumed here, or 313 K.
T_{LNA} = 72 K	The LNA (MAX2640) has an indicated noise figure of 0.9 dB [SLR 0138]; using T_0 given above and relations 2-4 and 2-5 this can be calculated to yield 72 K.
$T_{2ndStage}$ = 421 K	The IF amplifier (MAX2630) has an indicated noise figure of 3.7 dB [SLR 0139]. This can be calculated to yield 421 K.
G_{LNA} = 32	[SLR 0138] indicates a gain of 15.1 dB for the MAX2640, equaling a value of 32.
β = 0.15	After the first amplifier a filter and a mixer circuit are in place, next to negligible connections. This leads to a filter loss of -2 dB, as the same filter is used as before the first amplifier, and a mixer circuit loss of -6.3 dB according to the data sheet of the RMS-1 [SLR 0724]. The total of -8.3 dB equals a value of 0.15.

The resulting T_S or system noise temperature is ***467 K***. However, as said above a small margin for additional stage losses will be taken, so that the total system noise temperature shall be set at ***500 K***.

2.4.3 S-band downlink

For the S-band downlink, again the receiver ground segment is important for the noise calculation. [SLR 0014] presents the conceptual lay-out. Two stages shall be used in the noise calculation. The values of relevance:

α = 0.79	The sum of the receiver connector loss, transmission line loss, surge protector loss and pre-amplifier loss yields 1 dB total loss. This equals 0.79.
T_a = 200 K	SMAD [SLR 316] gives general antenna noise values of only 25 K above 2 GHz. This value however does not properly take into account man-made noise, in turn highest at low elevations. Also, the ISM-band is used for wireless internet applications, which might turn out to be a large source of noise on the ground station. Therefore, a conservative total 200 K antenna noise temperature is assumed.
T_0 = 290 K	The assumed physical temperature of the ground equipment.
T_{LNA} = 51 K	[SLR 0738] describes the use of low-noise amplifiers with a gain value of 30 dB and a noise figure of 0.7 dB. The latter then yields a noise temperature of 51 K.
$T_{2ndStage}$ = 500 K	Here again a large assumption is made. This second active device can be either a real amplifier, or the actual receiver. In case of the VHF downlink, the ICOM 910 had a noise temperature of 472 K, nearly 500 K. Therefore the latter value is assumed here, even though frequency bands are different.
G_{LNA} = 30	As argued above, following the low-noise amplifier specifics given in [SLR 0738].
β = 0.60	After the first amplifier some connectors and most notably a lengthy cable is in place. As with the UHF and VHF systems, some 20 m is estimated to be necessary. Given the attenuation per meter for the high quality ECOFLEX 15 cable of -0.10 dB/m, this yields -2.0 dB. This value is decreased to -3.2 dB to account for connector losses. -3.2 dB equals about 0.60.

The resulting T_S or system noise temperature is ***298 K***. Two large assumptions have been made:

- A 200 K antenna noise temperature to account for man-made noise
- A 28 K second stage noise temperature, resulting from an estimated 500 K noise temperature of the second active device; receiver or amplifier.

It can be seen that the impact of the second estimation is moderate, due to the large gain of the pre-amplifier. The antenna noise temperature is an important and tricky value. The link budget provided to the author by ISIS, describing their ground station implementation uses a more optimistic value of 100 K antenna noise. Furthermore, a 75 K amplifier noise value is used, but no more. The above calculation aims to be more detailed given the insecurities, but also stays on the safe side.

2.5 Received E_b/N_0

The power amplifier output multiplied by all consequent losses on the link basically yields signal strength at the receiver, or receiver power P_r. In case decibels are used, which is generally the case in a link budget, all losses are simply subtracted from (or added to, depending on the convention) the original power amplifier output. Taking into account the noise level at the receiver as well as the data rate of transmission, the overall received E_b/N_0 can be calculated using the following equation 2.3.

The received bit energy over noise spectral density E_b/N_0 is equal to [SLR 0316]:

$$\frac{E_b}{N_0} = \frac{P_r}{kT_sR} \tag{2-6}$$

Where:

k	Boltzmann's constant ($1.38 \cdot 10^{-23}$ J/K)
T_s	The system noise temperature in K
R	The data rate in symbols/s (Baud)

Equation 2.3 can be expressed in its decibel equivalent, see equation 2.4:

$$\frac{E_b}{N_0} = P_r - 10\,{}^{10}\log(kT_s) - 10\,{}^{10}\log R \tag{2-7}$$

This equation is the generally referred to as the *link equation* or *link budget*.

An implementation loss has been used to account for slight additional losses, but the current high detail of the budget and the vagueness of such a loss are no longer preferred. Instead, only a minimum link margin is incorporated, which is interpreted as a value added to the required E_b/N_0. These two parameters are introduced in the next two sections.

Finally, note that the data rate is given in symbols/s as opposed to bits per second. The reason for this is that Manchester coding requires two symbols per bit. Also, if forward error correction coding is added, there is a distinction between data bits and coding bits; together they can just be seen as symbols.

2.6 Required E_b/N_0

In order to properly demodulate a signal, a certain signal-to-noise ratio is required per bit duration, as introduced before. If the received E_b/N_0 is higher than the required E_b/N_0, the signal can be demodulated with *at least* the preferred bit error rate or *BER*. In this case, a *link margin* is said to remain.

The required E_b/N_0 values for different modulation schemes, as well as the choices for different modulation schemes have been commented on in [SLR 0014]. As such, for the three transmission links, being that of VHF downlink, UHF uplink and S-band uplink, the following values can be stated for a *BER* of 10^{-5}:

- VHF downlink, differentially encoded BPSK or DEBPSK: 10.3 dB
- UHF uplink, noncoherent BFSK: 13.3 dB
- S-band downlink, MSK: 9.6 dB

Finally forward error correcting codes can be applied, so that a *coding gain* can be achieved. This can lower the required E_b/N_0 sizeably as a number of dB's can be subtracted from it. For the STX, [SLR 0014] argues that 2-4 dB coding gain can be achieved for low-complexity coding schemes, and 7 dB for a more advanced concatenated coding scheme. As a result two options shall be kept for the S-band downlink:

- S-band downlink, simple coding scheme: -3 dB
- S-band downlink, concatenated coding scheme: -7 dB

2.7 Minimum link margin

Subtracting the required E_b/N_0 from the received E_b/N_0 gives conclusion on whether the link *closes* and how much margin results. However, in order to account for inaccuracies in the link budget, a contingency is used and added to the required E_b/N_0. This contingency is called the *minimum link margin*.

The approach used in this link budget is slightly different from the approach that is often taken by link budget designers. The common approach is to conclude that the link closes when the received E_b/N_0 plus the value that is called the minimum link margin is higher than the required E_b/N_0. For Delfi-n3Xt however it has been chosen to explicitly state the minimum link margin, which leads to having just one resulting *final link margin*.

The minimum link margin is taken to be 3 dB for all links (REQ.2.3-P.03).

For the VHF/UHF systems this is based on the fact that heritage is available and the system is no longer experimental.

As the STX is an experimental system, it might make sense for it to have a large link margin as the insecurities are larger. However, a larger value would excessively limit the presented achievements and is deemed over-conservative. More importantly, as its operations will include the option to change its data rate, a link budget that does not close can be fixed during the mission. The same can in fact be said for the 3 dB link margin already in place; nevertheless some unexpected losses are still expected.

3 Verification using Delfi-C^3

As there is in fact one representative measure of the correctness of the link budget (for the VHF downlink only), this measure is shortly presented here. This is of course Delfi-C^3.

In case of this downlink, little relevant changes are in place between the PTRX and RAP from a link budget point of view. The ground station set-up as of now is completely the same, and on the satellite some small component losses such as those incurred in the phasing circuit are slightly different. Nevertheless, these effects are minor.

Figure 3.1 below shows the frequency spectrum as measured on the ground station, *between the pre-amplifier and transceiver*.

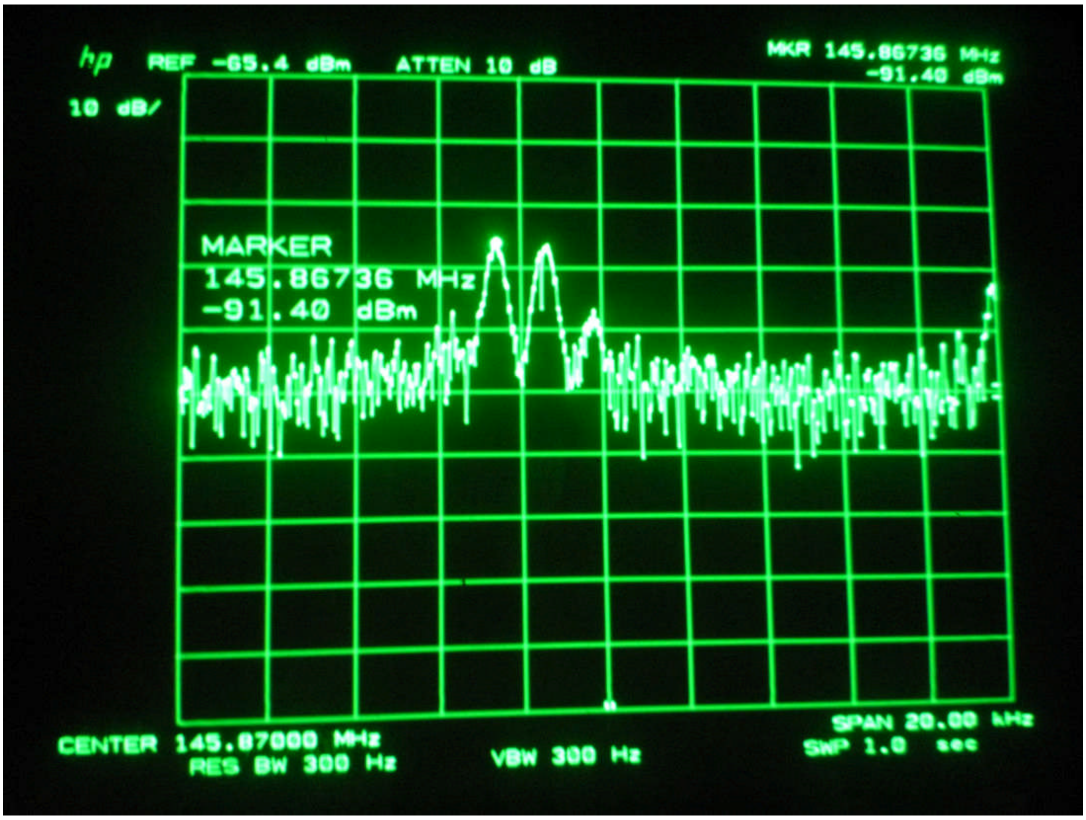


Figure 3.1: Measured frequency spectrum of the Delfi-C^3 transmission during a near-overhead pass

The bandwidth characteristics of the above measurement are discussed in [SLR 0014]. In this discussion, what is of interest is the measured power value of the peak(s). It is indicated to be ***-91.40 dBm***.

In case of the Delfi-n3Xt VHF downlink, a power value of *-147.25 dBW* results from the link budget, being the sum of the transmitted power, and the incurred channel and receiver losses. This is of course using a data rate of 1200 b/s. dBW and dBm are in fact related by a constant -30 dB difference; therefore the expected worst case power value presently in the link budget is ***-117.25 dBm***.

It can be seen, that the actual received power is much higher than the proposed *worst case* scenario. However, as is illustrated by Table 3.1 below, the difference is completely explainable.

Table 3.1: Theoretical range of signal strength values in the measurement case

Power value according to current link budget		**-117.25 dBm**
Differences in measurement case	Pre-amplifier gain	~ 13 dB instead of 0 dB
	Path loss (700 km)	-132.64 dB instead of -141.46 dB
	Atmosphere loss	-0.7 dB instead of -1.8 dB
Minimum theoretical signal strength value		**-94.33 dBm**
Variables	TX antenna directivity	-5 to 1.42 dB instead of -5 dB
	Pointing loss	-0.5 to 0 dB instead of -0.5 dB

In the table above a number of *differences in the measurement case* and *variables* are stated. The former values arise because:

- The satellite was in fact almost overhead. As the link budget is based on a minimum elevation angle of 10°, the path loss and atmosphere losses should be adapted for. A distance of 700 km is assumed for the adaption.
- A pre-amplifier increases the measured signal strength. In the link budget this value is not included as it also increases the noise signal strength. As explained before, an intermediate 13 dB is taken.

The variables in the link budget, for which worst-case values are used normally, arise because:

- The satellite attitude is unknown; its antenna might be oriented towards Delft with its point of maximum or minimum directivity. The actual maximum antenna directivity is shown in [SLR 0036].
- The on-ground antenna has a variable pointing loss.

In result, it can be seen that the theoretical signal strength value according to the link budget is in the range -**87.41 to -94.33 dBm**. The measured value of -91.40 dBm is almost perfectly in the middle of this range.

4 Next Steps

For the link budget, the next steps are logically related to further specification of gain, losses and noise values. As there are three links, the next steps are assessed per link.

Both the VHF and UHF communication link budget can be said to be of reasonably high accuracy and detail. No significant improvements are required as the links generally close with a sufficient margin. Nevertheless, a number of improvements can be suggested. The S-band link on the other hand is still subject to large insecurity. Although it will be closed for certain data rates and certain elevation angles, the exact data rates achievable are still unclear.

The link budget of the S-band downlink should therefore be further specified. The S-band downlink is discussed first, followed by the VHF downlink and UHF uplink respectively.

S-band downlink

The following variables can be said to have the highest insecurity given their impact:

- **Ground antenna directivity** and **pointing loss**
  Depending on the actual antenna (size), the antenna directivity can change by some 10-15 dB. The type of antenna combined with more information on the tracking system in turn can yield a better estimate of the pointing loss.
- **Coding gain**
  Given different coding schemes, a gain of between 3 and 7 dB can be expected. The exact coding scheme that is to be applied should be determined, and its actual performance should consequently be measured.
- **Satellite antenna directivity**
  A current -5 dB directivity is assumed, but depending on the actual antenna system performance and acceptable satellite attitudes this value might be somewhat different.
- **Antenna noise temperature**
  Most notably the effect of the many signals in the ISM band should be measured. The noise effects of these signals will also depend on the specific antenna (size). As data rates might be changed given the elevation angle of the satellite, its dependence on the latter is also important.
- **Coupler loss**
  Currently a coupler loss of -1.2 dB is assumed. This *bidirectional coupler* is likely useful to monitor the (power) performance of the patch antennas, given the experimental status of the STX. Nevertheless, it yields a large loss so its necessity should be assessed. At the very least it can be better specified, given its current specified range of -0.9 to -1.5 dB loss typically.
- **Pre-amplifier characteristics**
  The current assumptions with respect to the *noise figure* and *gain* of the amplifier have been based on a certain amplifier used in a different mission. The actual specifics of the final solution applied should be determined.

Finally the values of a number of variables with lower impact should be confirmed:

- **Actual power amplifier output**
- **Actual impedance mismatch loss**
- **Actual transmission line loss** based on the final cable lengths

VHF downlink

The accuracy of the VHF downlink link budget can be said to be high. Most of the components along the chain have been used in case of Delfi-C^3, and are thus available and proven. No remarkable discrepancies are apparent if the actual Delfi-C^3 power performance is compared to the VHF downlink link budget. The link budget closes for maximum achievable data rates on the VHF band, with an additional 2 dB for an 850 km altitude. The priority of the proposed changes below is therefore low.

Currently, the ground station on which the link budget is based does not support higher downlink data rates (than 1.2 kb/s). Therefore this variable should in fact be updated:

- **Noise temperature of the transceiver (ICOM 910)**
  Although the impact of this one variable is small (an added noise temperature value of 30 K), the component cannot be used to receive higher data rate signals. Therefore, the noise temperature of the future replacement should be determined.

A number of values belonging variables can be improved:

- **Satellite antenna directivity**
  The current (low) value of minimum antenna directivity has a large impact on the link budget. Therefore a better estimate can be determined, based on the actual antenna performance. For Delfi-C^3, functional tests have been performed for the integrated antenna system. The result of these measurements has been that antennas were shortened, due to conducting internal wires. The actual gain values over all angles were however not well measured or documented.
- **Antenna noise temperature**
  This value is currently taken to be 150 K, but is subject to large inaccuracy. More importantly, the value will change with elevation *and* azimuth angles. It would be very interesting to obtain (time-averaged) values of the antenna noise as a function of these angles.
- **Pointing loss**
  Although this might be a parameter that is very hard to determine, the actual pointing accuracy of the on-ground antenna system can be better specified. This would be a function of tracking errors, weather impacts and antenna directivity pattern.

UHF uplink

The accuracy of this link can similarly be said to be rather high. More importantly, the accuracy of the link budget for this link is less important given the huge link margin of over 20 dB.

Nevertheless, a number of values can be better specified:

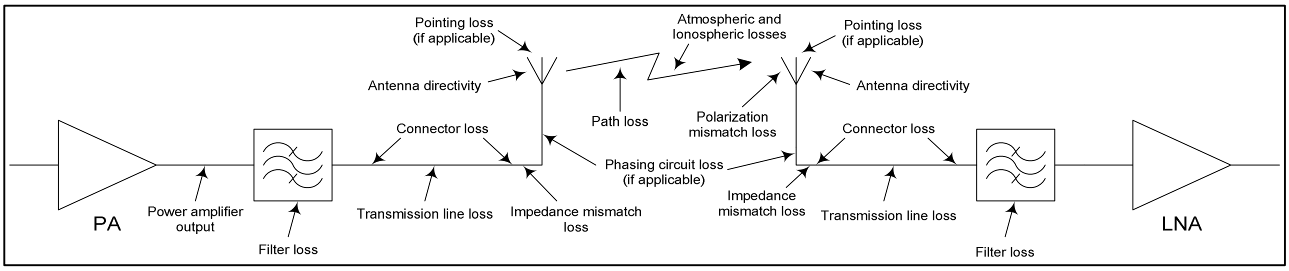
- **Satellite antenna directivity**
  The same holds as for the VHF downlink. In case of the UHF uplink however, a directivity of even -10 dB is assumed.
- **Pointing accuracy**
  See VHF downlink.
- **Antenna noise temperature**
  A value of 290 K has now been assumed. This value is subject to large guesswork and could be further specified. Given the fact that the satellite will logically operate in space, and applies an omnidirectional directivity pattern, its further specification is however very difficult.

PTRX downlink (600 km)

Link budget

Version: 3.0
Date: 13-6-2010

	Link parameters		
General	Transmission frequency	146	MHz
	Power amplifier output power	0,2	W
	Orbital altitude (circular orbit)	600	km
	Data rate	1200	symbols/s
	Modulation scheme	BPSK	
	Coding gain	0	dB



	Description	**Value**	**Unit**		
Transmitter	Power amplifier output	-6,99	dBW		
	*Coupler loss	-1,00	dB		
	Connector loss	-0,20	dB		
	Transmission line loss	0,05	dB		
	Phasing circuit loss	-1,70	dB		
	Antenna directivity (minimum)	-5,00	dBi		
	Transmitted EIRP			-14,84	dBW
Channel	Path loss	-141,46	dB		
	Atmosphere loss	-1,80	dB	+	
	Pointing loss	-0,50	dB		
	Polarization mismatch loss	0,00	dBi		
	Total channel loss			-143,76	dB
Receiver	Antenna directivity (maximum)	12,35	dB		
	Connector loss	-0,10	dB		
	Transmission line loss	-0,40	dB	+	
	*Surge protector loss	-0,20	dB		
	*Pre-amplifier loss	-0,20	dB		
	Received power			11,46	dB
	Total Eb *(Total signal power divided by data rate)*		=	-177,93	dBW
	Receiver noise temperature	264	K		
	Total N0 *(Noise temperature times Boltzmann's Constant)*			-204,39	dBW

Received Eb/N0	+	26,45	dB
Required Eb/N0 for BER = 10^(-5)	-	10,30	dB
Minimum link margin	-	3,00	dB

Conclusion Link closes with final margin of 13,15 dB

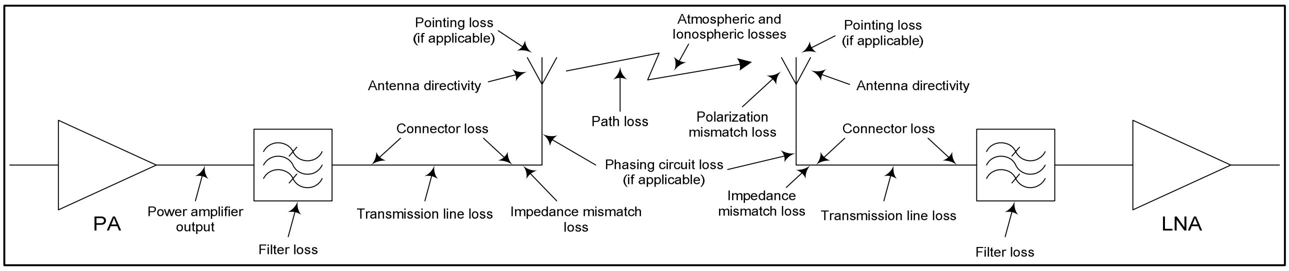
PTRX downlink (850 km)

Link budget

Version: 3.0
Date: 13-6-2010

Delfi-n3Xt

	Link parameters		
General	Transmission frequency	146	MHz
	Power amplifier output power	0,2	W
	Orbital altitude (circular orbit)	600	km
	Data rate	1200	symbols/s
	Modulation scheme	BPSK	
	Coding gain	0	dB



	Description	**Value**	**Unit**		
Transmitter	Power amplifier output	-6,99	dBW		
	*Coupler loss	-1,00	dB		
	Connector loss	-0,20	dB		
	Transmission line loss	0,05	dB		
	Phasing circuit loss	-1,70	dB		
	Antenna directivity (minimum)	-5,00	dBi		
	Transmitted EIRP			-14,84	dBW
Channel	Path loss	-143,59	dB		
	Atmosphere loss	-1,80	dB	+	
	Pointing loss	-0,50	dB		
	Polarization mismatch loss	0,00	dBi		
	Total channel loss			-145,89	dB
Receiver	Antenna directivity (maximum)	12,35	dB		
	Connector loss	-0,10	dB		
	Transmission line loss	-0,40	dB	+	
	*Surge protector loss	-0,20	dB		
	*Pre-amplifier loss	-0,20	dB		
	Received power			11,46	dB
	Total Eb *(Total signal power divided by data rate)*		=	-180,06	dBW
	Receiver noise temperature	264	K		
	Total N0 *(Noise temperature times Boltzmann's Constant)*			-204,39	dBW

Received Eb/N0	+	24,32	dB
Required Eb/N0 for BER = 10^(-5)	-	10,30	dB
Minimum link margin	-	3,00	dB

Conclusion Link closes with final margin of 11,02 dB

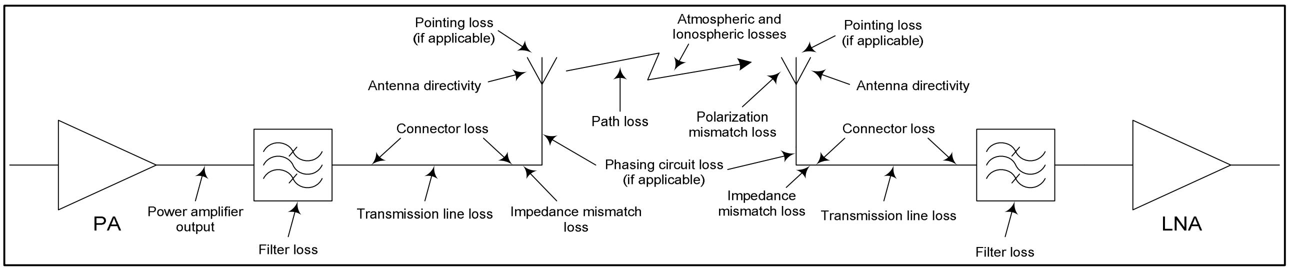
PTRX uplink (600 km)

Link budget

Version: 3.0
Date: 13-6-2010

Delfi-n3Xt

	Link parameters		
General	Transmission frequency	436	MHz
	Power amplifier output power	250	W
	Orbital altitude (circular orbit)	600	km
	Data rate	1200	symbols/s
	Modulation scheme	Man. FSK	
	Coding gain	0	dB



	Description	**Value**	**Unit**		
Transmitter	Power amplifier output	23,98	dBW		
	*Surge protector loss	-0,20	dB		
	Connector loss	-0,10	dB		
	Transmission line loss	-1,91	dB		
	Antenna directivity (maximum)	16,40	dBi		
	Transmitted EIRP			38,17	dBW
Channel	Pointing loss	-1,00	dB		
	Path loss	-150,96	dB	+	
	Atmosphere loss	-1,50	dB		
	Polarization mismatch loss	0,00	dBi		
	Total channel loss			-152,46	dB
Receiver	Antenna directivity (minimum)	-10,00	dB		
	Phasing Circuit loss	-2,15	dB		
	Impedance mismatch loss	0,00	dB	+	
	Connector loss	-0,20	dB		
	Transmission line loss	0,08	dB		
	Filter loss	-2,00	dB		
	Received power			-14,27	dB
	Total Eb *(Total signal power divided by data rate)*		=	-159,34	dBW
	Receiver noise temperature	500	K		
	Total N0 *(Noise temperature times Boltzmann's Constant)*			-201,61	dBW

Received Eb/N0	+	42,27	dB
Required Eb/N0 for BER = 10^(-5)	-	13,30	dB
Minimum link margin	-	3,00	dB

Conclusion Link closes with final margin of 25,97 dB

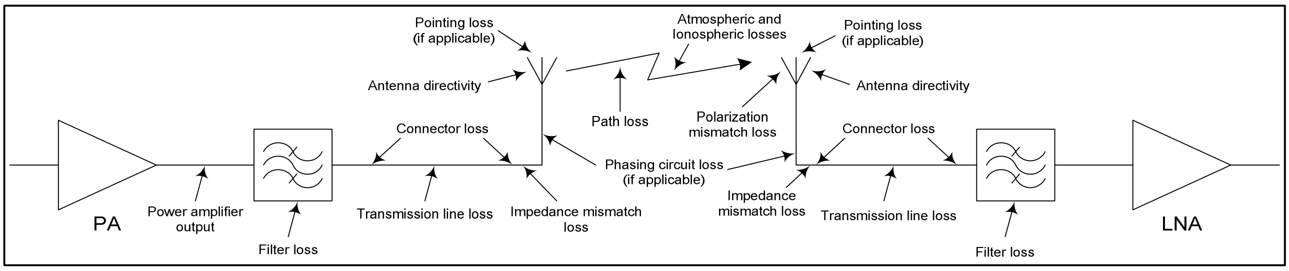
PTRX uplink (850 km)

Link budget

Version: 3.0
Date: 13-6-2010

Delfi-n3Xt

	Link parameters		
General	Transmission frequency	436	MHz
	Power amplifier output power	250	W
	Orbital altitude (circular orbit)	600	km
	Data rate	1200	symbols/s
	Modulation scheme	Man. FSK	
	Coding gain	0	dB



	Description	**Value**	**Unit**		
Transmitter	Power amplifier output	23,98	dBW		
	*Surge protector loss	-0,20	dB		
	Connector loss	-0,10	dB		
	Transmission line loss	-1,91	dB		
	Antenna directivity (maximum)	16,40	dBi		
	Transmitted EIRP			38,17	dBW
Channel	Pointing loss	-1,00	dB		
	Path loss	-153,09	dB	+	
	Atmosphere loss	-1,50	dB		
	Polarization mismatch loss	0,00	dBi		
	Total channel loss			-154,59	dB
Receiver	Antenna directivity (minimum)	-10,00	dB		
	Phasing Circuit loss	-2,15	dB		
	Impedance mismatch loss	0,00	dB	+	
	Connector loss	-0,20	dB		
	Transmission line loss	0,08	dB		
	Filter loss	-2,00	dB		
	Received power			-14,27	dB
	Total Eb *(Total signal power divided by data rate)*		=	-161,47	dBW
	Receiver noise temperature	500	K		
	Total N0 *(Noise temperature times Boltzmann's Constant)*			-201,61	dBW

Received Eb/N0	+	40,14	dB
Required Eb/N0 for BER = 10^(-5)	-	13,30	dB
Minimum link margin	-	3,00	dB

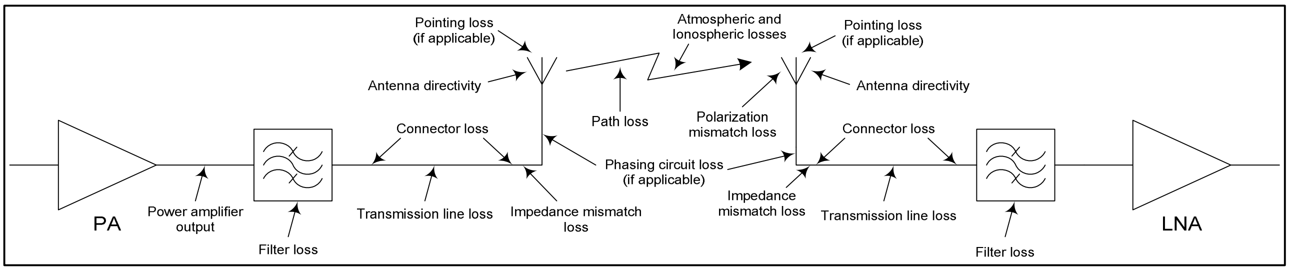
Conclusion Link closes with final margin of 23,84 dB

STX downlink

Link budget

Version: 3.0
Date: 13-6-2010

	Link parameters		
General	Transmission frequency	2400	MHz
	Power amplifier output power	0,315	W
	Orbital altitude (circular orbit)	600	km
	Data rate	93	ksymbols/s
	Modulation scheme	MSK	
	Coding gain	7	dB



	Description	**Value**	**Unit**		
Transmitter	Power amplifier output	-5,02	dBW		
	*Coupler loss	-1,20	dB		
	Connector loss	-0,20	dB		
	Transmission line loss	-0,27	dB		
	Impedance mismatch OR Phasing circuit loss	-0,20	dB		
	Antenna directivity (minimum)	-5,00	dBi		
	Transmitted EIRP			-11,89	dBW
Channel	Path loss (minimum elevation: 10° or 40°)	-165,77	dB		
	Atmosphere loss	-1,50	dB	+	
	Pointing loss (on-ground)	-3,20	dB		
	Polarization mismatch loss	0,00	dBi		
	Total channel loss			-170,47	dB
Receiver	Antenna directivity (maximum)	35,00	dB		
	Connector loss	-0,20	dB		
	Transmission line loss	-0,50	dB	+	
	*Surge protector loss	-0,20	dB		
	*Pre-amplifier loss	-0,30	dB		
	Received power			33,80	dB
	Total Eb *(Total signal power divided by data rate)*		=	-198,25	dBW
	Receiver noise temperature	298	K		
	Total N0 *(Noise temperature times Boltzmann's Constant)*			-203,86	dBW
	Received Eb/N0		+	5,61	dB
	Required Eb/N0 for BER = 10^(-5)		-	2,60	dB
	Minimum link margin		-	3,00	dB
	Conclusion	Link closes with final margin of 0,01 dB			